Novel Ring Resonator-Based Integrated Photonic Beamformer for Broadband Phased Array Receive Antennas—Part II: Experimental Prototype

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Abstract—An experimental prototype is presented that illustrates the implementation aspects and feasibility of the novel ring resonator-based optical beamformer concept that has been developed and analyzed in Part I [1]. This concept can be used for seamless control of the reception angle in broadband wireless receivers employing a large phased array antenna (PAA). The design, fabrication, and characterization of a dedicated chip are described, in which an 8 × 1 optical beamforming network, an optical sideband filter for single-sideband suppressed carrier modulation, and a carrier re-insertion coupler for balanced optical detection are integrated. The chip was designed for satellite television reception using a broadband PAA, and was realized in a low-loss, CMOS-compatible optical waveguide technology. Tuning is performed thermo-optically, with a switching time of 1 ms. Group delay response and power response measurements show the correct operation of the OBFN and OSBF, respectively. Measurements on a complete beamformer prototype (including the electro-optical and opto-electrical conversions) demonstrate an optical sideband suppression of 25 dB, RF-to-RF delay generation up to 0.63 ns with a phase accuracy better than \( \frac{\pi}{10} \) radians, and coherent combining of four RF input signals, all in a frequency range of 1–2 GHz.

Index Terms—CMOS-compatible optical waveguide technology, optical beamforming, optical ring resonators, phased array antennas, photonic integration, RF photonics.

I. INTRODUCTION

The motivation for phased array beamforming in the optical domain heavily relies on the integration of the beamforming functionality in optical chips. This results in a small size, low weight, low loss, low production and installation costs (when produced in mass), enhanced thermal and mechanical stability, and inherent immunity to electromagnetic interference.

The experimental prototype that is presented in Part II of this paper is based on the novel ring resonator-based beamformer concept that was presented in Part I [1]. This concept can be used for beam angle control in a broadband phased array receive antenna. The corresponding scheme is presented in Fig. 1. The signals from the antenna elements (AEs) are amplified by low-noise amplifiers (LNAs), and converted to the optical domain by means of optical single-sideband suppressed carrier (SSB-SC) modulation. The latter is performed using Mach–Zehnder modulators (MZMs) and a common optical sideband filter (OSBF). The actual beamforming is performed by means of ring resonator-based delays and coherent combining in an optical beamforming network (OBFN). After sideband-filtering, the output signal is converted back to the electrical domain by re-inserting the unmodulated carrier, and performing balanced coherent optical detection. The reason for using SSB-SC modulation is that the required number of optical ring resonators (ORRs) in the OBFN increases with increasing optical bandwidth. The main advantage of this system is that the ORR-based delay elements provide seamless beam angle control over a large instantaneous bandwidth, without beam squinting, using compact (integrated) devices. For a comparison with other existing approaches for photonic beamforming (such as the ones surveyed in [2]), more details about the operation principles of the ORR-based delay elements, the justification of using filter-based SSB-SC modulation and balanced coherent optical detection, and an extensive analysis of the beamformer’s impact of the performance of a phased array receiver system, we refer the reader to Part I of this paper [1].

As concluded in Part I [1], one of the key challenges in realizing such a system is keeping the losses in the optical part sufficiently low. Therefore, it is desirable to integrate as much functionality in one chip as possible. The chip that will be presented in this paper will contain an OBFN, an OSBF, and a carrier re-insertion coupler, as illustrated by the dashed box in Fig. 1. It is realized in the TriPleXTM waveguide technology of LionIX B.V. [3]–[5], as this provides the possibility of integrating relatively
complex (passive) functionality in one chip, with low waveguide propagation losses and low fiber-to-chip coupling losses.

In the past we published several papers with measurements on TriPleX™ chips for optical beamformers, showing different parts of the functionality mentioned above. In [6] we presented group delay measurements on a three-ring optical delay device, followed by group delay measurements on single-chip realizations of $4 \times 1$ [7] and $8 \times 1$ OBFNs [8], and time domain measurements on a $2 \times 1$ OB FN [9]. In [10], [12] we presented magnitude response measurements on a chip in which an OSBF was integrated together with a $4 \times 1$ OB FN. In [11] these results were extended to demonstrate that the filter successfully suppresses the undesired sideband, and that—when the OB FN is also properly tuned—the RF-to-RF transfer from modulator input to detector output indeed has a linear phase response in the RF range of interest, according to the desired delay. Coherent combining of multiple input signals was demonstrated in [13]. These measurements were performed on several different chips.

In Part II of this paper we present the complete trajectory from the design of a dedicated optical chip, to measurements on a complete beamformer setup. This includes the translation from RF-level system requirements into specifications of the structure and parameters of the optical chip. This is extensively described here as it is unique to the proposed beamformer concept, and—as will be shown—far from straightforward. The actual dimensioning of all the parameters in the beamformer presented here is based on an experimental demonstrator that was built within the framework of the SMART project [12], [13]. As described in Part I [1], the aim of this project was to investigate the feasibility of a broadband airborne receiver for satellite television reception, using a broadband conformal phased array, and a broadband optical beamformer. The design and theoretical performance study of the beamformer part have extensively been discussed in Part I [1]. The analysis showed that the required number of AEs in the array is in the order of 2000, and that the optical beamformer hardly affects the performance of the receiver—provided that proper amplifiers are used, and that the optical losses are kept sufficiently low.

As described in [12], [13], the experimental demonstrator that has been developed in the SMART project is a simplified version of the system that has been analyzed in Part I [1]. The array is a 64-element rectangular grid, comprising eight rows of eight AEs. (Since this array will not provide sufficient gain to receive signals from the satellite, the satellite signal was received by a dish and retransmitted towards the array when testing the final setup.) The antenna beam of the array is steered in the vertical direction by mechanically tilting the array, whereas the horizontal steering is provided by an $8 \times 1$ optical beamformer, which is designed according to the ring resonator-based beamforming concept. The latter is presented in Part II of this paper, as an experimental illustration of the feasibility of the beamformer concept that was proposed in Part I [1]. (Measurements on a complete system, including AEs and front-ends, will be presented in a later paper.)

The optical setup is based on an optical chip that contains the OB FN, OSBF, and carrier re-insertion coupler. The specifications of this chip are related to the applications for which it is designed (satellite television reception); this is further described in Section II. In Section III the actual mask design and realization of the chip are described, followed by its characterization in Section IV. The realization and testing of the complete optical beamformer setup are described in Section V. The paper ends with conclusions in Section VI.

## II. FUNCTIONAL DESIGN OF THE BEAMFORMER CHIP

This section describes the functional design of the optical chip, which involves, amongst others, a specification of the required number of ORRs in the OB FN, the free spectral range (FSR) of the ORRs, and the filter structure of the OSBF. (In Section III it is described how this functional design is translated into the actual layout of the chip, and how the device is fabricated.) The functional design is described in two steps. In Section II-A, the (RF-level) requirements of the complete beamformer are formulated, and translated into requirements on the optical chip. Then, in Sections II-B–II-D, the resulting functional designs for the OB FN, OSBF, and carrier re-insertion coupler are described, respectively.

### A. Overview of the Requirements

The specifications for the optical chip follow from the requirements of the complete beamformer at RF level. One of these requirements is the operating frequency range, which is 10.7–12.75 GHz for Digital Video Broadcasting by Satellite (DVB-S). The corresponding requirement for the optical frequency range follows from the optical spectrum that results after modulation. When the MZMs are properly biased, the optical carrier is suppressed, so that the spectrum of the modulated optical signal only contains two sidebands (see Section III-D in Part I [1]). The separation between the two sidebands equals twice the minimum frequency of the modulating signal (21.4 GHz), whereas the widths of the sidebands equal the bandwidth of the modulating signal (2.05 GHz). In our setup, however, the DVB-S signals are first down-converted using standard low-noise blocks (LNBs), to an IF range of 950–2150 MHz. In that case the separation between the sidebands is 1.9 GHz, and the widths of the sidebands are 1.2 GHz.

Nevertheless, at the time of designing the chip, it was not yet decided whether down conversion would be performed prior to the electro-optical (E/O) conversion or not. (The considerations on this will be further described in Section V.) Therefore, it was decided to design the optical chip in such a way that both scenarios could be supported. For the OB FN this implies that the delay elements should provide group delay responses with a
minimum bandwidth of 2.05 GHz. For the OSBF it implies that, for both scenarios, it should be possible to locate the passbands and stopbands of the filter in such a way that one sideband is entirely passed, whereas the other sideband is entirely suppressed.

The requirements on the delays that have to be supported follow from the geometry of the PAA and the required maximum scan angle. This was also discussed in the numerical example on DVB-S that was presented in Section V of Part I [1]. There it followed that the respective ports of the beamformer require linearly increasing delay tuning ranges that are integer multiples 68 ps.

As discussed in Section II of Part I [1], the group delay curves of the ORR-based delay elements will be rippled. The corresponding phase responses will hence be rippled as well. From simulations it followed that the effect of the phase errors on the corresponding beam pattern can be neglected when the phase errors are kept below $\pi/16$ radians. Therefore, we use this value to specify the phase accuracy of the beamformer chip.

B. Design of the Optical Beamforming Network

The 8 × 1 OBFN was chosen to have a binary tree structure. As discussed in Section II-C of Part I [1], this minimizes the required total number of ORRs. The resulting branch structure is shown in Fig. 2.

The optical group delays that need to be provided in each beamformer path are equal to the delays that are required at RF level [1]. Therefore, it follows from the specifications in the previous section that the delay elements in Stage 1, 2, and 3 require tuning ranges of 68, 136, and 272 ps, respectively. The bandwidths of the delay elements need to be at least 2.05 GHz, in order to accommodate one sideband in both system scenarios.

Earlier research [14] has shown that—for a given ripple specification—the required number of ORRs per delay element is roughly proportional to the product of required delay tuning range and the required optical bandwidth. This is irrespective of the FSR of the ORRs, as long as the optical bandwidth is well below the FSR. Furthermore, it is desirable to make the FSR as large as possible, as the FSR is inversely proportional to the circumference, so, the larger the FSR, the smaller the required total chip area. Our technology provides an upper bound to the FSR of 15 GHz (this will be explained in Section III); in that case, the required optical bandwidth is indeed well below the FSR. Therefore, the FSRs of the ORRs in the OBFN should be in the order of 15 GHz.

In Fig. 2 it can be seen that the delay elements in the first and second stage are chosen to consist of only one ORR, whereas the delay element in the third stage is chosen to consist of two ORRs. This is motivated as follows. The required settings of the power coupling coefficient $k_\phi$ and round-trip phase $\phi$ of an ORR depend on the required delay value and frequency range. For a delay element consisting of one ORR this relation is still analytically tractable, but a more complicated optimization problem arises when a delay element consists of more than one ORR. In that case the optimization has to be done numerically. This is considered in more detail in [15]. Using the optimization routines developed in [15], it was shown that delay elements consisting of one ORR, supporting an optical bandwidth of 2.05 GHz, can provide delay tuning ranges of 68 and 136 ps with a phase error below $\pi/16$ radians, whereas delays of 272 ps will result in phase errors of more than $\pi/16$. However, a delay element of two rings can provide a tuning range of 272 ps over 2.05 GHz of bandwidth with a phase error of less than $\pi/16$ radians. That is the reason why the delay elements in the first and second stage in Fig. 2 consist of one ring, whereas the delay element in the third stage consists of two rings.

The combiners in the OBFN are also made tunable (with power coupling coefficients $k_\alpha$ through $k_{12}$), for three reasons:
- it enables compensation for differences in AE gain, LNA gain, and optical propagation losses between the different branches;
- it enables amplitude tapering, which could be used for enhanced beam shaping (e.g., sideband suppression);
- it provides additional flexibility during the device characterization; see Section IV.

These tunable combiners (and also the tunable couplers inside the ORRs) can be realized as a symmetric Mach–Zehnder interferometer (MZI) with an optical phase shifter in one of its arms.

The remaining tunable optical phase shifters that are shown in Fig. 2 ($\phi_9$ through $\phi_{15}$) are required for synchronization of the optical phases of the SSB-SC-modulated optical signals that are combined in the OBFN, as described in Section III-E of Part I [1].

Obviously, the eight inputs of the OBFN should be accessible from the inputs of the chip. The rightmost output (Coupler 15) should be internally connected to the OSBF, and as many as possible of the remaining couplers should be made accessible at the outputs of the chip, for testing purposes. This is further considered in Sections III and IV.

C. Design of the Optical Sideband Filter

A logical and efficient design choice for the OSBF is to use a filter structure based on the same building blocks as the OBFN (couplers and ORRs). The OSBF can then be realized in the same technology as the OBFN (see Section III), and, hence, be integrated on the same chip. We chose to use an asymmetric MZI with an ORR in the shortest branch, as shown in the inset of Fig. 3. The round-trip time $\tau_d$ of this ORR should be taken twice as large as the path delay difference $\tau_{1d}$ of the MZI. The
advantage of such a filter is that it has flattened passbands and stopbands that can be made relatively wide, with steep transitions in between [16]. The response of this filter is periodic, with an FSR that is equal to the FSR of the MZI (1/τ). If we neglect propagation losses in the filter, assume that the couplers in the MZI are perfectly balanced (κ1 = κ2 = 1/2), and choose φd = 0 and φr = π, the power response from the upper input port to the lower output port of the filter in the inset of Fig. 3 can be proven to be given by [16]

\[ H_{12}(f) = \frac{\sqrt{1 - \kappa_r \cos(3\pi f \tau) + \cos(\pi f \tau)}}{2 - \kappa_r + 2\sqrt{1 - \kappa_r \cos(4\pi f \tau)}}. \]  

(1)

In Fig. 3 one period of this power response is plotted for three different values of κr. Apparently, choosing the value of κr is a trade-off between obtaining large passband and stopband widths (low κr), and large stopband suppression and low passband ripple (high κr).

The FSR and the widths of the passband and stopband should be chosen such that, irrespective of the scenario, one sideband of the modulated optical signal is passed, with sufficiently low passband ripple, and the other sideband is sufficiently suppressed. Without going into detail about the precise design procedure, we show here that this is possible when we choose the FSR as 6.70 GHz and κr as 0.78. The corresponding power response is shown in Fig. 4 (solid line), together with the spectrum of the modulated optical signal for the two system scenarios (dashed lines), for different values of κr.

![Fig. 3. One period of the theoretical power response of the filter shown at the inset (with τ = 2τ, κ1 = κ2 = 1/2, φd = 0, and φr = π), for different values of the power coupling coefficient κr.](image)

![Fig. 4. Theoretical power response of the designed OSBF (solid line), and spectrum of the modulated optical signal for the two system scenarios (dashed lines). The part that corresponds to the optical carrier is also plotted (dashed arrow at frequency f0), although it is supposed to be suppressed by the MZMs. The two inner sidebands (IF) correspond to the scenario where the AE signals are down-converted prior to electro-optical conversion, whereas the two outer sidebands (RF) correspond to the scenario where no down conversion is performed.](image)
The requirements on the FSR that were formulated in Section II are converted into waveguide lengths $L$ by

$$L = \frac{c_0}{N g \text{FSR}}$$

where $c_0$ is the speed of light in vacuum. $L$ can either denote a path length difference (in case of an MZI) or a circumference (in case of an ORR).

For the applied TriPleX™ waveguide geometry the minimum length of a directional coupler is approximately 1 mm. Together with the minimum bending radius and heater length it follows that a thermo-optically tunable ORR has a minimum circumference of 13 mm. Using (2) it follows that this corresponds to a maximum FSR of 15 GHz, as was already mentioned in Section II. For the ORRs in the OBFF a safety margin of 1 mm is added to the circumference, resulting in an FSR of 13.8 GHz.

In the OBFF the FSR of the MZI has to be 6.70 GHz, resulting in a required path length difference of 28.9 mm between the two branches. The FSR of the ORR in the OBFF should be exactly half the FSR of the MZI, so the required circumference of the ORR is 57.8 mm.

The carrier re-insertion coupler is simply implemented as a tunable MZI, including an optical phase controller.

A geometrical requirement to the optical waveguide layout that has been taken into account—which is not very critical for correct system operation but simplifies the chip characterization—is that all branches from the inputs to the OBFF until the output of the OBFF have equal optical path lengths, and, hence, equal optical losses.

To maximize the manufacturing yield, the number of chips per wafer should be as large as possible, so the footprint of the chip should be minimized. Another general method that we used to optimize the yield for such a complex chip as described here, was our step-by-step approach: first we made designs for basic building blocks (BBBs), then we fabricated and characterized these BBBS, and finally we used the resulting empirical data for our relatively complex systems built up from these BBBS. For example, the BBB of the binary-tree $8 \times 1$ OBFF in Fig. 2 is a $2 \times 1$ OBFF. This $2 \times 1$ OBFF was optimized by first making and characterizing its own BBBS, i.e., directional couplers, MZIs, bent waveguides of varying curvature and width, and straight waveguides of varying width.

The last requirement on the mask design is that the input and output waveguides of the chip should have a standard pitch of 250 $\mu$m, so that the chips can be pigtailed to commercially available connectorized fiber array units (FAUs), with polarization-maintaining fibers (PMFs).

An overview of the mask set designed for the complete beamformer chip (including OBFF, OBFF and carrier re-insertion coupler) is presented in Fig. 5, showing the waveguide layout, the position of the chromium heaters, the heater leads, and the bond pads. The chip footprint is 66.0 mm $\times$ 12.8 mm; five of these chips can be accommodated on a silicon wafer with a diameter of 10 cm. Each chip has 37 heaters connected to 41 bond pads, eight input waveguides, and ten output waveguides (of which eight have to be pigtailed). Underneath the lower row of bond pads there are some test structures (directional couplers and straight waveguides).

The optical waveguide layout was optimized for minimum chip footprint and for minimum optical path length differences and optical loss differences between the branches. The chip footprint was minimized firstly by flipping the branches and ORRs in the OBFF with respect to the structure shown in Fig. 2, as presented in close-up in Fig. 6, and secondly by folding the ORR and MZI branches of the OBFF, as presented in close-up in Fig. 7. The optical path length differences and optical loss differences between the branches were minimized firstly by making the design of the OBFF as symmetrical as possible. Secondly, the number of straight-to-bend transitions and the total angle of bent parts have been made equal for all optical branches in the OBFF and for the three optical paths in the OBFF (both MZI branches and the ring).

Note that it was not possible to avoid a few waveguide crossings. Waveguide crossings were only allowed between non-essential waveguides (which are useful for characterization but are not required for correct device operation); the structure shown in Fig. 7 contains three of such waveguide crossings. Additionally, in the center of the OBFF (Fig. 6) there are two waveguides that are not essential for device operation, but which could not be led to the output waveguide section without crossing essential waveguides. Therefore, they are terminated by stray light absorbers, which consist of an angle of 80 degrees, a linear taper to a very small waveguide width, and a metal disc on top in order to attenuate the light leaking away. Because the angle is smaller than 90 degrees, the light that is still present after the stray light absorber will not result in additional interference due to Fabry–Perot resonances of substrate modes at the chip edges.

B. Fabrication

For the realization of the beamformer chip the TriPleX™ waveguide technology of LionIX B.V. was used [3]. While in
most waveguide technologies the waveguides consist of a high-index core, the TriPleX™ technology realizes a waveguide with a shell of silicon nitride and a core of silicon oxide, grown by low-pressure chemical-vapor deposition (LPCVD), in a single lithography step [4]. The alternating sequence in which the silicon oxide and silicon nitride are stacked results in stress compensation, since the stresses of silicon oxide and nitride counteract. This enables the realization of higher-index waveguides. By varying the geometry of the waveguide, it can be tailored from completely polarization-dependent to polarization-independent. Furthermore, its loss can be made very low over a large wavelength range (from ultra-violet to infra-red) [5]. As described in Section I, these low losses are essential for the beamformer chip that is considered here, and the large wavelength range enables the use of relatively inexpensive optical sources in the 850 nm range.

Fig. 8 shows a schematic drawing and a scanning electron microscope image of a cross section of the box-shaped waveguide geometry that was used in this beamformer chip. This geometry was chosen for its low loss and low optical crosstalk properties. Because of the lack of a slab between the different waveguides on the chip, crosstalk between the different paths is kept to a minimum. Different other geometries can be realized by changing etch depth and layer thicknesses. The complete fabrication process is described in more detail in [4].

After the realization of the devices on silicon wafers the individual chips were diced from the wafer and packaged for characterization (Section IV) and integration in the complete system (Section V). For temperature stabilization, the chip was placed on a copper block, which acts as a heat sink. By means of a temperature-dependent resistor and Peltier elements, the temperature of the chip is controlled with an accuracy better than 0.01 K, with local and temporal variations even smaller than that. Polarization-maintaining FAUs were glued onto the chip (butt coupling) after active alignment of the fibers. The electrical connections to the bond pads (see Fig. 5) were made by means of probe pens, in order to control the heater voltages. Although small temperature fluctuations might result in shrinkage or expansion of the glue, the effect of the heater elements on the glue is negligible here because of the temperature stabilization.
Since there is only a very thin layer of glue between the FAU and the chip, the glue’s contribution to the insertion loss of the chip can be neglected as well.

C. Chip Control System

A dedicated control system was realized for tuning the chip to the desired settings. This control system converts the values of the desired delay settings into the required voltages over the chromium heaters. This is performed in four steps.

First, the desired delay settings are converted into the required values of the power coupling coefficients $k$ and the phase shifts $\phi$ in the ORRs, according to the theory that was described in Section II of Part I [1]. This conversion is performed in a microprocessor on a dedicated control mother board, by means of polynomial interpolation formulas. The coefficients of these polynomials are calculated offline by means of numerical optimization routines on a personal computer. These are described in further detail in [15].

Then, the desired values of the $k$s and $\phi$s are converted to the corresponding phase shifts in the chip by inverting the transfer function of the MZI-based tunable couplers. This is also done in the microprocessor.

The eventual phase shifts in the chip are proportional to the power that is dissipated in the heaters, which is proportional to the square of the voltage. For the heaters in our chip, a phase shift of $2\pi$ radians is achieved by a voltage of 25 V. The corresponding calculation of the required voltages is also carried out in the microprocessor.

Finally, the required heater voltages are supplied by means of dedicated digital-to-analog converters (DACs) and amplifiers, which were combined and realized on a printed circuit board (PCB). The precision of the DACs is 14 bits. One such a PCB which were combined and realized on a printed circuit board (PCB). The precision of the DACs is 14 bits. One such a PCB, which was combined and realized on a printed circuit board (PCB), can accommodate 32 heaters, and several PCB can be stacked on our control mother board. In this case, two PCBs were used to control the 37 heaters on this beamformer chip.

The power consumption of the heaters is relatively large: a phase shift of $2\pi$ radians requires a power dissipation of 700 mW. Hence, assuming a uniform distribution of the tuned phases over the OBFN, it follows that the average power consumption per heater is 350 mW, resulting in a total average power consumption for the entire chip of approximately 13 W (with a peak power consumption close to 26 W). We expect to largely reduce this in the future, by replacing the thermo-optical tuning by electro-optical tuning. This should have no impact on the fundamental beamforming concept, but will definitely reduce the switching time (now roughly 1 ms). Despite the large dissipation in the heaters, the heat sink and temperature stabilization turned out to result in very stable chip settings.

IV. CHARACTERIZATION OF THE BEAMFORMER CHIP

In this section the characterization of the beamformer chip is described, by presenting optical group delay response measurements on the OBFN, optical power response measurements on the OSBF, and measurements of the losses in the chip.

A. Characterization of the Optical Beamforming Network

As described in Section II, different parts of the chip can be measured separately by using the corresponding input and output port. For example, when a chip output is used that is connected to the lower output of Coupler 15 in Fig. 2, the OSBF is skipped and only the OBFN remains in the measured optical path.

The optical group delay response of each OB FN path was measured by means of the phase-shift method. Fig. 9 shows the corresponding measurement setup. The Agilent N5230A PNA-L network analyzer generated an electrical 50 MHz signal, which modulated the light from the EM4 high-power distributed feedback (DFB) laser by means of an Avanex PowerLog FA-20 MZM. The modulated optical signal was coupled into the OB FN chip. PMFs were used for the connections between the laser, MZM, and OB FN chip, such that the chip was investigated for transverse-electric-polarized light. Seven OB FN paths (Input 2 to 8 in Fig. 2) were subsequently measured by connecting the MZM to the corresponding input fibers of the chip. Because of the high losses in this particular chip (we will come back to this in Section IV-C), the output signal of the chip was amplified by means of an erbium-doped fiber amplifier (EDFA). The RF signal was detected from the optical output signal by means of an Emcore R2860E optical detector, and led to the network analyzer. The network analyzer measured the phase difference $\Delta \phi(\lambda)$ between the modulating RF signal and the detected RF signal for different laser wavelengths $\lambda$. The laser wavelength was varied by sweeping the laser injection current. The group delay response was estimated from these results by calculating $\tau_g(\lambda) \approx \Delta \phi(\lambda)/(2\pi f_{RF})$, where $f_{RF}$ is the frequency of the electrical signal (50 MHz in our case). With this setup the group delay response for each OB FN path was measured over one FSR of 14 GHz.

In Fig. 10 the results are shown for two different delay settings of the OBFN. When drawing the graphs, the minimum delay of each path was used as the zero delay reference of that path. Suppose this minimum delay setting corresponds to a negative reception angle. This can be realized by inserting proper (fixed) delay offsets elsewhere in the system, either in the RF part (between the AEs and the MZMs) or in the optical part (between the MZMs and the optical combiners in the OB FN), by using different physical lengths for the RF waveguides or optical waveguides, respectively. (For example, the first branch—which has no delay tuning element—should have the longest physical length. Linearly decreasing offsets are required for the other branches, and the last branch—which has the largest delay tuning range—requires no delay offset, so it has the shortest physical length.) The two delay settings shown in Fig. 10 were made in such a way that linearly increasing
delays were achieved for the seven OBFN paths over an optical bandwidth of 1 GHz. These two settings correspond to two different beam angles for a linear PAA. The delay steps for the two settings were approximately 40 ps and 80 ps, respectively, which correspond to distances of 1.2 cm and 2.4 cm in air for two neighboring AEs. Since the element spacing for the K_u-bandy array is approximately 1.2 cm [1], these roughly correspond to reception angles of 0 degrees (broadside) and +90 degrees, respectively.

Note that delay steps in the measurement slightly exceed the required maximum step of 68 ps for the designed antenna. This is one of the reasons why the bandwidth that was achieved in our measurements is smaller than the designed 2.05 GHz. The other reason is that the fixed directional couplers that constitute the MZI-based tunable couplers turned out to have power coupling coefficients that were not exactly equal to 0.5, due to fabrication inaccuracies. This resulted in a reduced tuning range of the tunable couplers, which, for the ORRs, resulted in a slight increase in the minimum delay of the ORR-based delay elements. Hence, the ORRs needed to provide higher maximum delays than in theory to achieve a desired delay tuning range, which eventually resulted in smaller delay bandwidths than the designed 2.05 GHz.

B. Characterization of the Optical Sideband Filter

The power response of the OSBF was measured separately by using the chip input that corresponds to the OBFN path without ORRs, and a chip output that is directly connected to one of the outputs of the OSBF. The measurement was performed by means of the same setup as in the previous subsection, now using the network analyzer to directly measure the power response.

The results are presented in Fig. 11, which shows the optimized OSBF response measured over an FSR of 6.7 GHz. The response is normalized to the response that was measured when the chip was configured to have no filter in the measured path. Moreover, the actual dB levels that are shown in Fig. 11 are halved with respect to the dB levels that were actually measured by means of the network analyzer, as the measured electrical response scales with the square of the optical response. Hence, the shown response is the optical power response of the filter. The passband, transition band, and stopband have widths of approximately 3 GHz, 1 GHz, and 1.7 GHz, respectively, and the stopband suppression is 25 dB, as desired.

The reason why the measured width of the stopband is slightly smaller than the designed width (2.05 GHz), is that the FSR of the ORR in the OSBF turned out not to be exactly half the FSR of the MZI, due to a fabrication inaccuracy. Therefore, the width of the stopband had to be slightly decreased to still achieve the desired 25 dB suppression. This was not really a problem in our system setup (Section V), as down conversion was performed prior to E/O conversion, so that the required bandwidth was only 1.2 GHz.

C. Optical Losses in the Chip

The minimum insertion loss of the chip appeared to be around 12 dB (with all combiners in the measured path tuned such that the splitting loss is minimized). It is estimated that roughly 10 dB should be attributed to fiber-chip coupling losses (5 dB per facet), based on simulations and on measurements on a straight waveguide fabricated in the same process. This high value is caused by the large mismatch in mode field diameter of the PMF (roughly 10 μm) and the waveguides in the chip (roughly 3 μm). The waveguide end faces of the chip were

Fig. 11. Measured optical power response of the OSBF in the chip.
not optimally tapered, in order to prevent the risk of a low fabrication yield for this novel device.

For small delays the waveguide loss in the ORRs in the wavelength range of interest is proportional to the effective waveguide length [6]. By varying the delay, the waveguide loss in the ORRs was estimated to be approximately 0.6 dB/cm. As a result, the maximum insertion loss occurs when the delay elements are tuned to their maximum setting. For example, when the delay elements in the path corresponding to Input 8 in Fig. 2 (i.e., the ORRs with numbers 4, 6, 7, and 8) are tuned to a total maximum delay of 0.56 ns (which corresponds to roughly 11 cm in the chip), the additional excess loss in the ORRs is roughly 7 dB. This was confirmed by measurements.

The excess loss of the OSBF follows from the measured optical power transfer in the passband (see Fig. 11), and is roughly equal to 2 dB. Therefore, the total excess loss (excluding intrinsic splitting and combining losses) in the modulated paths of chip can vary from roughly 14 to 21 dB.

Using the theory in Part I [1] it can be shown that such losses would result in an unacceptable performance degradation if the chip were applied in a complete receiver system. In the measurements that we present in this and the next section we overcame this by using an EDFA, but this would certainly not be desirable in a final product, because of the associated additional costs, but also because of the additional noise that would be introduced. However, recent improvement in our technology have resulted in chips for different purposes with waveguide propagation losses well below 0.1 dB/cm [3], [5]. By effectively tapering the devices we expect to lower the fiber-chip coupling losses to well below 1 dB per facet [5]. Therefore, we expect to fabricate new beamformer chips with sufficiently low waveguide propagation losses and fiber-chip coupling losses in the near future. An even better solution is to completely avoid fiber pigtailling and directly integrate light sources and detectors via a different packaging technology such as flip chip. This is also part of our on-going work.

Measurements on the MZI-based tunable combiners in the OBFP showed that the combiners are able to compensate power differences up to approximately 12 dB between two combining branches. This is more than enough for compensating loss differences, as the maximum difference in loss between combining branches in this chip is roughly 4 dB (for Combiner 15 in Fig. 2). The loss differences in future chips will even be much smaller because of lower waveguide propagation losses.

V. REALIZATION AND TESTING OF THE COMPLETE BEAMFORMER SETUP

After characterizing the chip it was used to build a prototype of a complete beamformer setup, according to the scheme in Fig. 1. The laser, MZMs, and detector were the same ones as used for the chip characterization. The measurements were performed with a single optical detector instead of a balanced detector, because we did not have a balanced detector available at the time of the measurement. The optical connections were made by means of PMF patch cords and PMF splitters.

As described in Section III-E in Part I [1], the phase shifts in each branch of the optical beamformer needed to be stabilized in order to enable constructive coherent optical combining in the OSBF. As part of the optical paths that are combined consists of fiber patch cords and splitters, this involved both mechanical and temperature stabilization. These were achieved by putting the PMF splitters and MZMs on top of a heavy metal plate, which was put inside a custom-made styrofoam box, together with the optical chip. Eventually this enabled us to stabilize the optical phases within acceptable limits for a few minutes, which turned out to be sufficient to be able to perform the measurements that we present here.

As mentioned in Section II-A, when applying this beamformer in a complete system, one of the design considerations is whether down conversion should be performed prior to the E/O conversion or not. Down conversion has the advantage that modulators and detectors with lower speeds can be used. However, the disadvantage is that a lot of LNBs are required, and, moreover, the down converters in these LNBs need to have a common oscillator signal in order to maintain correlation between the output signals of the LNBs. Eventually the decision was made to down-convert prior to beamforming, because it turned out that it would not be possible to include sufficient amplification between our AEs and optical modulators without introducing feedback oscillations, if no down conversion were performed. Although the AEs and front-ends were not included in the measurements presented here, it was therefore decided that all the measurements on the beamformer should be performed in the IF range (950–2150 MHz).

In the following three subsections the correct operation of the system is demonstrated by presenting measurements of the spectrum of the modulated optical signal, the RF-to-RF phase response from one beamformer input to the output, and the RF-to-RF power response when coherently combining four beamformer input signals, respectively.

A. Sideband Suppression

Based on the OSBF power response measured in Section IV-B, a demonstration of sideband-filtering and carrier suppression of a modulated optical signal was performed. An RF signal with a frequency sweeping from 1 to 2 GHz was applied to one of the MZMs. For this measurement the laser was directly connected to this MZM, without the fiber splitters. The OSBF response was shifted in frequency such that one sideband of the modulated optical signal fitted in the passband and the other sideband in the stopband. For this demonstration the optical heterodyning technique was used to measure the optical spectrum. This was done by amplifying the optical output signal of the beamformer chip by means of the EDFA, and combine it with the output signal of a Santec tunable laser, using a directional coupler. The wavelength of this laser was tuned close to the wavelength of the EM4 DFB laser, so that the spectrum of the output signal of the detector contained a part that is a frequency-shifted version of the spectrum of the modulated optical signal, where the frequency shift corresponds to the center frequency of the tunable laser. This spectrum was measured by means of an Agilent N9020A signal analyzer, using the ‘Maximum Hold’ option in order to visualize all modulating frequencies while sweeping the modulating signal from 1 to 2 GHz.

The results are shown in Fig. 12. The solid line corresponds to the spectrum that was obtained when the OSBF was tuned...
such that one sideband is suppressed, while the dashed line corresponds to the spectrum that was obtained when the OSBF was turned off (resulting in the spectrum before filtering). The dotted line is the measured optical power response of the OSBF. Even without sideband filtering, the measured upper sideband has a slightly lower magnitude than the lower sideband, because the responsivity of the optical detector decreases with increasing output frequency. The peak between the two sidebands is a shifted version of the optical carrier line of the DFB laser, which shows that the carrier is not fully suppressed by the MZM (due to a slight imbalance in the losses of the MZM paths, and a slight bias drift). The position of the peak (roughly 7 GHz) equals the difference between the optical carrier frequencies of the two lasers. (Apparently the center wavelength of at least one of the lasers is slightly drifting.) The results show that one sideband of the signal is suppressed by roughly 25 dB, whereas the other sideband remains unchanged.

**B. RF-to-RF Delay Generation**

For the demonstration of the RF-to-RF delay generation by the beamformer, an RF signal sweeping from 1 to 2 GHz was again applied to one MZM input, using the network analyzer. One directional coupler was placed between the laser and this MZM. The output of the MZM and the second output of this directional coupler were connected to Input 8 of the OBFN (see Fig. 2) and the carrier re-insertion input of the beamformer chip, respectively. One output of the chip was amplified by the EDFA and led to the optical detector, which was connected to the network analyzer. Three different delay settings were made in the eighth branch of the OBFN (corresponding to 0, 0.40, and 0.63 ns), such that the corresponding group delay responses covered the sideband of the modulated optical signal that passed through the OSBF. (The other sideband and carrier were suppressed, as shown in the previous section.) By means of coherent optical detection, signal recovery was performed after the combination of the delayed sideband and the unmodulated optical carrier. Using the network analyzer the phase difference between the modulating RF signal and the recovered RF signal at the detector output was measured as a function of frequency, for the three different delay settings.

The results, which can be considered as the RF-to-RF phase responses, are shown in Fig. 13 (solid lines). The dashed lines are the desired responses for the corresponding delay values. The minumum delay setting was again used as the zero reference. Besides some ripple, the obtained phase responses show a good match to the desired ones: the measured phase error is below $\pi/10$ radians. The ripples in the results are mainly attributed to Fabry–Perot resonances, due to the slight reflections between the various fiber connectors in the optical signal path. In the future implementation this will not be a problem, because the full beamformer will be integrated on a single chip, including splitters and modulators. The trends in the measured results suggest that the phase accuracy will be even better then.

Though not shown in the figure, the corresponding magnitude responses of the beamformer were measured to be flat over the signal band, but with larger loss for higher delay, because the optical loss in the ORR-based delays increases with delay value (as discussed in Section IV-C).

**C. RF Signal Combining**

For the demonstration of coherent combination of the SSB-SC-modulated optical signals in the OBFN, the output signal of the network analyzer was again swept from 1 to 2 GHz, split by means of three RF directional couplers, and applied to the RF inputs of four MZMs. The laser signal was split to the optical inputs of these four MZMs and the carrier re-insertion input of the chip, by means of four directional fiber couplers. The output optical power of the MZMs was connected to the OBFN inputs that correspond to Inputs 1, 2, 4, and 8 in Fig. 2. Hence, the modulated optical signals were combined in the OBFN, filtered, combined with the unmodulated optical carrier, detected by means of the photodetector, and routed back to the network analyzer. (The EDFA was not used in this measurement.) Consequently, four paths can be identified between the output and input of the network analyzer, all of them being partly in the electrical domain and partly in the optical domain. The differences between the lengths of these paths (mainly due to differences in RF cable lengths) were removed by properly tuning the ORR-based delay elements in the OBFN. Then, the optical phase shifters in the OBFN
were manually tuned such that the maximum output power was achieved after the combination of the delay-synchronized modulated optical signals. The power response from the input of the RF splitting network to the output of the beamformer was measured as a function of frequency, and is shown by the upper line in Fig. 14. Then, two outputs of the RF splitting network were disconnected from the MZMs and terminated, and the same measurement was repeated, including the manual tuning of the optical phase shifters in the OBPN. Hence, the RF signal was now applied to only two of the inputs of the beamformer, while maintaining the same input power level. This was repeated for the other two inputs, by reconnecting the outputs of the RF splitting network and moving the terminators. The two resulting curves are the second and third curve in Fig. 14.

Finally, the RF signal was subsequently applied to the four MZM inputs individually by connecting only one of the outputs of the RF splitting network and terminating the other three. The respective measurement results are shown by the four lower curves in Fig. 14. The measured curves illustrate that, after each time the number of combined signals was doubled, there was roughly 6 dB increase in RF power level at the output. This demonstrates that the RF signals were coherently combined in the beamformer. The ripples in the signal response were again caused by Fabry–Perot resonances, as described in the previous subsection.

VI. CONCLUSION AND FUTURE WORK

A novel optical beamformer for broadband receive PAAAs has been presented, employing filter-based optical SSB-SC modulation, ring resonator-based broadband time delays, coherent optical combining, and coherent optical detection. The main advantages of the proposed concept are its large instantaneous bandwidth, seamless tunability of the PAA reception angle, immunity to electromagnetic interference, and potential for integration as a compact, light-weight device.

The implementation aspects and feasibility of the proposed concept were illustrated by presenting design considerations and measurements on a dedicated optical chip, and on an experimental setup built around this chip. The chip contains an ORR-based OBPN, OSBF, and carrier re-insertion coupler, and was realized in the CMOS-compatible TriPleX™ technology of LioniX B.V. The chip is tuned thermo-optically, with a switching time of 1 ms. Measurements on the group delay response and power response illustrated the proper operation of the OBPN and OSBF in the chip, respectively. Measurements on the complete setup (including the E/O and O/E conversions) demonstrate an optical sideband suppression of 25 dB, time delay generation up to 0.63 ns (19 cm in air) with a phase accuracy better than $\pi/10$ radians, and coherent combining of four beamformer input signals, all in a frequency range of 1–2 GHz.

The measured insertion loss and waveguide propagation loss of the chip were 12 dB and 0.6 dB/cm, respectively, which are still too high for successful application in a complete receiver system employing the proposed beamformer concept. However, based on the demonstration of similar devices for different purposes [3], we expect to realize beamformer chips with sufficiently low waveguide propagation losses and fiber-chip coupling losses in the near future. Moreover, it is desirable to integrate the splitters and optical modulators on the same chip as the OBPN, OSBF, and carrier re-insertion coupler, and package this chip directly with the laser and detectors (i.e., without a fiber pigtail). This will further reduce the losses, simplify the optical phase synchronization, and avoid reflections in fiber connectors. This is also part of ongoing work.

Another challenge is the upscaling of the experimental demonstrator. This involves the modular design of a beamformer using multiple optical chips and possibly multiple lasers and detectors. (An example of such a modular system was given in Section V of Part I [1].) This will also involve further automation of the chip control system, and reduction of its power consumption. The latter requires the thermo-optical tuning in the chips to be replaced by a different tuning mechanism. All of these challenges are part of our ongoing research.

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