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# Orthogonal Versus Zero-Forced Beamforming in Multibeam Antenna Systems: Review and Challenges for Future Wireless Networks

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(Invited Paper)

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**ABSTRACT** Orthogonality in multibeam antennas is revisited. The difference between orthogonal beamforming and zero-forced beamforming is highlighted. The intriguing relation between orthogonality, reciprocity and losses is recapitulated. Different approaches on the design of orthogonal beamforming networks and implementation of zero-forced beamforming strategies are shown with various examples from the antenna-research-oriented literature. The use of orthogonal and zero-forced beamforming is discussed from the communication system perspective with relevant studies from diverse disciplines. Some of the future research challenges and potential benefits are outlined for the next generation satellite and cellular communication applications.

**INDEX TERMS** Beamforming, interference suppression, multibeam antenna, orthogonality, wireless systems, zero-forcing.

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## I. INTRODUCTION

Most arrays onboard multiple beam satellites so far, including Iridium, Globalstar, Iridium Next, have generated fixed/switched multiple beam footprints with several users per beam. The first example of an on-board array generating a beam per user is the Tracking and Data Relay Satellite System (TDRSS) [1]. The onboard S-band array was used, from 1983 onwards, to generate up to 20 simultaneous beams, each tracking a spacecraft. Signals received by the array elements were multiplexed and sent to the ground for beam forming. This is only possible with narrow band signals. The transmit Ka-band arrays of the SPACEWAY satellites [2] provided up to 24 simultaneous co-channel beams, hopped in well separated fixed footprints, with about 1 dB cross over gain loss.

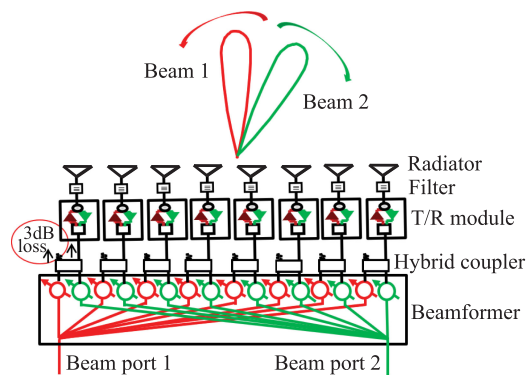
Adaptive arrays with few beams have been initially flown onboard US military satellites such as the MILSTAR,

Defense Satellite Communications System (DSCS), and their successors Advanced Extremely High Frequency (AEHF) and Wideband Global Satcom (WGS), with more beams, capacity and features such as jammer nulling. Limited details are available on their payloads. X-band adaptive arrays have been also developed for European dual-use satellites: SkyNet, Sicral, Syracuse and SpainSat. More recently, adaptive array technologies have been transferred to Ku-band commercial satellites: e.g. SmallGEO's RedSat and Quantum receive arrays. Adaptive beam forming by precoding with generation of nulls in known jammer or co-channel user directions is applied in radar and early Long-Term Evolution (LTE) mobile communication systems, generally with one beam per array at a time. Therefore, in a multi-user scenario, most systems have proposed a multiplicity of single beam antenna arrays, each serving a different user [3]. Later, with the development

of the Multi-User Multiple Input Multiple Output (MIMO) concept [4] and with the introduction of the Massive MIMO technology [5], the discussions on forming multiple flexible simultaneous co-frequency beams from a large single array on the transmit side have increased. Many works have emerged, especially in the signal-processing community, that study the formation of multiple beams sharing the same time-frequency-code resources with channel state information based precoding methods for interference management [6], [7].

From the design point of view, a recent and very comprehensive review on multiple beam arrays [8] reports various developments of truly multiple beam arrays including fully-digital, fully-analog and hybrid implementations commonly seen in applications such as radio astronomy [9], military radar [10], satellite communications [11], [12], personnel imaging [13], automotive radar [14], mobile communications [15], [16] and so on. Among these options, fully-digital schemes provide the most flexible, versatile and robust performance. However, they suffer from high complexity, power consumption and cost. Analog beamforming, and in particular, beamforming networks are an attractive solution to introduce multiple beam capability in low cost mass market products [17]. On the other hand, several hybrid beamforming strategies have recently been proposed in 5G as an appealing compromise, especially at the mm-wave bands [18], [19]. It is believed that different technologies will find their applications in different contexts [20], e.g. phased arrays in base stations where the power supply is not a major problem, and beamforming networks in user terminals or drones etc., which are battery driven and thus more constrained in power consumption.

So far, most commercial space communication systems use fixed or hopped multiple beams with limited co-channel interference control. However, it can be seen in the latest state-of-the-art that phased arrays are being considered for civilian satellite applications as well. For example, Starlink space segment is using phased array antennas [21], which is also the solution under development for O3b's mPower. Similarly, with the advent of 5G/6G and allocation of frequencies in the mm-wave range for more bandwidth, modern developments of phased array solutions have emerged in the terrestrial networks. Some state-of-the-art base station antenna examples include the active integrated arrays designed by Ericsson [22], UCSD [23], IBM [24], Nokia [25], NXP [26] and Qualcomm [27], which can all generate only a single beam at a time. The industrial high-volume mm-wave 5G market is still far from true multibeam antennas due to many practical factors such as cost, design complexity, cooling and computational burden. Although there are several examples of chipscale hybrid/digital beamforming mm-wave receivers with relatively small number of elements and concurrent beams [28]–[30], only very few practical developments of large-scale (i.e. > 64-element) truly multiple flexible beam arrays (in the form of digital beamforming [31], [32] and



**FIGURE 1. Lossy\* conventional analog single polarized active array beamforming network for two agile beams. (\*) The losses come from the beam combining (or splitting in receive) stage. The beamformer in Fig. 1 itself is not lossy, but the beamforming is (as in Blass matrices [43]). This is due to the couplers placed outside the beamformer. More discussions on definition of orthogonality and losses are given in Section II-C.**

hybrid beamforming [33], [34]) were reported for mm-wave cellular communications.

Therefore, the future of both space and cellular communications will need more flexibility, frequency re-use (through spatial multiplexing) and power efficiency at space and ground levels with low complexity [20], [35]. The antenna challenges are to limit gain and power efficiency losses due to array amplitude tapers, beam cross-overs, and to greatly reduce the side lobe interference accumulation at each user, occurring in current systems employing multiple beam forming matrices [8]. Furthermore, the need for efficient and flexible adaptive multiple beam forming with co-channel interference suppression in future satellite communications at space [36], [37] and ground [38] segments is well recognized [39]. Similarly, 5G mobile communications base stations and user equipment require multiple beams and low interference.

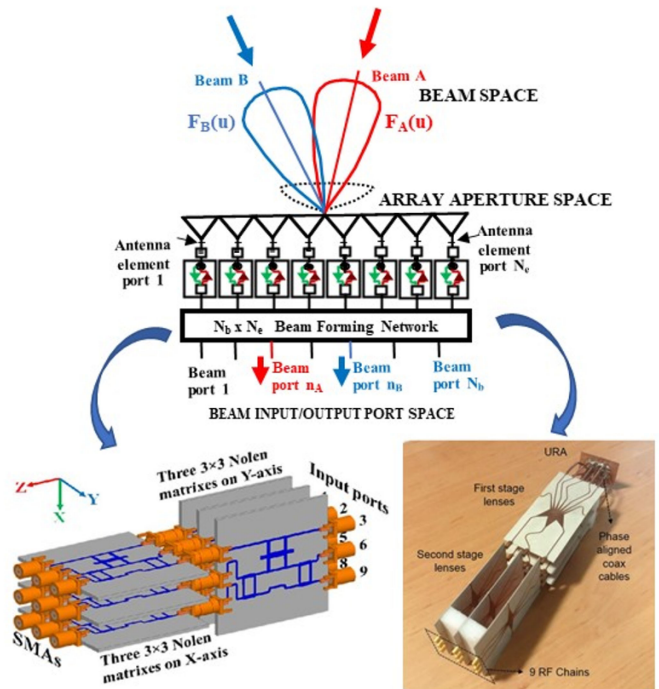
A straightforward way of designing a multiple beam former is to have a dedicated beamforming network for each separate beam port and to use power dividers/combiners behind each antenna element. This topology is generally referred to as the “fully-connected architecture” in the literature [18], [40]. A two beam network illustration of such a configuration is given in Fig. 1. Despite being flexible on adjusting the array excitation weights, the fully-connected architecture has a major disadvantage of being lossy, which could be severe especially with more than a few beams. The combining/dividing losses (equal to  $10 \log(N_b)$  for  $N_b$  beams) have to be compensated by an increase of amplifier gain at the element level [3].

A necessary condition to obtain a lossless (ideally) multiple beam forming network is that the beams are orthogonal (in a complex inner product sense) [41]. However, the fact that the beams are orthogonal is not enough to guarantee that the beam forming network is lossless. In other words, a lossy network can generate orthogonal beams as well since vectors do not need to have equal magnitudes to be orthogonal. Then, the

transfer block of the scattering matrix (i.e. the block matrix representing beam-port to array-port transfer coefficients) will not be unitary anymore, but its column vectors will still be orthogonal, corresponding to the distinction between orthogonal and orthonormal vectors terminology<sup>1</sup>. Besides, orthogonality does not necessarily mean that for each beam, signals incident from, or transmitted towards, the other beam peak directions will be zero (as with a Butler matrix [42]). The interference-free system is generally achieved and maintained via flexible (ideally) zero-forced, or very low sidelobe, beams, for which the peak of one beam is at the same angle with (or close to) the null of all the other beams. In general, it is required to have flexibility on the main beam angles while maintaining accurate zeros at the other beam peaks. This requires reconfigurability in the beamforming network. At the current stage, for more flexibility on the angular position of each zero-forced beam, either digital beamforming with increased complexity/cost is employed or highly lossy networks (as in Fig. 1, with combining losses of  $10 \log(N_b)$  dB for  $N_b$  beams) are used in analog beamforming.

It is noticeable that the use of the term “orthogonal” is misleading in many cases in the literature and the authors from different disciplines may actually refer to different scenarios defined by orthogonality. In this paper, the aims are: (i) to highlight the differences between orthogonal beams and zero-forced beams, (ii) to briefly explain the relation between the three concepts extensively used in linear microwave circuits; orthogonality, losses and reciprocity, from an antenna perspective, (iii) to review the literature on orthogonal beamforming networks and flexible zero-forced beamforming architectures, (iv) to show the potential use of orthogonal beamforming and zero-forced beamforming in the signal processing domain for communication system studies, and (v) to provide the reader with insights, future research directions and foreseen challenges.

It is important to be aware of the fact that many studies in the literature wrongly refer to the zero-forced beams as orthogonal beams, which may cause confusion. On the other hand, the zero forced beam phrasing used in this paper has a direct similarity with the zero-forcing terminology that is widely used in the array signal processing community as a precoding strategy [44]. The correspondence is valid in line-of-sight scenarios and when the beam spacing is large enough (more than the angular resolution of the beam,  $\approx \lambda/D$  where  $\lambda$  is the wavelength at the operating frequency and  $D$  is the



**FIGURE 2.** A generalized multiple beam former and two examples of implementations: using microwave circuit components [47] (on the left), and using Rotman-like lenses [48] (on the right).

array length) to guarantee sufficient antenna gain at the main beam positions [45], [46].

The rest of the paper is organized as follows. Section II is dedicated to the comparison of orthogonality and zero-forcing. In Section II-A, the concept of orthogonality in multi-beam antennas is revisited. In Section II-B, the difference between orthogonal beams and zero-forced beams is explained and illustrated. In Section II-C, the intriguing relation between orthogonality, beamforming network losses and reciprocity is discussed from the antenna perspective. In Section III and Section IV, many distinct examples are shown for orthogonal beamforming networks and flexible zero-forced beamforming strategies, respectively. In Section V, the use of orthogonal beamforming and zero-forced beamforming in a communication system is described via several key studies in the literature. In Section VI, future research directions and challenges on the topic are described. Finally, in Section VII conclusions are drawn.

## II. ORTHOGONALITY VS. ZERO-FORCING

### A. DEFINITION OF ORTHOGONALITY

In linear algebra, orthogonality is clearly defined. In Hilbert spaces [49], [50], two vectors  $\mathbf{a}$  and  $\mathbf{b}$  are orthogonal when their inner product  $\langle \mathbf{a}, \mathbf{b} \rangle = 0$ . For a multiple beam array antenna as shown in Fig. 2, this can be applied in the antenna beam port space, its radiating aperture space and its far field radiation beam space, which are Hilbert spaces linearly mapped from one to the other by beam forming and by

<sup>1</sup>In this respect it is worth noting that there are subtle differences in the mathematical terminology referred to inner product spaces and their representing matrices,  $\mathbf{T}$ , corresponding to the vectors projection on a complete orthonormal base which span all the space. If the inner product is defined with complex conjugation, the representing matrix of set of orthonormal vectors is a unitary matrix  $\mathbf{T}\mathbf{T}^H = \mathbf{I}$ . On the other hand, if the vectors in the inner product are not complex conjugated, the representing matrix of set of orthonormal vectors is an orthogonal matrix,  $\mathbf{T}\mathbf{T}^T = \mathbf{I}$ . In the following we will address the properties of orthogonal beams and relevant beamforming networks. Orthogonality of the beams and relevant excitations should not be confused with the orthogonality, in matrix sense, of the transfer block of the scattering matrix.

radiation/scattering operations. The norm  $\|\mathbf{a}\|$  of a vector, is derivable from  $\|\mathbf{a}\|^2 = \langle \mathbf{a}, \mathbf{a} \rangle$ . Using appropriate inner products,  $\|\mathbf{a}\|^2$  can represent the power applied (or received) at a beam port, reaching the antenna aperture surface or radiated by a particular beam. Linearity/superposition, reciprocity and energy conservation mechanisms can be included in these transformations, for microwave circuit or lens-based beamformers (as illustrated in Fig. 2) and taking into account losses as well as the gain of the amplifiers. Advantages and limitations of Rotman lens-based beamformers are discussed in [48]. Their main advantage over microwave circuit based ones is the bandwidth of the lens, due to its use of time delay rather than phasing. The larger insertion and spill over losses of the lens can be compensated by increased amplification at the element level.

Still, there is no standard (IEEE) definition of antenna beam orthogonality. Most past related work applies to fixed multiple beams, not to adaptive ones and authors vary in their definitions.

According to [41], true orthogonal beamforming is achieved with a lossless (ideally) simultaneous multiple beamformer, with zero (ideally) coupling between the beam ports. According to [51], a multibeam system is truly orthogonal when both criteria below are satisfied:

- (C-i) No internal coupling between antenna ports (generally termed as isolated/decoupled ports in the literature), which means that the pairwise complex dot-products of the excitation vectors for different beams are zero.
- (C-ii) No coupling or interference through radiated field, which means that the average complex inner product of the pairwise radiation patterns is zero.

The above beam coupling terminology and its relation to interference is rather confusing. If the beam weight vectors of two beams using the same radiators, are orthogonal, they can originate losslessly each from one of two separated beam ports, through a cascade of lossless, isolated and completely matched  $2 \times 2$  directional couplers, each equipped of a phase shifter. In this case, the inner product of the beams will automatically be zero.

It can be seen that the conditions, (C-i) and (C-ii), are connected through a Fourier Transform (FT) and thus, can be represented by the same mathematical constraint. Let us consider an aperture of length  $L$  with the one-dimensional aperture distribution of  $g(x)$ . In this case, the radiated field,  $F(u = \sin \theta)$  is computed as

$$F(u) = \int_{-L/2}^{L/2} g(x) e^{jkxu} dx \quad (1)$$

where  $k$  is the wavenumber.

As represented in Fig. 2, to generate two (similar-shaped) beams  $A, B$  with the peaks towards  $\pm u_0 = \sin \alpha$  for an arbitrary angle  $\alpha$ , the required aperture distributions are represented as

$$h_A(x) = g(x) e^{jkxu_0}, \quad h_B(x) = g(x) e^{-jkxu_0} \quad (2)$$

According to (C-ii), the orthogonality is achieved when the following relation holds (assuming  $u \approx \theta$  [41])

$$\int F_A(u) F_B^*(u) du = 0 \quad (3)$$

where  $F_A$  and  $F_B$  are the patterns corresponding to the excitation from the ports of the beams  $A$  and  $B$ , respectively. The sign “\*” denotes the complex conjugate. Note that this condition is valid for beams close to boresight and is linked to angles close to boresight, with the approximation  $\theta \approx \sin \theta$  being valid for small values of  $\theta$ . This also requires the beams to be narrow, so that the angular range considered with the main beam remains relatively close to boresight.

The patterns  $F_A$  and  $F_B$  in (3) are given by

$$\begin{aligned} F_A(u) &= \int_{-L/2}^{L/2} g(x) e^{jkx(u+u_0)} dx, \\ F_B(u) &= \int_{-L/2}^{L/2} g(y) e^{jky(u-u_0)} dy \end{aligned} \quad (4)$$

Then, (3) can be expressed<sup>2</sup> as [41],

$$\int_{-\infty}^{\infty} \int_{-L/2}^{L/2} \int_{-L/2}^{L/2} [g(x) g^*(y) e^{jku(x-y)} e^{jku_0(x+y)}] dx dy du = 0 \quad (5)$$

As the integral with respect to  $u$  gives a Dirac delta function, (5) can be written in a simpler form as

$$\int_{-L/2}^{L/2} g(x) g^*(x) e^{jkx2u_0} dx = 0 \quad (6)$$

It can be seen that (6) is the first condition of orthogonality (C-i) given in [51]. For array antennas, (6) can be interpreted in a discrete form as the complex dot-product of the excitation coefficients for the two beams  $A$  and  $B$ . Thus, (C-i) and (C-ii) represent the same concept, which lies behind the general definition of orthogonality in inner spaces.

By comparing with (1), it is observed that the integral in (6) shows the pattern of the aperture distribution of  $|g(x)|^2$  at the angular location  $2u_0$ . Therefore, to have the integral equal to zero, the inter-beam spacing of  $2u_0$  should be located at the null of the pattern corresponding to the aperture distribution of  $|g(x)|^2$ . For uniform illumination, i.e.  $|g(x)| = 1$ , the orthogonality condition results in the angular inter-beam spacing of  $n\lambda/L$  (for  $n = 1, 2, 3, \dots$ ). This is illustrated in Fig. 3.

There are several other key studies in the literature that focused on the theory of orthogonality in linear and planar multibeam antennas [50]. In [52], it was also concluded that to form simultaneous multiple beams in a lossless manner, the beams must be Hermitian orthogonal, which puts a strict constraint on the excitation coefficients. Later, in [53], [54], it was shown that the truly orthogonal beam positions must satisfy

<sup>2</sup>in the simplifying assumption that the integration of the radiation pattern can be extended outside the visible space,  $|u| \leq 1$ , to the reactive energy domain,  $|u| > 1$ .

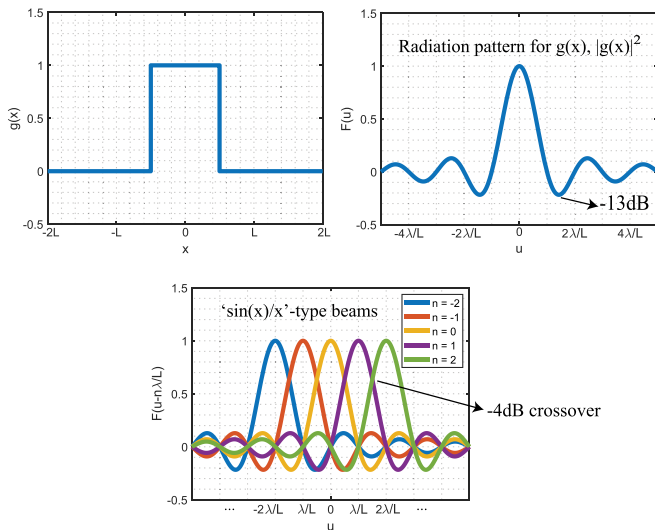


FIGURE 3. Orthogonality properties for uniform illumination.

certain reciprocal lattice conditions and that the phase steering vectors of the lossless beamforming network must correspond to the multi-dimensional Discrete Fourier Transform (DFT).

### B. DOES ORTHOGONAL MEAN ZERO-FORCED?

An important, and commonly misinterpreted, aspect of orthogonality in the recent publications on multibeam antenna systems for future wireless networks is the inter-beam interference. In order to achieve (mutually) interference free transmissions in a one beam per user multibeam system, it is required that, for each beam, signals incident from or transmitted towards the other beams' (peak) directions be zero. Due to the mixed terminology used in the literature, such interference-free beams are renamed as Zero-Forced (ZF) beams in this paper. For the case of uniform illumination with  $|g(x)| = 1$  discussed before, the orthogonal beams  $A$  and  $B$  are spaced in such a way that the spacing is coincident with the zeros of the basic radiation pattern  $F(u)$ . However, this is not always the case since the orthogonality condition (C-i,-ii) only define the beam coupling in terms of excitations' or patterns' overlap. For the complex-valued radiation fields used in (3), the fast variations in phase in the integrand may yield a zero result [55].

As a straightforward but not widely seen example, consider that the beam ports of orthogonal/decoupled/isolated beams  $A$  and  $B$  are connected to a 3 dB hybrid coupler and the two output ports are properly phased to obtain the new beam ports, let us define them as  $C = (A + jB)/\sqrt{2}$  and  $D = (A - jB)/\sqrt{2}$ . In this configuration illustrated in Fig. 4, the aperture distributions (in accordance with (2), for  $|g(x)| = 1$ ) for beam  $C$  and  $D$  are given by

$$\begin{aligned} h_C(x) &= (e^{jkxu_0} + je^{-jkxu_0})/\sqrt{2} \\ h_D(x) &= (e^{jkxu_0} - je^{-jkxu_0})/\sqrt{2} \end{aligned} \quad (7)$$

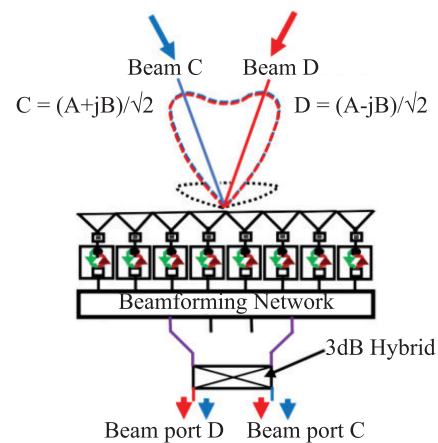


FIGURE 4. Formation of orthogonal (yet spatially overlapped) beams.

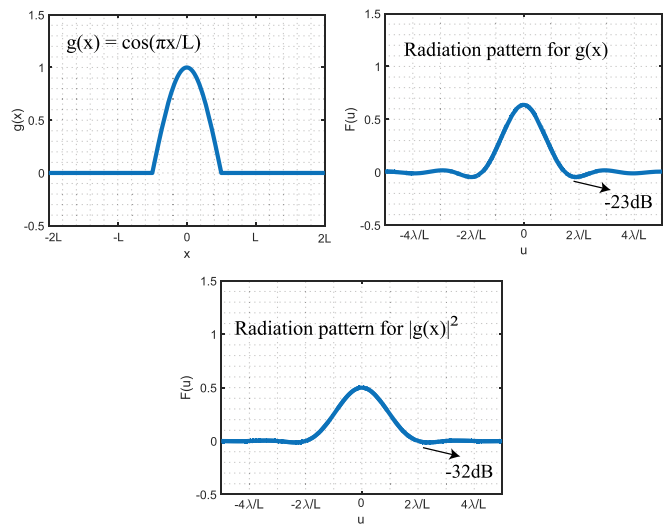


FIGURE 5. Orthogonality properties for cosine-type illumination.

In line with the formulation of (6), the new orthogonality condition for the two aperture distributions is expressed as

$$\int_{-L/2}^{L/2} \cos(kx2u_0)dx = 0 \quad (8)$$

which is again true when the beam spacing of  $A$  and  $B$ ,  $2u_0$ , is equal to an integer multiple of  $\lambda/L$ . Thus, the beams  $C$  and  $D$  are also orthogonal/decoupled/isolated, but this time not ZF. In fact, the two new beam ports produce the exact same gain patterns. A related reference on this is the dual mode antenna using shaped oversized reflectors [56].

Another example could be to assume a non-uniform aperture distribution (i.e.  $g(x) \neq 1$ ). For instance, the pattern nulls for  $g(x) = \cos(\pi x/L)$  correspond to  $u = (n + 0.5)\lambda/L$ , while the orthogonal beam spacing is equal to  $u = (n + 1)\lambda/L$ , for  $n = 1, 2, 3, \dots$ . This phenomenon is shown in Fig. 5. Following the same reasoning, it can be seen that truly orthogonal multiple beam formers that suppress the sidelobes generally

result in low beam crossovers (for the  $g(x) = \cos(\pi x/L)$  distribution, we obtain  $-23$  dB sidelobe level but with  $-9.5$  dB crossover level). Therefore, for the applications that require low sidelobes and high crossovers, lossy networks are generally used, and the losses are compensated by the power amplifiers at the element level [3], [8], [57].

Some disconnect has been observed between definitions and emphasis in the microwave/antennas and the digital communications/signal processing literature. This is apparent in the above discussion on orthogonal and zero-forcing beam forming. Array microwave beamforming networks (as shown in Fig. 1) can simultaneously produce multiple independently agile pencil beams (potentially with Rayleigh angular resolution), as well as reconfigurable shaped or multi-lobed beams. For each beam, weights with adjustable (usually quantified) amplitudes and phases, or time delays, can be computed and refreshed to meet service requirements of several co-channel and other users simultaneously. To meet these needs, microwave beamforming can generate beam excitation functions ranging from uniform (power efficient) amplitudes or tapered ones (for low sidelobes) with planar phase fronts for pencil beams to slightly tapered amplitudes with nearly planar phase fronts for zero forced beams, or synthesized amplitude and phase excitations for shaped or multi-lobed beams.

Unlike time shifters, phase shifters typically provide the same weight phases within each beam sub-band. This creates a problem in applications with multiple paths of unequal lengths, and therefore transit times, both in the receive and transmit modes. Creating, in the directions of the resolvable paths with the strongest signals, pencil beams, preferably zero-forced in all other co-channel path directions, and then time synchronizing, weighing and constructively combining the signals to and from those beam ports is a complex solution. Moreover, in the transmit mode, the loss of combining  $N_b$  beam RF signals into each antenna element power amplifier,  $10 \log(N_b)$ , becomes prohibitive for high numbers of users and multiple paths, and must be compensated by more power amplifier gain.

In principle, digital beam forming allows to implement any beam excitation function achievable with the analog beamformer, without its combining loss. In addition, digital beam forming, with more flexibility, allows to integrate in the precoded signals the time synchronisation and constructive combination of same user signals along their different paths as well as the zero-forcing to reduce co-channel interference. The generic analog solution, as described in Fig. 1, can indeed produce any amplitude/phase excitation required. Constraints are introduced by specific types of microwave beamformers, such as Butler matrices, etc., with the benefit that these specific solutions do not introduce recombination losses. Thus, one ends up with the typical trade-off between flexibility and performance, with higher flexibility leading to higher losses.

However, while analog beamforming solutions require one converter per beam, the digital solutions require one per element. When the number of elements is much larger than the

number of beams, this brings a clear penalty in the power consumption and cost, partly offset by the suppression of the phase shifters (although their consumption is generally quite small, but again magnified by the number of elements, this may be non-negligible). Hybrid beam forming, with a reduced number of RF up/down conversion chains, each connected to subsets of the complete array elements via analog precoders, cannot provide the same coverage performance as the fully-digital or even the analog beamforming but might be a competitive solution for the mid-term.

### C. ORTHOGONALITY, LOSSES, AND RECIPROCITY

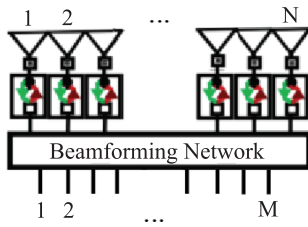
The strong constraints on the beam shape, side lobe level, beam pointing, etc. that are imposed by a theoretically lossless orthogonal beamformer design were discussed in Section II-A.

It is important to notice that all beamforming networks using matched and isolated  $n$ -port couplers and phase shifters are automatically producing orthogonal beams, if each beam port is connected to array elements and not to loads [58]. In these networks, there are three spaces: the beam port excitation space, the antenna/array surface illumination space and the beam space. Transformations between these are typically rotations. Thus, orthogonality in the beam port space is preserved at illumination and at beam levels. The square beamforming matrices with such property are commonly referred to as orthogonal beamforming matrices in the literature. On the other hand, there is no standard terminology used for the rectangular counterparts which are called orthogonal beamforming matrices, orthonormal beamforming matrices or beamforming matrices with orthonormal/orthogonal columns/rows in different sources.

The connection between orthogonality and losses is not obvious. One can maintain perfectly orthogonal (and ZF in the case of a Butler matrix) beams, while inserting equal attenuators before each element port. In other words, a beamforming network producing orthogonal beams will still maintain the orthogonality if attenuators are added at the element ports. Similarly, the attenuators can be placed at the beam ports, which do not need to be equal for the beams to remain orthogonal.

For reciprocity, let us consider a system/circuit that includes the receive beam ports,  $r = 1 \dots R$  and transmit beam ports,  $t = 1 \dots T$ . Let us focus on a cellular communication scenario with a base station and phone users. On transmit, the base station sends the signal  $a_t$  from each port  $t$ , resulting in a beam, or outgoing spectrum of plane waves all around. The user  $r$  receives one plane wave, of which he captures/receives some local power and a signal voltage  $b_r$  ( $b_r = \sum_t S_{rt} a_t$ , where  $S_{rt}$  is the base station to users transfer block of the scattering matrix of the system<sup>3</sup> and not that of the base station's beam forming network). On receive, it is the user  $r$  who transmits a signal  $a_r$ . The user does not send back a plane wave, as he had received, but his own beam pattern plane wave spectrum,

<sup>3</sup>known also as base station to user channel matrix.



**FIGURE 6.** A generalized multiple beam forming network.

of which the base station receives only one bit of one plane wave and beam port  $t$  captures some power and a voltage  $b_t$  ( $b_t = S_{tr}a_r$ ). Reciprocity says that  $S_{tr} = S_{rt}$ . If now we allow the user to change his direction vector seen from the base station, moving on a sphere around it, and pointing at it,  $S_{rt}$  becomes  $S_{rt}(u)$ , the transmit pattern of that base station beam port. Similarly on receive  $S_{tr}(u)$  will be the receive beam of port  $t$  of the base station. It is because, for each direction  $u$ ,  $S_{tr} = S_{rt}$  so that the transmit and receive patterns of the base station for a beam port  $t$  are the same. Thus, antenna transmit and receive operations are not reciprocal, although reciprocity applies to them. What would be really reciprocal of transmitting out a signal through an (array) beam would be to reverse its propagation flow, recreating in this receive mode the conjugate of the transmit (array) illumination back.

A detailed discussion on reciprocity and losses in terms of the behavior of the beamforming network in the transmit and receive modes was presented in [59]. As the most critical observation of the discussion, it was indicated that there is a possibility to have different levels of internal losses in the network depending on the operation mode, which is yet consistent with the reciprocity theorem, such that the internal losses represent the image of the radiated power variations in the transmit mode. This way, the link budget is balanced and reciprocity is verified. Thus, according to the definition proposed in [59], a lossless network in the transmit mode can be lossy in the receive mode. This is readily understood in the case of a  $M \times N$  beamformer (see Fig. 6), where  $M$  is the number of beam ports and  $N$  the number of radiating element ports with  $N > M$ . This beamformer may be a lossless and matched  $N \times N$  Butler matrix with  $(N - M)$  beam ports terminated with loads. The decoupling between beam ports and between element ports, and the orthogonality of the lossless  $N \times N$  beamformer result in lossless operation in transmit for the  $M$  active beams. In receive, some power from incident plane waves is dissipated in the loaded ports, unless the wave comes from one of the  $N$  beam peak directions.

At this point, it is useful to remember that a spectrum of plane waves (beam) is transmitted and that separate plane waves are received which might interfere at the beam ports or not. The apparent discrepancy between transmit and receive losses in [59] comes from the fact that transmitting from decoupled beam ports through a unitary (i.e. with Hermitian orthonormal columns) scattering matrix (with no loss of power

inside the matrix formed by couplers and phase shifters) generates orthogonal array illuminations and radiation patterns. All the input power is radiated, with or without illumination taper. On receive, because of reciprocity, only incident array illuminations that are conjugate of the transmitted ones can focus in the corresponding beam ports, without any loss in the other beam or terminated ports. This is clearly not the case when the incident illumination comes from a single uniform plane wave between beam peak directions or if the matrix creates an illumination taper. The transmit and receive gains are equal in all directions because the beam port receives the convolution (or discrete inner product) of the incoming plane wave array illumination generated on transmit by that beam port. The same phenomenon is seen with scan loss. On receive, from off-boresight with a large element/subarray, some of the incident intercepted power is reflected, whereas a perfect match on transmit can be achieved with all the power radiated, but the antenna is not lossy on receive. A perfect match is also achieved on receive if the incoming plane wave spectrum is the conjugate of the outgoing one on transmit (i.e. the radiation pattern).

Next in this section, a brief complementary mathematical description on the behavior of a multiple beam forming network is given. To prevent any confusion, it is worth to note the following: a feed network is a junction with  $N$  ports and an  $N \times N$  square scattering matrix  $S$  with elements denoted as  $S_{ij}$ . Some,  $N_b$ , are beam ports, some,  $N_e$ , are element ports (note that normally  $N_b \leq N_e$ ) and some,  $N_l$ , usually loaded, can be unused beam or coupler ports. If the network is lossless, it means that all power coming in comes out, whichever way it comes in, i.e. that all  $N$  vectors  $S_j$  (with components  $S_{ij}$ ,  $i = 1$  to  $N$ , possibly including non-zero  $S_{ii}$ 's, or non-zero  $S_{ij}$ 's between beam ports or element ports) are unit norm complex vectors. The beamforming network transmit and receive transfer matrices  $T$  (with elements  $T_{ij}$ ) and  $T^t$  (with elements  $T_{ji}$ ) relate signals coming out of the  $N_e$  radiating element ports,  $i$ , to signals applied to the  $N_b$  individual beam ports,  $j$ , and vice-versa, with simply  $T_{ij} = S_{ij}$ . Unlike its scattering matrix, the beamforming network transfer matrices involving only beam and element ports are rectangular if  $N_b < N_e$ . Then, while it is possible to have all power applied to one beam port come out at element ports, i.e.  $|T_j| = 1$  for beam ports  $j$ , it is clear that  $|T_i^t| = 1$  will not apply, because part of the power applied to element ports will come out of unused ports. The network with  $N_b + N_e$  ports only with the  $N_l$  unused ports excluded, will be lossy and will not have a unitary scattering matrix.

In the above, if the  $N_l$  unused ports are kept and treated as beam ports, then it becomes possible to have  $|T_j| = 1$  and  $|T_i^t| = 1$  ( $T$ ,  $T^t$  and  $S$  can be unitary matrices). As an example, let us take an  $8 \times 8$  Butler matrix (later given in Fig. 9) of which we decide to use only two beams, i.e. two beam ports (assume port 1 and 5 in Fig. 9) and 8 element ports. We can remove the 5 unused couplers and 2 phase shifters, then we will have 6 unused ports which are still part of the lossless junction. The  $16 \times 16$  scattering matrix is still unitary and users



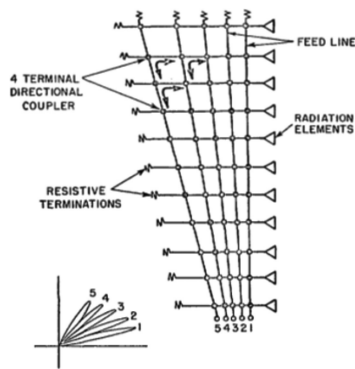


FIGURE 7. The Blass matrix topology [43].

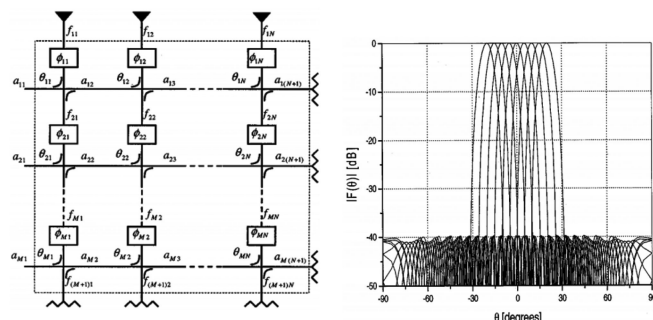


FIGURE 8. The lossy Blass matrix beam-forming network in [60] and corresponding array factor patterns.

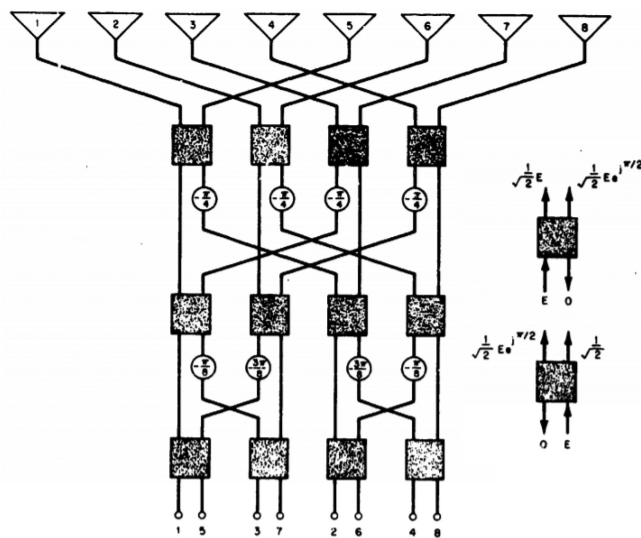


FIGURE 9. An 8-element Butler beamforming network [61].

at the peaks of beams 1 and 5 still have a two-way lossless operation. The power coming from other users will end up in the 2 beam ports and in the 6 unused (loaded) ports with no loss.

However, it is also commonly accepted that the loaded ports are not a part of the transfer matrix, and the matrix is not necessarily square. In other words, the number of input (beam ports) and output (element ports) may differ. In fact, as

explained above, the rectangular transfer matrix is a subset of the complete junction (square) scattering matrix. Therefore, all the information is not there, particularly for receive. More details on such a formulation strategy can be found in [59].

The above considerations on orthogonality, reciprocity, losses and zero-forcing are straightforward, but they are relevant and useful when passive linear analog multiple beamforming networks are used to establish initial designs and trade-offs for multiple beam array designs for transmission of multiple co-channel data streams. In practice, various hardware imperfections and errors from passive and active components and structures, such as amplitude and phase or time delay quantification, non-linearities, thermal effects, variable loading due to coupling for amplifiers or in-phase and quadrature-phase (IQ) imbalance in digital beamforming, will require careful study and calibration schemes to correct their impact on the quality of service.

### III. ORTHOGONAL BEAMFORMING NETWORKS

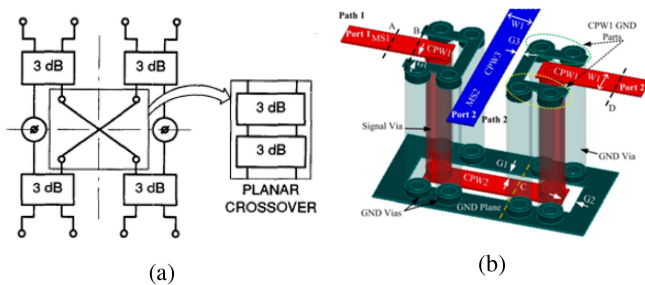
In this section, a review of the most widely known orthogonal beamforming networks is given from the circuit realization point-of-view.

First, the non-orthogonal Blass matrix was introduced in [43] with no constraints on the output excitations and resulting lossy behavior. A general Blass matrix architecture is given in Fig. 7. Due to the lossy behavior, such a network might be applicable in use cases where power consumption is not so critical [60]. A sample network formulation and resulting multibeam array pattern of a Blass matrix are shown in Fig. 8.

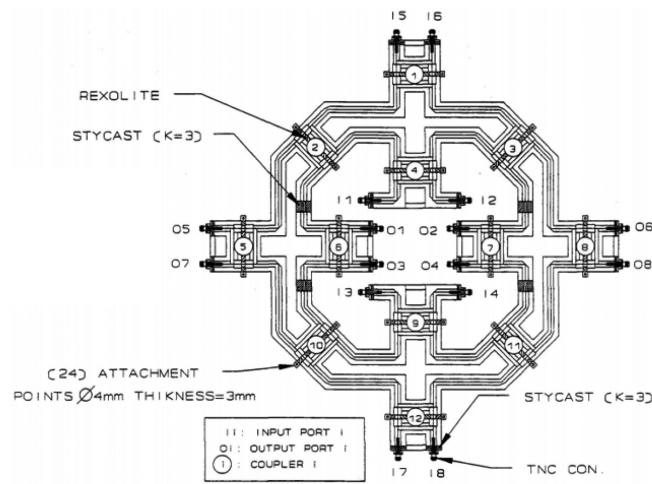
After the introduction of the Blass matrix, there was a growing interest towards the lossless orthogonal matrices [61]. Among them, the well-known conventional Butler matrix was introduced in [42], which typically has the same number of inputs and outputs, equal to an integral power of two. A sample implementation of a standard 8x8 Butler beamforming network is illustrated in Fig. 9.

The standard Butler matrices in planar configurations lead to crossovers in realization, the number of which gets higher with the size of the Butler matrix. The crossover function in a Butler matrix is usually addressed by using two cascaded hybrids [62], [63], which results in an incompact design and larger component count. Motivated by this, more compact single-substrate crossover designs with a broadband operation were also proposed in the literature [64], as seen in Fig. 10. However, single layer 4x4, 8x8 Butler matrix designs without any crossings have been successfully demonstrated as well [65] (see Fig. 11).

It is worth of note that non-standard or oversized Butler matrix realizations can be found in the literature which overcome these design rules. Rectangular Butler matrices with less beam ports than elements are easily obtained by suppressing unused couplers and terminating unused isolated ports [66], [67]. Moreover 3x3, 5x5, 6x6 etc. hybrid couplers and Butler matrices have been introduced in two-dimensional [68]–[71] and three-dimensional [58], [62] configurations. For better



**FIGURE 10.** Sample crossover designs: (a) cascaded hybrids [62], (b) microstrip to coplanar waveguide transition [64].



**FIGURE 11.** A single layer  $8 \times 8$  Butler matrix design at L-band without cross-overs [65].

visualization and understanding, several examples of the non-standard/modified Butler matrices are provided in Fig. 12.

For relatively higher frequencies, the use of Substrate Integrated Waveguide (SIW) technology (advantages and applications of which reviewed extensively in [72]) in the realization of Butler matrices has been proposed. In [73], a low-cost and wideband  $4 \times 4$  Butler matrix with no crossovers was designed and integrated with slot array elements on the same substrate for radar applications at 77 GHz. In [74], a compact, two-layer, wideband  $4 \times 4$  Butler matrix in SIW technology was proposed for Ku-band applications centered at 12.5 GHz. The two SIW-based Butler matrix designs in [73], [74] are shown in Fig. 13 for better visualization.

In addition to the conventional Butler matrices with one-dimensional beamforming, Butler matrix networks that are compatible with planar arrays and have the ability to do beamforming in both azimuth and elevation have been proposed in various types. One straightforward way to achieve this is to employ two orthogonal stacks of Butler matrices [75], [76], as illustrated in Fig. 14. An alternative method is the direct two-dimensional beamforming approach using two-dimensional coupler and crossover components [77], [78], which reduces the length of the network significantly. An example from [78] is given in Fig. 15. Such networks generate a square

(or rectangular) lattice of beams, which is not optimal in terms of the crossover between the beams. Recently, in [79], two-dimensional Butler matrices generating a triangular lattice of beams (see Fig. 16 for visualization) was proposed for minimizing the gain roll-off.

A few years later than the introduction of the Blass and Butler matrix, the Nolen matrix was proposed in [80] as an appealing alternative in the sense that it is lossless like the Butler matrix and realized with a serial feeding network without crossovers like the Blass matrix. The Nolen matrix presents a generalized form of an orthogonal multiple beamforming network as it has no limitation on the number of ports. The general architecture of a Nolen matrix is shown in Fig. 17.

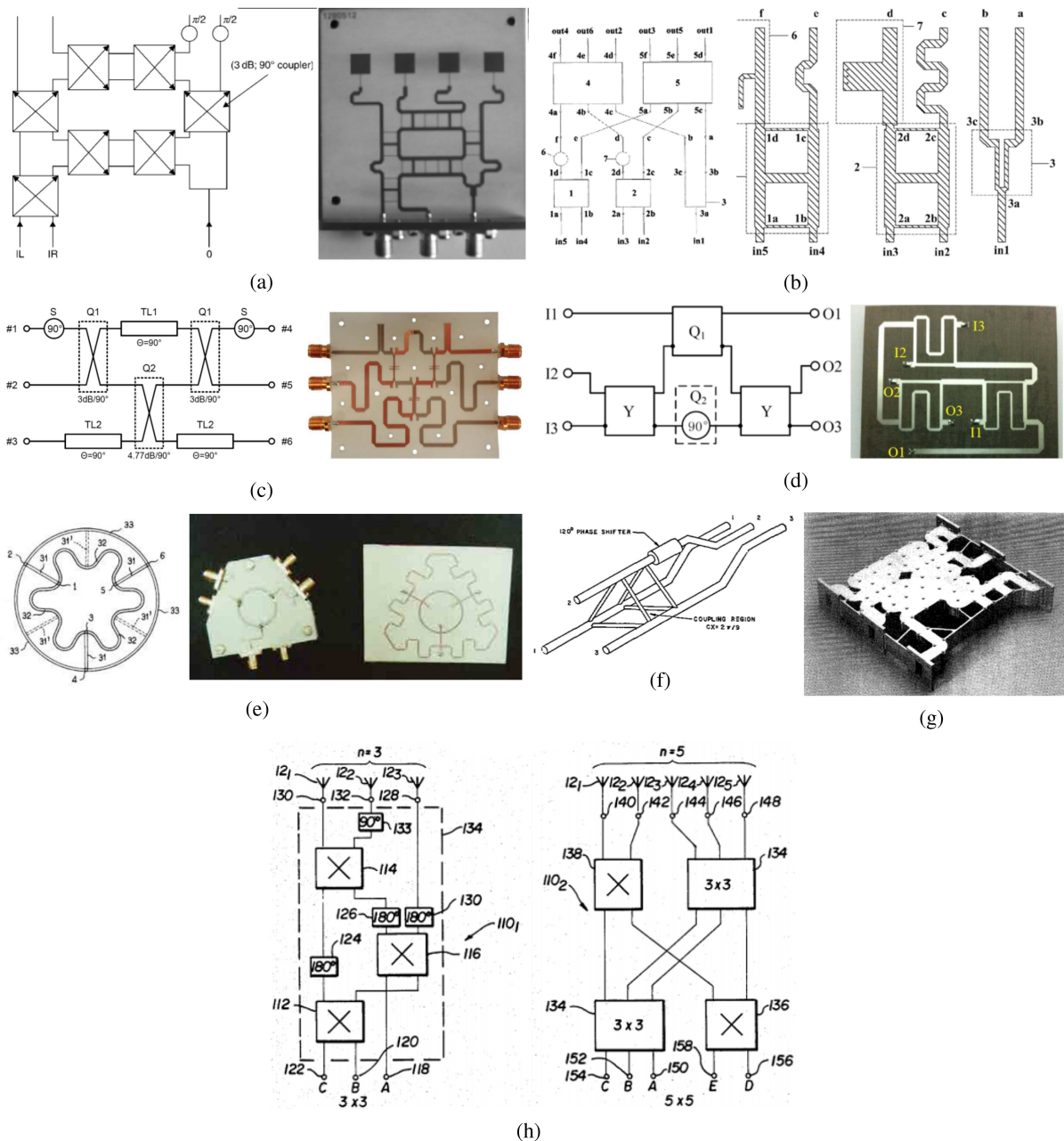
Although Nolen-like matrices can provide compact designs, they generally show a relatively narrowband response due to the dispersive behavior with unequal paths from ports to the antenna elements. A sample  $4 \times 4$  Nolen matrix [81] with the corresponding radiation pattern change in the frequency range of 2.1–2.3 GHz is shown in Fig. 18. Later, in [82], the design rules for cancellation of beam squint with frequency were discussed with a focus on a single-beam four element linear array. The extension of concept to Blass and Nolen matrices was also discussed and it was found that the main limitation originates from the different insertion phase of different directional couplers over frequency.

For higher frequencies, similarly to the Butler matrix implementations, SIW-based Nolen matrix designs have been examined. In [83], a Ku-band  $4 \times 4$  Nolen matrix (given in Fig. 19) was designed and experimentally validated in SIW technology for the first time. Furthermore, to address the beam squint issue, more “parallel” SIW-based Nolen matrices based on coupler delay compensation were proposed. An example is the broadband  $4 \times 4$  Nolen matrix presented in [84], the structure of which is shown in Fig. 20 for visualization. In a recent work [85], a compact modified Nolen matrix topology generalized to one-dimensional parallel switching matrices with an arbitrary number of beams was proposed. Its configuration and realization in SIW for a  $5 \times 5$  parallel matrix is shown in Fig. 21.

Besides the linear array implementations, as seen in the Butler matrices, two-dimensional Nolen matrix networks connected to planar arrays and generating multiple unique radiation beams on azimuth and elevation were also proposed. This function can be achieved by stacking and cascading multiple Nolen matrices. An example of such a design was presented in [47] and shown here in Fig. 22, in which six  $3 \times 3$  are combined to generate 9 beams.

There are several other works that studied size reduction in Nolen matrices. It was recently shown in [86] that one way to achieve this is to realize each coupler with lumped-elements (i.e. capacitors and inductors). A  $3 \times 3$  ultra-compact (with 90% size reduction as compared to the conventional designs) Nolen matrix proposed in [86] is provided in Fig. 23 for concept visualization.

In addition to the above-mentioned various types of beamforming matrices with different design goals, modified Butler

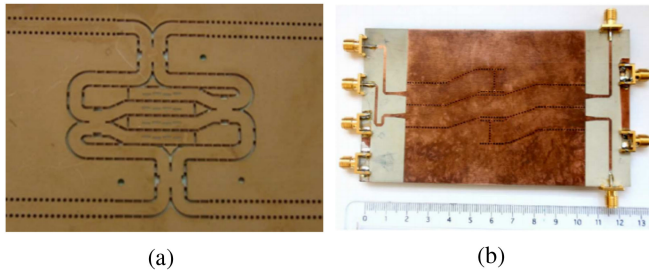


**FIGURE 12.** Non-standard or oversized modified Butler matrix realizations: (a) a 3×4 matrix with a broadside beam [66], (b) a 5×6 matrix [67], (c) a broadband 3×3 matrix [68], (d) a 3×3 matrix [69], (e) a printed 3×3 matrix [70], (f) a 3×3 3D matrix [58], (g) a 3×3 matrix in waveguide [62], (h) a 5×5 matrix architecture [71].

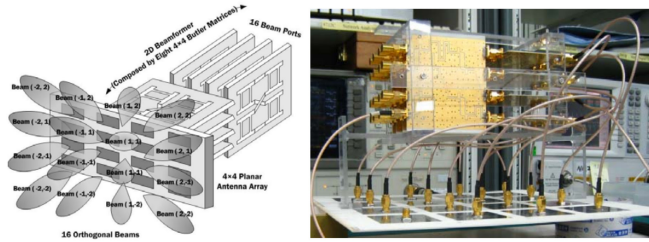
and Nolen matrix configurations with reduced side lobes were presented in the literature. In the case of the low side lobe Butler matrices, unbalanced power dividers with proper phase compensations are attached to a standard Butler matrix. This way, the number of outputs is increased and desired amplitude law is achieved. A sample modified 4×8 Butler matrix design with less than -20 dB side lobes was presented in [87]. The design and pattern results are shown in Fig. 24 for visualization. Another example with an amplitude-tapered Nolen matrix beamforming network was presented in [88] and shown here in Fig. 25. In this case, the matched loads (located at the

bottom four ports in the matrix topology in Fig. 25) dissipate some power in the receive mode with a plane wave incidence. The level of “losses” depends on the angle of incidence (for further information, please see Fig. 5 in [88]).

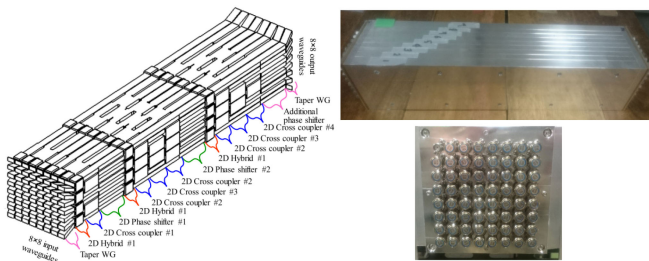
The multiple beam forming circuits revisited in this section are mainly implemented on printed circuit boards with bulky couplers, except the reduced size matrix realizations. However, for highly-integrated RF systems, there is a growing need for silicon-based implementations. Low-cost is absolutely key for the industry involved, and planar printed antenna with silicon-based RF and processing circuits, like RFCMOS and



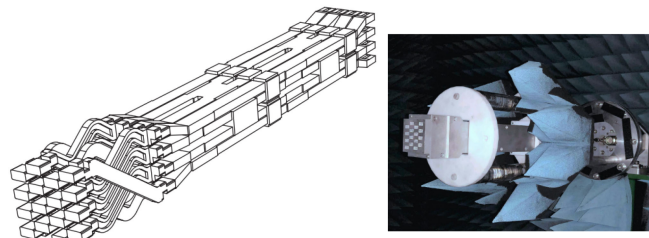
**FIGURE 13.** Butler matrix realizations in SIW technology: (a) a low-cost and wideband  $4 \times 4$  matrix at 77 GHz feeding a slot antenna array [73], (b) a compact and wideband two-layer  $4 \times 4$  matrix at 12.5 GHz [74].



**FIGURE 14.** Two stacks of  $4 \times 4$  Butler matrices at 2.45 GHz generating 16 beam states [75].

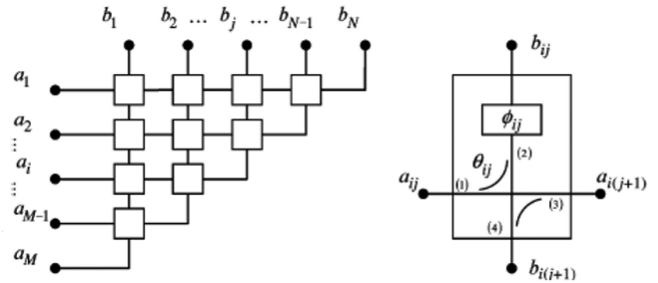


**FIGURE 15.** A  $64 \times 64$  two-dimensional hollow waveguide Butler matrix at 20 GHz using two-plane short slot couplers [78].

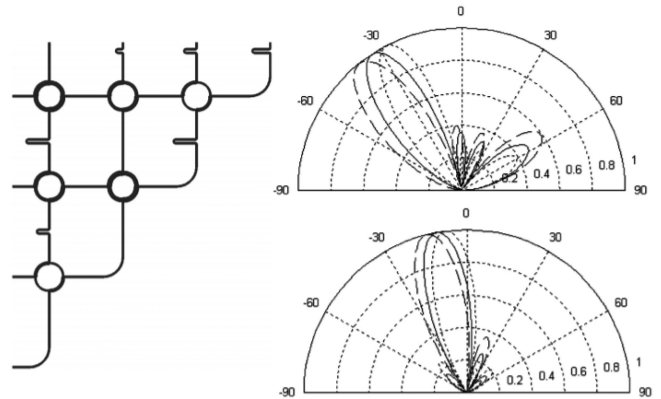


**FIGURE 16.**  $16 \times 16$  connecting network and two-dimensional Butler matrix at 20 GHz generating a triangular lattice of beams [79].

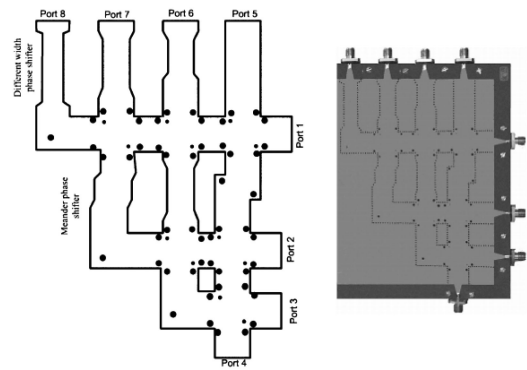
BiCMOS, are a clear option [35]. To this aim, several successful monolithic CMOS Butler matrix designs were reported in the literature [89]–[91]. Fig. 26 gives a visualization of several relevant examples. As mentioned in Section V-B, such matrices can be used as building blocks in a hybrid beamforming system, as proposed in [92] (active sub-arrays with analog beam forming chips, with digitally elaborated sub-array weights at baseband with true time delays). This greatly



**FIGURE 17.** The Nolen matrix topology [81].



**FIGURE 18.** The  $4 \times 4$  Nolen matrix in [81] and its radiation pattern (for two beam ports) in the range of 2.1–2.3 GHz.



**FIGURE 19.** A  $4 \times 4$  Nolen matrix designed in SIW technology at 12.5 GHz [83].

reduces the number of converters and decreases processing complexity as compared to fully-digital beamforming, while improving the field-of-view, gain and operational bandwidth as compared to the existing analog and hybrid multibeam systems [3].

It is convenient to mention here that multiple approximately-orthogonal beams can also be realized by using lens-fed arrays. A well-known network is the Rotman lens [93], [94]. The original Rotman lens configuration is given in Fig. 27. The interested readers are referred to [95] for a detailed discussion on the development of Rotman lenses. An example lens with an almost half-cosine aperture

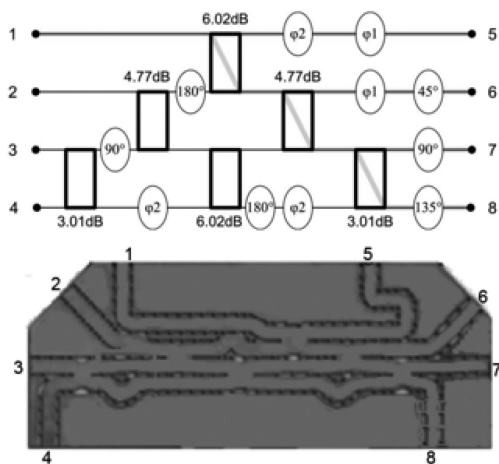


FIGURE 20. A 4 × 4 SIW broadband Nolen matrix designed at 77 GHz [84].

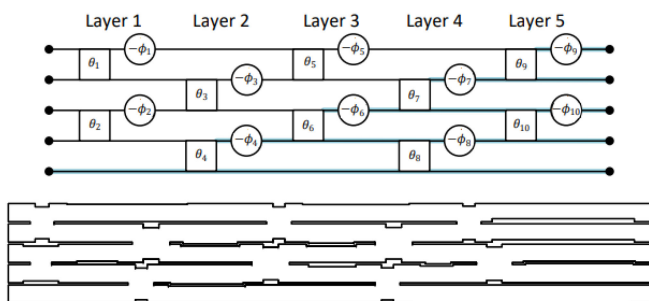


FIGURE 21. A 5 × 5 parallel switching matrix configuration and its realization in SIW technology at 76 GHz [85].

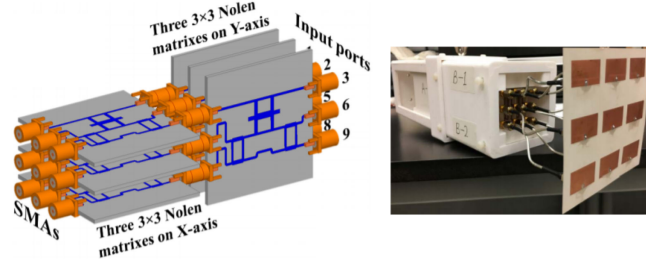


FIGURE 22. The two-dimensional Nolen matrix network connected to a 3 × 3 planar array generating 9 beams at 5.8 GHz [47].

distribution,  $-23$  dB sidelobes and  $-9.5$  dB beam crossovers (recall Fig. 5) was given in [41]. There are several works in this domain which trade off some loss to achieve a higher beam crossover level. Such non-orthogonal beam sets can be achieved with the addition of beam overlap networks as in [96], [97]. More recently, a modified Rotman lens (called as the Fourier Rotman lens) that realizes the Discrete Fourier Transform (DFT) operation (similar to the Butler matrix) was introduced in [98], [99]. The designed Rotman lens structure in [99] and the pattern results are shown here in Fig. 28 for illustration purposes. Such networks are useful as they can provide relatively lower complexity as compared to a Butler

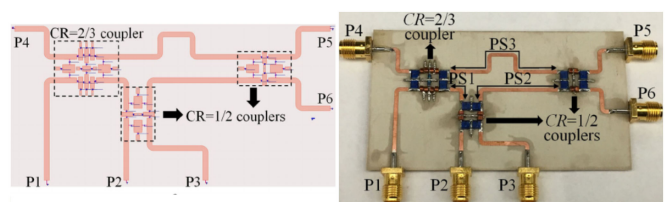


FIGURE 23. An ultra-compact 3 × 3 Nolen matrix design at 1 GHz with lumped-element couplers [86].

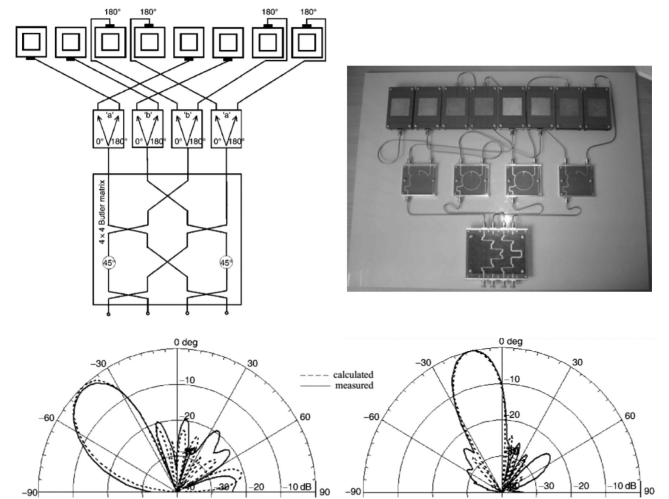


FIGURE 24. A modified 4 × 8 Butler matrix for reduced side lobes [87].

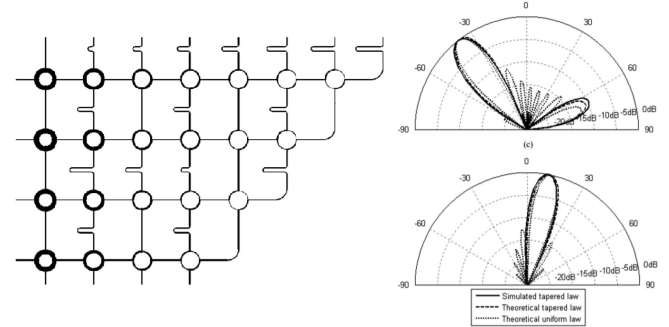
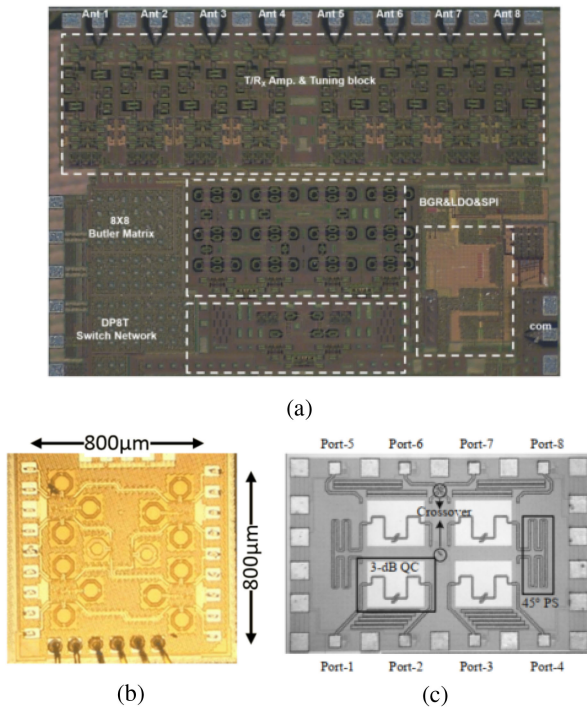


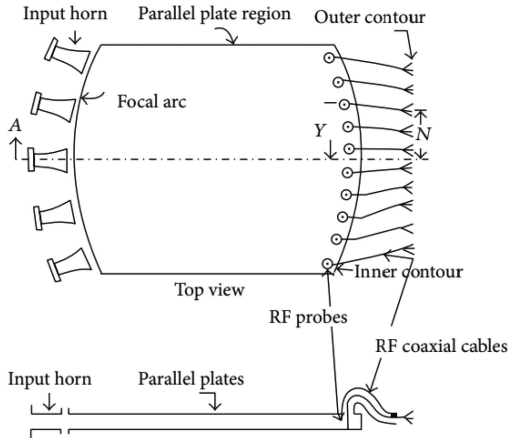
FIGURE 25. A modified 4 × 8 Nolen matrix for reduced side lobes [88].

matrix counterpart when the order of the matrix becomes large.

As a final remark in this section, it is useful to note that the term “quasi-orthogonality” is also encountered in the literature. In these works, the phrase “quasi-” is used to indicate that the orthogonality condition in (6) is partially satisfied as the integral gives a small, yet non-zero, value. In [100], a multibeam mm-wave reflector antenna design was presented for 5G communications. Five quasi-orthogonal beams were synthesized based on the orthogonality criterion such that the inner product of the radiation pattern of two adjacent beams is less than a desired threshold. Fig. 29 shows an illustration of the design and the radiation patterns. In [102], the use of mutually orthogonal radio beams that have a low beam coupling

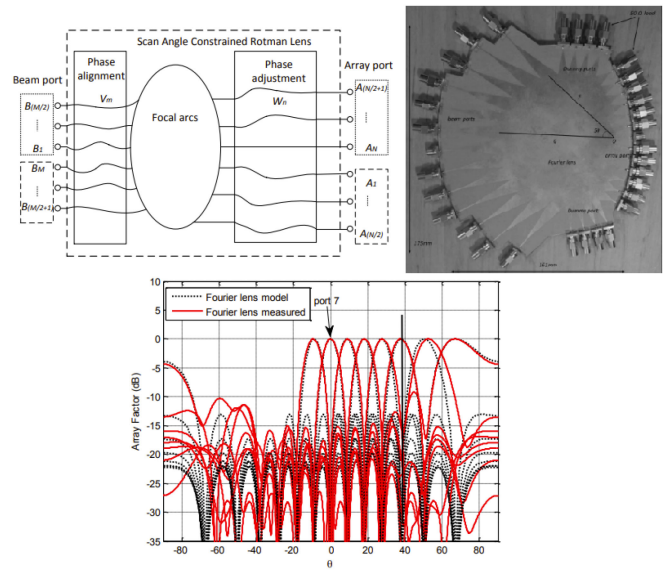


**FIGURE 26.** CMOS Butler matrix implementations: (a) a 28 GHz  $8 \times 8$  Butler matrix beamformer in 65 nm CMOS process [89], (b) a 9-13 GHz  $4 \times 4$  Butler matrix in 32 nm CMOS Silicon-on-Insulator [90], (c) a 24 GHz  $4 \times 4$  Butler matrix in 0.18- $\mu\text{m}$  CMOS technology [91].

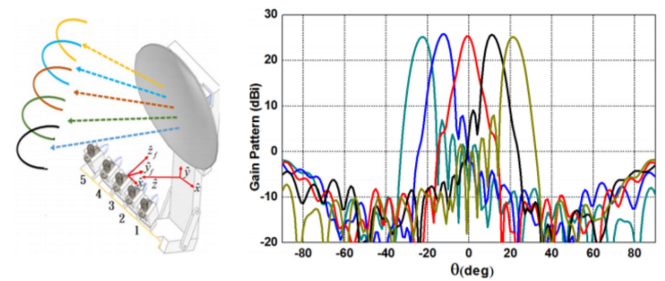


**FIGURE 27.** The original Rotman lens scheme [94].

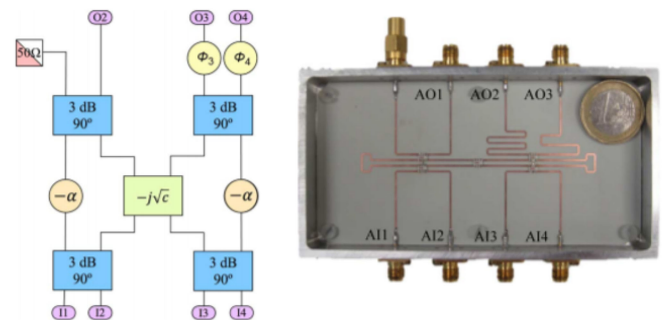
factor was proposed for radio communication systems. The orthogonal beam space was divided into orthogonal channels, each composed of a few (2 or 3) orthogonal beams. In [101], considering the intrinsic limitations of the orthogonal lossless networks (i.e. maximal number of beams is equal to the number of radiating elements, with an imposed relation between them), the possibility of using dissipative (or lossy) networks with more beams than the elements was investigated. The beamforming network is given in Fig. 30 for visualization. In the final design, a four quasi-orthogonal switching multibeam network (with three orthogonal beams at a desired elevation



**FIGURE 28.** The Fourier Rotman lens [99].



**FIGURE 29.** A multibeam mm-wave reflector antenna with 5 quasi-orthogonal beams [100].

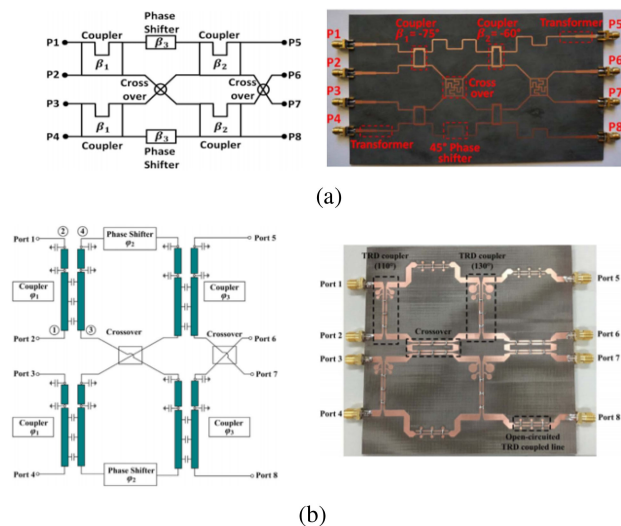


**FIGURE 30.** A  $4 \times 3$  dissipative network for 3 orthogonal beams at desired elevation and 1 beam at broadside [101].

angle and a beam at broadside) was proposed for a triangular subarray of three radiating elements.

#### IV. FLEXIBLE ZERO-FORCED BEAMFORMING NETWORKS

In Section II-B, it was explained that orthogonal beamforming does not necessarily result in zero-forced beams which is crucial to cancel the interference at the beam ports. In Section III, it was seen that standard Butler matrices (or equivalent networks) provide simultaneous orthogonal and



**FIGURE 31.** Flexible phase difference  $4 \times 4$  Butler matrices: (a) [103], (b) [104].

zero-forced beams, yet with strict conditions on the beam spacing.

Several relatively flexible Butler matrices were proposed in the literature [103], [104] which are able to move all Butler beams together without losing the orthogonality. The flexibility is very limited as the beam spacing is kept the same. Besides, the potential use of tunable couplers were mentioned, but not implemented. Two examples of such “relatively-flexible” Butler matrices are given in Fig. 31 for visualization.

In this section, different examples from the literature on the realization of the flexible zero-forced beams are given. Zero-forcing precoding under the domain of digital signal processing is covered in Section V.

In [105], a wideband multibeam MIMO receiver with 4 reconfigurable zero-forced beams was designed (in 22-nm fully-depleted silicon-on-insulator technology) with a flexibly programmable analog beamforming by exploiting a vector modulator. The aim was arbitrary analog interference rejection by spatial notch filtering. The proposed beamforming architecture is given in Fig. 32 for better visualization.

In [106], a generic multibeam architecture (given in Fig. 33) was presented for a satellite ground terminal which is capable of using multiple zero-forced beams to simultaneously communicate with multiple different orbital satellites operating at the same frequency band. Broadly, it was mentioned that the beamforming network can provide fixed, reconfigurable or dynamic beams for tracking and be built in an analogue or digital fashion.

In [107], the zero-forced beam generation concept was verified experimentally by using two well-separated pencil beams in the azimuthal plane at the receiver side. The experimental setup is as shown in Fig. 34. Following the estimation of the angular locations of the transmitters, the dedicated beamforming networks (labeled as OBFM in Fig. 34) were adjusted to receive interference-free information from the

two transmitters exploiting the same time-frequency resource block. Later, the same authors extended their study with an application of zero-forced vertical beams in an indoor communication scenario [108].

In [28], a partially overlapped 64-67 GHz four-element hybrid beamforming receiver fabricated in a 130-nm SiGe BiCMOS process was presented. The null steering performance was demonstrated for the reception of two concurrent streams with two overlapped clusters of elements. The realization of the proposed beamformer architecture and the formation of two simultaneous zero-forced beams are shown in Fig. 35 for visualization.

In [109], a 10 GHz 65-nm CMOS four-element digital beamforming receiver front-end was reported with spatial cancellation of co-channel interferers (of more than 20 dB) at RF. The beamformer topology and its implementation are shown in Fig. 36. The concept was also extended to the cancellation of multiple spatial interference signals which may occur in more realistic scenarios. In [110], a 4-element 0.1-to-3.1 GHz digital beamforming receiver prototype was implemented in 65-nm CMOS technology and the formation of multiple arbitrary notches was demonstrated. The proposed array configuration is given in Fig. 37. A similar example can be found in [111], where a scalable 0.1-to-1.7 GHz spatio-spectral-filtering 4-element digital receiver array implemented in 65-nm CMOS technology was presented.

In [29], a 28-/37-GHz hybrid beam forming receiver (in 65-nm CMOS technology) was designed with four antenna inputs and two baseband output streams. A new technique called time-multiplexed least-mean-square (LMS) was introduced in an RF beamformer for the first time to perform symbol-by-symbol LMS adaptation of the antenna pattern. The adaptation technique was then used to steer both the main lobe and a null when two beams are simultaneously directed to the receiver from two transmitters. The corresponding beam former design and beam patterns are shown in Fig. 38. In [30], the same authors presented a 25–30 GHz fully-connected hybrid beamforming receiver (in 65-nm CMOS technology) with eight antenna inputs and two baseband output streams. To demonstrate the interference cancellation capability, a signal was applied to one element, and a phase shifted version was applied to another element. By combining the two channels, a peak-to-null ratio of 20 dB was achieved.

It is worth to mention that several existing and most related multibeam examples with the demonstration of zero-forced beams have been included here. The list can be broadened with other antenna systems using chipscale multibeam front-ends [112], [113] or hybrid architectures employing analog beamforming integrated circuits [114]–[116], as they can be used for the purpose of flexible zero-forcing with proper array element excitations.

## V. ORTHOGONAL BEAMFORMING AND ZERO-FORCING IN COMMUNICATION SYSTEM STUDIES

In this section, as a complementary and closely related subject, the focus will be more on the signal processing aspects in

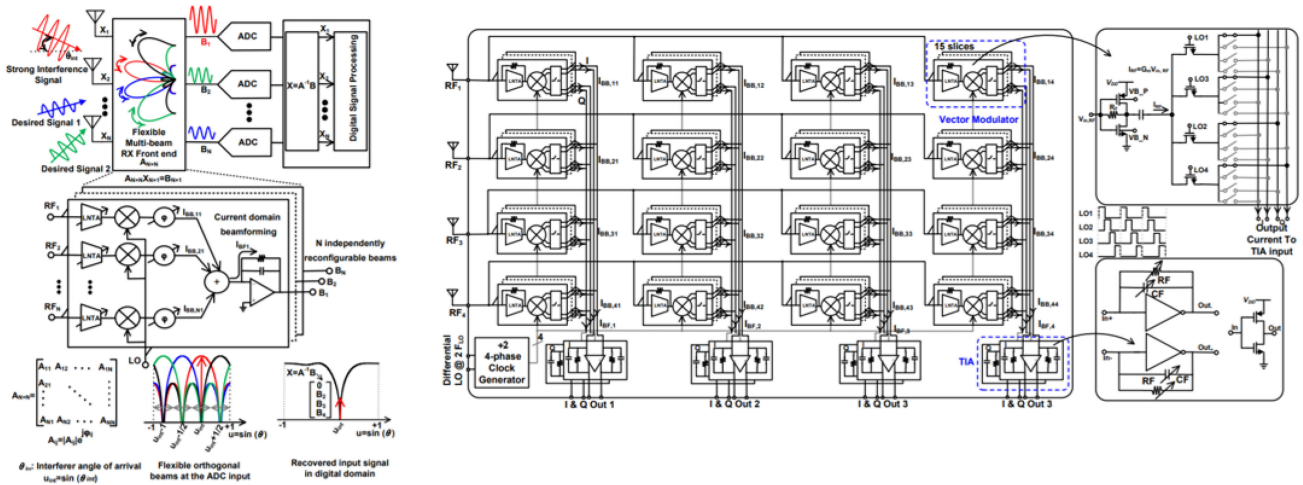


FIGURE 32. A 4 × 4 MIMO receiver with 4 reconfigurable zero-forced beams [105].

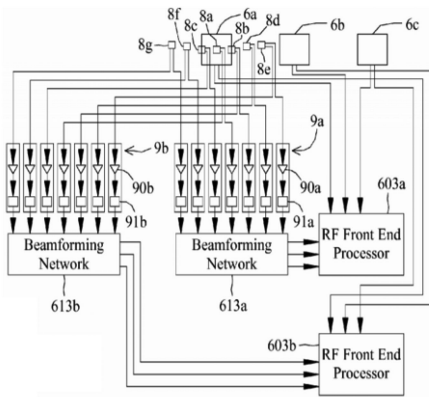


FIGURE 33. Multiple zero-forced beam forming network for a satellite ground terminal [106].

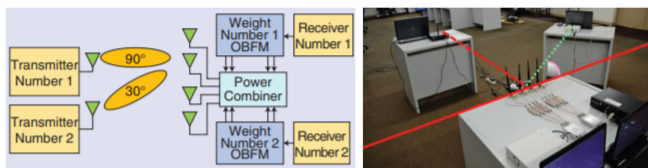


FIGURE 34. Zero-forced beam forming setup for a communication system serving two simultaneous co-frequency users [107].

a communication system employing orthogonal and/or zero-forced beams.

### A. THE USE OF ORTHOGONAL BEAMS

It is known that for conventional precoders (matched filtering, zero-forcing, minimum mean square error), once the user scheduling is performed, the transmitter computes the excitation weights to determine each user’s achievable Signal-to-Interference-plus-Noise-Ratio (SINR). In other words, to precisely know the SINR of a user (especially in the case of totally random selection of users), the beamformers need to be

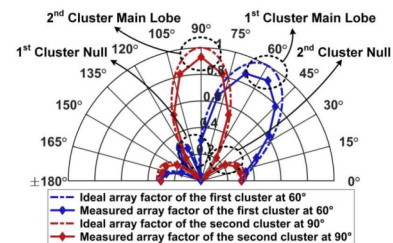
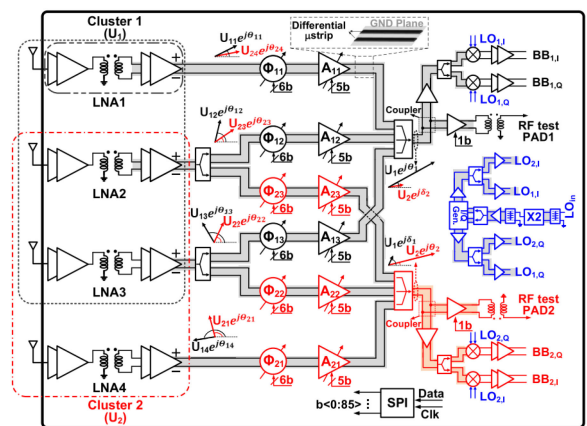
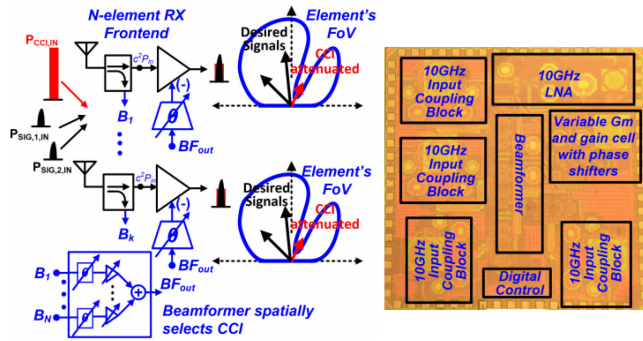


FIGURE 35. The realization of a partially overlapped 64-67 GHz four-element hybrid beamforming CMOS receiver and a sample formation of two zero-forced beams [28].

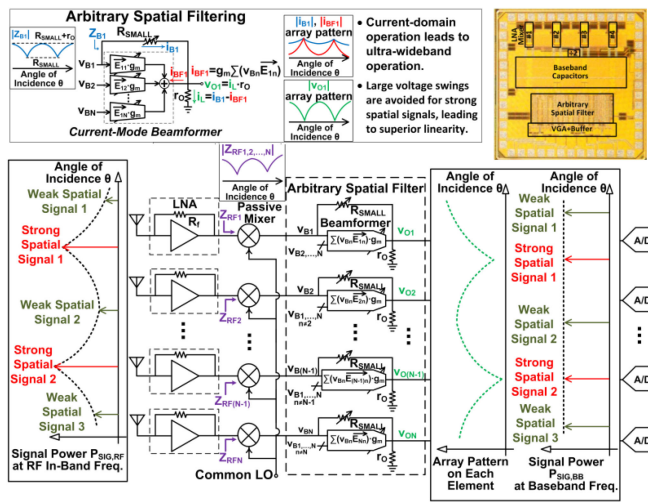
computed first, which is generally a time consuming and computationally expensive operation. Therefore, with orthogonal beamforming in the signal processing area, it is proposed that the transmitter will use an orthogonal set of excitations for which the inter-user interference and the precise knowledge of SINR can be easily obtained. This way, the aim is to find the optimal user set easily, without computing the complex beamformers.

Following this motivation, in [117], an Opportunistic Space Division Multiple Access (SDMA) algorithm (OSDMA) was

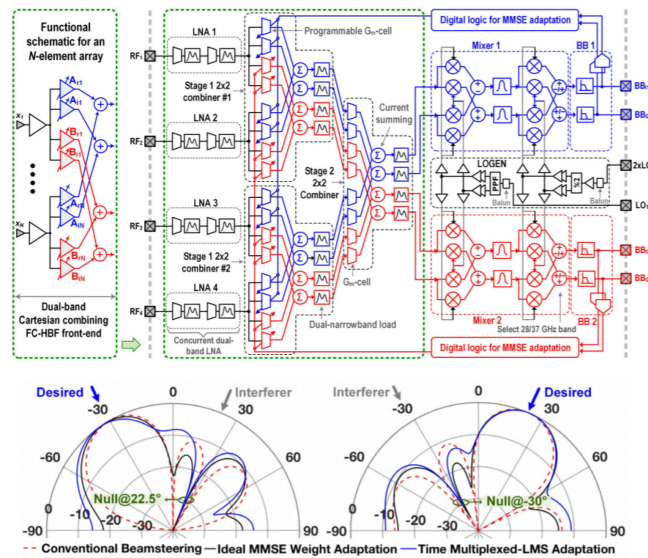




**FIGURE 36.** A 10 GHz four-element digital beamforming CMOS receiver front-end with out-of-beam co-channel interference cancellation at RF [109].



**FIGURE 37.** A 4-element 0.1-to-3.1 GHz digital beamforming CMOS receiver forming multiple arbitrary spatial notches [110].



**FIGURE 38.** A 28-/37-GHz hybrid beam forming CMOS receiver with four antenna inputs and two baseband output streams and the corresponding beam forming patterns assuming a uniform linear array under different beam forming strategies [29].

proposed for MIMO broadcast channels in which the base station sends a number of random orthonormal beams using as many random beamformers as the transmit antennas. Each user feeds back its best beam and its corresponding SINR to the base station. Based on the received SINR's, the base station schedules transmissions to a part of the users. Such a technique was proven to be efficient for a large number of active users. In [118], [119], the idea of random orthogonal beamforming was extended with an adaptive scheme to dynamically control the number of active beams according to the operation environment, which is seen to be useful in sparse networks with low number of users or in the case of uneven spatial distribution of the users.

In [120], the use of orthogonal beamforming vector constraint for transmission (referred to as orthogonal linear beamforming, OLBf) was exploited in MIMO broadcast channels using SDMA. Tailored to that, a low-complexity (as compared to the exhaustive search) user selection technique was proposed to achieve the highest sum-rate, based on the exact knowledge of SINR at the transmitter with the orthogonal transmission. Two different algorithms (based on the selection strategy of the basis vector of the orthonormal precoding set obtained via Gram-Schmidt orthogonalization) were proposed to jointly decide on a smart set of the orthogonal vectors to be assigned to the associated users. Through parametric studies, the advantages of OLBf in terms of the user scheduling complexity and throughput as compared to the zero-forcing and matched filtering precoding were shown for low-SNR and low number of user scenarios.

Complementary to the work in [120], the optimal choice of the unitary orthogonal vectors regarding the number of feedback bits, the amount of latency, and the sum capacity was investigated later in [121]. The use of partially orthogonal beamforming weights was proposed in [122].

In [123], orthogonal beamforming in space division multiplexing broadcast channels was studied under the name per user unitary and rate control (PU2RC), which was proposed as a 3GPP-LTE standard. The main feature of PU2RC is the orthogonal precoding constraint where each user selects a beamformer weight from a codebook of multiple orthonormal bases, which, from the antenna-system perspective, presents major similarities with the fixed and orthogonal grid-of-beams approach. This is also referred to as limited-feedback OS-DMA (LF-OSDMA) in the relevant literature. Through simulation results, the superior throughput performance and robustness of PU2RC as compared to zero-forcing beamforming was demonstrated for sufficiently large number of users.

Following the work in [123], an improved (in terms of fairness) user scheduling approach for orthogonal beamforming was presented in [124]. The method is based on a round-robin scheduling which uses the CSI to arrange the transmit power among the users so that a given SINR constraint is fulfilled. Through the system simulations, the approach was found to be effective for time-varying channels and in real-time multimedia traffic.

## B. THE USE OF ZERO-FORCED BEAMS

Due to their simplicity from the system management perspective, switched-beam schemes based on fixed orthogonal and zero-forced beams have been widely used in the conventional satellite communication systems and cellular systems. Some examples include ISM-band wireless communications [125] (fixed or hybrid fixed/adaptive beamforming in elevation/azimuth), wideband code division multiple access networks [126] and mm-wave communications [127], [128].

Considering the performance limitations of purely fixed orthogonal and zero-forced beams, several architectures with beam reconfigurability were proposed in the literature. Based on a combination of fixed orthogonal beams and a reconfigurable beamforming network, relatively low-complexity hybrid architectures were introduced, adapted to a direct radiating array [129], [130], to a focused array with multipoint amplifiers [129] and to overlapped subarrays [131]. In [132], an efficient, modular and scalable design solution for reconfigurable beamforming networks was given.

Recently, hybrid beamforming approaches have also been studied that combine the fixed orthogonal zero-forced multiple beamforming with digital precoding for performance improvement [133]. Some examples from the literature employ a large Butler network [134] or a Rotman lens [48] connected to a baseband processor.

In a recent work on a hybrid mm-wave system [135], a single-user scenario was considered with one data stream per each RF chain. The problem of simultaneous clustering of antenna elements both at the transmitter and receiver with optimal precoding associated to the clusters was addressed. The aim was to maximize the signal to inter-subarray (i.e. between the selected transmit and receive subarrays) interference ratio. It was mentioned the study is limited to point-to-point communication and isolated sensor network applications, yet with potential to be extended to multi-user environments including inter-user interference aspects.

In [136] a hybrid beamforming system based on interleaved subarrays was proposed for multibeam multiplexing in arbitrary beam directions. The aim was to achieve much narrower beam widths with the interleaved structure, while avoiding high side lobes and grating lobes through varying/fixed antenna spacing with an associated analog weighting of the subarray elements and a simple digital inter-subarray coding scheme. Through joint optimization of analog and digital coefficients in sample linear and planar arrays, the formation of two and three simultaneous angularly well-separated beams with low interference levels (below  $-20$  dB) were reported, which can potentially be improved by implementing zero forcing based optimization techniques.

Yet, the full-power of zero-forced beams appears when created via fully digital beamforming. For example, in [137], multiple zero-forced beams (up to 5, with full frequency reuse at the same time) were formed digitally at the transmitter side for indoor multiuser wireless communications with a deterministic channel in line-of-sight and in the presence of active

scattering devices. In [138], the “one-beam-per-user” concept was proposed in satellite communications with dispersed and flexible coverage. In this scenario, a large direct radiating array complemented with an adaptive beamforming mechanism provides a maximal gain available by the aperture towards the user of interest, which can improve the link budget by up to 3–4 dB at the beam crossings as compared to the traditional fixed multibeam coverage. At the same time, the co-user (or jamming station) interferences can be rejected effectively with the adaptive zero-forced beamforming. To achieve this in a relatively efficient way, three iterative and fast-converging algorithms (namely conjugate-gradient beamformer, auxiliary-vector beamformer and random beamspace processing) were proposed, the performances of which depend on the presence and the number of interferers. In [139], zero-forced beams were used to optimize the quality-of-service in line-of-sight based multiuser SDMA systems. The impact of decreasing the number of zeros (i.e. complexity reduced zero-forcing [140]) on the SINR and the processing burden was also investigated.

There are also studies which combine the advantages of orthogonal beamforming and zero-forcing precoding in an alternating fashion depending on the changes in the environment. Such a strategy was shown to outperform each individual scheme in all ranges of SNR and number of users [141].

Flexible zero-forced beams are, in general, hard to synthesize and maintain. The system non-idealities, quantization errors and fabrication tolerances may have a strong negative impact on the level of nulls, which may significantly deteriorate the interference suppression performance of the multibeam array. In this case, the throughput performance of the system approaches to the one of the matched filtering, or simple adaptive beam steering [45]. Furthermore, in [142], it was pointed out that the performance (in terms of the bit error rate) of the switched-beam system approaches that of the adaptive beam steering (with more computational complexity) when the spatial beam spacing becomes very small. That is why the grid-of-beams technique with large crossover level is proposed (mostly with digital beamforming due to high losses in the case of analog beamforming), which selects the excitation vector from a pre-defined set [143]–[146]. In the grid-of-beams approach, the beams are not required to be orthogonal or zero-forced and, in principle, an arbitrary number of beams can be generated. In [57], the grid-of-beams method was investigated as a pragmatic approach for massive MIMO in broadband communication satellites. The detailed system analyses presented include critical radio resource management aspects such as the optimal beam spacing, minimum user spacing, per user power normalization, array element spacing, number of array elements, number of simultaneously active users and resource slicing in time and frequency sub-bands. The appealing performance-complexity trade-off with the proposed approach was shown in comparison to the conventional precoding algorithms.

It is worth noting that the study in [57] and most of the current work on the matched filtering precoding or grid-of-beams

use periodic array topologies with high side lobes on average. However, it can be seen that literature is very rich regarding the aperiodic array synthesis techniques [147]–[155] and the multidisciplinary system advantages (on the link quality, computational burden, thermal management etc.) brought by the array layout irregularity [45], [156]–[162]. Motivated by this, recently, the idea of employing optimized coverage-specific aperiodic multibeam array antenna layouts with low side lobes ( $< -30$  dB) and optimal power efficiencies was proposed in [163]. The set of beams generated by such multibeam array layouts can be considered as “quasi zero-forced” due to the very low level of interference between its beam ports. Thus, the computational complexity in precoding can be greatly reduced (with a small impact on the throughput) by using, instead of the zero-forced beams, grid-of-beams or matched filtering combined with the quasi zero-forced beam generating layouts.

## VI. FUTURE RESEARCH DIRECTIONS, BENEFITS AND CHALLENGES

To improve the system capacity and network quality in the future satellite and cellular communication systems, it is required that: (i) the antenna gain is maximized and the total interference is minimized simultaneously towards several beam positions using the same frequency, and (ii) power amplifier efficiency is kept close to optimum to limit consumption. Therefore, the ultimate aim with a flexible multiple beamforming architecture is to generate beams minimizing cumulated co-channel interference of users with arbitrary (but well separated) angular positions over some bandwidth, while approaching to a (theoretically) lossless realization with maximum efficiency. Such an architecture would remove the combining losses of current analogue multibeam active array solutions, while optimizing the array gain, SINR's and power efficiency.

The potential applications include:

- Future GEO, MEO and LEO satellite payloads for communication to and from mobile or fixed users with high frequency re-use and power efficiency requirements.
- Users of the above communication satellites, in particular for connection to 5G and 6G systems, for which higher gain and therefore agile narrower beam(s) will be needed (e.g. individual satellite tracking and smooth-handover functionality in large LEO (OneWeb, Starlink) or MEO (O3b) constellations).
- Gateways, using one or a few multiple beam array(s) instead of one pointable reflector per satellite.
- Base stations and mobile users of the future 5G and 6G infrastructures.
- Radar systems, in particular to suppress multipath and interferer signals.

Although there are several works that focused on modified Butler matrices with relatively flexible phase differences, to the authors knowledge, these techniques allow all the Butler beams to be moved together; otherwise, the peak gain of one beam no longer corresponds to the nulls of all the other beams

and large interference may occur. Therefore, there is some interest in synthesizing flexible zero-forced beams, realized with a low-loss and non-dispersive matrix, and with reconfigurability on the multiple beamforming network. The major challenges foreseen are in the synthesis of such a beamforming matrix (including practical design factors, mutual coupling), realization in low cost integrated circuit technology and calibration of the network.

## VII. CONCLUSION

In summary, multiple beam generation networks are reviewed from the antenna, electronics and communication system perspectives, with a focus on the distinct properties of orthogonal and zero-forced beamforming. As the terminology is generally confused in the literature, the precise definitions of orthogonal and zero-forced beamforming in multibeam antennas are given. The intriguing relation between orthogonality, losses and reciprocity is explained.

Various examples of orthogonal and flexible zero-forced beamforming networks are shown. Some useful insights on their operation principles, design/implementation, applications and advantages/disadvantages are provided. The review is also tailored to the system level studies including the signal processing aspects.

Furthermore, the applicability of the current technologies in the next-generation satellite and cellular communication applications is discussed. It is pointed out that the demanding high-gain, low-interference requirements of the future wireless systems call for the development of low-loss, non-dispersive and reconfigurable multiple zero-forced beamforming matrices that can be realized in low-cost integrated circuit technology.

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