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Electro-Optic beamforming in seamless wireless-to-photonic receiver arrays

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Abstract—This paper develops the theory behind electro-optic beamforming in seamless wireless-to-photonic receiver arrays. We show that the modulated phase of a traveling optical carrier signal in a wireless-to-photonic receiver array is governed by the direction of wireless reception, forming the basis for seamless electro-optic beamforming. We define a parameter referred to as the electro-optic array factor, which essentially governs the shape of the resultant electro-optic beam pattern. The analytical and physical interpretations behind maximas and nulls in seamless electro-optic beamforming are presented, and the design equations for controlling the electro-optic beam pattern are derived. In addition, the inadvertent nature of the presence of secondary electro-optic maximas is highlighted. Two distinct configurations of seamless electro-optic beamforming i.e. parallel- k and anti-parallel- k modes are introduced and compared. Further, a seamless electro-optic beamforming configuration that enables spatial-to-wavelength division multiplexing is proposed towards the end of the paper. The core theoretical contributions presented in this paper are crucial for further investigation and practical realization of seamless electro-optic beamforming in the next-generation communication systems like B5G and 6G.

Index Terms—Seamless wireless-to-photonic receiver, Seamless electro-optic beamforming, Electro-optic array factor, Electro-optic phase matching, Spatial-to-wavelength division multiplexing.

I. INTRODUCTION

The proliferating demand for ultra-high bandwidth based applications has been driving the pursuit of Beyond 5G (B5G) [1]–[3] and 6G [4]–[8] communication networks. The relatively uncongested mmWave and THz regimes [9] are being considered for the implementation of B5G and 6G. However, wireless transmission losses due to free-space fading and atmospheric attenuation become dominant at such high-frequency regimes. This can cause a major hindrance in practically implementing high data-rate based long-distance communication links. This challenge can be tackled by integrating short-range wireless links with long-range photonic links [10]–[12]. Such communication systems are called wireless-to-photonic links. In such communication systems, a wireless signal serves short-range cellular areas called pico/femto cells [13]. Wireless-to-photonic receivers [14]–[17] located within such pico/femto cells capture such wireless signals and map them to the equivalent optical domain. The resultant optical signals are then carried away by long-range optical fiber networks. Such optical signals can be re-converted back to the wireless domain by photonic-to-wireless converters [18]–[25].

Wireless-to-photonic links carry the potential for making high data-rate based long-distance transmission practically viable.

Wireless-to-photonic converters consist of conversion units called electro-optic modulators [26]. Over the last few years, electro-optic modulators based on Lithium Niobate (LiNbO_3) have proven to be worthy candidates, owing to their appreciably high electro-optic conversion efficiency [27]. Quite recently, an electro-optic modulator based on a thin-film LiNbO_3 with an undercut structure was presented [28]. In [29], a hybrid integration of Silicon photonic components and thin-film LiNbO_3 was demonstrated. A novel transmission line architecture in a thin-film LiNbO_3 platform was proposed, designed and simulated [30]. A breakthrough experimental demonstration for overcoming the voltage-bandwidth tradeoff in LiNbO_3 -based modulators using multi-structured electrodes was reported in [31].

Conventional wireless-to-photonic receivers consist of electronic RF-downconversion chains, in addition to the electro-optic modulation block. The received high-frequency wireless signal is down-converted to the electrical baseband regime. An electro-optic modulator then modulates an optical carrier signal using the electrical baseband signal. However, the presence of an electronic RF-downconversion chain might increase the system latency. Therefore, such types of receivers will not be suitable for the ultra-fast future-generation communication systems [32]. Antenna-coupled electro-optic modulator based wireless-to-photonic receivers, making no use of electronic RF-chains, have much lower latency compared to the conventional ones. We refer to them as seamless wireless-to-photonic receivers. Very recently, a W-band gap-embedded patch antenna array integrated with a non-linear electro-optic polymer was reported [33]. In [34], an antenna-based electro-optic modulation technique using ferroelectric crystals with polarization-reversed in a LiNbO_3 substrate was proposed. An extensive investigation of an electro-optic modulator based on multiple-quantum-well structures was presented in [35]. The importance of seamless wireless-to-photonic receivers in photonic electric-field sensing based applications have also gained prominence lately [36]–[38]. Extensive research is also being pursued on plasmonic based seamless wireless-to-photonic receivers in the recent times [39]–[41].

One of the salient features of high-capacity based communication systems is digital modulation, where digital baseband data is encapsulated within high-frequency carrier signals. Recently, we reported several novel implications of mapping digital phase-levels from the wireless domain to the photonic domain in slotted patch antenna array-based seamless wireless-to-photonic receivers [42]. Another salient feature of high-

capacity based modern communication systems is beamforming. Extensive research has been carried out on beamforming in conventional wireless receivers. For instance, Intelligent Reflecting Surfaces (IRS) based beamforming scheme has been sought out by researchers in the recent times [43]–[45]. A rigorous discussion on elevation beamforming for cellular networks using full-dimension massive MIMO antenna arrays was reported in [46]. A detailed survey on hybrid beamforming techniques in 5G has been presented in [47]. However, the aspect of beamforming in wireless-to-photonic receivers remains unexplored so far. Therefore, it is imperative to develop the core theory behind seamless electro-optic beamforming.

This paper introduces seamless electro-optic beamforming. We show that the slotted patch antenna array based seamless wireless-to-photonic receiver investigated in [42] can function as a seamless electro-optic beamformer. In section II, the fundamental theory behind seamless electro-optic beamforming has been introduced. The analytical expression of the instantaneous phase of an electro-optically modulated optical wavefront traveling in a seamless wireless-to-photonic receiver is derived. We show that the electro-optically modulated phase of an optical wavefront is governed by the wireless reception angle, thereby forming the basis for seamless electro-optic beamforming. Section III derives the analytical conditions for obtaining electro-optic maximas and nulls. We show that electro-optic maximas can be oriented along any desired direction by suitably choosing the array periodicity of the receiver. We also show that multiple inadvertent secondary electro-optic maximas exist in some scenarios. The physical interpretations behind the presence of maximas and nulls in seamless electro-optic beamforming have been explained. Section IV classifies seamless electro-optic beamforming into two categories: parallel- k and antiparallel- k modes. We show that in anti-parallel- k mode, the array is more compact, and the electro-optic beamwidth is broader, as compared to parallel- k mode. We also show that no secondary electro-optic maximas exist when the seamless electro-optic beamformer operates in the anti-parallel- k mode. Based on this observation, we propose a seamless electro-optic beamforming configuration that enables spatial-to-wavelength division multiplexing. The novel contributions reported in this paper will lay the foundation of seamless electro-optic beamforming for future-generation communication systems like B5G and 6G.

II. SEAMLESS ELECTRO-OPTIC BEAMFORMING

This section will introduce the fundamental theory behind seamless electro-optic beamforming. The structure of an antenna array based seamless wireless-to-photonic receiver is shown in Fig. 1(a). The receiver consists of centrally slotted metallic patch antennas of dimension $W \times L$, cascaded with array periodicity D . Additionally, a single-mode LiNbO₃ optical waveguide is routed under the array. The individual array elements are suitably designed to receive a y -polarized wireless signal $E_{mw}(t)$. The free-space phase-constant of $E_{mw}(t)$ is $k_{mw} = \omega_{mw}/c$. Here ω_{mw} is the frequency of $E_{mw}(t)$ and c is the speed of light in air. We consider $E_{mw}(t)$ to be obliquely incident on the receiver array, at an angle α with the z -axis,

as shown in Fig. 1(a). The free-space vector phase-constant of $E_{mw}(t)$ can be expressed as $k_{mw} \sin \alpha \hat{x} + k_{mw} \cos \alpha \hat{z}$. Therefore, the expression of $E_{mw}(t)$ incident on the receiver array located at $z=0$ is,

$$E_{mw}(t) = E_m \sin(\omega_{mw}t - k_{mw} \sin \alpha x) \quad (1)$$

where E_m is the electric field-strength of $E_{mw}(t)$. When $E_{mw}(t)$ excites the receiver array, an electric-field $E_{slot}(t)$ is generated in the slots of the individual array elements.

Fig. 1. (a) The structure of a 2-element slotted patch antenna array based seamless wireless-to-photonic receiver. (b) A 30GHz electric-field $E_{slot}(t)$ of strength $|E_{slot}| = 6.49$ kV/m is induced in the $1\mu\text{m}$ wide slot of a $2.9\text{mm} \times 3.57\text{mm}$ patch antenna upon receiving a y -polarized 30GHz wireless mmWave signal of strength $|E_m| = 1\text{V/m}$. (c) Variation of $|E_{slot}|$ induced in a single array element with respect to the direction of wireless reception α .

We designed a patch antenna array, with each array element having a $1\mu\text{m}$ wide centrally located slot. A single-mode LiNbO₃ optical waveguide of dimension $343\text{nm} \times 183\text{nm}$ was routed under the array. The effective optical refractive index η_{op} of the optical waveguide was computed to be 1.73. The physical length L of the array elements were chosen to be around 3.57mm , so as to achieve 30GHz wireless reception. The width W of the array elements were selected as 2.9mm , as per Eq. (19) of Appendix A, in order to ensure optimum seamless electro-optic phase modulation [42] for 30GHz wireless reception. The designed structure was then excited by a y -polarized 30GHz plane-wave $E_{mw}(t)$ of strength $E_m = 1\text{V/m}$ in HFSS. Consequently, it was found that a y -polarized 30GHz electric-field $E_{slot}(t)$ of strength $|E_{slot}| = 6.49$ kV/m gets induced in the slots of the individual array elements, as shown in Fig. 1(b). This implies that upon reception, the incident wireless signal-strength gets enhanced by a factor of 6490 in this design. The received field-strength $|E_{slot}|$ in the individual array elements is a function of the direction of reception of $E_{mw}(t)$, as shown in Fig. 1(c).

When the angle of wireless reception is α , the phases of the electric-fields induced in two successive array elements separated by D , differ by $k_{mw} \sin \alpha D$ radians. Therefore, the instantaneous phase-difference between the electric-fields induced in the n^{th} and $(n-1)^{\text{th}}$ array element is,

$$\angle E_{slot^n}(t) - \angle E_{slot^{n-1}}(t) = -k_{mw} \sin \alpha D \quad (2)$$

TABLE I
PARAMETERS AND THEIR NOTATIONS

Sr.	Notation	Parameter
1	$E_{mw}(t)$	Incident wireless mmWave E-field
2	$E_{slot^n}(t)$	Induced mmWave E-field in the n^{th} array element
3	$E_{op}(t)$	Optical carrier wavefront
4	η_{op}	Effective optical refractive index of waveguide
5	r_{33}	Pockel's electro-optic coefficient of waveguide
6	L	Individual array element length
7	W	Individual array element width
8	D	Array periodicity
9	T_W	Time for $E_{op}(t)$ to travel W inside waveguide
10	T_D	Time for $E_{op}(t)$ to travel D inside waveguide
11	α	Angle of reception of $E_{mw}(t)$
12	$\theta_{op^N}(t)$	Net phase of $E_{op}(t)$ in N-element receiver array
13	ψ	Electro-optic phase-matching factor
14	$\delta\theta$	Electro-optic array factor
15	α_{pm}	Direction of primary electro-optic maxima
16	α_{sm}	Direction of secondary electro-optic maxima
17	α_n	Direction of electro-optic null

The instantaneous electric-field $E_{slot^n}(t)$ induced in the n^{th} array element is,

$$E_{slot^n}(t) = |E_{slot}| \sin \left(\omega_{mw} t - (n-1) k_{mw} \sin \alpha D \right) \quad (3)$$

The electric-field induced in an array element causes an instantaneous shift in the effective refractive index, η_{op} , of the optical waveguide segment located under it, by virtue of Pockel's effect [48]. The instantaneous refractive index shift $d\eta_{op^n}(t)$ [48] of the waveguide segment located under the n^{th} array element is,

$$d\eta_{op^n}(t) = -\frac{r_{33}}{2} \eta_{op}^3 E_{slot^n}(t) \quad (4)$$

where r_{33} is the Pockel's electro-optic coefficient of LiNbO₃ [49]. When a y -polarized optical carrier wavefront $E_{op}(t)$ travels a distance dx through the n^{th} array element, it gets electro-optically phase modulated by $E_{slot^n}(t)$. The phase $d\theta_{op^n}(t)$ acquired by $E_{op}(t)$ by virtue of seamless electro-optic phase-modulation is,

$$d\theta_{op^n}(t) = k_{op} d\eta_{op^n}(t) dx \approx k_{op} d\eta_{op^n}(t) \frac{c}{\eta_{op}} dt \quad (5)$$

where k_{op} is the phase-constant of $E_{op}(t)$. The net instantaneous phase $\theta_{op^N}(t)$ acquired by $E_{op}(t)$ in an N-element receiver array has been derived in Appendix B. The expression of $\theta_{op^N}(t)$ is,

$$\begin{aligned} \theta_{op^N}(t) &= \sum_{n=1}^{n=N} \int_{t+(n-1)T_D}^{t+(n-1)T_D+T_W} k_{op} d\eta_{op^n}(t) \frac{c}{\eta_{op}} dt \\ &= \delta\theta |E_{slot}| \cos \left(\omega_{mw} t + \frac{(N-1)\psi}{2} \right) \end{aligned} \quad (6)$$

where $\delta\theta$ is the electro-optic array factor, ψ is the electro-optic phase-matching factor, T_W is the time elapsed for $E_{op}(t)$ to travel W i.e. width of a single array element, and T_D is the time elapsed for $E_{op}(t)$ to travel D i.e. the distance between

two successive array elements. The expressions of ψ and $\delta\theta$ are given in Eq. (7) and Eq. (8), respectively.

$$\psi = k_{mw}(\eta_{op} - \sin \alpha)D \quad (7)$$

$$\delta\theta = -\frac{k_{op} \eta_{op}^2 r_{33} c}{\omega_{mw}} \left(\frac{\sin \left(\frac{N\psi}{2} \right)}{\sin \left(\frac{\psi}{2} \right)} \right) \quad (8)$$

The electro-optic array factor $\delta\theta$ essentially represents the physical quantity of electro-optic modulation-depth. It can be noted from Eq. (7) and Eq. (8) that $\delta\theta$ is governed by the direction of wireless reception α . In other words, the extent by which $E_{op}(t)$ gets electro-optically phase modulated is dependent on α . We shall see later that this forms the basis of seamless electro-optic beamforming. It must however be highlighted that the directional dependence of $\delta\theta$ is not governed by the directional gain-profile of the array elements. This can be verified from Eq. (8), which shows that $\delta\theta$ is independent of $|E_{slot}|$. Therefore, it can be concluded that the directional dependence of $\theta_{op^N}(t)$ persists even if the receiver array is assumed to have isotropic gain.

Fig. 2. Schematic representation of (a) Conventional analog beamforming performed in the electrical-amplitude domain, and (b) Electro-optic beamforming performed in the optical-phase domain.

It is important to highlight the fundamental difference between conventional analog beamforming [50] and seamless electro-optic beamforming before moving on to the next section. As depicted in Fig. 2(a), conventional analog beamforming functions in the electrical-amplitude domain. In conventional analog beamforming, the resultant beam pattern depicts the directional dependence of the received signal-strength in a wireless receiver array. On the other hand, seamless electro-optic beamforming functions in the optical-phase domain, as depicted in Fig. 2(b). In seamless electro-optic beamforming, the resultant beam pattern indicates the directional dependence of the electro-optic modulation-depth $\delta\theta$.

The notations of the various parameters which are referred to in this paper are enlisted in Table I. In the upcoming section, we will discuss several characteristics of seamless electro-optic beamforming in detail.

III. SHAPING THE ELECTRO-OPTIC BEAMPATTERN

As discussed earlier, the electro-optic array factor $\delta\theta$ is inherently governed by the direction of wireless reception α . Naturally, there will exist certain directions along which $\delta\theta$ is maximum. We refer to those as the directions of electro-optic maximas. The optical carrier wavefront $E_{op}(t)$ gets optimally phase-modulated by wireless signals received along the directions of electro-optic maximas. Likewise, it is also possible to get electro-optic nulls along certain directions. The optical carrier wavefront $E_{op}(t)$ remains blind to the reception of wireless signals along the directions of electro-optic nulls. In this section, we will discuss the analytical and physical interpretations behind electro-optic maximas and nulls in detail.

A. Electro-optic maximas

1) *Analytical interpretation:* From Eq. (8), it can be found that $\delta\theta$ is maximum when the electro-optic phase-matching factor ψ is an integral multiple of 2π . Since, the electro-optic phase-matching factor is a function of D and α , from now on we will denote it as $\psi(D, \alpha)$.

$$\psi(D, \alpha) = k_{mw}(\eta_{op} - \sin \alpha)D = 2\pi m \quad m = 1, 2, 3.. \quad (9)$$

The above equation represents the condition of electro-optic phase-matching. In practical scenarios, it might often be necessary to have maximas oriented along the desired directions. We refer to those as the directions of primary maximas, $\alpha = \alpha_{pm}$. To have a primary maxima oriented along α_{pm} , the condition of electro-optic phase-matching must be satisfied. From Eq. (9), it can be found that electro-optic phase-matching can be achieved for wireless reception along α_{pm} by choosing the array periodicity $D = D_{\alpha_{pm}}$ as,

$$D_{\alpha_{pm}} = \frac{2\pi}{k_{mw}(\eta_{op} - \sin \alpha_{pm})} \quad (10)$$

Fig. 3. (a) Obtaining a Primary Electro-Optic (EO) maxima along α_{pm} by setting the array periodicity as $D_{\alpha_{pm}}$. (b) The prescribed choices of array periodicities $D_{\alpha_{pm}}$ corresponding to various directions of primary maximas α_{pm} for 30GHz wireless reception.

By setting the array periodicity as $D_{\alpha_{pm}}$, as per Eq. (10), the resultant electro-optic beam pattern will have a primary maxima oriented along α_{pm} . This is shown in Fig. 3(a). Fig. 3(b) shows the prescribed choices of array periodicities $D_{\alpha_{pm}}$, corresponding to various directions of primary maximas α_{pm} for 30GHz wireless reception.

Fig. 4(a) shows an electro-optic beam pattern with the primary maxima oriented along $\alpha = \alpha_{pm} = -30^\circ$, obtained by setting the array periodicity as $D_{\alpha_{pm}} = D_{-30^\circ} = 4.5\text{mm}$. The variation of the corresponding electro-optic phase-matching factor $\psi(D_{-30^\circ}, \alpha)$ with respect to α has been shown in Fig. 4(b). It can be found that $\psi(D_{-30^\circ}, \alpha)$ is equal to 2π when $\alpha = \alpha_{pm} = -30^\circ$. This indicates that the condition of electro-optic phase-matching is satisfied for $\alpha = \alpha_{pm} = -30^\circ$, resulting in a primary maxima along that direction.

Fig. 4. (a) Electro-optic beam pattern with the primary electro-optic maxima oriented along $\alpha = \alpha_{pm} = -30^\circ$, obtained by setting the array periodicity as $D_{-30^\circ} = 4.5\text{mm}$. (b) Variation of $\psi(D_{-30^\circ}, \alpha)$ with α .

Fig. 5. (a) Electro-optic beam pattern with the primary electro-optic maxima oriented along $\alpha = \alpha_{pm} = 30^\circ$, obtained by setting the array periodicity as $D_{30^\circ} = 8.1\text{mm}$. (b) Variation of $\psi(D_{30^\circ}, \alpha)$ with α .

Fig. 5(a) shows an electro-optic beam pattern with the primary maxima oriented along $\alpha = \alpha_{pm} = 30^\circ$, obtained by setting the array periodicity as $D_{30^\circ} = 8.1\text{mm}$. However, unlike the previous case, in addition to the expected primary maxima oriented along 30° , there is an additional maxima i.e. a secondary maxima oriented along $\alpha = \alpha_{sm} = -47^\circ$. The reason behind this can be found from Fig. 5(b). It can be observed that $\psi(D_{30^\circ}, \alpha)$ is equal to 2π when $\alpha = \alpha_{pm} = 30^\circ$, due to which there is a primary maxima along 30° . This is quite expected. However, in addition to this, it can be noted that $\psi(D_{30^\circ}, \alpha)$ is equal to 4π at $\alpha = \alpha_{sm} = -47^\circ$. This shows that the condition of electro-optic phase-matching is additionally satisfied for $\alpha = \alpha_{sm} = -47^\circ$, resulting in a secondary maxima along that direction.

Fig. 6(a) shows an electro-optic beam pattern with the primary maxima oriented along $\alpha = \alpha_{pm} = 60^\circ$, obtained by setting the array periodicity to be $D_{60^\circ} = 11.58\text{mm}$. It can be seen that in addition to a primary maxima along $\alpha = \alpha_{pm} = 60^\circ$, there exist two secondary maximas along $\alpha = \alpha_{sm1} = 0^\circ$ and $\alpha = \alpha_{sm2} = -60^\circ$. This happens because $\psi(D_{60^\circ}, \alpha)$ is equal to 4π and 6π , for

$\alpha = \alpha_{sm1} = 0^\circ$ and $\alpha = \alpha_{sm2} = -60^\circ$, respectively. This gives rise to two secondary maximas along those directions.

Fig. 6. (a) Electro-optic beam pattern with the primary electro-optic maxima oriented along $\alpha = \alpha_{pm} = 60^\circ$, obtained by setting the array periodicity as $D_{60^\circ} = 11.58\text{mm}$. (b) Variation of $\psi(D_{60^\circ}, \alpha)$ with α .

From the above discussion, we can segregate the analytical conditions of electro-optic phase-matching for primary and secondary maximas. Let us consider a receiver with array periodicity $D_{\alpha_{pm}}$. The corresponding electro-optic phase-matching factor $\psi(D_{\alpha_{pm}}, \alpha_{pm})$ for wireless reception along α_{pm} can be expressed as,

$$\psi(D_{\alpha_{pm}}, \alpha_{pm}) = k_{mw}(\eta_{op} - \sin \alpha_{pm})D_{\alpha_{pm}} = 2\pi \quad (11)$$

If the above mentioned condition is satisfied, we get a primary maxima along α_{pm} . Now, if the same receiver receives a wireless signal along a direction α_{sm} , such that the electro-optic phase-matching factor $\psi(D_{\alpha_{pm}}, \alpha_{sm})$ is an integral multiple 2π (except 2π itself), as shown in Eq. (12), we get a secondary maxima along α_{sm} .

$$\psi(D_{\alpha_{pm}}, \alpha_{sm}) = k_{mw}(\eta_{op} - \sin \alpha_{sm})D_{\alpha_{pm}} = 2\pi m \quad m \neq 1 \quad (12)$$

The analytical conditions mentioned in Eq. (11) and Eq. (12) have been schematically represented in Fig. 7(a). The analytical relationship between α_{sm} and α_{pm} can be derived by substituting the expression of $D_{\alpha_{pm}}$ from Eq. (10) to Eq. (12).

Fig. 7. (a) Electro-optic phase-matching condition for primary and secondary electro-optic maximas. (b) Obtained electro-optic maximas corresponding to the desired primary electro-optic maximas.

$$\alpha_{sm} = \sin^{-1} \left(m \sin \alpha_{pm} + (1 - m)\eta_{op} \right) \quad m \neq 1 \quad (13)$$

From the above equation, it can be understood that once we orient the primary maxima along α_{pm} , the possible directions

of α_{sm} get locked. In other words, we have no independent control over the existence of secondary maximas. Fig. 7(b) shows a plot of all possible secondary maximas corresponding to different primary maximas.

The solutions of Eq. (13) give us all possible directions of secondary maximas. If a solution of Eq. (13) does not exist for a given α_{pm} , no secondary maximas will exist, much like the case of Fig. 4. On the other hand, Fig. 5 represents the scenario where a single solution of Eq. (13) exists. Similarly, Fig. 6 represents the scenario where two solutions of Eq. (13) exist.

Next we explore the physical interpretation behind maximas in seamless electro-optic beamforming.

2) Physical interpretation: This subsection explains the physical interpretation behind maximas in seamless electro-optic beamforming. It was mentioned earlier that the electro-optic array factor $\delta\theta$ represents the physical quantity of electro-optic modulation depth. In other words, the extent by which $E_{op}(t)$ gets electro-optically phase-modulated by the received wireless signal is determined by $\delta\theta$. For wireless reception along maximas i.e. along α_{pm} or α_{sm} , the resultant electro-optic modulation-depth is maximum. The reason behind this is as follows. For wireless reception along α_{pm} or α_{sm} , the electric fields induced in successive array elements are in phase with each other when $E_{op}(t)$ enters them. Due to this, $E_{op}(t)$ accumulates equal phases while traveling through each array element by virtue seamless electro-optic phase-modulation. Therefore, the net phase acquired by $E_{op}(t)$ is maximum in this case. This explains the reason behind the electro-optic modulation depth being maximum for wireless reception along maximas.

Fig. 8. Physical interpretation behind the presence of maximas in seamless electro-optic beamforming.

We can elaborate the above explanation by considering a 2-element receiver array with array periodicity $D_{\alpha_{pm}}$, as shown in Fig. 8. The width W of the array elements are approximately chosen as the half-guided wavelength of $E_{slot}(t)$, as per Eq. (19) of Appendix A. Due to this, $E_{op}(t)$ gets exposed to $E_{slot1}(t)$ and $E_{slot2}(t)$ over their half-cycle periods while traveling through element-1 and element-2, respectively, as shown in Fig. 8. For wireless reception along α_{pm} or α_{sm} , $\angle E_{slot2}(T + T_{D_{\alpha_{pm}}}) - \angle E_{slot1}(T)$ is equal to the integral multiples of 2π . Here T and $T_{D_{\alpha_{pm}}}$ are the instants at which $E_{op}(t)$ enters element-1 and element-2, respectively. Under such conditions, $E_{op}(t)$ gets exposed to exactly similar half-cyclic variations of $E_{slot1}(t)$ and $E_{slot2}(t)$ while traveling

through element-1 and element-2, respectively. Due to this, $E_{op}(t)$ acquires equal phases while traveling through each array element. Therefore, the net phase acquired by $E_{op}(t)$ by the time it exits element-2 at $t=(T+T_{D_{\alpha_{pm}}}+T_W)$ is $2\theta_{op}$, as shown in Fig. 8. Here θ_{op} can be referred to as the phase acquired by $E_{op}(t)$ while traveling under the individual array elements. Similarly, in an N-element receiver array, the net phase acquired by the $E_{op}(t)$ would be equal to $N\theta_{op}$, for wireless reception along α_{pm} or α_{sm} .

B. Electro-optic nulls

1) *Analytical interpretation:* We now discuss the analytical interpretation behind nulls in seamless electro-optic beamforming. For nulls to be present along direction $\alpha=\alpha_n$, $\delta\theta$ must be equal to zero. From Eq. (8), it can be found that $\delta\theta$ is equal to zero when,

$$\psi(D, \alpha) = k_{mw}(\eta_{op} - \sin \alpha)D = 2\pi m/N \quad m \neq N, 2N, 3N... \quad (14)$$

The above equation represents the condition of electro-optic phase-mismatch for nulls provided m is any integer except a multiple of N . In other words, if m is an integral multiple of N , the above equation boils down to the condition of electro-optic phase-matching for maximas. From Eq. (14), all the possible directions of nulls along $\alpha=\alpha_n$ can be found as,

$$\alpha_n = \sin^{-1} \left(\eta_{op} - \frac{2\pi}{k_{mw}D} \left(\frac{m}{N} \right) \right) \quad m \neq N, 2N, 3N... \quad (15)$$

The possible solutions of the above mentioned equation for a given D and N represent all the possible orientations of nulls. The relationship between α_n and α_{pm} can be derived by substituting the expression of $D_{\alpha_{pm}}$ from Eq. (10) to Eq. (15).

$$\alpha_n = \sin^{-1} \left(\left(\frac{m}{N} \right) \sin \alpha_{pm} + \left(1 - \frac{m}{N} \right) \eta_{op} \right) \quad m \neq N, 2N... \quad (16)$$

From the above equation, it can be deduced that the choice of α_{pm} locks the orientations of all possible nulls in the electro-optic beampattern.

2) *Physical interpretation:* This subsection explains the physical significance behind nulls in seamless electro-optic beamforming. For wireless reception along α_n , the net phase acquired by the $E_{op}(t)$ is zero. This implies that whenever a wireless signal is received along a null, it has no net effect on $E_{op}(t)$.

We now elaborate this considering a 2-element receiver with array periodicity $D_{\alpha_{pm}}$, as shown in Fig. 9. As discussed earlier, W is chosen such that $E_{op}(t)$ gets exposed to $E_{slot1}(t)$ and $E_{slot2}(t)$ over their half-cycle periods while traveling through element-1 and element-2, respectively. However, in this case the angle of wireless reception is α_n . Due to this, $\angle E_{slot2}(T+T_{D_{\alpha_{pm}}}) - \angle E_{slot1}(T)$ is equal to π . As a result, $E_{op}(t)$ gets electro-optically phase modulated by cyclic variations $E_{slot1}(t)$ and $E_{slot2}(t)$ having exactly opposite

polarities, as shown in Fig. 9. Therefore, the net modulated phase of $E_{op}(t)$ when it exits element-2 at $t=(T+T_{D_{\alpha_{pm}}}+T_W)$ is zero.

Fig. 9. Physical interpretation behind the presence of nulls in seamless electro-optic beamforming.

IV. PARALLEL- k AND ANTIPARALLEL- k CONFIGURATIONS OF SEAMLESS ELECTRO-OPTIC BEAMFORMING

In this section, we classify seamless electro-optic beamforming into two categories. They are referred to as parallel- k mode and anti-parallel- k mode. In order to fundamentally define these two modes, we make use of Fig. 10 (a) and (b). We consider the propagation of $E_{op}(t)$ to be along the $+x$ -direction. So, the phase-constant of $E_{op}(t)$ can be expressed as k_{op}^{+x} . As per our convention, when α_{pm} is positive, the x -component of the phase-constant of $E_{mw}(t)$ i.e. k_{mw}^{+x} , is aligned along k_{op}^{+x} , as shown in Fig. 10(a). This configuration is called the parallel- k mode of seamless electro-optic beamforming. As shown in Fig. 10(b), when α_{pm} is negative, the x -component of the phase-constant of $E_{mw}(t)$ i.e. k_{mw}^{-x} is aligned opposite to k_{op}^{+x} . This configuration is called the anti-parallel- k mode of seamless electro-optic beamforming.

We now highlight the key differences between parallel- k and anti-parallel- k modes, as summarised in Table II. The first notable difference is regarding their prescribed array periodicities. It can be found from Fig. 3(b) that $D_{\alpha_{pm}}$ is relatively lower for anti-parallel- k mode compared to its parallel- k counterpart. This happens because in anti-parallel- k mode, $\angle E_{slot^n}(t)$ leads $\angle E_{slot^{n-1}}(t)$, unlike in parallel- k mode. Due to this, the condition of electro-optic phase-matching is satisfied for a relatively lower value of $D_{\alpha_{pm}}$ in anti-parallel- k mode, i.e.,

$$D_{\alpha_{pm}} > D_{-\alpha_{pm}} \quad (17)$$

Therefore, receiver arrays operating in anti-parallel- k mode are relatively more compact compared to the ones operating in parallel- k mode.

The second notable difference is regarding the electro-optic beamwidth. The primary electro-optic beam is narrower for parallel- k mode as compared to that of its anti-parallel- k counterpart. The reason behind this can be explained as follows. Electro-optic beamwidth depends on the rate of change of $\delta\theta$ with respect to α . In other words, a sharper change in $\delta\theta$ with respect to α gives rise to a narrower electro-optic beamwidth. From Eq. (8) it can be deduced that electro-optic beamwidth is proportional to the rate of change of $\psi(D_{\alpha_{pm}}, \alpha)$ with respect to α .

Fig. 10. (a) Parallel- k , and (b) Anti-parallel- k configuration of seamless electro-optic beamforming. (c) Comparison between the respective electro-optic beamwidths of parallel- k mode ($\alpha_{pm}=30^\circ$) and anti-parallel- k mode ($\alpha_{pm}=-30^\circ$). (d) Comparison between the slope of $\psi(D_{\alpha_{pm}}, \alpha)$ in parallel- k mode ($\alpha_{pm}=30^\circ$) and anti-parallel- k mode ($\alpha_{pm}=-30^\circ$) of seamless electro-optic beamforming.

$$\psi'(D_{\alpha_{pm}}, \alpha) = \frac{d\psi(D_{\alpha_{pm}}, \alpha)}{d\alpha} = -k_{mw}D_{\alpha_{pm}} \cos \alpha \quad (18)$$

It can be observed from the above equation that $\psi'(D_{\alpha_{pm}}, \alpha)$ is proportional to $D_{\alpha_{pm}}$. Since $D_{\alpha_{pm}} > D_{-\alpha_{pm}}$, it can be concluded that $\psi'(D_{\alpha_{pm}}, \alpha) > \psi'(D_{-\alpha_{pm}}, \alpha)$. This shows that $\delta\theta$ varies more rapidly with respect to α in parallel- k mode. Due to this, the electro-optic beamwidth in parallel- k mode is relatively narrower. This argument can be verified by citing the example of Fig. 10(c). It can be observed that the electro-optic beam is narrower in the parallel- k mode with $\alpha_{pm}=30^\circ$, as compared to its anti-parallel- k counterpart with $\alpha_{pm}=-30^\circ$. The reason behind this can be attributed to fact that the slope of $\psi(D_{30^\circ}, \alpha)$ is steeper compared to that of $\psi(D_{-30^\circ}, \alpha)$, as shown in Fig. 10(d).

TABLE II
COMPARISON BETWEEN PARALLEL- k AND ANTIPARALLEL- k
CONFIGURATIONS OF SEAMLESS ELECTRO-OPTIC BEAMFORMING

Sr.	Parallel- k	Antiparallel- k
1	$k_{mw} \uparrow \uparrow k_{op}$	$k_{mw} \uparrow \downarrow k_{op}$
2	Array less compact	Array more compact
3	Secondary maximas possible	Secondary maximas not possible
4	EO beamwidth narrower	EO beamwidth broader

The third key difference between parallel- k mode and anti-parallel- k mode is regarding secondary maximas. It can be seen from Fig. 7(b) that no secondary maximas exist in anti-parallel- k mode. The reason behind this can be understood from Eq. (13), according to which no solutions of α_{sm} exist for any negative value of α_{pm} . However, when α_{pm} is positive, multiple solutions of α_{sm} may exist. This shows that while secondary maximas are always absent in the anti-parallel- k mode, they may be present in parallel- k mode.

The absence of secondary maximas in anti-parallel- k mode has an interesting implication, which will be discussed in the upcoming subsection. It will be shown that a seamless beam-former operating in anti-parallel- k mode can independently modulate two counter-propagating optical carriers of different wavelengths using wireless signals incident along two different directions.

A. Seamless spatial-to-wavelength division multiplexing

The schematic of a seamless electro-optic beamformer operating in anti-parallel- k mode is shown in Fig. 11. A Laser Diode LD-1, generates an unmodulated optical signal $E_{op1}(t)$ of wavelength λ_1 , that propagates along the $+x$ -direction. The optical signal $E_{op1}(t)$ is circulated from port-1 to port-2 of circulator-1, after which it enters the receiver array. We consider the receiver array to be excited by two wireless signals $E_{mw1}(t)$ and $E_{mw2}(t)$ along direction α_{pm1} and α_{pm2} , respectively. It must be noted from Fig. 11, that $\alpha_{pm1} = -\alpha_{pm2}$. Therefore, $E_{op1}(t)$ forms an anti-parallel- k pair with $E_{mw1}(t)$, and forms a parallel- k pair with $E_{mw2}(t)$. The array periodicity is chosen as $D_{\alpha_{pm1}}$. Due to this, the receiver operates in anti-parallel- k mode. So, $E_{op1}(t)$ gets optimally phase-modulated by $E_{mw1}(t)$ received along α_{pm1} , indicating a primary maxima in that direction. Since, no secondary maximas exist in anti-parallel- k mode, $E_{op1}(t)$ will not get phase-modulated by $E_{mw2}(t)$, which is incident along α_{pm2} . Therefore, $E_{op1}(t)$ remains isolated from the incident $E_{mw2}(t)$. The modulated version of $E_{op1}(t)$ exits the receiver array through port-2 of circulator-2. It is then circulated from port-2 to port-3 of circulator-2, to one of the input ports of a 2:1 optical coupler.

Fig. 11. Schematic representation of the proposed seamless spatial-to-wavelength division multiplexing.

The same structure can be used to modulate a counter-propagating optical carrier using the wireless signal $E_{mw2}(t)$ received along α_{pm2} direction. As shown in Fig. 11, an unmodulated optical signal $E_{op2}(t)$ of wavelength λ_2 , propagating along the $-x$ -direction is generated by Laser Diode LD-2. The optical signal $E_{op2}(t)$ enters the receiver array after being circulated from port-1 to port-2 of circulator-2. In this case, $E_{op2}(t)$ forms an anti-parallel- k pair with $E_{mw2}(t)$, and a parallel- k pair with $E_{mw1}(t)$. Since, $\alpha_{pm1} = -\alpha_{pm2}$, it can be deduced that $D_{\alpha_{pm1}} = D_{-\alpha_{pm2}}$. Due to this, even in this case

the receiver operates in anti-parallel- k mode. Consequently, $E_{op2}(t)$ gets optimally phase-modulated by $E_{mw2}(t)$ received along α_{pm2} , indicating a primary maxima along that direction. Also, $E_{op2}(t)$ remains unaffected by $E_{mw1}(t)$ incident along α_{pm1} . The modulated version of $E_{op2}(t)$, then exits the receiver array through port-2 of circulator-1. It is then circulated from port-2 to port-3 of circulator-1, to the other input port of the 2:1 optical coupler. The optical coupler multiplexes $E_{op1}(t)$ and $E_{op2}(t)$ into a single optical fiber channel, as shown in Fig. 11. The net electro-optic beam pattern of this configuration is shown in Fig. 12.

Fig. 12. Electro-optic beam pattern for spatial-to-wavelength division multiplexing.

Therefore, in this configuration, a wireless signal incident along a particular direction can independently modulate an optical carrier of a particular wavelength. So, two wireless signals incident along two different directions can independently modulate optical carriers of two different wavelengths. This configuration enables seamless mapping of spatial wireless radiation diversity to optical wavelength division multiplexing. This proposed scheme can therefore be referred to as spatial-to-wavelength division multiplexing.

V. CONCLUSION

In this paper, the theory behind seamless electro-optic beamforming has been developed. It has been shown that a slotted patch antenna array based wireless-to-photonic receiver can function as a seamless electro-optic beamformer. We derive the electro-optic array factor, which essentially determines the characteristics of seamless electro-optic beamforming. The electro-optic array factor was found to be independent of the gain of the receiver array. We report the analytical and physical interpretations of maximas and nulls in seamless electro-optic beamforming. The design guidelines for controlling the seamless electro-optic beam pattern have been discussed. The inadvertent nature of secondary maximas have been highlighted. We classify seamless electro-optic beamforming into two categories: Parallel- k and Antiparallel- k modes and compare them. We also propose a seamless electro-optic beamforming configuration that enables spatial-to-wavelength division multiplexing. The seamless electro-optic beamformer presented in this paper does not consist of an electronic RF-downconversion chain, unlike conventional analog beamformers. Due to this, the latency of seamless electro-optic beamformers is expected to be much lower, thereby making them more suitable for next-generation communication systems like B5G and 6G. The theoretical contributions and the design guidelines mentioned in this paper will be necessary for further

investigation and the practical realization of seamless electro-optic beamformers.

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APPENDIX A

OPTIMUM WIDTH OF THE ARRAY ELEMENTS

It was shown in [42] that the width W of the individual array elements in a seamless wireless-to-photonic receiver must be chosen as per Eq. (19) to achieve optimum seamless electro-optic phase modulation.

$$W = \frac{\pi}{k_{mw}\eta_{op}} = \frac{\lambda_{mw}}{2\eta_{op}} \quad (19)$$

where k_{mw} is the phase constant of the received wireless signal of wavelength λ_{mw} , and η_{op} effective optical refractive index of the LiNbO₃ waveguide.

The time T_W taken by $E_{op}(t)$ to travel the distance W within the optical waveguide can be approximately found as,

$$T_W \approx \frac{\eta_{op}}{c} W = \frac{\lambda_{mw}}{2c} \quad (20)$$

Similarly, the time T_D taken by $tE_{op}(t)$ to travel the distance D can be found as,

$$T_D \approx \frac{\eta_{op}}{c} D \quad (21)$$

APPENDIX B

DERIVATION OF THE INSTANTANEOUS OPTICAL PHASE

Here we mention the details of the derivation of Eq. (6). To derive $\theta_{op^N}(t)$, we firstly integrate Eq. (5) in the time interval $t + (n-1)T_D$ to $t + (n-1)T_D + T_W$, where $n = 1, 2, 3, \dots, N$. The integrals for every value of n are summed up to get $\theta_{op^N}(t)$.

$$\theta_{op^N}(t) = \sum_{n=1}^{n=N} \int_{t+(n-1)T_D}^{t+(n-1)T_D+T_W} k_{op} d\eta_{op^n}(t) \frac{c}{\eta_{op}} dt \quad (22)$$

Putting the expression of $d\eta_{op^n}(t)$ from Eq. (4) to the above equation, we get,

$$\begin{aligned} \theta_{op^N}(t) &= \sum_{n=1}^{n=N} \int_{t+(n-1)T_D}^{t+(n-1)T_D+T_W} -k_{op} \frac{r_{33}}{2} \eta_{op}^3 E_{slot^n}(t) \frac{c}{\eta_{op}} dt \\ &= \sum_{n=1}^{n=N} \int_{t+(n-1)T_D}^{t+(n-1)T_D+T_W} -k_{op} \frac{r_{33}}{2} \eta_{op}^2 c E_{slot^n}(t) dt \end{aligned} \quad (23)$$

We substitute the expression of $E_{slot^n}(t)$ from Eq. (3) into the above equation. The resultant equation can be solved as follows,

$$\begin{aligned}
\theta_{op^N}(t) &= \sum_{n=1}^{n=N} \int_{t+(n-1)T_D}^{t+(n-1)T_D+T_W} -k_{op} \frac{r_{33}}{2} \eta_{op}^2 c |E_{slot}| \sin \left(\omega_{mw} t - (n-1) k_{mw} \sin \alpha D \right) dt \\
&= -\frac{k_{op} \eta_{op}^2 r_{33} c}{2} |E_{slot}| \sum_{n=1}^{n=N} \int_{t+(n-1)T_D}^{t+(n-1)T_D+T_W} \sin \left(\omega_{mw} t - (n-1) k_{mw} \sin \alpha D \right) dt \\
&= -\frac{k_{op} \eta_{op}^2 r_{33} c}{\omega_{mw}} |E_{slot}| \sum_{n=1}^{n=N} \sin \left(\frac{\omega_{mw} T_W}{2} \right) \sin \left(\omega_{mw} t + (n-1) \omega_{mw} T_D + \frac{\omega_{mw} T_W}{2} - (n-1) k_{mw} \sin \alpha D \right)
\end{aligned} \tag{24}$$

We put the expression of T_W from Eq. (20) in Eq. (24),

$$\begin{aligned}
\theta_{op^N}(t) &= -\frac{k_{op} \eta_{op}^2 r_{33} c}{\omega_{mw}} |E_{slot}| \sum_{n=1}^{n=N} \sin \left(\frac{\pi}{2} \right) \sin \left(\omega_{mw} t + (n-1) \omega_{mw} T_D + \frac{\pi}{2} - (n-1) k_{mw} \sin \alpha D \right) \\
&= -\frac{k_{op} \eta_{op}^2 r_{33} c}{\omega_{mw}} |E_{slot}| \sum_{n=1}^{n=N} \cos \left(\omega_{mw} t + (n-1) (\omega_{mw} T_D - k_{mw} \sin \alpha D) \right)
\end{aligned} \tag{25}$$

We now put the expression of T_D from Eq. (21) in Eq. (25),

$$\begin{aligned}
\theta_{op^N}(t) &= -\frac{k_{op} \eta_{op}^2 r_{33} c}{\omega_{mw}} |E_{slot}| \sum_{n=1}^{n=N} \cos \left(\omega_{mw} t + (n-1) (k_{mw} (\eta_{op} - \sin \alpha) D) \right) \\
&= -\frac{k_{op} \eta_{op}^2 r_{33} c}{\omega_{mw}} |E_{slot}| \left(\frac{\sin (N k_{mw} (\eta_{op} - \sin \alpha) D/2)}{\sin (k_{mw} (\eta_{op} - \sin \alpha) D/2)} \right) \cos \left(\omega_{mw} t + (N-1) k_{mw} (\eta_{op} - \sin \alpha) D/2 \right) \\
&= \delta \theta |E_{slot}| \left(\frac{\sin (\frac{N\psi}{2})}{\sin (\frac{\psi}{2})} \right) \cos \left(\omega_{mw} t + \frac{(N-1)\psi}{2} \right)
\end{aligned} \tag{26}$$

REFERENCES

- [1] K. Samdanis and T. Taleb, "The road beyond 5G: A vision and insight of the key technologies," *IEEE Network*, vol. 34, no. 2, pp. 135–141, 2020.
- [2] V. Ziegler, T. Wild, M. Uusitalo, H. Flinck, V. Räisänen, and K. Hätönen, "Stratification of 5G evolution and beyond 5G," in *2019 IEEE 2nd 5G World Forum (5GWF)*, 2019, pp. 329–334.
- [3] X. Chen, D. W. K. Ng, W. Yu, E. G. Larsson, N. Al-Dhahir, and R. Schober, "Massive access for 5G and beyond," *IEEE Journal on Selected Areas in Communications*, vol. 39, no. 3, pp. 615–637, 2020.
- [4] C. De Lima, D. Belot, R. Berkevns, A. Bourdoux, D. Dardari, M. Guillaud, M. Isomursu, E.-S. Lohan, Y. Miao, A. N. Barreto, *et al.*, "Convergent communication, sensing and localization in 6G systems: An overview of technologies, opportunities and challenges," *IEEE Access*, 2021.
- [5] H. Tataria, M. Shafi, A. F. Molisch, M. Dohler, H. Sjöland, and F. Tufvesson, "6G wireless systems: Vision, requirements, challenges, insights, and opportunities," *Proceedings of the IEEE*, 2021.
- [6] W. Jiang, B. Han, M. A. Habibi, and H. D. Schotten, "The road towards 6G: A comprehensive survey," *IEEE Open Journal of the Communications Society*, vol. 2, pp. 334–366, 2021.
- [7] J. R. Bhat and S. A. Alqahtani, "6G ecosystem: current status and future perspective," *IEEE Access*, vol. 9, pp. 43 134–43 167, 2021.
- [8] M. Alsabah, M. A. Naser, B. M. Mahmmod, S. H. Abdulhussain, M. R. Eissa, A. Al-Baidhani, N. K. Noordin, S. M. Sait, K. A. Al-Utaibi, and F. Hashim, "6G wireless communications networks: A comprehensive survey," *IEEE Access*, vol. 9, pp. 148 191–148 243, 2021.
- [9] M. Z. Chowdhury, M. Shahjalal, S. Ahmed, and Y. M. Jang, "6G wireless communication systems: Applications, requirements, technologies, challenges, and research directions," *IEEE Open Journal of the Communications Society*, vol. 1, pp. 957–975, 2020.
- [10] S. Rommel, D. Dodane, E. Grivas, B. Cimoli, J. Bourderionnet, G. Feugnet, A. Morales, E. Pikasis, C. Roeloffzen, P. van Dijk, M. Katsikis, K. Ntontin, D. Kritharidis, I. Spaleniak, P. Mitchell, M. Dubov, J. B. Carvalho, and I. T. Monroy, "Towards a scaleable 5G fronthaul: Analog radio-over-fiber and space division multiplexing," *J. Lightwave Technol.*, vol. 38, no. 19, pp. 5412–5422, Oct 2020. [Online]. Available: <http://www.osapublishing.org/jlt/abstract.cfm?URI=jlt-38-19-5412>
- [11] C. Lim, Y. Tian, C. Ranaweera, T. A. Nirmalathas, E. Wong, and K.-L. Lee, "Evolution of radio-over-fiber technology," *J. Lightwave Technol.*, vol. 37, no. 6, pp. 1647–1656, Mar 2019. [Online]. Available: <http://www.osapublishing.org/jlt/abstract.cfm?URI=jlt-37-6-1647>
- [12] K. Zeb, X. Zhang, and Z. Lu, "High capacity mode division multiplexing based MIMO enabled all-optical analog millimeter-wave over fiber fronthaul architecture for 5G and beyond," *IEEE Access*, vol. 7, pp. 89 522–89 533, 2019.
- [13] T. S. Rappaport, Y. Xing, O. Kanhere, S. Ju, A. Madanayake, S. Mandal, A. Alkhateeb, and G. C. Trichopoulos, "Wireless communications and applications above 100 GHz: Opportunities and challenges for 6G and beyond," *IEEE access*, vol. 7, pp. 78 729–78 757, 2019.
- [14] P. T. Dat, Y. Yamaguchi, A. Kanno, M. Motoya, S. Oikawa, J. Ichikawa,

- and N. Yamamoto, "Millimeter-wave radio-over-fiber system using optical phase modulation and photonic downconversion for uplink fronthaul transmission," *Optics Letters*, vol. 46, no. 10, pp. 2493–2496, 2021.
- [15] F. Mokhtari-Koushyar, E. Heidari, H. Dalir, C.-J. Chung, X. Xu, V. J. Sorger, and R. T. Chen, "Wideband multi-arm bowtie antenna for millimeter wave electro-optic sensors and receivers," *Journal of Lightwave Technology*, vol. 36, no. 16, pp. 3418–3426, 2018.
  - [16] D. Park, V. R. Pagán, P. S. Cho, J. Luo, A. K.-Y. Jen, and P. V. Petruzzi, "RF photonic downconversion of vector modulated signals based on a millimeter-wave coupled electrooptic nonlinear polymer phase-modulator," *Opt. Express*, vol. 25, no. 24, pp. 29885–29895, Nov 2017. [Online]. Available: <http://www.osapublishing.org/oe/abstract.cfm?URI=oe-25-24-29885>
  - [17] Y. N. Wijayanto, H. Murata, and Y. Okamura, "Electrooptic millimeter-wave–lightwave signal converters suspended to gap-embedded patch antennas on low- k dielectric materials," *IEEE Journal of Selected Topics in Quantum Electronics*, vol. 19, no. 6, pp. 33–41, 2013.
  - [18] P. T. Dat, Y. Yamaguchi, K. Inagaki, M. Motoya, S. Oikawa, J. Ichikawa, A. Kanno, N. Yamamoto, and T. Kawanishi, "Transparent fiber–radio-fiber bridge at 101 GHz using optical modulator and direct photonic down-conversion," in *Optical Fiber Communication Conference*. Optical Society of America, 2021, pp. F3C–4.
  - [19] K. Kasai, T. Sato, T. Hirooka, M. Yoshida, and M. Nakazawa, "64 Gbit/s, 256 QAM transmission through coherent optical-wireless link at 61 GHz using simple and high OSNR carrier frequency converter," in *2021 Optical Fiber Communications Conference and Exhibition (OFC)*. IEEE, 2021, pp. 1–3.
  - [20] Y. Xia, L. Valenzuela, A. Maharry, S. Pinna, S. Dwivedi, T. Hirokawa, J. Buckwalter, and C. Schow, "A fully integrated O-band coherent optical receiver operating up to 80 Gb/s," in *2021 IEEE Photonics Conference (IPC)*. IEEE, 2021, pp. 1–2.
  - [21] M. G. Ahmed, D. Kim, R. K. Nandwana, A. Elkholy, K. R. Lakshmikumar, and P. K. Hanumolu, "A 16-Gb/s-11.6-dBm OMA sensitivity 0.7-pJ/bit optical receiver in 65-nm CMOS enabled by duobinary sampling," *IEEE Journal of Solid-State Circuits*, 2021.
  - [22] H. Li, J. Sharma, C.-M. Hsu, G. Balamurugan, and J. Jaussi, "A 100Gb/s-8.3 dBm-sensitivity PAM-4 optical receiver with integrated TIA, FFE and direct-feedback DFE in 28nm CMOS," in *2021 IEEE International Solid-State Circuits Conference (ISSCC)*, vol. 64. IEEE, 2021, pp. 190–192.
  - [23] G. Keiser, "Optical receiver operation," in *Fiber Optic Communications*. Springer, 2021, pp. 267–302.
  - [24] C. Castro, R. Elschner, T. Merkle, C. Schubert, and R. Freund, "100 Gb/s real-time transmission over a THz wireless fiber extender using a digital-coherent optical modem," in *Optical Fiber Communication Conference*. Optical Society of America, 2020, pp. M4I–2.
  - [25] T. Harter, S. Ummethala, M. Blaicher, S. Muehlbrandt, S. Wolf, M. Weber, M. M. H. Adib, J. N. Kemal, M. Merboldt, F. Boes, *et al.*, "Wireless THz link with optoelectronic transmitter and receiver," *Optica*, vol. 6, no. 8, pp. 1063–1070, 2019.
  - [26] T. A. Maldonado, "Electro-optic modulators," *Handbook of optics*, vol. 2, pp. 13–11, 1995.
  - [27] M. Zhang, C. Wang, P. Kharel, D. Zhu, and M. Lončar, "Integrated Lithium Niobate Electro-optic modulators: when performance meets scalability," *Optica*, vol. 8, no. 5, pp. 652–667, 2021.
  - [28] X. Liu, B. Xiong, C. Sun, J. Wang, Z. Hao, L. Wang, Y. Han, H. Li, and Y. Luo, "Sub-terahertz bandwidth capacitively-loaded thin-film Lithium Niobate electro-optic modulators based on an undercut structure," *Optics Express*, vol. 29, no. 25, pp. 41 798–41 807, 2021.
  - [29] J. Cai, C. Guo, C. Lu, A. P. T. Lau, P. Chen, and L. Liu, "Design optimization of silicon and Lithium Niobate hybrid integrated traveling-wave Mach-Zehnder modulator," *IEEE Photonics Journal*, vol. 13, no. 4, pp. 1–6, 2021.
  - [30] X. Huang, Y. Liu, Z. Li, H. Guan, Q. Wei, Z. Yu, and Z. Li, "Advanced electrode design for low-voltage high-speed thin-film Lithium Niobate Modulators," *IEEE Photonics Journal*, vol. 13, no. 2, pp. 1–9, 2021.
  - [31] P. Kharel, C. Reimer, K. Luke, L. He, and M. Zhang, "Breaking voltage–bandwidth limits in integrated Lithium Niobate modulators using micro-structured electrodes," *Optica*, vol. 8, no. 3, pp. 357–363, 2021.
  - [32] P. Yang, Y. Xiao, M. Xiao, and S. Li, "6G wireless communications: Vision and potential techniques," *IEEE Network*, vol. 33, no. 4, pp. 70–75, 2019.
  - [33] T. Kaji, I. Morohashi, Y. Tominari, N. Sekine, T. Yamada, and A. Otomo, "W-band optical modulators using electro-optic polymer waveguides and patch antenna arrays," *Optics Express*, vol. 29, no. 19, pp. 29 604–29 614, 2021.
  - [34] H. Murata, H. Yokohashi, S. Matsukawa, M. Sato, M. Onizawa, and S. Kurokawa, "Antenna-coupled electrode electro-optic modulator for 5G mobile applications," *IEEE Journal of Microwaves*, vol. 1, no. 4, pp. 902–907, 2021.
  - [35] Y. Miyazeki, H. Yokohashi, S. Kodama, H. Murata, and T. Arakawa, "InGaAs/InAlAs multiple-quantum-well optical modulator integrated with a planar antenna for a millimeter-wave radio-over-fiber system," *Optics express*, vol. 28, no. 8, pp. 11 583–11 596, 2020.
  - [36] J. Zhang, F. Chen, and B. Liu, "Integrated photonic electric field sensor operating more than 26 GHz," *IEEE Microwave and Wireless Components Letters*, vol. 30, no. 10, pp. 1009–1012, 2020.
  - [37] F. A. Juneghani, M. G. Vazimali, G. S. Kanter, and S. Fathpour, "Integrated electro-optical sensors for microwave photonic applications on thin-film Lithium Niobate," in *2021 International Topical Meeting on Microwave Photonics (MWP)*. IEEE, 2021, pp. 1–4.
  - [38] C.-J. Chung, X. Xu, Z. Pan, F. Mokhtari-Koushyar, R. Wang, H. Yan, H. Subbaraman, and R. T. Chen, "Silicon-based hybrid integrated photonic chip for k_u band electromagnetic wave sensing," *Journal of Lightwave Technology*, vol. 36, no. 9, pp. 1568–1575, 2018.
  - [39] M. Burla, C. Hoessbacher, W. Heni, C. Haffner, Y. Fedoryshyn, D. Werner, T. Watanabe, H. Massler, D. L. Elder, L. R. Dalton, *et al.*, "500 GHz plasmonic Mach-Zehnder modulator enabling sub-THz microwave photonics," *APL Photonics*, vol. 4, no. 5, p. 056106, 2019.
  - [40] J. E. Muench, A. Ruocco, M. A. Giambra, V. Misiakis, D. Zhang, J. Wang, H. F. Watson, G. C. Park, S. Akhavan, V. Sorianello, *et al.*, "Waveguide-integrated, plasmonic enhanced graphene photodetectors," *Nano letters*, vol. 19, no. 11, pp. 7632–7644, 2019.
  - [41] Y. Salamin, B. Baeuerle, W. Heni, F. C. Abrecht, A. Josten, Y. Fedoryshyn, C. Haffner, R. Bonjour, T. Watanabe, M. Burla, *et al.*, "Microwave plasmonic mixer in a transparent fibre–wireless link," *Nature photonics*, vol. 12, no. 12, pp. 749–753, 2018.
  - [42] N. Ghosh and S. Pendharker, "Wireless to optical phase mapping in a seamless digital wireless-photonic link," *Journal of Optics*, vol. 23, no. 12, p. 125702, Nov 2021. [Online]. Available: <https://doi.org/10.1088/2040-8986/ac31d5>
  - [43] B. Zheng, C. You, and R. Zhang, "Double-IRS assisted multi-user MIMO: Cooperative passive beamforming design," *IEEE Transactions on Wireless Communications*, 2021.
  - [44] Q. Wu, X. Zhou, W. Chen, J. Li, and X. Zhang, "IRS-Aided WPCNs: A new optimization framework for Dynamic IRS Beamforming," *IEEE Transactions on Wireless Communications*, pp. 1–1, 2021.
  - [45] H. U. Rehman, F. Bellili, A. Mezghani, and E. Hossain, "Joint active and passive beamforming design for IRS-assisted multi-user MIMO systems: A vamp-based approach," *IEEE Transactions on Communications*, vol. 69, no. 10, pp. 6734–6749, 2021.
  - [46] A. Kammoun, M.-S. Alouini, *et al.*, "Elevation beamforming with full dimension MIMO architectures in 5G systems: A tutorial," *IEEE Communications Surveys & Tutorials*, vol. 21, no. 4, pp. 3238–3273, 2019.
  - [47] I. Ahmed, H. Khammari, A. Shahid, A. Musa, K. S. Kim, E. De Poorter, and I. Moerman, "A survey on hybrid beamforming techniques in 5G: Architecture and system model perspectives," *IEEE Communications Surveys & Tutorials*, vol. 20, no. 4, pp. 3060–3097, 2018.
  - [48] D. Marcuse, *Principles of quantum electronics*. Elsevier, 2012.
  - [49] R. Weis and T. Gaylord, "Lithium niobate: summary of physical properties and crystal structure," *Applied Physics A*, vol. 37, no. 4, pp. 191–203, 1985.
  - [50] J. Litva, "Digital beamforming in wireless communications," *Artech house*, 1996.