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Low Cross-Polarization Vivaldi Arrays

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LOW CROSS-POLARIZATION VIVALDI ARRAYS

A Dissertation Presented

by

JOHN T. LOGAN

Submitted to the Graduate School of the
University of Massachusetts Amherst in partial fulfillment
of the requirements for the degree of

DOCTOR OF PHILOSOPHY

September 2016

Electrical and Computer Engineering

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A Dissertation Presented

by

JOHN T. LOGAN

Approved as to style and content by:

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to my mom, Sue, my dad, Tom, and my brother, Alex

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Vivaldi arrays. He has been an invaluable resource throughout this experience and I am excited and grateful to personally work with him in the future at NRL.

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ABSTRACT

LOW CROSS-POLARIZATION VIVALDI ARRAYS

SEPTEMBER 2016

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Ultra-wideband (UWB) electronically scanned arrays (ESA) with high efficiency, excellent polarization agility, and wide-scan matching remain essential for servicing multifunctional RF front-ends and other communications, sensing, and jamming or countermeasure systems. To this day, the most popular antenna array element in modern UWB-ESA systems is the Vivaldi, or flared notch, due to its superior wide-scan wide impedance bandwidth, well-known design guidelines, and practical embodiment versatility. Despite their popularity, these arrays tend to radiate unacceptably high cross-polarization levels, thus encouraging a great research opportunity.

This dissertation presents the theory and design of a new class of UWB-ESAs, termed Sliced Notch Antenna (SNA) arrays, that remedy the high cross-polarization problems in Vivaldi arrays while maintaining their desirable impedance performance. The critical enabling insight of this work lays in revealing the nature of polarization

purity or cross-polarization ratio (CPR) degradation in Vivaldi arrays to arise from a highly imbalanced ratio of longitudinal and transverse currents within the element. This work introduces a novel design strategy that intrinsically balances these currents over a UWB operating band, achieving decade-order (10:1) bandwidths and low cross-polarization. Moreover, the design approach is simple, intuitive, and can be implemented in a manner that does not inflate cost expenditures. In fact, the proposed topology can facilitate significantly reduced costs and manufacturing times in Vivaldi arrays by permitting electrically disconnected elements over a large portion of the original Vivaldi fin. Following the presentation of the theoretical and operational principles, a few infinite SNA array design implementations are proposed that achieve decade-order bandwidths and low cross-polarization (e.g. active VSWR < 2.1 out to $\theta_0 = 60^\circ$ scans and CPR < -15 dB for D-plane $\theta_0 = 45^\circ$ scans) across different embodiments (i.e. single-polarized and dual-polarized configurations in standard printed circuit board fabrication or machined stock-metal hybrid builds).

A single-polarized 19×19 SNA array and its Vivaldi counterpart are designed, fabricated, assembled, and measured to comparatively demonstrate similar wideband impedance behavior over a 1.2-12 GHz (10:1) operating range that covers near the entirety of four popular RF bands (L, S, C, and X); center embedded element impedance measurements show good agreement with finite array simulations out to $\theta_0 = 60^\circ$ scans. The SNA array embedded element pattern offers better than 15 dB cross-polarization level improvements over the Vivaldi array out to wide angles in the D-plane. Active finite array scan simulations suggest both arrays closely track the directivity in the principal planes (e.g. less than 0.5 dB co-polarized gain drop at broadside), while the SNA array offers up to 25 dB improvements in co-polarized gain for non-principal plane scans in and around the D-plane with approximately 15 dB improvement on average over the Vivaldi array in the mid-band and high-band.

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CHAPTER 1

INTRODUCTION

1.1 Problem Statement

This work develops a new class of low cross-polarization, ultra-wideband (UWB) electronically scanned arrays (ESA) to meet the needs of modern multifunctional RF systems [1], high-throughput or low-power communication systems [2–4], high resolution radars and sensing imagers [5–8], polarimeters and radio telescopes [9–11], and electronic warfare systems [12–14]. Although a number of accomplished UWB-ESAs have been proposed over the past decades [15–23], *the quest for low-cost, high-efficiency UWB-ESAs with greater than 6:1 bandwidth (bandwidth ratio, $BWR = \frac{f_{high}}{f_{low}} : 1$) and low cross-polarization over wide scan volumes (i.e. $> 45^\circ$) continues.*

Most high-efficiency (low-loss) UWB-ESA apertures, certainly those that archive BWRs wider than 6:1 such as the Vivaldi array [24], are subject to an inherent bandwidth versus cross-polarization trade-off. A qualitative display of that trade-off is presented in Fig. 1.1(a). At one extreme, low-profile arrays such as capacitively coupled dipole or connected arrays exhibit excellent polarization properties under scanning (high polarization purity) but at the expense of reduced bandwidth, whereas high-profile arrays such as Vivaldi arrays occupy the opposite extreme with the widest bandwidths but poor polarization properties under scanning. Quantitative comparisons of a typical low-loss, state-of-the-art low-profile capacitively coupled dipole UWB-ESA [25] and a state-of-the-art Vivaldi UWB-ESA [18] are given in Figs. 1.1(b)-(c) where the maximum active VSWR levels within a $\theta_0 = 60^\circ$ scan cone and

the cross-polarization ratio (CPR) patterns (where $\text{CPR} = |\mathbf{E}_{\text{cx}}|^2 / |\mathbf{E}_{\text{co}}|^2$ is evaluated in the high frequency band end) are presented. It is apparent that Vivaldi arrays have a clear edge in terms of bandwidth and impedance (VSWR) performance out to wide scan angles for over 10:1 bandwidths but at the expense of high amounts of undesired cross-polarization in the non-principal scan planes, whereas low-profile arrays offer good wide-scan polarization but at the expense of reduced bandwidth and degraded impedance matching. Thus, system-level performance of modern UWB-ESA-based systems is inherently limited by this bandwidth/matching and cross-polarization compromise at the aperture level. This work will address this important research and technological opportunity in the wideband antenna array area by introducing a new state-of-the-art antenna array topology termed the Sliced Notch Antenna (SNA) array that does not trade bandwidth/matching for polarization while maintaining high efficiency.

Figure 1.1: Bandwidth vs. polarization trade-offs in typical low-loss state-of-the-art UWB-ESAs. (a) Qualitative bandwidth vs. cross-polarization trade-offs; (b) Maximum active VSWR out to $\theta_0 = 60^\circ$ scans versus frequency plot of a state-of-the-art low-profile array and Vivaldi array; (c) Cross-polarization ratio (CPR(θ, ϕ) = $|\mathbf{E}_{cx}(\theta, \phi)|^2 / |\mathbf{E}_{co}(\theta, \phi)|^2$) comparison between the same arrays (blue desirable, red undesirable).

1.2 Antenna Array Figures of Merit and Useful Definitions

Before proceeding into the significance of the problem at hand, it will be useful to renew some basic definitions and relations that will help quantify some of the impedance bandwidth and polarization figures of merit (FoM) for dual-polarized arrays used repeatedly throughout this dissertation. The choices are consistent with IEEE standards [26] and other influential resources [27–30]. The treatment of polarization will be restricted to purely monochromatic waves, but the interested reader can refer to [27] for partially polarized wave considerations. Frequency and scan direction (θ_0, ϕ_0) dependencies are implicit. FoMs and definitions are divided into two parts: terms related to (i) bandwidth and impedance and (ii) polarization.

1.2.1 Bandwidth and Impedance

(Active) Voltage Standing Wave Ratio (VSWR): is the measure of input power reflected from an array element, when all elements of a particular polarization are excited, and is a function of the active reflection coefficient, Γ_{ii}^a , over a range of frequencies at a specified scan angle as defined by [28]

$$\Gamma_{ii}^a = \frac{Z_A - Z_0}{Z_A + Z_0}, \quad (1.1)$$

$$\text{VSWR} = \frac{1 + |\Gamma_{ii}^a|}{1 - |\Gamma_{ii}^a|}, \quad (1.2)$$

where Z_A is the active input impedance of an element and Z_0 is the characteristic impedance of the antenna interface or transmission line. For various reasons that deal with both the antenna electronics and the radiation medium, the characteristic impedance of the transmission line often takes the standardized value of 50Ω , but in some applications it can deviate from that value to benefit matching. This work will only use $Z_0 = 50\Omega$. In wideband antenna array design, the range of frequencies over which the active $\text{VSWR} < 2$ at broadside scan in an infinite array environment typically defines the *bandwidth ratio*.

(Instantaneous) Bandwidth Ratio (BWR): is the range of continuous frequencies where the performance of an array conforms to a specified antenna FoM, most often some standard related to active VSWR or gain, as defined by the ratio of the highest-to-lowest frequencies of that band, i.e. $\text{BWR} = \frac{f_H}{f_L} : 1$, where f_H and f_L are the highest and lowest frequency of operation. This is illustrated in Figs. 1.2(a) and (b) for an infinite array of periodic element spacing D in which the BWR is designated over the band with broadside scan active $\text{VSWR} < 2$. BWR will conform to this FoM in this work.

It is noteworthy that for ESAs scanning in the upper hemisphere, f_H may not be equal to the grating lobe frequency, f_g , set by the array periodicity D (typically half-wavelength spacing, $f_g = \frac{c_0}{2D}$ where c_0 is the speed of light in a vacuum, to avoid grating lobes). Often

Figure 1.2: Array bandwidth definition. (a) Illustration of an active infinite array; (b) Bandwidth ratio (BWR) definition.

times, this rule is overlooked as it can significantly ease the design of UWB-ESAs, but at the cost of non-ideal *aperture sampling* which increases the number of feed electronics and, thus, cost expenditures.

Fractional Bandwidth (FBW): is bandwidth defined as $FBW = 200 \cdot \frac{f_H - f_L}{f_H + f_L}$. FBW is related to BWR by $BWR = \frac{200 + FBW}{200 - FBW}$.

Bandwidth (BW): is bandwidth defined as $BW = f_H - f_L$.

Aperture Sampling: is the rate f_H/f_g at which a designed array element achieves the ‘Nyquist’ (or half-wavelength) interval spacing. It is desirable for f_H/f_g to be equal to 1 or 100%. If $f_H < f_g$, the aperture becomes oversampled and the elements must scale downwards in size to operate at the designed f_g , leading to more elements required to fill a given area and, consequently, more T/R modules. This effect is charted in Fig. 1.3, where more than a 50% increase in T/R modules results from 80% aperture sampling as one such example. Thus, it is important to maintain $f_H = f_g$ within the scan volume for cost considerations, although it typically can be a challenging task for the designer to achieve due to various impedance-disrupting phenomena.

(Active) Orthogonal Port Isolation: is a measure referring to dual-polarized arrays when two ports, one for each orthogonal polarization, co-exist within the array periodic unit cell. It quantifies the coupling between orthogonal ports when all the array elements of one polarization are excited. It is desirable to minimize this figure as it leads to undesired cross-polarization radiation and scan or polarization dependent active VSWR. Typical benchmark values for orthogonal port isolation are less than -10 dB.

1.2.2 Polarization

Co-polarization field (Co-pol.): is the radiated (or received) far electric field of the antenna in an intended (reference) direction, i.e. $\mathbf{E}_{co} = E_{co}\hat{\mathbf{e}}_{co}$. It is noted that the co-polarization field of an array is a function of the observation angle and frequency.

Figure 1.3: Effects of aperture sampling on the number of required T/R modules for a rectangular grid array.

Cross-polarization field (Cx-pol.): is the polarization orthogonal to a specified co-polarization i.e. $\mathbf{E}_{cx} = E_{cx}\hat{\mathbf{e}}_{cx}$ where $\hat{\mathbf{e}}_{cx} \perp \hat{\mathbf{e}}_{co}$ and $\hat{\mathbf{e}}_{cx} \perp \hat{\mathbf{k}}$, with $\hat{\mathbf{k}}$ being the radiation propagation (or receiving) direction. This definition is general enough, but does not lead to a unique co-polarization field; many co-polarizations are possible depending upon the coordinate system used [31].

Polarization efficiency (p): is the ratio of the power received by an antenna from a given plane wave of arbitrary polarization to the power that would be received by the same antenna from a plane wave of the same power flux density and direction of propagation in which the state of polarization has been adjusted for maximum received power [27], i.e.

$$p = \frac{P_{rcv}}{S^{inc}A_e}, \quad 0 \leq p \leq 1 \quad (1.3)$$

where P_{rcv} is the receive power, S^{inc} is the incident field power, and A_e is the effective aperture of the receiving antenna. p is thus the polarization loss from the wave-antenna interaction at a certain direction of propagation, and is commonly written as

$$p = |\hat{\mathbf{e}}_w \cdot \hat{\mathbf{e}}_a^*|^2 \quad (1.4)$$

where $\hat{\mathbf{e}}_w$ and $\hat{\mathbf{e}}_a$ are the wave and antenna polarization unit vectors, respectively.

Cross-Polarization Isolation (XPI): is the ratio of wanted power level to unwanted power level in the same communication channel when the transmitting antenna is radiating some nominally orthogonally polarized fields in the $\hat{\mathbf{u}}$ and $\hat{\mathbf{v}}$ directions. This is indicated by the blue lines in Fig. 1.4 and represents the amount of interference on a given channel (in this case receive channel 1 i.e. XPI_1) due to polarization degradation stemming from a depolarizing propagation medium or radiated cross-polarization. The receive channel XPIs are written in terms of output voltages by

$$XPI_1 = \frac{|V_{1u}|^2}{|V_{1v}|^2}, \quad (1.5) \quad XPI_2 = \frac{|V_{2v}|^2}{|V_{2u}|^2}. \quad (1.6)$$

where $V_{1u}(V_{1v})$ and $V_{2u}(V_{2v})$ respectively indicate the output voltages on receive channels 1

Figure 1.4: Cross-polarization isolation (XPI) and cross-polarization discrimination (XPD) in dual-polarized systems. The transmit channel radiates nominally orthogonal polarized fields of arbitrary bases $\hat{\mathbf{u}}, \hat{\mathbf{v}}$.

and 2 due to some field polarized in the $\hat{\mathbf{u}}(\hat{\mathbf{v}})$ direction. Higher values of XPI are desirable as they indicate less interference on a channel. Although a useful parameter, XPI is typically difficult to measure and *cross-polarization discrimination* is more commonly measured.

Cross-Polarization Discrimination (XPD): is the ratio of the received output co-polarized power on a given communication channel to that of the cross-polarized power on the orthogonally polarized channel. The red lines in Fig. 1.4 represent XPD_u where $\hat{\mathbf{u}}$ is assumed to be co-(cross-)polarized with receiver antenna 1(2). The receive channel XPDs are written in terms of output voltages by

$$\text{XPD}_u = \frac{|V_{1u}|^2}{|V_{2u}|^2}, \quad (1.7)$$

$$\text{XPD}_v = \frac{|V_{2v}|^2}{|V_{1v}|^2}. \quad (1.8)$$

Higher values of XPD are desirable as they indicate lower cross-polarization. XPD is more directly measurable than XPI and lends itself more to typical chamber antenna measurements.

Furthermore, XPD and XPI can be generalized if the transmit and receive channel hardware are balanced (i.e. equal channel gains) by $\text{XPD}_u = \text{XPD}_v$ and $\text{XPI}_1 = \text{XPI}_2$, and, in addition, XPD and XPI can be shown to be equivalent if the propagation medium is symmetric [32] (i.e. free-space), resulting in

$$\text{XPI} = \text{XPI}_1 = \text{XPI}_2 = \text{XPD} = \text{XPD}_u = \text{XPD}_v. \quad (1.9)$$

Lastly, it is noteworthy to mention that both XPD and XPI are communication system FoMs since they involve end-to-end voltage quantities that depend not only on the antennas but also on the wireless communication channel and the transmit/receive chain channels.

Figure 1.5: Qualitative graphical depiction of the calculation of the cross-polarization ratio (CPR) pattern in dB polar format from the difference between cross-polarization (cx-pol.) and co-polarization (co-pol.). Red colors indicate higher magnitude.

Cross-Polarization Ratio (CPR): is the ratio of the power density in the cross-polarized component of the radiated far-field wave (S_{cx}) to that in the co-polarized component (S_{co}) [27]:

$$\text{CPR} = \frac{p_{cx}}{p_{co}} = \frac{S_{cx}}{S_{co}} = \frac{|E_{cx}|^2}{|E_{co}|^2}, \quad 0 < \text{CPR} < \infty \quad (1.10)$$

where $p_{co(cx)} = S_{co(cx)}/S_w$ and S_w is the total power density in the wave¹. Lower CPR is desirable with typical values ranging from -15 dB and sometimes as low as -35 dB. If we assume that both receive channel antennas are orthogonal and the communication medium is symmetric, then it can be shown that [27]

$$\text{CPR} = \frac{1}{\text{XPD}}. \quad (1.11)$$

CPR is thus very similar to XPD, but does not include all system effects; in other words it is purely an antenna FoM. Frequently throughout this dissertation, CPR will be written in terms of its operating frequency and observation or scan angles (θ, ϕ) by

$$\text{CPR}(f, \theta, \phi) = \frac{|\mathbf{E}_{cx}(f, \theta, \phi)|^2}{|\mathbf{E}_{co}(f, \theta, \phi)|^2} \quad (1.12)$$

and this quantity will be graphically expressed for the upper hemisphere in dB as in Fig. 1.5 with qualitative example radiation patterns in polar format at a fixed frequency. The azimuthal and radial directions correspond to the observation (scan) angles ϕ and θ , respectively, and areas with deeper shades of red indicate higher levels of cross-polarization.

¹Assuming the same effective aperture for two antennas, the co- and cross-polarization efficiencies sum to unity i.e. $1 = p_{co} + p_{cx}$ [27].

Axial Ratio (AR): is the ratio of the major to minor axes of a polarization ellipse [26]. AR is a useful FoM for circular polarization (CP) to quantify cross-polarization, and can be related to CPR through [27]

$$|\text{AR}| = \left| \frac{1 + \sqrt{\text{CPR}_{\text{CP}}}}{1 - \sqrt{\text{CPR}_{\text{CP}}}} \right| \quad (1.13)$$

where CPR_{CP} restricts CPR to CP. Furthermore, p and XPD can be explicitly written in terms of AR for arbitrary elliptically polarized waves by [27]

$$p = \frac{1}{2} + \frac{4\text{AR}_a\text{AR}_w + (\text{AR}_a^2 - 1)(\text{AR}_w^2 - 1) \cos 2\Delta\tau}{2(\text{AR}_a^2 + 1)(\text{AR}_w^2 + 1)} \quad (1.14)$$

$$\text{XPD} = \frac{\text{AR}_{cx}^2 + 1}{\text{AR}_{co}^2 + 1} \cdot \frac{(\text{AR}_{co}^2 + 1)(\text{AR}_w^2 + 1) + 4\text{AR}_{co}\text{AR}_w + (\text{AR}_{co}^2 - 1)(\text{AR}_w^2 - 1) \cos 2\Delta\tau_{co}}{(\text{AR}_{cx}^2 + 1)(\text{AR}_w^2 + 1) + 4\text{AR}_{cx}\text{AR}_w + (\text{AR}_{cx}^2 - 1)(\text{AR}_w^2 - 1) \cos 2\Delta\tau_{cx}} \quad (1.15)$$

where $\text{AR}_{a(w)}$ is the AR of the antenna(wave), $\text{AR}_{co(cx)}$ is the AR of the co(cross)-polarization component of the antenna, and $\Delta\tau$ is the relative tilt angle of the polarization ellipse (i.e. alignment angle). Despite being fairly involved, (1.15) can be reduced in one form by considering an ideal dual linearly polarized receiving system (i.e. $|\text{AR}_a| = |\text{AR}_{co}| = |\text{AR}_{cx}| = \infty$) in which the antenna is aligned with the incoming wave ($\Delta\tau_{co}=0$) to become [27]

$$\text{XPD} = |\text{AR}_w|^2. \quad (1.16)$$

It is noteworthy to point out that AR, in addition to preceding cross-polarization FoMs, are all related (i.e. by (1.9), (1.11), (1.16), etc.), such that in the case of linear polarization: lower CPR $\leftrightarrow$ higher XPD $\leftrightarrow$ higher AR $\leftrightarrow$ nearer unity p_{co} . In the case of CP, the former changes where unity AR is optimal. Thus, the recurring statement of “low cross-polarization” can be extended with similar interpretation across different FoMs e.g. an antenna engineer may interpret low cross-polarization in terms of low CPR whereas a communications engineer may think in terms of high XPR or XPI. In this dissertation only linear polarizations will be considered, but the findings in the work can be further extended for CP considerations through the provided AR relationships.

Up until this point, the previous polarization FoMs have been suited for communication systems, whereas the following polarization definitions are more common for radars and sensing systems.

Jones Matrix (J): is the transfer function that relates an incident electric field of arbitrary orthogonal polarization bases ($E_u\hat{\mathbf{e}}_u$ and $E_v\hat{\mathbf{e}}_v$) to the output voltages (v_1 and v_2) of the dual-polarized receiver channels by [29]

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} J_{11} & J_{12} \\ J_{21} & J_{22} \end{bmatrix} \begin{bmatrix} E_u \\ E_v \end{bmatrix}. \quad (1.17)$$

From this, the unknown incident field state of polarization can be found by inverting the Jones Matrix, $\mathbf{J}$. XPD and XPI can be written as a function of elements of $\mathbf{J}$ by [30]

$$\text{XPD}_u = \frac{|J_{11}|^2}{|J_{21}|^2} \quad (1.18)$$

$$\text{XPD}_v = \frac{|J_{22}|^2}{|J_{12}|^2} \quad (1.19)$$

$$\text{XPI}_1 = \frac{|J_{11}|^2}{|J_{12}|^2} \quad (1.20)$$

$$\text{XPI}_2 = \frac{|J_{22}|^2}{|J_{21}|^2} \quad (1.21)$$

which plainly show that common antenna polarization FoMs (including CPR) vary as a function of the off-diagonal elements and thus change depending upon the coordinate system of the incident wave polarization.

Intrinsic Cross-Polarization Ratio (IXR): is a FoM to characterize cross-polarization effects in dual-polarized phased array polarimeters that is independent of coordinate systems as defined by [30]

$$\text{IXR} = \left(\frac{\kappa(\mathbf{J}) - 1}{\kappa(\mathbf{J}) + 1} \right)^2, \quad 0 < \text{IXR} < 1 \quad (1.22)$$

where $\kappa(\mathbf{J})$ is the spectral condition number [33] of the Jones matrix in (1.17). IXR constitutes a measure of the degree of orthogonality between the dual-polarized channels and ultimately indicates the total relative error of the *fully calibrated polarimeter in the presence of system noise and channel cross-coupling*; the upper bound of this error can be shown to be [30]

$$\frac{\|\Delta \mathbf{e}\|}{\|\mathbf{e}\|} \lesssim \left(1 + 2\sqrt{\text{IXR}} + \dots \right) \left(\frac{\|\Delta \mathbf{J}\|}{\|\mathbf{J}\|} + \frac{\|\Delta \mathbf{f}\|}{\|\mathbf{f}\|} \right) \quad (1.23)$$

where the measured output voltage vector is $\mathbf{f} = [v_1 \ v_2]^T$ and the estimated Jones vector is $\mathbf{e} = [E_u \ E_v]^T$ i.e. from (1.17), respectively the output receiver voltages and the polarization state of the incident electric field². In other words, regardless of inaccuracies associated with $\mathbf{J}$ and $\mathbf{f}$ through errors in measurement, modeling, etc., an array with lower (good) IXR will lead to a more accurate measurement than one with worse IXR if the same calibration/correction was applied to both arrays in the same system³. Consider three such examples where $\text{IXR} = -7.5 \text{ dB}$, -15 dB and -30 dB ; this leads to $2\sqrt{\text{IXR}} = 0.84, 0.35$, and 0.06 and indicates that the measurement total relative error upper bound of (1.23) will be magnified by 84%, 35% and 6%. *Thus, lower values of IXR are more desirable and indicate better calibratability; lower IXR can only be achieved with better design of the polarimeter and the array CPR (e.g. maintaining better channel orthogonality via low aperture cross-polarization or better channel isolation).* IXR is ultimately a more convenient FoM for situations where the polarization of the received wave by the antenna is not known a priori (i.e. radars, radiometers, etc.) and when polarimetric calibration can be used to compensate for polarization degradation.

²Note that the definition of IXR used in (1.22) results in negative dB values to be consistent with the definition of CPR i.e. lower IXR and CPR both indicate lower cross-polarization. This is different from [30] where IXR yields positive dB values (like XPD and XPI); this consequently leads to a small modification in (1.23).

³By extension, this also indicates that an array with bad IXR at a particular scan angle will also lead to worse CPR post-calibration at that scan angle.

1.3 Significance

A high-efficiency UWB-ESA with greater than 6:1 BWR and very low CPR or IXR within a wide scan cone will have broad impacts on modern and future RF and millimeter-wave systems. This section will detail the importance and impact of (i) wide bandwidth and good matching and (ii) good polarization properties on an antenna array for communication, sensing, countermeasure, and multifunctional RF systems.

1.3.1 Significance of Bandwidth and Matching

1.3.1.1 Communications

ESAs remain in high demand for communication systems on satellite, maritime, aeronautical, and land mobile terminal applications due to their ability to form multiple independently high-speed scanned beams electronically with the potential for low-weight and low-profile platforms [34]. Typical communication systems utilize reflector antenna systems that are heavy/bulky, vulnerable to the inertia of the platform, and thus have a higher chance of failure due to moving parts.

Aside from these innate advantages, the wide bandwidth of a UWB-ESA further benefits these systems by enabling increased throughput communications [2, 3] in addition to low-power wideband transmission [4]. This can be seen upon considering the capacity of a communication channel, i.e. the upper-bound on the rate of information that can be reliably transmitted over that communication channel. For the most straightforward channel model, the additive white Gaussian noise (AWGN) communication channel capacity, C_{AWGN} , is given by [35]

$$C_{AWGN} = BW \log_2 \left(1 + \frac{P}{N_0 \cdot BW} \right) \text{ [bits/sec]}, \quad (1.24)$$

where P is the average received power and N_0 (W/Hz) is the noise power spectral density. The term $\frac{P}{N_0 \cdot BW}$ is equivalent to the signal-to-noise ratio (SNR) of the

Figure 1.6: The effect of bandwidth BW on channel capacity C_{AWGN} for different P/N_0 . (a) Bandwidth-limited regime; (b) Power-limited regime.

channel. A relationship between SNR and BW then exists as depicted in Figs. 1.6(a)-(b) for resulting channel capacities C_{AWGN} as a function of bandwidth BW for two different received power-to-noise spectral density values ($\frac{P}{N_0} = \{0.5, 1\} \cdot 10^9$). Fig. 1.6(a) shows the ‘bandwidth-limited regime’ where the capacity rapidly increases with the available bandwidth (thus its name) while (b) observes a wider operating bandwidth, ΔBW_2 , in the ‘power-limited regime’ where the power is the determining factor of the capacity. A few conclusions can be drawn from these figures:

- In (a), increased BW results in increased throughput and enables higher BW to be traded for lower transmit power (indicated by the dotted horizontal markers as one such example).
- In (b), wider BW yields more throughput per unit bandwidth for higher power transmissions (indicated by the change in capacity ΔC_{AWGN} over the same BW (black markers) for different $\frac{P}{N_0}$).

Thus, minimal system losses (i.e. good antenna matching, high-efficiency, etc.) represented by P in Eq. (1.24) and high BW are of critical importance to maximize channel throughput and/or enable low power communications.

Furthermore, if we consider a mobile receive platform and a stationary transmitting UWB-ESA platform (i.e. a communications satellite in geostationary orbit) establishing a point-to-point link over free-space conditions, the receiver carrier-to-noise ratio (C/N, or SNR) is given by [12]

$$C/N = \frac{P_r}{kT_{sys}BW} = (P_t G_t) \left(\frac{G_r}{T_{sys}} \right) \left(\frac{1}{kBW} \right) \left(\frac{\lambda}{4\pi R} \right)^2, \quad (1.25)$$

where $P_{t(r)}$ is the transmit(receive) power, $G_{t(r)}$ is the transmit(receive) antenna gain, T_{sys} is the equivalent system noise temperature, k is Boltzmann's constant, R is distance, λ is the operating wavelength, and any polarization or atmospheric absorption losses are neglected. BW has a negative impact on C/N and reduces the maximum link range, while good matching levels and high-efficiency antennas increase the EIRP ($P_t G_t$) and maximize the G_r over T_{sys} ratio through reduced system noise and losses to ultimately help compensate for higher BWs. It is important to retain strong matching within the scan volume so the link is not compromised due to poor C/N.

1.3.1.2 Active Sensing and Radars

Bandwidth and matching inherently play significant roles in active sensing and radar systems through controlling range and detection in the radar range equation via the system SNR given for a monostatic radar by [12]

$$SNR = \frac{P_r}{kT_{sys}BW} = (P_t G) \left(\frac{G}{T_{sys}} \right) \left(\frac{\sigma}{kBW} \right) \frac{\lambda^2 10^{-0.2\alpha R}}{(4\pi)^3 R^4}, \quad (1.26)$$

where $G = G_t = G_r$ is the single transmit/receive antenna gain, α is the one-way atmospheric attenuation coefficient, and σ is the average target radar cross-section (RCS). As with communication systems, BW has adverse effects on SNR, while good matching levels facilitate better SNR through improved EIRP and G/T ratios.

Despite BW having unfavorable impacts on the system SNR, UWB radars and imagers offer improved target identification by providing increased range resolution (ΔR) through [36]

$$\Delta R = \frac{c_0}{2BW} \text{ [m]}. \quad (1.27)$$

where c_0 is the speed of light in a vacuum. Effectively, higher BW (shorter pulses) leads to heightened detected target range measurement accuracy. Shorter pulses also offer innate advantages such as more detailed target characteristics and operation stability at low observation elevation angles [37], multipath resilience [38], and more robust clutter rejection capabilities [39]. For the aforementioned reasons, UWB-ESA radar technologies find extended use in through-wall detection applications such as imaging for first-responder rescue operations [7], ground penetrating radars for underground exploration [40–42], and underground unexploded ordinance detection (UXO) using Synthetic Aperture Radar (SAR) [43, 44]. Considerable attention has been drawn to the utilization of UWB-ESAs for the detection of early-stage breast cancer [8, 45].

UWB radars also offer advantages through improved secretiveness (i.e. covertness) and immunity against external threats. Wider bandwidths facilitate low-probability-of-intercept (LPI) radars with reduced detectability by spreading the transmitted signal’s energy over a broader spectrum (reducing the power emanated per unit bandwidth) [12]. Likewise, operation over a wider bandwidth forces noise jammers to spread their energy over a much broader band (as opposed to focusing high power density over a narrow band or single-frequency i.e. spot jamming), resulting in enhanced jamming resilience. A demonstration of this resilience for an air-surveillance radar in the presence of barrage (wideband noise) jamming was presented in [38] with results for the cases of narrowband reflector and wideband planar array systems in Figs. 1.7(a)-(b), respectively. In this instance, the jammer is more effective if fewer targets are seen due to overwhelming the receiver threshold in the presence of ex-

Effect of Jamming on Target Detection Ability in Narrowband/Wideband Systems

Figure 1.7: The effect of barrage jamming on the field of views of narrowband and wideband radar system. Note that in this instance the jammer is more effective if fewer targets are seen. (a) Narrowband parabolic reflector system; (b) Wideband planar phased array system. (**SOURCE:** M. Skolnik, G. Linde, and K. Meads, “Senrad: An Advanced Wideband Air-Surveillance Radar,” *IEEE Trans. Aerosp. Electron. Syst.*, vol. 37, pp. 1163-1175, Oct. 2001.) Adapted to include text.

ternal noise; the narrowband system fails to detect the majority of targets while the wideband system successfully retains the ability to detect several targets.

1.3.1.3 Passive Sensing

Significant interest has been placed in utilizing UWB-ESAs for radio astronomy over the years (specifically development efforts towards the Square Kilometre Array project) [9–11]. Such interest stems from the ability of ESAs to rapidly scan large field of views with multiple independent beams that can be adjusted on the fly by controlling beamforming coefficients in the signal processing receive hardware. UWB-ESAs, UWB-ESA reflector feeds, and Focal Plane Arrays (FPA) can further offer inherent benefits in terms of sensitivity as can be seen from the ideal radiometer equation [12]

$$\frac{\Delta T}{T_{sys}} = \frac{1}{\sqrt{\tau \cdot BW}} \quad (1.28)$$

Figure 1.8: UWB-ESAs for modern astronomical radio telescopes. (a) 20,000 element dual-polarized Vivaldi array for radio astronomy; (b) 8×7 element dual-polarized Vivaldi array as a phased array feed for a parabolic dish radio telescope. (**SOURCE:** K. Warnick, R. Maaskant, M. Ivashina, D. Davidson, and B. Jeffs, “High-sensitivity phased array receivers for radio astronomy,” *Proc. IEEE*, vol. 104, pp. 607-622, March 2016.)

where ΔT is the standard deviation of the output temperature (sensitivity) and τ is the receiver integration time. Large arrays with wide bandwidths improve system sensitivity for the detection of weaker astronomical signals. Conversely, increased bandwidth can increase survey speed while maintaining the same system sensitivity. To minimize system noise penalties in T_{sys} and also maximize SNR, active array impedances must remain low-loss and well-matched to the front end amplifiers when scanning [46]. An ongoing research challenge is to improve ESA impedance matching as it is typically poorer than traditional horn feed systems [11].

Practical implementations of modern UWB-ESAs for radio astronomy are shown in Fig. 1.8(a-b), where in (a) 20,000 dual-polarized Vivaldi elements are employed as a multi-beam aperture array and in (b) an 8×7 dual-polarized Vivaldi array is used as a phased array feed for a radio telescope [11].

The concept of covert passive radars has also been proposed to detect and track objects with ESAs through the use of very small reflections from third-party illuminators such as commercial broadcasting and GPS satellites [47]. Recently, the Air Force

Research Laboratory has released a request for the design of a 10:1 dual-polarized phased array antenna for covert passive radar technology [48].

1.3.2 Significance of Polarization

1.3.2.1 Communications

In communication systems, the transmitter and receiver antennas are designed to radiate or receive polarized beams with polarizations that are set a priori. Due to this, it is necessary to maintain strong polarization efficiency p , c.f. Eq. (1.4), within a given link to maximize the received power P_r as given by the Friis equation,

$$\frac{P_r}{P_t} = G_t G_r \left(\frac{\lambda}{4\pi R} \right)^2 p, \quad (1.29)$$

where a point-to-point link in free-space conditions is considered⁴. Minimizing cross-polarization (p closer to unity) to maximize SNR, throughput, and at times avoid loss of service, is thus a critical concern. However, there can be a plethora of unknowns or dynamic factors in the propagation path of a signal from transmitter to receiver that can deteriorate p such as depolarizing mediums (i.e. rain, ionospheric Faraday rotation, etc.) or random handset orientation.

Polarization diverse systems (i.e. dual-polarized systems) are employed to help combat fading in such circumstances and retain adequate SNR [27, 49]. The quality of polarization diverse links can be characterized by several cross-polarization FoMs such as XPD and XPI (c.f. Fig. 1.4), etc. that involve quantifying the amount of interference on a certain channel as discussed in Section 1.2; low cross-polarization was shown to be critical in each case to improve XPD and XPI (i.e. minimize CPR) and reduce channel interference for proper signal reception. Furthermore, polarization

⁴In this instance, near-unity p generally implies the polarization efficiency of the co-polarization, p_{co} , is near-unity. Low CPR facilitates p_{co} and p_{cx} being closer to unity and zero, respectively c.f. (1.10); CPR and p are also related to XPD and AR as discussed in Section 1.2.

diverse systems offer the same channel capacity benefits as a spatial diversity (antennas separated by space) system but while saving space and cost due to co-located orthogonal antennas [50]. This can alternatively be thought of as increasing channel capacity by as much as a factor of two as compared to a single-polarized channel by employing two simultaneous orthogonally polarized channels (frequency reuse).

1.3.2.2 Radars and Sensing

In contrast with communication systems, the implications of polarization on radar and sensing systems are not as straightforward because only the receiver system is polarized a priori while the sources of radiation (incident to the receiver) are generally either complex depolarizing scatterers illuminated by a polarized wave, or natural sources of unpolarized or partially polarized electromagnetic or thermal black-body radiation (i.e. terrestrial emissions, celestial signals from astronomical bodies such as “radio stars,” pulsars, etc.). Thus, the incident wave polarization is typically *not known a priori* and may be irrelevant. This is depicted in Fig. 1.9 for the case of a dual-polarized radiometer (polarimeter) or sensor where the polarized portion of the source is of arbitrary orthogonal bases $\hat{\mathbf{u}}$ and $\hat{\mathbf{v}}$ with respect to that of the V and H basis of the receive antennas. Due to this type of ambiguity, general cross-polarization FoMs that apply to communication systems are not typically relevant for radar and sensing systems, and it is not as clear as to how to interpret the concept of cross-polarization.

To better quantify cross-polarization for polarimetric radiometers, the intrinsic cross-polarization (IXR, c.f. (1.22)) FoM was developed in [30]. In an ideal world, such systems can be perfectly calibrated (by adjusting beamforming weights) for the beam at a certain scan direction and correct for any unwanted polarization degradation effects through precise knowledge of the transfer Jones matrix $\mathbf{J}$ [29]. However, $\mathbf{J}$ has to be deduced from a known scatterer through measurements that is susceptible

Figure 1.9: Depiction of a polarimetric sensing system in which the receiver is not aware of the polarization bases ($\hat{\mathbf{u}}, \hat{\mathbf{v}}$) a priori as the transmitter is a natural source of energy or an arbitrary scattered return.

to noise and measurement/system errors (e.g. low noise amplifiers, array impedance mismatch, mutual coupling between elements, etc.), ultimately limiting the effectiveness of calibration (i.e. increased total relative errors c.f. (1.23)) and thus the ability to correct the receiver system cross-polarization. As discussed in Section 1.2, IXR is a measure of the resulting polarimeter calibratability, and is directly related to the degree of orthogonality between the dual-polarized channels⁵. As a result, low antenna array CPR is an extremely important factor to ensure good IXR (and high channel orthogonality) for high measurement accuracy.

Polarization diversity in radars also enables the acquisition of the target return complete polarization state, improved target detection in the presence of clutter and jammers, and enhanced target discrimination [51]. A dual-polarized 9:1 Vivaldi array was recently used for full polarimetric airborne measurements of snow-over-sea-ice [6].

⁵It is noteworthy to point out that IXR is invariant to the incident polarization coordinate system and the calibration procedure by nature of its definition, unlike typical cross-polarization FoMs used in dual-polarized communication systems that will inevitably change due to such factors like CPR, XPD, XPI, etc.

Figure 1.10: Depiction of an Electronic Warfare (EW) scenario.

1.3.3 Electronic Warfare

UWB-ESAs find significant interest for use in electronic warfare (EW) systems and its classical divisions: electronic counter measures (ECM), electronic support measures (ESM), and electronic counter countermeasures (ECCM). ESAs are generally attractive for EW as they can achieve high gain, multiple independent beams, low side lobe levels, and adaptive nulling for jammer cancellation [12, 36]. Considering an EW scenario as depicted in Fig. 1.10 where a transmitter and receiver are establishing some unknown communications or bistatic radar functionality over a channel, direct benefits pertaining to UWB-ESAs in EW can be outlined:

- ECM intends to deliberately degrade or destroy the functionality by jamming the receiver. As the electrical characteristics of the function to be jammed are not generally known a priori and can be dynamic (i.e. frequency and polarization agile), dual-polarized UWB-ESAs with good polarization control (low cross-polarization) offer high value. Additionally, minimizing losses in the ECM system (i.e. good matching) is important for high power to saturate the receiver.
- ESM passively eavesdrops RF emissions to detect, intercept, identify, and locate threats. As these emissions are not known a priori, it is again desirable to have

wideband, polarization diverse operation with large dynamic range. In this manner, the ESM could be able to recognize and locate the ECM threat to aid efforts in marginalizing its effectiveness.

Furthermore, low cross-polarization UWB-ESAs provide substantial ECCM benefits (actions taken to protect the radar from ECM). UWB performance inherently provides ECCM by enabling frequency agility to circumvent spot (single-frequency) jamming and force the jammer to spread its power over a wide operating range; this results in improved jamming resilience as discussed in Section 1.3.1.2. Additionally, low cross-polarization helps mitigate the potential of cross-polarization jammers to exploit systems utilizing antennas/arrays that have high cross-polarized gains at certain scan angles or side lobes that could render ECCM techniques such as side lobe blanking (SLB) or side lobe cancelling (SLC) ineffective [36]. For example in the case of a monopulse radar, the system can trigger false target indications due to cross-polarized jamming (an issue exploited in reflector antenna systems due to cross-polarized side lobes referred to as “Condon lobes”) [52]. Moreover, if the radar was utilizing a constant false alarm rate (CFAR) algorithm, the receive subsystem threshold could automatically be raised when high amounts of external noise enter through a cross-polarized side lobe to ultimately impair the ability to detect targets.

1.3.4 Multi-Band and Multifunctional Systems

Modern mobile military platforms are required to support a growing number of RF functions that include advanced radar/sensing, communications, signals intelligence (SIGINT), and EW as can be seen in the case of a naval platform in Fig. 1.11(a). Currently, these functionalities are supported through densely-packed groups of antennas (otherwise coined “antenna forests” or “antenna farms”) c.f. Fig. 1.11(b) in which each antenna is typically responsible for a specified narrowband mission. This trend is particularly problematic as space, weight, and low observable considerations

come at a premium. Additionally, radio-frequency interference (RFI) between antennas poses unforeseen performance issues as the antennas are typically designed to operate in the absence of one another, and in-band or even out-of-band coupling from a transmit system in close proximity to a receive antenna could easily saturate or even burn the receiver system.

A high-performance low-loss UWB-ESA system with excellent wide-scanning capabilities and polarization agility plays a critical role in alleviating these issues as it is capable of simultaneously supporting arbitrary independent mission capabilities in an omni-azimuth (full 360°) field-of-view (FOV). This approach to integrating RF systems is often referred as a multifunctional RF system (MFRFS), and aims to remedy such concerns by consolidating RF mission functionalities into adaptable/programmable RF electronics systems that are serviced by a single or reduced number of UWB-ESAs as depicted in Fig. 1.11(c). In this MFRFS, common RF operations across the electromagnetic spectrum are jointly serviced by a single adaptable transceiver unit (such as an RF-FPGA [53]) and a dual-polarized UWB-ESA. In this manner, the UWB-ESA not only enables integrated system functionality, but also multi-band adaptation to accommodate dynamic environments such as harsh weather conditions, ECM conditions, etc. Although a very attractive notion, this concept heavily requires electrical properties that currently elude state-of-the-art UWB-ESAs: extreme amounts of bandwidth, excellent impedance matching at wide scan, high radiation efficiency, and low cross-polarization.

1.3.5 Summary

We will briefly re-cap some critical points.

Wide bandwidth enables:

1. Low-power transmission and higher channel capacity in communication systems
2. Improved radar range resolution, secretiveness, and clutter/multipath resilience

Figure 1.11: (a) Conceptual multifunctional UWB-ESA application servicing various naval warship missions (**SOURCE:** Aviation Intel. (2013), “The B-2 bomber of the seas: Thoughts on the Zumwalt class destroyer and the Navy’s broken procurement strategy.” [Online]. Available: <http://aviationintel.com/ddg-1000-close-weapon-system-zumwalt-class-ponderings/>); (b) Typical naval vessels host more than one hundred antennas to perform various radar, sensing, communication, and electronic surveillance operations (**SOURCE:** The Economist. (2011), “Liquid radio.” [Online]. Available: <http://www.economist.com/node/18007506>); (c) An illustration of the multifunctional RF concept in which various RF functionalities spanning across the electromagnetic spectrum are jointly serviced by a single wideband adaptable transceiver unit (such as an RF-FPGA) and a UWB-ESA.

3. Increased radiometer sensitivity and faster survey speeds
4. Jamming resilience and increased defense capabilities
5. System integration

Low cross-polarization enables:

1. Low polarization loss factors in communication systems
2. More accurate measurements and beam calibration in sensing systems
3. Improved defense capabilities (jammer resilience, reduced false target indications, improved target detection, etc.)

As will be shown in sequel, most popular high-performance UWB-ESAs with greater than 6:1 BWR have alarming CPR levels (greater than +10 dB) in their main scanning beam throughout much of their scan volume and sometimes even in their side lobes. A wide-scan UWB-ESA with low cross-polarization throughout its scanning volume helps remedy these issues and delivers significant performance benefits.

1.4 Previous Work

This section will provide a brief overview of the most important research efforts culminating into various available state-of-the-art UWB-ESAs. Although this study is by no means complete, it will be sufficient to make the case that no state-of-the-art array can meet the demanding polarization and BW/matching FoM requirements needed for some of the future systems applications outlined in the previous section.

For clarity, UWB-ESAs are grouped into two categories: (i) Vivaldi end-fire type planar arrays and (ii) low-profile planar arrays.

1.4.1 Vivaldi (Flared-Notch or Tapered-Slot) Arrays

To the best of our knowledge, there exists no other high-efficiency (low-loss) antenna array that can compete with Vivaldi arrays in terms of bandwidth, matching level, and wide-scan impedance performance; bandwidths in excess of one decade ($>10:1$) are achievable with broadside VSWR <1.6 and $\theta = 45^\circ$ scan VSWR <2 [18, 54–56]. Decades of research and development [15, 18, 54–62] have produced a plethora of practical embodiments and design guidelines that have established Vivaldi arrays as the array-of-choice for many UWB-ESA systems [6, 11, 15, 55, 61]. A typical Printed Circuit Board (PCB) based Vivaldi antenna element is depicted in Fig. 1.12(a) highlighting some of its key features along with a fabricated dual-polarized design in Figs. 1.12(b). Arguably the most defining characteristic is the exponentially tapered slot (or fins) that acts as a smooth impedance transition from a relatively low impedance to free-space impedance by varying the width of the slot. Another critical component is the Knorr wideband balun/transition [63] that consists of a slotline cavity and open-circuit stub connected to a feed line to provides a balanced slotline excitation while transitioning to an unbalanced 50Ω feed stripline. Other embodiments utilizing this antenna element or similar variations of its components include successful machined stock-metal embodiments [56, 62] shown in Figs. 1.13(a) and (b) and hybrid builds that combine more than one manufacturing technology e.g. all-metal or conductive molded plastic with PCB feeding for the Mechanotch [64] or 3D printing with PCB in [65], shown in Figs. 1.14(a) and (b).

In addition to their diverse structural options, another desirable feature of Vivaldi arrays is their well-known and robust design guidelines. For instance, one relates the array depths (or profile), H , to the impedance bandwidth ratio, by [24] $\text{BWR} = \frac{f_H}{f_L} \approx \frac{2H}{D}$ where D is the unit-cell dimension (recall $D = \lambda_H/2$ for ideal aperture sampling i.e. $\lambda_H = \lambda_g$). This relationship is a convenient design feature as the experienced designer can achieve a desired bandwidth by adjusting the depth of the element and

Figure 1.12: PCB-based Vivaldi arrays. (a) Typical Vivaldi array antenna element; (b) An 8×8 5:1 dual-pol. PCB array (picture courtesy of Prof. Dan Schaubert, UMass Amherst).

Figure 1.13: All-metal Vivaldi arrays. (a) A 17×9 5:1 dual-polarized Body-of-Revolution (BOR) array manufactured with metal lathe (**SOURCE:** H. Holter, “Dual-polarized broadband array antenna with BOR-elements, mechanical design and measurements,” *IEEE Trans. Antennas Propag.*, vol. 55, pp. 305-312, Feb. 2007); (b) An 8×8 12:1 dual-polarized array manufactured with electrical discharge machining (**SOURCE:** R. Kindt and W. Pickles, “Ultrawideband all-metal flared-notch array radiator,” *IEEE Trans. Antennas Propag.*, vol. 58, pp. 3568-3575, Nov. 2010).

Figure 1.14: Hybrid Vivaldi arrays. (a) 4:1 hybrid ‘Mechanotch’ array (picture courtesy of Northrop Grumman); (b) 5:1 Broadband Additively Manufactured Array Antenna (BAMAA) array (**SOURCE:** V. Sanchez, T. Spence, and M. Cooley, “Broadband additively manufactured array antennas (BAMAAs) and planar-fed stepped notch (PFSN) arrays,” *Antenna Applications Symposium, Allerton Park, Monticello, IL*, pp. 520-532, Sep. 2015).

tuning around that point. Despite this design convenience, this very relationship also shows that deep element profiles are a consequence of high BWR Vivaldi designs by $H \approx \frac{\text{BWR}}{4} \lambda_H$; as two such examples, 5:1 and 10:1 BWR designs would respectively net profiles of approximately $1.25\lambda_H$ and $2.5\lambda_H$. As will be shown in more detail in Chapter 2, hindered polarization properties under wide scanning are engendered as a direct repercussion of this wide BWR and high profile design compromise in Vivaldi arrays.

Despite their high bandwidths and excellent impedance matching even under wide scan, Vivaldi arrays suffer from significant degradation of cross-polarization (or CPR) when scanning in the non-principal planes (most commonly associated with the diagonal-plane) [24, 66–69]. To highlight the severity of this problem, we plot the center embedded element pattern of a 19×19 10:1 Vivaldi array around mid-band at the E-plane, H-plane, and D-plane in Fig. 1.15(a). We note that the main beam

Figure 1.15: Persisting challenges in all state-of-the-art Vivaldi technologies and designs. (a) High cross-polarization at non-principal planes, in this case the D-plane cut of the center embedded element pattern; (b) Non-modular design and assembly due to the need of electrical connection across the entire depth of neighboring elements.

polarization of an ESA at a given scan direction is the same as its embedded element pattern at the respective observation angle [66]. It is seen that the cross-polarization in the principal planes remains very low even out to wide angles, while the D-plane alarmingly increases with elevation angle early on. This is particularly problematic as a loss of polarization control incurs away from the principal radiation planes (E-/H-planes). Therefore, in the absence of correctional measures, system operation would be effectively compromised when scanning off-axis in the non-principal planes due to high cross-polarization levels as elucidated in Section 1.3.2.

Calibration procedures incorporating excitation re-weighted feeding schemes have been investigated to create cross-polarization nulls at desired scan angles and frequencies for dual-polarized Vivaldi array configurations [69]. The procedure is virtually the same as to that which was discussed in Sections 1.2 and 1.3.2.2 for polarimetric calibration by virtue: a desired polarization is synthesized at a certain scan direction using beamforming weights derived from the ideal Jones matrix c.f. (1.17). Although providing successful correction/calibration of cross-polarization at desired scan an-

gles, these correctional procedures often do not meet all Vivaldi UWB-ESA system needs as they are:

1. Inherently narrowband
2. Tolerance prone or inaccurate (poor IXR)
3. Applicable only to dual-polarized configurations
4. Detrimental to side lobe cross-polarization levels
5. Time-consuming

The interested reader can find a more detailed discussion on these procedures and design examples in Appendix A.

Aside from these correctional procedures, research into incorporating a combline structure along a ‘Bunny Ear’ fin to further reduce cross-polarization has been conducted, but, despite offering a unique low-profile 3:1 design, has unfortunately resulted in little-to-no cross-polarization improvement [70, 71]. A clever variation of the Vivaldi antenna has also been developed in which electrically disconnected conductive strips are fed by a modular resonant element (i.e. a half-wave dipole), offering (with respect to traditional Vivaldi antennas) improved mechanical convenience and increased forward gain over a narrow band i.e. around 22% fractional bandwidth (1.25:1 BWR) [72]. The modular feeding enables several stacked planar layers to form a wideband reconfigurable antenna, but, due to the resonant narrowband nature of each reconfigured antenna, a significant number of layers would be necessary to attain similar BWRs as those of traditional Vivaldi antennas where >5:1 BWRs are readily achievable [57].

In addition to electrical performance considerations, significant interest has been placed upon reducing the fabrication complexity of dual-polarized Vivaldi arrays. Vivaldi antenna elements require electrical connection between adjacent elements to

avoid impedance resonances [73]. This can lead to complicated assembly in dual-polarized arrangements as seen in that of a PCB-based design in Fig. 1.15(b), where soldering was necessary along the large element depths at every intersection of orthogonally-polarized cards. Fabrication and assembly can thus become very expensive, and several recent research initiatives have revolved around eliminating the soldering process altogether by typically incorporating machined stock-metal to embody the high-profile element fins as can be seen in previous embodiments in Figs. 1.13 and 1.14. Although these designs simplify the construction of the dual-polarized flared notch portions, they require special machining and feeding techniques that affect cost, weight, and fabrication at high frequencies (above X-band). This challenge is highlighted here because as it will become clear in the later chapters i.e. Chapter 3, a beneficial byproduct of the proposed Sliced Notch Antenna (SNA) array is the avoidance of such electrical contact requirements between most of the upper part of the Vivaldi fins.

1.4.2 Low-Profile UWB Arrays

A bulk of UWB-ESA research over recent years has focused upon low-profile ($H \approx \lambda_H/2$) UWB-ESAs that improve scanning cross-polarization and have the potential of low-cost, high-volume manufacturing and assembly. However, as will become clear in the next sections, none of these technologies have yet to demonstrate the quality of wideband wide-scan impedance matching of Vivaldi arrays but have indeed solved the problem of high cross-polarization. An overview of prior work will be provided, and, for convenience, these arrays have been classified into three low-profile UWB array categories: vertically-integrated, “quasi-planar,” and planar; the differences between each category lay in the radiating structures, feeding schemes, and/or balun structures. Each category will be considered as follows in this dissertation:

- *Vertically-integrated*: the radiating structure i.e. dipoles (but not necessarily the feeding structure) are printed or machined in the form of planar cards that are then oriented vertically (longitudinal to the broadside scan direction) during assembly to form a planar array,
- *Quasi-planar*: the radiating structure resides on one or multiple transverse (with respect to the broadside scan direction) printed layers, whereas the feeding structure does not and typically invokes three-dimensional feeding structures,
- *Planar*: both the radiating and feeding structures are manufactured with planar fabrication techniques i.e. standard microwave PCB fabrication.

1.4.2.1 Vertically-Integrated UWB Arrays

Several low-profile ($H \leq \lambda_H/2$), vertically-integrated arrays have been developed in the past decade to deliver wide-scan UWB performance. These UWB arrays typically consist of capacitively coupled horizontal dipoles or connected dipole (i.e. current sheet) radiators placed above a ground plane, many of which are covered by superstrate dielectrics. Vertically-integrated arrays include but are not limited to the Bunny-ear [67, 74], connected-dipole arrays [20], balanced antipodal Vivaldi antennas (BAVA) [75–77], and printed dipoles with integrated baluns [22, 78, 79].

The Bunny-ear array displayed in Fig. 1.16(a) is able to achieve bandwidths of 5:1 while requiring external baluns/hybrid feeding schemes in addition to adjacent element electrical connection [67]. Alternatively, film resistors are connected between the Bunny ear elements and the ground plane to dissipate problematic resonances while being fed through external baluns/hybrids in [74], resulting in reduced efficiency and power handling; a dielectric-free aluminum-based implementation of this array in Fig. 1.16(b) was demonstrated recently with similar performance to meet demand needs for the Square Kilometre Array [80]. Similarly, earlier versions of BAVA and its doubly-mirrored extension (DmBAVA) are able to maintain up to 5:1

bandwidths while also utilizing external baluns/hybrids in the feeding network [76]. To eliminate the necessity of external baluns/hybrids in these arrays, the Banyan Tree Antenna (BTA) [81] array was developed to achieve a 4:1 bandwidth when fed directly with an unbalanced 50Ω feed line by mitigating in-band common-mode resonances with shorting strips connected to the radiating arms as shown in Fig. 1.17. A dual-polarized connected-dipole array with unbalanced feeding and interleaving orthogonal polarization cards was recently developed for mobile satellite terminals with approximately 30% FBW [82]. Further advancements in unbalanced-fed UWB-ESAs include a modified BAVA array that offers 10:1 bandwidth [19]. This array provides the widest reported bandwidth for low-profile arrays, and a very convenient, modular assembly at some expense in VSWR (<2.75 broadside, <4.2 at 45°) and some additional mechanical machining for the newly introduced metallic U-shaped channelized posts highlighted within pictures of the array in Fig. 1.18. Although these low-profile, vertically-integrated arrays maintain a $\text{CPR} < -12$ dB at $\theta < 45^\circ$ scan across the band, their impedance bandwidth is limited to approximately 6:1 with a modest $\text{VSWR} < 2$ at broadside scan.

Dipole arrays with integrated baluns have also made significant advancements over the last years. A dual-polarized coincident phase dipole array with an integrated Marchand balun shown in Fig. 1.19(a) was reported in [78] to obtain approximately 3:1 bandwidth in an infinite array environment. A single-polarized printed tightly-coupled dipole array (TCDA) shown in Fig. 1.19(b) was able to attain to 7.35:1 in [22] with slightly higher cross-polarization levels in the D-plane ($\text{CPR} \approx -8$ dB at $\theta = 45^\circ$), 93% aperture sampling, and an increased element profile $H \approx 0.8\lambda_g$. This array also relies on a Wilkinson power divider seen in the insert of Fig. 1.19(b) because each array unit cell includes two dipoles radiators. The use of Wilkinson power dividers makes this array more complex (placing and soldering a large number of chip resistors), lossy, prone to high-power damage, and increases the true array

Figure 1.16: Isometric views of low-profile vertically-integrated Bunny-Ear arrays. (a) An early 9×9 Bunny-Ear array operating over 1-5 GHz (**SOURCE:** J. J. Lee, S. Livingston, and R. Koenig, “A low-profile wide-band (5:1) dual-pol array,” *IEEE Antennas Wireless Propag. Lett.*, vol. 2, no. 1, pp. 46-49, 2003); (b) A recent 8×8 dielectric-free 0.2-1 GHz Bunny-Ear array (**SOURCE:** L. Garcia-Munoz et al., “Broadband active differential array for the mid-frequency SKA band,” *IEEE Antennas Propag. Mag.*, vol. 56, no. 2, pp. 27-38, Apr. 2014.).

Figure 1.17: Isometric views of Banyan Tree Antenna (BTA) arrays utilizing shorting strips connected to the radiating arms. (a) Single-polarized configuration; (b) Dual-polarized configuration. (**SOURCE:** S. Holland and M. Vouvakis, “The Banyan Tree Antenna Array,” *IEEE Trans. Antennas Propag.*, vol. 59, no. 11, pp. 4060-4070, Nov. 2011.)

Figure 1.18: A 10:1 dual-polarized U-channel BAVA array. (a) 8×9 prototype of a dual-polarized configuration (**SOURCE:** W. Elsallal and J. Mather, “An ultra-thin, decade (10:1) bandwidth, modular “BAVA” array with low cross-polarization,” *Proc. IEEE Antennas and Propagation Society Int. Symp.*, pp. 1980-1983, Jul. 2011.); (b) Close-up view showing the U-shaped channel metal post responsible for the strong interelement capacitive coupling (**SOURCE:** W. Elsallal et al., “Characteristics of decade-bandwidth, Balanced Antipodal Vivaldi Antenna (BAVA) phased arrays with time-delay beamformer systems,” *Proc. IEEE Phased Array Tech. Int. Symp.*, pp. 111-116, Jul. 2013).

profile to $H \approx 1.6\lambda_g$. Recently, a single-polarized 6.1:1 bandwidth TCDA with a parasitic superstrate has indicated $VSWR < 3.2$ scanning out to $\theta = 60^\circ$ in the H-plane and $\theta = 75^\circ$ in the E-plane through infinite array simulations with demonstrated low cross-polarization [79]; this comes at the expense of a truncated unit cell dimension to improve E-plane scan matching (i.e. 82% aperture sampling rate) and an increased profile of $H = 0.9\lambda_g$.

1.4.2.2 “Quasi-Planar” UWB Arrays

Several notable arrays were developed to elucidate the demand for maximal bandwidth with a planar architecture. The current sheet antenna (CSA) array [16, 83] is one of the first dipole-based arrays that leverages Wheelers’ current sheet concept [84] to attain wide bandwidths. CSAs can achieve up to 9:1 bandwidth by using a dual-polarized tightly-capacitively coupled horizontal dipoles in a coincident phase center arrangement as shown in Fig. 1.20. These printed dipoles are a quarter-wavelength apart from the ground plane at midband, thus they counteract the inductive nature of

Figure 1.19: Depiction of vertically-integrated dipole arrays with integrated baluns. (a) A coincident phase center dipole array with orthogonally integrated Marchand baluns (**SOURCE:** W. Pickles and W. Dorsey, “Proposed coincident phase center orthogonal dipoles,” *Antenna Applications Symposium, Allerton Park, Monticello, IL*, pp. 106-124, Sep. 2007.); (b) A single-polarized tightly-coupled dipole array with integrated balun (TCDA-IB) (**SOURCE:** J. Doane, K. Sertel, and J. Volakis, “A wideband, wide scanning tightly coupled dipole array with integrated balun (TCDA-IB),” *IEEE Trans. Antennas Propag.*, vol. 61, no. 9, pp. 4538-4548, Sep. 2013.).

the ground plane at lower frequencies and lead to increased bandwidth due to better matching at low frequencies. A planar interwoven spiral array has also demonstrated broadside 10:1 performance, but is limited to circular polarization only and scanning performance has yet to be addressed [85].

Other planar arrays such as the fragmented aperture array (FAA) [86] in Fig. 1.21 and the long slot and thumbtack arrays [17, 87–89] in Fig. 1.22 use connected electric or “magnetic” elements, respectively, that could yield infinite bandwidth if radiating in free space and infinite array configurations, but when unidirectional (half-space) radiation is needed, bandwidths are limited by resonances due to the introduction of a ground plane beneath the radiating layer. The elements in these arrays are contiguously connected to form continuous current sheets that can be modeled as an array of center-fed periodically-spaced (typically $\lambda_H/2$) dipoles attached at the adjoining ends. The long slot and thumbtack arrays achieve 4:1 bandwidth with low cross-polarization (CPR < -15 dB) with non-lossy ground planes [88, 89].

Figure 1.20: Current Sheet Antenna (CSA) arrays. (a) Single-polarized CSA array highlighting the interdigitated capacitors to achieve high inter-element coupling; (b) Dual-polarized CSA array aperture and feed network designed for operation over 2-18 GHz. (**SOURCE:** M. Jones and J. Rawnick, “A new approach to broadband array design using tightly coupled elements,” *Proc. IEEE Military Commun. Conf. (MILCOM)*, pp. 1-7, Jun. 2007.)

Figure 1.21: A Fragmented Aperture Array (FAA) with 33:1 bandwidth operation. This array uses a resistive/lossy screen between the planar radiating layer and the ground plane to mitigate scan dependent resonances. (**SOURCE:** P. Friederich et. al., “A new class of broadband planar apertures,” *Antenna Applications Symposium, Allerton Park, Monticello, IL*, pp. 561-587, Sep. 2001.)

Figure 1.22: Pictures of Long slot arrays. (a) A single-polarized long slot array designed for operation over 0.2 to 2 GHz while backed by a lossy ferrite loaded ground plane (**SOURCE:** J. J. Lee, S. Livingston, and D. Nagata, “A low profile 10:1 (200-2000 MHz) wide band long slot array,” *Proc. IEEE Antennas and Propagation Society Int. Symp.*, pp. 1-4, Jul. 2008.); (b) Dual-polarized coincident phase “Thumbtack” array operating from 0.5-2 GHz (**SOURCE:** S. Livingston and J. J. Lee, “A wide band low profile dual-pol “Thumbtack” array,” *Proc. IEEE Phased Array Tech. Int. Symp.*, pp. 447-483, Oct. 2010).

Dual-polarized TCDAs have also been proposed in which the radiating dipoles are printed on planar layers and utilize vertically-integrated interlocking feed boards consisting of 2-way Wilkinson power dividers and Marchand baluns that are soldered to a dipole layer [90–92]. Implementations of such arrays are shown in Figs. 1.23(a) and (b). The array of Fig. 1.23(a) achieves 6:1 VSWR<3:1 out to 60° scans with high efficiency and 82% aperture sampling [91], while in 1.23(b) a 13.1:1 bandwidth is demonstrated with VSWR<3 out to 45°, 88% aperture sampling and a profile $H = 1.79\lambda_g/2$, and 40-50% efficiency across the majority of the band (partially due to resistive losses from a resistive card placed between the radiating layers and ground plane) [92]. Like earlier TCDA implementations in the prior section, these arrays offer UWB performance but generally at the expense of increased complexity, deeper profiles, lower efficiency, and potential power handling issues from feed board chip resistors.

Figure 1.23: Depiction of vertically-integrated dipole arrays with integrated baluns. (a) A 6:1 dual-polarized TCDA-IB (**SOURCE:** M. Novak and J. Volakis, “Ultrawideband antennas for multiband satellite communications at UHF- K_u frequencies,” *IEEE Trans. Antennas Propag.*, vol. 63, no. 4, pp. 1334-1341, Apr. 2015); (b) Top and bottom views of a lossy 13.1:1 dual-polarized TCDA-IB (**SOURCE:** D. Papantonis and J. Volakis, “Dual Polarized Tightly Coupled Array with Substrate Loading,” *IEEE Antennas Wireless Propag. Lett.*, vol. 15, pp. 325-328, 2016).

Although a few of the discussed arrays can exhibit UWB performance with $>8:1$ BWRs, they also require the use of additional lossy devices and components. Such devices include lossy wideband external baluns and lossy screens and R-cards [88, 93] that are respectively used to ensure balanced feed lines and to suppress resonances from the ground plane. As a result, these factors can decrease the radiation efficiency (typically between 2-3 dB of losses) and array power handling. Without such components, the array bandwidths become compromised due to unbalanced feed line currents and ground plane resonances, and can be reduced to 4:1 for one instance [94]. Additionally, as mentioned earlier, although each of the addressed array layers are fully printable and planar, external vertically-integrated or three-dimensional feeding support mechanisms are typically required to maintain their wideband performance. Aside from external baluns and matching circuits, “cable organizers” [83] are sometimes utilized to shield the vertical field lines to prevent scan-induced resonances [95] as shown in Fig. 1.24. These three-dimensional components complicate manufacturing as they must be mechanically and electrically connected to the ground plane,

Figure 1.24: Cross-sectional view of a typical quasi-planar UWB array (Section 1.4.2.2) compared with that of planar UWB arrays (Section 1.4.2.3).

and it is difficult to connect the vertical cables drawn through the organizers to the dipoles on the planar layer. All of the factors discussed within this section lead to architectures that could impair low-cost manufacturing or in some cases radiation efficiency over the operating band.

1.4.2.3 Planar UWB Arrays

To circumvent fabrication difficulties, lack of aperture modularity, and performance-limiting feeding methods that plague quasi-planar UWB arrays, the Planar Ultrawideband Modular Antenna (PUMA) array [96] was developed. The PUMA array is fully manufactured with planar etched circuits and plated vias without the need of external baluns and cable organizers to enable a low-cost multilayer PCB fabrication process. This simplified construction was based on an unbalanced (single-ended) feeding scheme of the balanced dipole radiators that used shorting vias at the fed-dipole arm to mitigate common-modes that would otherwise develop within the operating band. The array can be constructed modularly and retains a low-profile with a total depth of about $\lambda_H/2$. The first class of PUMA arrays, shown in Fig. 1.25(a), demonstrated low VSWR and good scan performance over a 3:1 bandwidth up to 21

(a)

(b)

Figure 1.25: Planar Ultrawideband Modular Antenna (PUMA) arrays. (a) Three assembled 8×8 PUMA modules that operate over 7-21 GHz; (b) A 1-5 GHz PUMA unit cell with a backplane matching network. (**SOURCE:** S. S. Holland, *Low-profile, modular, ultra-wideband phased arrays*. Ph.D. thesis, University of Massachusetts Amherst, 2011.)

GHz [23]. Despite exhibiting good performance, a hard bandwidth limit was imposed by the low-end loop modes spurring from the introduction of shorting vias [97]. To overcome this, a planar matching network was printed on the opposite side of the ground plane to yield a 5:1 bandwidth [97] like that shown in Fig. 1.25(b). Although the bandwidth was enhanced, the highest frequency was limited to 5 GHz due to stringent design dimensions restricting manufacture with standard PCB technologies and the aperture sampling was 85%.

In the last half-decade, efforts were undertaken to enhance PUMA array bandwidth up to 6:1 without the use of external components and also improve their manufacturing repeatability [25, 98–103]. The culmination of this work has currently led to a low-cost, low-risk, efficient 6:1 dual-polarized UWB-ESA. A 3×3 module topological depiction of this array is illustrated in 1.26(a) and a fabricated 8×8 dual-polarized design is shown in 1.26(b). A critical enabling innovation has hinged upon the reconfiguration of the shorting via into a capacitively loaded via to mitigate a broadside common mode without inhibiting low-frequency performance [98]. Despite

Figure 1.26: 6:1 dual-polarized PUMA arrays. (a) Top-view of a 3×3 module depiction without dielectric cover (**SOURCE:** M. Lee, “6:1 PUMA arrays: Design and finite array effects,” University of Massachusetts Amherst, *M.S. Thesis*, 2016); (b) An 8×8 array operating over 3.5-21 GHz (**SOURCE:** R. Kindt et al., “A 6:1 bandwidth PUMA array at 7mm scale,” presented at IEEE Phased Array Tech. Int. Symp. 2016).

these research initiatives, state-of-the-art PUMA array bandwidths have resulted in a maximum of 6:1 bandwidth and, like its fellow low-profile arrays, can achieve a low CPR (< -15 dB out to $\theta = 45^\circ$) but with a modest $\text{VSWR} < 2$ bandwidth at broadside scan and ultimately cannot yet compete with Vivaldi array bandwidths and matching levels.

The Folded Notch array offers 10:1 bandwidth with good scanning impedance performance ($\text{VSWR} < 3.5$ out to $\theta = 60^\circ$) in infinite array simulations [104, 105]. Its design concept revolves around stacking thin, alternating, electrically disconnected planar metal plates upon compact, laminated foam or dielectric layers as shown in Fig. 1.27(a) to effectively embody a ‘folded notch’ (flattened flared notch approximation) impedance transformer. Consequently, the number of layers becomes proportional to the bandwidth of the structure and can result in a large amount of layers for

Figure 1.27: Isometric and cross-sectional side views of a dual-polarized Folded Notch array. (SOURCE: M. Cooley et al., “Planar-fed folded and stepped notch arrays: Novel wideband array technologies for multi-function AESAs,” presented at IEEE Phased Array Tech. Int. Symp. 2016.)

higher BWRs. The cross-polarization performance has yet to be addressed in the non-principal scanning planes, and measured results have yet to be made available.

1.5 Summary of Previous Work

As elaborated in prior sections, UWB-ESAs are generally limited in terms of either bandwidth or cross-polarization performance. A chart comparing instantaneous bandwidth and D-plane $\theta_0 = 45^\circ$ CPR at the highest frequency of some of the previously discussed arrays is shown in Fig. 1.28. It should be noted that this chart is only including high-efficiency (low-loss) arrays with broadside VSWR < 2 and little-to-no external components (relaxing the VSWR requirements, introducing lossy materials, and/or utilizing external hardware such as baluns/hybrids will broadband array performance at the expense of deteriorated efficiency and/or increased cost/complexity). The top-right corner indicates an ‘ideal’ goal of a decade or more bandwidth design with good polarization performance. From this chart, two conclusions can be drawn:

Figure 1.28: Summary of bandwidth and cross-polarization ratio (CPR) trends of some state-of-the-art low-loss high efficiency UWB-ESAs with VSWR<2 broadside and little-to-no external components [17, 19, 55, 56, 89, 91, 103].

- (1) All low-profile arrays solve cross-polarization issues, but limit their bandwidth operation to about 6:1 broadside matching i.e. VSWR<2 and low losses (efficiencies > 85%)
- (2) High-profile arrays such as Vivaldi arrays exhibit superior bandwidth for the same VSWR and efficiency as stated above, but their polarization is rather poor at scan angles wider than $\theta = 15^\circ$

Typical UWB-ESA design over the past decades has realized these two trends and focused upon two different research directions (denoted as 1 and 2 in Fig. 1.28) to achieve the ideal UWB-ESA; those directions are:

- (1) Focus efforts upon maximizing bandwidth while keeping the $\lambda_H/2$ profile in low-profile arrays. However, it has been shown repeatedly that low-profile arrays are fundamentally limited in the maximum bandwidth they can achieve due to the small volume they occupy (currently $\approx 6:1$ in state-of-the-art designs) [106–108],

and efforts to increase the bandwidth settle for relaxing VSWR requirements and/or introducing lossy materials to reduce the array Q factor, altogether leading to decreased radiation efficiency and increased system noise figure.

- (2) By far the less common research is that of addressing polarization issues in high-profile arrays without destroying their unrivaled bandwidths and matching levels. This has been mainly pursued by using appropriate calibration and look-up-table (LUT) based array excitation re-weighted feeding schemes [69], but these types of remedies are particularly prone to calibration errors, time-consuming, inherently narrowband, applicable only to dual-polarized configurations, deteriorate side lobe cross-polarization, and offer limited correction that often does not meet system needs. It is highlighted that this is a general “software-based” approach in the sense that any array, low or high profile, can be calibrated to improve its polarization performance.

1.6 Contributions

This work takes a radically different approach in design methodology to develop a UWB-ESA with both strong bandwidth and polarization performances. As recent design efforts have typically focused upon the low-profile array design corner and their bandwidth enhancement, this dissertation will do the opposite—it will instead embrace the superior impedance bandwidth of Vivaldi arrays and address their poor polarization issues with an elegant “hardware”-based solution. Specifically, the contributions of this work can be summarized as:

- (1) **Unveiling the true cause of high cross-polarization in Vivaldi arrays through a simple theoretical model:** A common belief in the UWB-ESA community is that the profile of the array controls the polarization performance. In other words, the higher the profile, the worse the CPR will be. It will be

shown that this is a misconception and instead the actual controlling factor is the vertical-to-horizontal current length ratio within the Vivaldi array unit cell. This is a critical understanding as it *states that even high-profile arrays can maintain good polarization performance as long as this ratio is properly balanced.*

- (2) **Propose a re-configured Vivaldi array element, termed the Sliced Notch Antenna (SNA), that is theoretically proven to simultaneously have excellent impedance bandwidth, matching levels, and CPR:** Using the model and theory developed in (1), a topological modification of the Vivaldi element is proposed that elegantly solves the long-standing problem of poor polarization in Vivaldi arrays when scanned in the non-principal planes. The enabling technical innovation hinges upon the re-configuration of the high-profile continuous metal Vivaldi fin into a series of disconnected yet strongly capacitively coupled vertically stacked metal strips that follow the envelope of the original Vivaldi fin. Impedance and far-field theory are developed to prove wideband operation and good polarization properties.
- (3) **Design, manufacture, and test of a Sliced Notch Antenna (SNA) array, with 10:1 BWR, low CPR, excellent matching, and high efficiency:** To the best of our knowledge, the SNA array is the first UWB-ESA technology to achieve decade-order (10:1) bandwidth with low cross-polarization properties and high efficiency. Using the strategy proposed in (2), SNA arrays effectively control current contributions responsible for high cross-polarized radiation without significantly affecting the input impedance of the array. The SNA array concept can be implemented in a variety of manufacturing technologies and embodiments (PCB, all-metal, and hybrid builds) similar to Vivaldi arrays. A single-polarized PCB prototype of an SNA array is comparatively

evaluated through impedance and far-field measurements against its equivalent Vivaldi array counterpart to showcase performance differences. Dual-polarized SNA design simulations and measurements across various embodiments will also be shown to indicate excellent behavior with some offering structural conveniences such as modular/solderless assembly. Ultimately, the SNA array will fill the high bandwidth and low cross-polarization design void in the top-right corner of Fig. 1.29(a) by retaining (or even improving) the wideband wide-scan (out to 60°) performance of the Vivaldi array and fixing the polarization performance, as shown in Figs. 1.29(b) and (c), respectively.

1.7 Dissertation Overview

Chapter 2 establishes a theoretical foundation for impedance and polarization performances in Vivaldi arrays. A simple single-mode equivalent circuit model at broadside scan will be developed as a foundation for the proposed design work in later chapters. A simple model will also be developed that characterizes the far-field performance through a simple vectorial superposition of the vertical and horizontal currents within the array unit cells. This method will shed light into the poor cross-polarization behavior in Vivaldi arrays as a function of their radiating currents. Critically, it will be shown that even a high-profile Vivaldi element can produce low cross-polarization as long as the vertical-to-horizontal current-length ratio within its element is controlled appropriately to produce a pattern with broader co-polarized beamwidth. This developed theory provides the groundwork for the design work in later chapters.

Chapter 3 will utilize the concepts and insights of the preceding chapter to introduce a new class of Vivaldi arrays capable of decade-order (10:1) bandwidths and low cross-polarization. These arrays, termed Sliced Notch Antenna (SNA) arrays, effectively control current contributions responsible for high cross-polarized radiation,

Figure 1.29: Performance comparisons for state-of-the-art UWB-ESAs and the proposed SNA array. (a) Qualitative bandwidth vs. cross-polarization trade-offs; (b) Instantaneous bandwidth comparison for maximum active VSWR out to $\theta_0 = 60^\circ$ scans between a state-of-the-art Vivaldi array and the proposed SNA array; (c) Cross-polarization ratio (CPR) comparison between the same arrays.

without significantly affecting the input impedance of the array. The driving design concept behind the development of SNA arrays will be presented that stems from a realization in Chapter 2. The chapter aims to familiarize the reader with the SNA array topology and also provide examples of both single-polarized and dual-polarized designs in different embodiments alongside their Vivaldi array counterparts in addition to proof-of-concept simulations.

Once the reader is familiarized with the proposed SNA array, Chapter 4 will develop the theoretical framework and operating principles for impedance and polarization performance. Equivalent circuit models will be developed and validated based upon that of the infinite Vivaldi array in Chapter 2 to characterize the impedance performance. The improved polarization performance will be established through extending the method in Chapter 2 to incorporate beneficial effects due to the sliced topology. After grasping the theoretical foundation, design guidelines will be presented to understand the basis for the nature of past and future SNA array designs.

Chapter 5 will present the design, fabrication, and measurements of an exemplary single-polarized 19×19 SNA array and its equivalent Vivaldi array counterpart. Each array operates over 1.2-12 GHz (10:1 bandwidth) and incorporates the same feeding, profile, and mechanical complexity, but the SNA array employs the modified sliced fin topology discussed in depth previously Chapter 3 to enable improved polarization behavior. Impedance and far-field performance of each array prototype in 1×19 , 5×19 , and 19×19 configurations are evaluated in infinite array and finite array simulations and compared with measurements to show excellent agreement. Results suggest the SNA array offers the same 10:1 bandwidth as its Vivaldi array counterpart yet offers >15 dB improvement in polarization performance on average and >30 dB in worst Vivaldi scan angles.

Chapter 6 concludes the dissertation with an outlook on future work.

CHAPTER 2

VIVALDI ARRAYS

The proposed UWB-ESA, the Sliced Notch Antenna (SNA) array, aims to overcome the common misconception that a high-profile array will have poor cross-polarization performance when scanning off-broadside in the non-principal planes. This ideology largely stems from the hindered polarization characteristics exhibited by high-profile Vivaldi array elements. This chapter will enlighten the reasoning behind this misconception and reveal the true nature of cross-polarization deterioration in Vivaldi arrays. Additionally, it will provide a synopsis of Vivaldi array impedance design behavior while developing approximate circuit models to be used as baseline references for the SNA array impedance theory in Chapter 4.

2.1 Introduction

As elucidated to in Chapter 1, wideband arrays of Vivaldi antenna elements remain highly popular due to their excellent wideband wide-scan impedance performance, making the realization of instantaneous bandwidths in excess of three octaves ($>8:1$) fairly straightforward to the experienced designer. One relates the array element depths (or profile), H , to the impedance bandwidth ratio by [24]

$$\frac{f_{high}}{f_{low}} \approx \frac{2H}{D} \quad (2.1)$$

where D is the unit-cell dimension (typically half-wavelength at the highest operation frequency). Despite their good impedance behavior and well-known design guidelines,

Figure 2.1: Bandwidth and polarization ratio trade-offs for varying element aspect ratios (H/D) of well-optimized single-polarized PCB-based Vivaldi infinite arrays.

Vivaldi arrays exhibit poor polarization characteristics when scanned off-broadside in non-principal planes [24, 66–69].

To demonstrate the severity of the polarization problem in wideband Vivaldi arrays and to shed light into the associated design compromise, a plot of the instantaneous bandwidth and worst cross-polarization ratio (recall (1.12); $\text{CPR} = |\mathbf{E}_{\text{cx}}|^2 / |\mathbf{E}_{\text{co}}|^2$) over the $\theta_0 \leq 45^\circ$ scan volume at the highest operation frequency versus element aspect ratio (depth-to-periodicity, H/D) for various well-optimized single-polarized PCB-based Vivaldi infinite arrays is presented in Fig. 2.1. A clear trade-off between bandwidth and CPR may be observed; deeper designs yield higher bandwidth, but at the expense of CPR, e.g. a 10:1 design has CPR near 10 dB (positive dB values indicate situations where the cross-polarized fields are larger than the co-polarized). From observation, the profile of the array is indeed linked to cross-polarization deterioration.

This troublesome trend severely hinders the radiation properties of Vivaldi arrays and has led researchers to investigate the low-profile topologies and calibration procedures discussed in Chapter 1. So far no low-profile technology has yet to demonstrate the quality of high-efficiency (low-loss) wideband wide-scan impedance matching of Vivaldi arrays. Additionally, polarization calibration procedures are routinely used to correct polarization using external re-weighted feeding schemes to implement some degree of correction in Vivaldi arrays [69], but these types of remedies are restricted to dual-polarized arrangements, narrowband waveforms, correction only narrowly around the main beam, raise side lobe cross-polarization levels, and are prone to error from system noise and tolerancing issues. Despite known weaknesses, Vivaldi arrays still continue to see widespread use and directly fixing their polarization drawbacks could provide measurable benefits for legacy and future UWB-ESA systems utilizing such arrays; the proposed SNA array introduced in Chapter 3 pursues this route.

Before presenting the SNA array, this chapter will provide an analysis of Vivaldi array impedance and polarization operating principles. An approximate single-mode circuit model for infinite Vivaldi arrays at broadside scan will be presented and validated to provide a review of Vivaldi array impedance behavior fundamentals and, more importantly, also establish familiarized references for SNA array impedance theory in Chapter 4. A simple theoretical model will also be developed that shows high cross-polarization in Vivaldi arrays is due to a relationship between vertical and horizontal currents propagating along the flare region, and not merely the high profile of the element. From the theoretical modeling in this chapter, it will be shown that even a high-profile element can produce low cross-polarization as long as its vertical and horizontal currents are controlled appropriately to produce an element pattern with broader beamwidth. This understanding plays a critical role in the development and operation of the SNA arrays throughout the rest of this dissertation.

2.2 Impedance Analysis

This section will develop an approximate single mode equivalent circuit model for an infinite Vivaldi array element at broadside scan. Although the findings and results may come as a review to some readers, the proposed SNA arrays descend from Vivaldi arrays and it thus becomes a natural precursor step to validate a Vivaldi array model before establishing an SNA element approximate circuit. Analysis of the model against full-wave simulations will follow to establish trust in the method.

Each element will be composed of stepped characteristic impedance slotline transmission lines backed by a slotline cavity; together they form a stepped matching transformer that matches a low feed impedance (100Ω) to a high free-space broadside scan resistance (377Ω). The characteristic impedance of each slotline section is numerically calculated from an infinite array analysis tool (ANSYS® HFSS® [109]). Various element cases will be analyzed that incorporate both a different number of stepped transmission lines and element thicknesses to provide rudimentary design insight. Once validated, these circuit models will serve as baseline references for the proposed SNA array in Chapter 4.

2.2.1 Approximate Circuit Model

An approximate single mode equivalent circuit model will be developed for the single-polarized infinite two-dimensional Vivaldi array over a ground plane of Fig. 2.2 at broadside scan¹ for varying element thicknesses t . As seen in Chapter 1, modern state-of-the-art Vivaldi arrays range from embodiments utilizing thin PCB trace metallizations and thick stock-metal structures; element thickness plays a significant

¹The model can be extended to include off-broadside scanning by expressing the periodic structure through a plane wave expansion with a finite number of propagation waves and infinite number of evanescent waves (altogether forming a spectral series of Floquet modes) [110]. Array elements with non-planar structures (i.e. Vivaldi elements) lead to active array impedance expressions involving fairly obscure nested integrals [110, 111]. The scope of this work instead employs a single-mode circuit model to approximate active element impedance at broadside scan; broadside scan behavior is known to serve as a good benchmark indication of the overall scan impedance performance.

Figure 2.2: An infinite two-dimensional single-polarized Vivaldi array.

role in the impedance tuning of the array element. The unit cell dimensions are set to $D = D_E = D_H = \lambda_g/2 = \frac{c_0}{2f_g}$, where $\lambda_g/2$ is the half-wavelength corresponding to the grating lobe frequency, $f_g = 12$ GHz.

The progression to the single mode circuit model is shown in Figs. 2.3(a-c). The Vivaldi array element in Fig. 2.3(a) is chiefly characterized by its tapered slotline to achieve a smoother impedance transition to high impedance free-space from a low impedance feed presented at the delta-gap excitation location (red arrow) above the terminated slotline cavity. It is easier to approximate the continuous taper with a discrete number of steps following the original taper as in Fig. 2.3(b). The taper is decomposed into a total of $N+1$ of transmission lines with characteristics impedances Z_n (where $n = 0, 1, 2, \dots, N$) controlled by the the taper step slot widths ($2d_n$) and

Figure 2.3: Development of Vivaldi element approximated single-mode circuit model at broadside scan. (a) Continuous tapered slot; (b) Stepped approximation of tapered slot; (c) Approximate circuit model of (b).

respective electrical lengths $\beta_n \ell_n$ that are connected in series to form the circuit² in Fig. 2.3(c).

The Z_n of each step is numerically calculated through an infinite array slotline analysis. The electrical length of each step is chosen to be a fixed value of $\approx \lambda/8$ at mid-band (≈ 6 GHz) to minimize the analysis degrees of freedom and also provide realistic transformer lengths for wideband matching. The resistance R_A is the radiation resistance of free-space at broadside scan (377Ω). The feed impedance is set to $Z_{\text{feed}} = 100\Omega$; typically this value is around $75\text{--}100\Omega$ and is commonly transitioned

²The taper step parameters for the total amount of steps, $N + 1$, and the taper step slot width, $2d_n$, are chosen as they are convenient for the SNA impedance theory definitions as will become clear in Chapter 4.

to an unbalanced 50Ω line with a Knorr balun approach³ [63]. The slotline cavity is rectangular for simplicity and is represented by a short-circuited transmission line with parameters $Z_{cav}, \beta_{cav}\ell_{cav}$ that is terminated in an inductance L_{cav} . This inductance incorporates the reactance of the slotline cavity that has been the subject of several parametric studies [58, 112]; larger cavities result in improved low-frequency performance by providing increased antenna resistance and inductive reactance. This value was found to be negligible for designs with small operating bandwidth ($<3:1$) in which the cavity is small. Lastly, it was necessary to include a series inductance L_{feed} in the feeding to compensate for the partial inductance induced by the finite length of the lumped gap port used in full-wave simulations (as also noted in [113] for a dipole array analysis). The values for the inductances were found through an optimization process, but may be found through more complicated deembedded parasitic extraction methods [114]. The evaluation of this single mode circuit yields an approximation of the active impedance of an infinite array element at broadside scan.

2.2.2 Equivalent Circuit Analysis

ANSYS[®] Designer[®] [109] was used to evaluate the equivalent circuit in Fig. 2.3(c) by comparison with numerical results from ANSYS HFSS full-wave simulations of various single-polarized Vivaldi infinite arrays. No dielectric materials are included. Each Vivaldi array design was first optimized in the full-wave solver to produce a reasonable good broadside scan impedance. As previously stated, *the electrical length of each taper step ($\beta_n\ell_n$) is fixed so that a larger number of steps ($N + 1$) effectively means a deeper element profile H .*

³This circuit does not include a Knorr balun and rather focuses upon the element impedance behavior isolated from additional feeding transition techniques; such methods can of course be implemented to facilitate better matching.

Figure 2.4: Full-wave simulation models in ANSYS® HFSS® of doubly periodic, single polarized Vivaldi arrays with different numbers of taper step sections $N+1$ (not drawn to scale). (a) One step case; (b) Three step case; (c) Ten step case.

2.2.2.1 Number of Steps

The number of taper slot steps $N + 1$ in the stepped Vivaldi element is varied for three cases where $N + 1 = \{1, 3, 10\}$ as seen in the full-wave simulation models shown in Figs. 2.4(a-c). These elements have zero thickness. The values of N were chosen to provide a wide variety of impedance performance. A Floquet port is assigned to the top of each unit cell, and periodic boundary conditions (PBC) are placed on the side faces of the unit cell for infinite array conditions. A ground plane is assigned to the bottom of the unit cell in which the slotline cavity is terminated. A lumped port is excited directly above the slotline cavity. The values for the approximate circuit model parameters and physical dimensions of the full-wave simulations for the $N + 1 = \{1, 3, 10\}$ taper step models can be respectively found in Tables 2.1, 2.2, and 2.3.

The input impedance loss at the feed point for the $N + 1 = \{1, 3, 10\}$ taper section steps is respectively presented in Smith charts of Figs. 2.5(a-c) over a 0.5-13

Table 2.1: Dimensions of an infinite Vivaldi array element with one taper slot step c.f. Fig. 2.4(a) and its Circuit Approximation Parameters at broadside scan

Physical Dimensions	Circuit Approximation
$Z_0 = 100\Omega$	$Z_0 = 100\Omega$
$D_E, D_H = 12.5\text{mm}$	$R_A = 377\Omega$
$2d_{cav} = 0.5\text{mm}$	$Z_{cav} = 136\Omega$
$2d_0 = 0.5\text{mm}$	$Z_0 = 136\Omega$
$\ell_{cav} = 10\text{mm}$	$\beta_{cav}\ell_{cav} = 72^\circ$ at 6 GHz
$\ell_n = 7.5\text{mm}$	$\beta_n\ell_n = 54^\circ$ at 6 GHz
	$L_{feed} = 0.5nH$
	$L_{cav} = 0nH$

Table 2.2: Dimensions of an infinite Vivaldi array element with three taper slot steps c.f. Fig. 2.4(b) and its Circuit Approximation Parameters at broadside scan

Physical Dimensions	Circuit Approximation
$Z_0 = 100\Omega$	$Z_0 = 100\Omega$
$D_E, D_H = 12.5\text{mm}$	$R_A = 377\Omega$
$2d_{cav} = 3\text{mm}$	$Z_{cav} = 216\Omega$
$2d_0 = 0.5\text{mm}$	$Z_0 = 136\Omega$
$2d_1 = 1.5\text{mm}$	$Z_1 = 170\Omega$
$2d_2 = 3.5\text{mm}$	$Z_2 = 228\Omega$
$\ell_{cav} = 9\text{mm}$	$\beta_{cav}\ell_{cav} = 64.8^\circ$ at 6 GHz
$\ell_n = 7.5\text{mm}$	$\beta_n\ell_n = 54^\circ$ at 6 GHz
	$L_{feed} = 0.7 \text{ nH}$
	$L_{cav} = 0.62 \text{ nH}$

Table 2.3: Dimensions of an infinite Vivaldi array element with ten taper slot steps c.f. Fig. 2.4(c) and its Circuit Approximation Parameters at broadside scan

Physical Dimensions	Circuit Approximation
$Z_0 = 100\Omega$	$Z_0 = 100\Omega$
$D_E, D_H = 12.5\text{mm}$	$R_A = 377\Omega$
$2d_{cav} = 8\text{mm}$	$Z_{cav} = 321\Omega$
$2d_0 = 0.25\text{mm}$	$Z_0 = 119\Omega$
$2d_1 = 0.5\text{mm}$	$Z_1 = 136\Omega$
$2d_2 = 1\text{mm}$	$Z_2 = 157\Omega$
$2d_3 = 1.5\text{mm}$	$Z_3 = 170\Omega$
$2d_4 = 2\text{mm}$	$Z_4 = 192\Omega$
$2d_5 = 2.5\text{mm}$	$Z_5 = 201\Omega$
$2d_6 = 3\text{mm}$	$Z_6 = 211\Omega$
$2d_7 = 4.25\text{mm}$	$Z_7 = 244\Omega$
$2d_8 = 5.5\text{mm}$	$Z_8 = 268\Omega$
$2d_9 = 6.75\text{mm}$	$Z_9 = 297\Omega$
$\ell_{cav} = 8.25\text{mm}$	$\beta_{cav}\ell_{cav} = 59.4^\circ$ at 6 GHz
$\ell_n = 7.5\text{mm}$	$\beta_n\ell_n = 54^\circ$ at 6 GHz
	$L_{feed} = 0.5 \text{ nH}$
	$L_{cav} = 3.5 \text{ nH}$

GHz range. The solid red lines represent results from the approximate circuit model whereas the blue dashed lines represent the full-wave simulation from ANSYS HFSS. Good correlation is observed between simulations for all cases, but there are some differences due to the circuit model not incorporating full-wave effects. As the total number of steps is increased, the bandwidth and impedance matching is improved because the element depth H increases; this is expected as more sections in a multi-section matching transformer can provide a smoother impedance match between two unequal loads and facilitate better low-frequency matching [28]. This can perhaps more clearly be seen in VSWR format in Fig. 2.6, where a ten-step Vivaldi array achieves approximately a 12:1 bandwidth with $\text{VSWR} < 2.1$ while a three-step Vivaldi array reduces to a 4.8:1 bandwidth with $\text{VSWR} < 2$. Respectively, these two bandwidths approximately indicate the upper bandwidth bounds achievable by modern low-loss Vivaldi and low-profile UWB-ESAs. Essentially, deeper elements are critical for low-frequency performance as they are capable of more gracefully supporting circulating current loops between adjacent array elements when operating at longer wavelengths i.e. lower frequencies. This bandwidth-enhancing mechanic is unique to Vivaldi-type arrays and allows physical element profile to be traded for bandwidth.

2.2.2.2 Element Thickness

The circuit approximation is now evaluated for different element thicknesses, t , when the number of tapered slot steps is kept to $N + 1 = 10$. The full-wave simulation models of interest are shown in Figs. 2.7(a-c) for cases where $t/D_H = \{0, 0.165, 0.5\}$; the thickness is expressed as a ratio to the fixed unit cell dimension D_H where, for example, infinitely thin PEC and thicker metal spanning half the unit cell correspond respectively to $t/D_H = \{0, 0.5\}$. These values are typical for actual Vivaldi designs ranging from thin metallizations in PCB to thick metallizations in stock-metal designs. The circuit parameters and physical dimensions for the infinitely thin sheet

Figure 2.5: Broadside scan return loss ($Z_0 = 100\Omega$) as predicted by the proposed equivalent circuit and full-wave simulations for $N + 1 = \{1, 3, 10\}$ taper step infinite Vivaldi arrays. (a) One step case; (b) Three step case; (c) Ten step case.

Figure 2.6: Comparison of equivalent circuit model prediction with full-wave simulations for the active VSWR ($Z_0 = 100\Omega$) of three single-polarized infinite Vivaldi arrays with $N + 1 = \{1, 3, 10\}$ tapered line steps. (a) One step case; (b) Three step case; (c) Ten step case.

model are the same as before in Table 2.3, and those for the $t/D_H = \{0.165, 0.5\}$ cases can be found in Appendix Tables B.1 and B.2. It is noteworthy from these tables and observations from Figs. 2.7(a-c) that thicker elements require wider slotline gaps to maintain well-matched impedance performance (relative to thinner elements). This can be reasoned by considering the characteristic impedance of a lossless transmission line, $Z_0 = \sqrt{L/C}$, where L and C are inductances and capacitance per-unit-length; thicker slotlines generate increased C in the taper slot gap which is compensated by increasing the slot width size to realize the original Z_0 .

The real/imaginary input impedance vs. frequency plots for these cases are charted in Figs. 2.8(a-c). Results generally show good agreement overall, with the infinitely thin sheet ($t/D_H = 0$) providing the best correlation and increasing thickness full-wave simulations deviating more from the circuit approximation. This is reasonable as thicker elements diverge more from an ideal slotline as the metalliza-

Figure 2.7: Full-wave simulation models in ANSYS® HFSS® of three single-polarized infinite all-metal Vivaldi arrays with ten taper slot steps and different thicknesses. (a) $t/D_H=0$; (b) $t/D_H=0.165$; (c) $t/D_H=0.5$.

(a)

(b)

(c)

Figure 2.8: Comparison of equivalent circuit model prediction with full-wave simulations for the input impedance (Z_{in}) of three single-polarized infinite Vivaldi arrays with ten taper slot steps and varying thicknesses. (a) $t/D_H = 0$ (infinitely thin conductor); (b) $t/D_H = 0.165$; (c) $t/D_H = 0.5$.

tion becomes large with respect to the guide wavelength and different current modes are supported within the full-wave model. It can also be seen that for the thickest case of Fig. 2.8(c), the imaginary component becomes entirely inductive over the operating band. This is due to thicker elements requiring wider slot widths to realize the same characteristic impedances of thinner elements, and in turn needing elongated lumped ports that produce larger self-inductances. A wave port implementation could be employed to eliminate this effect. Nonetheless, the proposed Vivaldi infinite array approximate circuit and full-wave simulations suggest good agreement with performance trends and will provide suitable reference baselines for the SNA array in Chapter 4.

2.3 Vivaldi Element Far-Field Analysis

This section will analyze the far-field performance of a Vivaldi antenna element. The focus of this analysis is to use a basic approach by approximating the currents on the Vivaldi fin in an infinite array environment and employing the free-space Green's function to calculate the far-field polarization performance. The goals are to (i) demonstrate a reasonable approximation of the cross-polarization ratio (CPR) for an infinite Vivaldi array and (ii) express the CPR as a simple function of controlling factors—namely to be shown as scan angle, frequency, and an important relationship between element currents.

2.3.1 Analysis Set-up and Assumptions

The theoretical work to be discussed makes several assumptions about the element analysis. Firstly, we consider an element within an infinite single-polarized array that is perfectly matched at all frequencies and scan angles and phased to radiate in the upper half-space towards (θ_0, ϕ_0) as depicted in the coordinate system of Fig. 2.9(a). The element unit cell dimensions, D_E and D_H , are assumed to be equal to the half-

Figure 2.9: Configuration of the Vivaldi antenna element far-field analysis. (a) Array element coordinate system; (b) Flowing currents along the Vivaldi array.

wavelength grating lobe spacing so only the fundamental Floquet (0,0) mode exists in the upper half-space. The structure is restricted to a smooth Vivaldi exponential taper of a certain taper rate and only forward travelling waves on the tapered structure are assumed to propagate. The metallized element is also shown to be embedded in a dielectric of permittivity ϵ_r within the tapered slot area (not a slab across the cell) for completeness, but only free-space scenarios and propagation will be considered. Furthermore, each element structure is infinitely thin, oriented in the $\hat{e}_H$ -plane, and is excited with a delta-gap excitation that enforces perfectly differential currents on both fin edges of the tapered slot; these currents form some aperture distribution dictated by the physical parameters of the Vivaldi element. Different from an isolated Vivaldi antenna element in which edge diffraction effects become significant [115], the infinite array environment engenders an array of periodic looping currents to manifest like in Fig. 2.9(b). Lastly, the effect of mutual coupling between neighboring elements on the element aperture distribution is neglected.

The analysis is limited in the sense that the polarization characteristics of an embedded element will inevitably be affected by additional current mode excitations, impedance mismatch, and mutual coupling in an array environment. Also, the model does not incorporate realistic feeding effects or element thickness that could otherwise contribute to cross-polarized components. However, results will still be shown to demonstrate reasonable agreement for the interested case.

The electric field is excited parallel to the plane of the slot ($\hat{e}_E$ -plane) to generate linear polarization. Ludwig's 3rd definition of polarization [31] is assumed as it is a convenient decomposition of linear-polarized fields; the resulting far-zone co-polarized and cross-polarized fields are then expressed as

$$\mathbf{E}_{\text{co}}(\theta, \phi) = \mathbf{E}_\theta(\theta, \phi) \sin \phi + \mathbf{E}_\phi(\theta, \phi) \cos \phi \quad (2.2)$$

$$\mathbf{E}_{\text{cx}}(\theta, \phi) = \mathbf{E}_\theta(\theta, \phi) \cos \phi - \mathbf{E}_\phi(\theta, \phi) \sin \phi \quad (2.3)$$

and the cross-polarization ratio (CPR) is recalled to be

$$\text{CPR}(\theta, \phi) = \frac{|\mathbf{E}_{\text{cx}}(\theta, \phi)|^2}{|\mathbf{E}_{\text{co}}(\theta, \phi)|^2} \quad (2.4)$$

where any frequency-dependence in terms is omitted for brevity. Approximate far-field equations will be expressed using elementary far-field radiation superposition of infinitesimal current sources following the aforementioned assumptions.

2.3.2 Unit Cell Model

The far-field components $\mathbf{E}_\theta$ and $\mathbf{E}_\phi$ are first found. The aperture distribution along the Vivaldi element fin is decomposed into a number of infinitesimal current radiators i.e. their spatial variation is constant. In this manner, the travelling wave current density along the element, $\mathbf{J}$, is restricted to the fin edges and broken into $N + 1$ sections with each comprised of a pair of infinitesimal horizontal ($\hat{e}_E$) and

Figure 2.10: Approximate current analysis of a Vivaldi array element. (a) Typical Vivaldi array currents; (b) Approximated currents; (c) Current map of a single section along the fin.

vertical ($\hat{e}_z$) current sources as shown in Fig. 2.10(b). The total number of sections is chosen so that the currents are small enough to remain valid under infinitesimal approximations. The currents within a section n can be modeled as shown in Fig. 2.10(c) with constant horizontal(vertical) current amplitudes, $I_{h(v)n}$, that are dictated by the travelling wave amplitude (i.e. changes upon the location of the section and the operating frequency). Assuming the horizontal (i.e. transverse or h) and vertical (i.e. longitudinal or v) current pairs are displaced from the $\hat{e}_z$ -axis by a distance $D_E/2 - d_n$ (where $0 \leq d_n \leq \frac{D_E}{2}$) and nearly co-located⁴ their two-element array factors (AF) are

$$AF_{h_n} = 2 \cos[kd_n \sin \theta \sin \phi] e^{jkz_n \cos \theta}$$

$$AF_{v_n} = 2j \sin[kd_n \sin \theta \sin \phi] e^{jkz_n \cos \theta}$$

where k is the wavenumber and z_n is the center of the section along the $\hat{e}_z$ -direction. Projecting these currents into the far-field, the resulting components are

⁴This implies that both vertical and horizontal current pairs are located at the same z_n and that the section is electrically small.

$$\mathbf{E}_{\theta_{h_n}}(\theta, \phi) \approx j \frac{\eta k}{4\pi r} e^{-jkr} I_{h_n} \ell_{h_n} \text{AF}_{h_n} \cos \theta \sin \phi \hat{\mathbf{e}}_\theta$$

$$\mathbf{E}_{\theta_{v_n}}(\theta, \phi) \approx -j \frac{\eta k}{4\pi r} e^{-jkr} I_{v_n} \ell_{v_n} \text{AF}_{v_n} \sin \theta \hat{\mathbf{e}}_\theta$$

$$\mathbf{E}_{\phi_{h_n}}(\theta, \phi) \approx j \frac{\eta k}{4\pi r} e^{-jkr} I_{h_n} \ell_{h_n} \text{AF}_{h_n} \cos \phi \hat{\mathbf{e}}_\phi$$

$$\mathbf{E}_{\phi_{v_n}}(\theta, \phi) = 0$$

Substituting these back into (2.4), the resulting CPR for a single Vivaldi element section is

$$\text{CPR}_n(\theta, \phi, \rho_{viv_n}) = \frac{|\cos \theta \sin \phi \cos \phi - \rho_{viv_n} \sin \theta \cos \phi - \cos \phi \sin \phi|^2}{|\cos \theta \sin^2 \phi - \rho_{viv_n} \sin \theta \sin \phi + \cos^2 \phi|^2} \quad (2.5)$$

where ρ_{viv_n} is the ratio of vertical-to-horizontal (v-to-h) components in a Vivaldi element section defined by

$$\rho_{viv_n} = \rho_{I_n} \rho_{\text{AF}_n}. \quad (2.6)$$

The ratios of vertical-to-horizontal (v-to-h) current-lengths and AFs, ρ_{I_n} and ρ_{AF_n} , are written as

$$\rho_{I_n} = \frac{I_{v_n} \ell_{v_n}}{I_{h_n} \ell_{h_n}} \quad (2.7)$$

$$\rho_{\text{AF}_n} = \frac{\text{AF}_{v_n}}{\text{AF}_{h_n}}. \quad (2.8)$$

The far-zone components of the entire Vivaldi element can be then found after a vectorial summation of the total contributions from each individual section n by

$$\mathbf{E}_{[\theta, \phi]}(\theta, \phi) = \sum_{n=1}^N \mathbf{E}_{[\theta, \phi]_n}(\theta, \phi, \rho_{viv_n}) e^{jkz_n \cos \theta} \quad (2.9)$$

where z_n is the center of a section n along the $\hat{\mathbf{e}}_z$ -direction. The CPR is then computed from (2.2-2.4).

In this manner, the CPR of a Vivaldi element is controlled in some way by the performance of its section CPR in (2.5), which is shown to depend upon:

1. A well-known dependency upon observation (or scan) angle
2. A well-known frequency (wavenumber) dependence
3. A somewhat obscure relationship between vertical and horizontal current components, ρ_{viv_n} .

An element with non-planar structure thus results in added dependencies from vertical currents that affect the polarization performance. It has been pointed out or implied over the past decade that the high-profile of Vivaldi elements supports harmful vertical currents which consequently have adverse effects on polarization in the non-principal planes [24, 66, 67, 69]; this observation motivated the design of several successful low-profile arrays discussed in Chapter 1 whose polarization properties ideally approximate that of a planar element i.e. they are dependent only upon scan angle and reach the aperture projection limit on the far-zone sphere via (2.5) when ρ_{viv_n} is zero and no vertical currents exist. The findings here support these implications, insights, and conclusions. However, the result here also indicates that there may perhaps be a more definitive culprit at hand, and rather a quantity involving the *ratio of vertical-to-horizontal current-lengths* plays a more complete role on polarization impacts. The next sections will shed light upon the usefulness of this result by first analyzing theoretical CPR trends within a single Vivaldi section n and then the CPR of the full Vivaldi element.

2.3.3 Single Section Results

The CPR from a section n along the Vivaldi element fin in (2.5) is seen to be controlled by observation (or scan) angle in addition to a vertical-to-horizontal current-length ratio, ρ_{viv_n} . For a section n , ρ_{viv_n} varies by the ratio of v-to-h magnitudes and lengths (ρ_{I_n}) and by the ratio of the AFs (ρ_{AF_n}). Consider the sources within a unit cell being maximally far apart separated by $d_n = D_E/2$ and $\lambda = \lambda_g$ ($kd_n = \pi/2$),

the CPR_n is swept as a function of ρ_{I_n} i.e. $(I_{v_n}\ell_{v_n})/(I_{h_n}\ell_{h_n})$ in Eq. 2.5 for various off-axis non-principal plane scan angles to result in Fig. 2.11(a). In general, the CPR_n begins to increase rapidly (the cross-polarization increases) when $\rho_{I_n} > 0.1$, and then stagnate at different rates depending upon the scan angle when $\rho_{I_n} > 2$. More specifically,

- For small ρ_{I_n} i.e. the vertical current portion is much smaller than the horizontal current portion, the CPR_n converges to that of a horizontal element (planar dipole) in an ‘ideal region’
- For large ρ_{I_n} i.e. the vertical current portion is much larger than the horizontal current portion, the CPR_n converges to that of a vertical element (monopole) in the ‘worst-case region’
- For ρ_{I_n} between these two regions, the CPR_n is in a ‘transition region’ controlled by the vertical-to-horizontal current-length ratio

It is reasonable to understand from these relations that a Vivaldi element has vertically-dominated current-length components due to the nature of its topology lending towards high element aspect ratios (H/D) to yield high bandwidths. *Therefore, it can be further speculated that a Vivaldi element yields poor polarization performance due to its innate geometry engendering high vertical-to-horizontal current-length ratios.*

It is also interesting to note that the D-plane ($\phi_0 = 45^\circ$) is not always guaranteed to be the worst scanning plane for a single-polarized array. In fact, for increasing ρ_{I_n} , the worst plane shifts towards the H-plane ($\phi_0 = 0^\circ$). Lastly, regardless of ρ_n , E-plane scans ($\phi_0 = 90^\circ$) and H-plane scans ($\phi_0 = 0^\circ$) theoretically achieve ideal CPR (0, or $-\infty$ dB), which is a well-known conclusion that a Vivaldi array suffers no polarization degeneracy while scanning in the principal planes. In practice, inevitable manufacturing tolerances and element imperfections will induce some cross-polarized fields in both principal planes, but non-principal scan planes still remain the most

heavily influenced due to the nature of the aperture projection on the far-zone sphere in (2.5).

The influence of the fin taper upon ρ_{AF_n} is also an important parameter in determining CPR. Assuming $\rho_{I_n} = 2$ for one such example, we can evaluate (2.5) from the minimum possible kd_n (i.e. 0) and maximum possible kd_n ($\frac{2\pi}{\lambda_g} \frac{D_E}{2} = \frac{\pi}{2}$). These results are charted for the same scan angles in Fig. 2.11(b). For larger kd_n (the current pairs are far from one another), the CPR_n always degrades, while for smaller kd_n (the current pairs are close to one another), the CPR_n always improves. This is expected as the vertical currents experience more cancellation for small kd_n , with the CPR_n results in that of an ideal horizontal planar element for perfect cancellation ($kd_n=0$) regardless of the value of ρ_{I_n} . It is also important to note that kd_n is frequency-dependent i.e. lower frequencies equate to smaller kd_n . To summarize,

- kd_n is small for close physical spacing and/or lower frequencies; this leads to improved CPR_n due to more vertical current cancellation
- kd_n is large for wide physical spacing and high frequencies; this leads to degraded CPR_n due to less vertical current cancellation

In the design of Vivaldi elements, ρ_{AF_n} is inherently set by the designer while optimizing the tapered fin region for impedance matching considerations. Lastly, although only an analysis with linear polarization was conducted, the procedure could be extended to dual-polarized arrays with circular polarization as well with similar conclusions as done for coincident and noncoincident phase dipole arrays in [116].

2.3.4 Example Element Results

An example Vivaldi element geometry generated in MATLAB is shown in the left side of Fig. 2.12(a) with $H/D_E=5.2$ where each blue circle represents a sample point of vertical and horizontal currents within a section n . Shown beside it, an

(a)

(b)

Figure 2.11: Resulting theoretical CPR_n (eq. 2.5) in dB for varying ρ_{viv_n} and its components. (a) CPR_n as a function of ρ_{I_n} (eq. 2.7) with $kd_n = \pi/2$ ($d_n = D_E/2$); (b) CPR_n as a function of ρ_{AF_n} (eq. 2.8) with $\rho_{I_n} = 2$.

infinite array full-wave HFSS model is identically created with a PEC sheet, delta-gap excitation, periodic boundary conditions (PBC), and a perfectly matched layer (PML). The CPR is computed for various scan angles across a 10:1 bandwidth (1.2-12 GHz) with full-wave simulations and theoretical calculations using eqs. (2.2-2.4) and (2.9). The resulting comparison between theory predictions and infinite array HFSS full-wave simulations is shown in Fig. 2.12(b) for $\phi_0 = 30^\circ$, $\theta_0 = \{45, 60\}^\circ$ scan angles i.e. the worst scanning plane for this array. Results suggest good agreement overall with some differences that are expected since the theoretical predictions do not take into several factors i.e. coupling between adjacent elements, radiation from any sources or current modes besides those along the fins, and an approximated aperture distribution. Despite this, the curves still follow the same trends with good correlation—the CPR generally degrades with increasing frequency and elevation scan angle. For the case of $\theta_0 = 60^\circ$ at high frequencies, the CPR actually begins to improve due to constructive/destructive interference between sections at this scan angle. In general, it is hard to fully predict the polarization behavior across the entire frequency band without using full-wave analytical models, but the same trends are largely followed and trust can be placed in the devised theoretical approach.

The significance of this result stems beyond achieving good accuracy with a full-wave solver. After all, it is elementary that the superposition of fields from several discrete current sources can yield a reasonable approximation of the aggregate far-zone characteristics in this case. *The intellectual merit resides in that the CPR of an element was written as a function of a ratio between the vertical and horizontal current-lengths within its element cell. It can then be plainly stated that the ratio of vertical-to-horizontal current-lengths has a large impact on the CPR within the scan volume, and not exclusively the profile of the element.*

Figure 2.12: Full Vivaldi element analysis comparisons between theoretical predictions and full-wave simulations. (a) Models with $H/D_E = 5.2$; (b) Polarization ratio vs. frequency for various scan angles.

2.4 Conclusions

This chapter provided theoretical analyses behind impedance and polarization behavior in single-polarized infinite Vivaldi arrays. A single-mode circuit approximation at broadside scan was carried through to review the impedance design behavior of Vivaldi arrays, and also develop baseline reference models for SNA array analysis in Chapter 4. A theoretical approach was also devised that analyzed the far-field characteristics of the Vivaldi element unit cell and related its polarization performance as a function of several factors: scan angle, frequency, and a relationship between the ratio of vertical-to-horizontal current-lengths. Critically, it was shown that high element vertical-to-horizontal current-length **ratios** (ρ_{viv_n}) are responsible for poor cross-polarization ratio (CPR) in Vivaldi arrays. It is fairly well-known in the wide-band array community that excessive vertical (longitudinal) currents stemming from the deep Vivaldi element profile yield high non-principal plane cross-polarization, but perhaps a more complete statement is that *the CPR is not strictly controlled by the element profile but rather the element vertical-to-horizontal current-length ratio.*

Plainly, the high-profiles utilized in Vivaldi elements for inherent wideband impedance matching become the enablers of problematic vertical-to-horizontal current ratios that deteriorate polarization performance, and, in essence, an element with a more well-balanced or near-zero ratio could conversely provide good polarization properties.

From these observations, it also becomes evident that reducing the profile of the element to minimize the vertical-to-horizontal current-length ratio is the most clear-cut solution to improve CPR. However, as elucidated to previously, low-profile arrays have yet to demonstrate the same quality of low-loss wide-scan wideband impedance matching as a Vivaldi array. This fact, in conjunction with the theory and conclusions developed within this chapter, will serve as insightful foundations for the development of the SNA arrays in the remaining chapters that can offer Vivaldi-like bandwidths with good CPR.

CHAPTER 3

SLICED NOTCH ANTENNA (SNA) ARRAYS

In Chapter 2, it was shown that the balance of vertical-to-horizontal current-length ratios within an array element is critical in controlling polarization performance. Traditional Vivaldi arrays exhibit poor polarization behavior in the non-principal planes due to their inherent high-profile design topologies engendering excessive vertical currents that progressively degrade this ratio for increasing element profiles (H). To overcome this issue, typical UWB-ESA design has focused upon low-profile array technologies as they intrinsically enable improved polarization characteristics due to their reduced profiles, but these arrays have yet to be able to demonstrate the efficient impedance performance achievable by Vivaldi arrays. Additionally, Vivaldi arrays continue to see widespread use despite known polarization weaknesses, and a direct fix of these issues could have significant benefits with minimal change and downtime for legacy and future UWB-ESA systems employing Vivaldi-type elements.

In light of the bandwidth and polarization trade-offs in modern UWB-ESAs, this chapter introduces the Sliced Notch Antenna (SNA) array. The SNA array incorporates an elegant solution that corrects the imbalanced current ratios in high-profile Vivaldi arrays by reconfiguring their deep, contiguous metal fins through the introduction of strategic discontinuities. In doing so, the SNA array favorably balances the vertical-to-horizontal current-length ratio within the element as a function of the modified slice implementation. With careful design, the modified SNA array topology is able to retain the impedance performance of a Vivaldi array while simultaneously providing good polarization performance.

This chapter will serve as a means to introduce and familiarize the reader with the proposed SNA array topology by presenting the driving design concept, example embodiments, and associated numerical results from full-wave simulations. For each design, the SNA array will be compared with its Vivaldi counterpart to more easily gauge differences in both structure and electrical performance. Chapter 4 will then present SNA array operation theory and guidelines that helped facilitate not only the construction of the SNA arrays from this chapter, but also the first fabricated SNA array proof-of-concept in Chapter 5.

3.1 Design Concept

The motivation behind the SNA array concept largely stems from a realization in the theory developed in Chapter 2. A comparison of D-plane $\theta = 45^\circ$ scan cross-polarization ratios (CPR) for a single section of vertical-to-horizontal current-length ratios ($I_{v_n}\ell_{v_n}/I_{h_n}\ell_{h_n}$) with current sources separated by $\lambda_g/2$ is shown in Fig. 3.1. It can be understood that for a typical high-profile Vivaldi array, the ℓ_{v_n} (vertical length of the current) term dominates due to the deep profiles and the polarization properties are consequently degraded. On the other hand, modern low-profile arrays solve this situation by intrinsically reducing the ℓ_{v_n} term which results in improved polarization properties.

Although low-profile arrays take a design route that solves polarization problems, the reduced profiles and smaller occupied volumes lead to significant challenges in achieving wideband wide-scan impedance bandwidths due to the difficulty in supporting large vertical (and looping) currents that are critical for low-frequency matching. *Thus, low-profile arrays essentially shift the design paradigm from that of a polarization problem to a bandwidth and impedance matching problem.* As a result, a vast amount of research efforts in the past decades have focused primarily around enhancing the bandwidth of a low-profile array as elucidated in Chapter 1. Despite several

Figure 3.1: D-plane $\theta = 45^\circ$ cross-polarization ratio (CPR) versus vertical-to-horizontal current-length ratios and modern UWB-ESA design placement. Large vertical current-lengths (ℓ_{v_n}) lead to poor CPR in high-profile Vivaldi arrays, while low-profile arrays have smaller ℓ_{v_n} to yield good CPR.

successful demonstrations of low-profile UWB-ESA technologies, none of these arrays have yet demonstrated the impedance prowess of a high-performance Vivaldi array without resulting in significant losses due to impedance mismatch, resistive loading, or external devices [93].

This work will take a different approach from this line-of-thought. A closer look at Fig. 3.1 reveals that there are other factors to consider that affect the CPR besides that of the vertical current length (ℓ_{v_n}), namely the relative magnitude of vertical and horizontal current-length i.e. $I_{v_n}\ell_{v_n}/I_{h_n}\ell_{h_n}$. With other properties to control the CPR, it seems plausible that if a systematic method of controlling the current-length ratio was devised then even a high-profile array could achieve good polarization performance. Retaining the high-profile would be desirable because it would enable wider bandwidths at improved matching levels. This leads to the ultimate research question of this dissertation: **Is there a systematic way to control the vertical-to-horizontal current-length ratio on the fin of a high-profile array?**

The SNA array affirmatively answers this question. In Fig. 3.2(a), a portion of a typical Vivaldi array is shown to contain vertically dominated currents leading to poor CPR as a consequence of its design supporting large looping currents for improved low-frequency matching. To balance the vertical and horizontal currents in a more favorable way, it would be desirable for the currents to take paths like those in Fig. 3.2(b). The multiple small current loops would theoretically provide a reduced vertical-to-horizontal current-length ratio, but how can this be practically implemented? The SNA array [117] answers this in Fig. 3.2(c) by strategically “slicing” the otherwise contiguous Vivaldi metal fin to enforce these well-balanced current paths that in turn result in better polarization performance while also retaining the high-profile necessary for excellent wideband matching. The remainder of this chapter will more deeply familiarize the reader with the SNA array topology while also initially validating performance with full-wave simulations.

Figure 3.2: Design methodology depictions behind the Sliced Notch Antenna (SNA) array concept. (a) Vertically dominated currents in Vivaldi arrays yield poor cross-polarization ratios (CPR); (b) Fictional well-balanced currents in Vivaldi arrays that would result in good CPR; (c) Well-balanced currents in the SNA array resulting in good CPR.

3.2 Design Preface

The primary purpose of the following sections are to familiarize the reader with the SNA array topology across various embodiments and its resulting performance benefits over Vivaldi arrays before proceeding to a deeper theoretical analysis in Chapter 4. For each design, topological and electrical behavior will be analyzed between a proposed SNA array and its equivalent Vivaldi array counterpart to highlight similarities and differences with brief sets of diagrammatic models and infinite array full-wave simulations. An ‘equivalent Vivaldi array counterpart’ (i.e. Vivaldi counterpart) means that, in reference to a specific SNA array, the elements of both arrays have the same:

1. Depth-to-periodicity ratio (H/D)
2. Structural embodiment basis (i.e. PCB, stock-metal, etc.)
3. Operational bandwidth
4. Operating high frequency

Essentially, the proposed SNA array and its Vivaldi counterparts are equivalent in terms of structure except for the modified fin region.

The following sections will first introduce two different types of SNA arrays through PCB-based embodiments. Following this, exemplary single-polarized and dual-polarized PCB-based designs of different types will be presented along with full-wave results to initially grasp electrical behavior. Predicted impedance and polarization performances of the arrays are simulated in HFSS’s full-wave infinite array environment and with in-house finite array Domain Decomposition Finite Element Method (DD-FEM) codes [118]. The proposed designs serve as examples that can be extended to other embodiments; early glimpses of hybrid stock-metal builds will be shown as some such examples.

Figure 3.3: Single-polarized PCB arrays. (a) Typical Vivaldi array; (b) SNA array (Type 1); (c) SNA array (Type 2).

3.3 Types of SNA Arrays

Single-polarized arrangements of a traditional PCB-based Vivaldi array and two exemplary PCB-based SNA array designs are shown in Fig. 3.3(a-c). All of the arrays share several identical design features: stripline-based PCB layouts with metallic vias, Knorr balun feed transitions, 50Ω end-launch SMA connectorization points, dielectric PTFE substrates, and flare and cavity parameters. The chief difference between the SNA and Vivaldi arrays can be seen in the upper fin regions, where the SNA array fin is composed of a series of vertically stacked metal strips that progressively outline some taper above a contiguous base. In addition to strong mutual coupling to neighboring elements, the gaps between the strips are strongly capacitively coupled and are used to facilitate better cross-polarization while not impeding matching.

Two ‘types’ of SNA arrays, Type 1 and Type 2, are shown respectively in Figs. 3.3(b-c). The Type 2 SNA array differentiates itself from Type 1 by vertically splitting the SNA fin down the center so that adjacent elements are electrically disconnected

Figure 3.4: Single-polarized SNA antenna element parameters. (a) Exploded view of a single-polarized SNA element; (b) Definition of parameters for a 10:1 SNA design.

in that region. Exploded views of each type are depicted in Fig. 3.4(a-b). Notably, Type 1 employs coupled metallic patches that increase the complexity of the design by requiring additional printed layers for support; these are used to help achieve ideal aperture oversampling by tuning out a problematic high-frequency resonance of whose origin will become clear in Chapter 4 (without the patches, $>90\%$ aperture sampling is still possible). Some benefits of Type 2 include:

- No sharp high-frequency cut-offs
- Two-layer PCB fabrication (down from four) and no parasitic patches
- Simpler extrapolation to most dual-polarized arrangements
- No gap-induced resonances [73] in the SNA fin area
- No plated vias necessary in the SNA fin area

However, as compared to the Type 1 SNA array, the Type 2 array makes electrical performance compromises and has:

- Slightly worse impedance matching levels (but still on par with the Vivaldi array)
- A scan-induced high-frequency surface wave is manifested at wide scan angles due to the periodic structure

The different types of single-polarized SNA arrays can be translated to other embodiments and dual-polarized arrangements; likewise some dual-polarization examples to be presented can be deconstructed to single-polarized versions. The choice of which type to use is highly dependent upon which is more convenient in a desired method of fabrication—the electrical performances will be on par with one another with optimized tuning. The following designs to be presented serve as examples. The Type 2 SNA array was chosen to be fabricated and will be included in Chapter 5 measurement analysis, while the Type 1 will be analyzed numerically in the following section. Lastly, any SNA array, regardless of type, will be referred to plainly as an ‘SNA array’ for generality.

3.4 A Single-Polarized SNA Array

The definition of parameters of a single-polarized SNA array and exemplary single-polarized 10:1 (1.2-12 GHz) 1×19 PCB-based Vivaldi and SNA arrays are presented in Figs. 3.5(a-c), respectively. The 1×19 linear cards are used to form a square 19×19 array. This array size was chosen as it is the largest array size that can fit comfortably on a standard PCB panel size ($12'' \times 18''$) for the designed operating band and it also has an odd number of elements to designate a true center element. Rogers 5880LZ ($\epsilon_r = 1.96$) is utilized for the substrates. Each array operates over 1.2-12 GHz (10:1) with the highest operating frequency equivalent to the half-wavelength grating lobe frequency ($f_H = f_g$) for ideal aperture sampling. It can be seen that the arrays practically share the same design parameters and critical dimensions. Metallic

Figure 3.5: Single-polarized SNA antenna element parameters. (a) Exploded view of a single-polarized SNA element; (b) Definition of parameters for a 10:1 SNA design.

blocks/spacers are inserted between adjacent rows to provide the appropriate unit cell dimensions, finite ground plane, and mechanical support (as can be seen in Chapter 5).

The primary differences are localized in the upper flared fin region, where the SNA, Figs. 3.5(a,c), incorporates several horizontal slices to embody a structure described previously in this chapter. Each slice is periodically spaced by 150 mils, and the sizes of the slices increase from 20 mils to 80 mils ($\approx \lambda_g/50 - \lambda_g/10$). The slice parameters (i.e. location, gap size, etc.) are chosen by design rule-of-thumbs to be discussed in Chapter 4 that essentially dictate the frequency-selectivity at each discontinuity junction by restricting or allowing portions of the vertical looping current to flow upwards. This is synonymous with the slices being thought of as cascaded low-pass filter sections as a function of element profile; each slice can be tuned to allow

the majority of low-frequency current to pass to form large (and looping) currents critical for low-frequency operation, while also restricting high-frequency current from forming unnecessarily large loops that deteriorate cross-polarization. The structure effectively reconfigures the contiguous Vivaldi fin into a series of vertically stacked metal strips that progressively taper to follow the envelope of the Vivaldi flare.

3.4.1 Active Impedance Parameters

Infinite array VSWR (referenced to $Z_o = 50\Omega$) versus frequency is plotted in Figs. 3.6(a) and (b) for E- and H-plane scans, respectively, at broadside, $\theta = 45^\circ$, and $\theta = 60^\circ$. The results suggest over 10:1 bandwidth for both arrays (bounded respectively by the vertical low-/high-frequency markers f_L and f_g , where f_g is equivalent to the half-wavelength grating-lobe frequency) with broadside VSWR <2 and H-plane $\theta=45^\circ$ scan VSWR <2.1 . Notably, the E-plane scan VSWR for the proposed SNA array is better than the Vivaldi, with significant advantages at $\theta_0 = 60^\circ$. The high-band broadside VSWR of the Vivaldi array is better than that of the proposed SNA array due to the effects of slicing as to be further discussed in Chapter 4, although both VSWRs remain below 2 across the entire band. The SNA arrays also have a noticeably sharper cut-off at the high-band edge for each scan due to a resonant length associated with slicing; a major design objective is to eliminate this resonance with careful slice implementation. Altogether, a well-tuned SNA array offers VSWR <2.25 out $\theta=60^\circ$ while the Vivaldi array offers VSWR <3.2 . A deeper investigation into finite array size effects on the active VSWR will be conducted in Chapter 5.

3.4.2 Far-Field Performance

Although exhibiting good impedance performance, the primary focus of the SNA array remains to address the degraded polarization performance of Vivaldi arrays in the off-broadside non-principal planes and maintain low CPR. Ludwig's 3rd definition of cross-polarization [31] is employed as it naturally suits single-polarized arrays.

Figure 3.6: Infinite array active VSWR simulations of SNA and Vivaldi array elements from Figs. 3.5(b-c). (a) E-plane scans; (b) H-plane scans.

Figure 3.7: Infinite array D-plane $\theta = 45^\circ$ scanning co-/cross-polarization levels versus frequency for a unit cell with 1W incident power for the proposed SNA array and its Vivaldi array counterpart.

The infinite array polarization characteristics of both arrays versus frequency when scanned at $\theta = 45^\circ$ in the D-plane ($\phi = 45^\circ$) with 1W of incident power at the port is shown in Figs. 3.7. For the Vivaldi array case, the cross-polarized radiated power increases with frequency, to the point that the cross polarization is higher than the desired co-polarization at the upper half of the frequency band. More specifically, the CPRs at the marked frequencies of $f_{LB} = 3$ GHz, $f_{MB} = 7.2$ GHz, and $f_{HB} = 11$ GHz are -8 dB, 0 dB, and +3.5 dB, respectively. The proposed SNA array, however, produces improved CPRs at the same frequencies of -13 dB, -12 dB, and -8 dB, respectively, due to the nature of its operating principles. Overall, the SNA array nets over 10 dB in CPR improvements in the $\theta = 45^\circ$ D-plane throughout the upper half of the band.

This improvement is further verified by the infinite array and finite 19×19 array embedded element CPR polar patterns throughout the upper half-space ($0 \leq \theta \leq 90$) at the high-band frequency ($f = f_{HB}$) for both arrays in Fig. 3.8. Here, more red levels indicate regions of higher cross-polarization (CPR), whereas blue are ‘cooler’ regions with good CPR. For both infinite and finite cases, the Vivaldi array (left column) suffers from extensive CPR deterioration within the marked $\theta \leq 45^\circ$ scan volume, and even more so at wider angles towards $\theta = 60^\circ$. On the other hand, the SNA array (right) greatly alleviates the issue in the $\theta \leq \{45^\circ, 60^\circ\}$ scan volumes, in addition to vastly improving some outside areas. Strong correlation exists between finite and infinite array results, with the only large exceptions being outside the $\theta \leq 60^\circ$ due to finite ground plane and finite array truncation effects. It is interesting to note that for the single-polarized Vivaldi array, the worst polarization is not at the D-plane, but rather around $\phi = 30^\circ$ for this case as also seen in Chapter 2 for high-profile single-polarized Vivaldi arrays. Preliminary full-wave investigations thus suggest that this single-polarized SNA array is capable of offering the same bandwidth as its Vivaldi

Figure 3.8: Infinite array and finite 19×19 array embedded element CPR patterns at the high-band frequency ($f = f_{HB} = 11$ GHz). More red levels indicate higher levels of cross-polarization. (a) Vivaldi array column; (b) Proposed SNA array column.

array counterpart but with significant polarization improvements in the non-principal scan planes.

3.5 A Dual-Polarized SNA Array Design

A dual-polarized PCB-based design is depicted in Fig. 3.9. The topology employs two interweaved orthogonally-polarized SNA cards (of Type 2) that are electrically disconnected to adjacent elements in the SNA array fin region. This is particularly convenient as the conventional required soldering process is not necessary within this region. Illustrations of a single-row assembly are shown in Fig. 3.10 that highlight a preferred assembly process by dividing the element into the top disconnected SNA array section and the bottom contiguous base. In this case, a tongue-and-groove

Figure 3.9: Depictions of a dual-polarized SNA array highlighting the electrically disconnected SNA fin region. (a) 4x4 isometric view; (b) Zoomed isometric view.

modular assembly is utilized by forming a ‘tongue’ at the base of the SNA piece that is inserted into a cut groove at the top of the bottom piece and held together by an adhesive. This exploratory type of SNA array implementation can altogether offer:

1. No adjacent element electrical connection in the SNA fin region
2. Easier access for any required soldering
3. Absence of vias in the SNA region
4. Modular ‘tongue-and-groove’ assembly option
5. Separate top/bottom pieces enabling hybrid fabrication choice options

The last point will be explored in upcoming sections incorporating hybrid build methods with PCB and stock-metal methods.

3.5.1 Active Impedance Parameters

Infinite array active VSWR for broadside and E-/H-plane scans out to $\theta_0 = 60^\circ$ for both the decade-bandwidth dual-polarized SNA array and its Vivaldi array counterpart are charted in Fig. 3.11. Both arrays exhibit good behavior across their band;

Figure 3.10: Drawings of a single row from a PCB-based dual-polarized SNA array. The fabrication can be split into two different sections that are assembled with a tongue-and-groove process and adhesive. (a) Pre-assembly parts; (b) Assembled row.

notably the in-band VSWR levels at broadside is <1.7 for the SNA array and <1.5 for the Vivaldi array. The SNA array scans better in the E-plane as seen repeatedly for all cases presented previously with $\text{VSWR} < 2.05$ as opposed to $\text{VSWR} < 2.4$ for the Vivaldi. The infinite array inter-element active port coupling ($|S_{12}|$) for D-plane $\theta = \{45, 60\}^\circ$ scans is plotted in Fig. 3.12. Again, both arrays exhibit very good and similar performance with maximum dB levels of -20/-17 and -12.5/-15 for Vivaldi/SNA arrays at $45^\circ/60^\circ$, respectively. The port coupling for broadside and principal plane scans are omitted but remain below 50 dB across the entire band for each array.

3.5.2 Far-Field Performance

The realized total gains of the dual-polarized SNA and its Vivaldi counterpart are charted in Fig. 3.13(a) for broadside and D-Plane $\theta = \{45, 60\}^\circ$ scans when one polarization is excited and the other is terminated. The standard reference directivity at the respective elevation scan angles is represented by the black-dashed lines. As both arrays exhibit minimal losses to input mismatch, port coupling, and do not utilize resistive materials, the realized gains track very closely to the aperture limit.

Figure 3.11: Infinite array VSWR simulations of the PCB-based dual-polarized SNA array and its Vivaldi counterpart. (a) Vivaldi array; (b) SNA array.

Figure 3.12: Infinite array inter-element port coupling ($|S_{12}|$) of the PCB-based dual-polarized SNA array and its Vivaldi counterpart for D-plane scans. (a) D-plane $\theta=45^\circ$; (b) D-plane $\theta=60^\circ$.

Figure 3.13: Infinite array realized total gain ($G_{co} + G_{cx}$) and CPR versus frequency for dual-polarized decade-bandwidth SNA and Vivaldi arrays at D-Plane and broadside scans. One polarization is excited, the other is terminated. (a) Realized total gain with reference directivities at respective elevation scan angles; (b) Cross-polarization ratio for D-plane $\theta_0 = \{45, 60\}^\circ$ scans.

The Vivaldi array exhibits maximum losses of about 0.7 dB around 3.5 GHz whereas the SNA array drops by only 0.25 dB around 9 GHz.

A different story is told upon analyzing polarization behaviors. The CPR vs. frequency is plotted in Fig. 3.13(b) for both D-Plane $\theta_0 = \{45, 60\}^\circ$ scans. Typical of Vivaldi arrays, the CPR degrades with increasing frequency and reaches as high as 8.75 dB at f_g for 45° case and near 10 dB at 10.5 GHz for 60° . On the contrary, the SNA array CPR stays relatively constant throughout the entire operating band at values of -15 dB and -9.3 dB for $\theta = \{45, 60\}^\circ$, respectively, and offers improvements >20 dB over its Vivaldi counterpart. These infinite array results can be further used as approximate CPR benchmarks for the main beam polarization in finite array scans like for the 19×19 array D-plane $\theta_0 = \{45, 60\}^\circ$ scans in Figs. 3.14(a-b). The co-polarized gain of the SNA array tracks the standard reference directivity until it begins to become underestimated in the low-frequency bands due to the electrically

(a)

(b)

Figure 3.14: 19×19 array D-plane scan gains for proposed SNA array and its Vivaldi counterpart. One polarization is excited, the other is terminated. (a) $\theta_0 = 45^\circ$ scan; (b) $\theta_0 = 60^\circ$.

small array size¹. While SNA array CPR levels stay better than -15 dB for both cases throughout the band, the Vivaldi array CPR significantly degrades with frequency. In fact, for the case of the Vivaldi array D-plane $\theta_0 = 60^\circ$ scan at 10 GHz, the CPR is +30 dB while the corresponding SNA array CPR is -27.5 dB, suggesting a 57.5 dB improvement. Lastly, it has also been generally observed that finite array CPR is better than infinite array CPR in this instance and other works [25, 119].

To provide a perspective of the full scanning volume, co-polarized and cross-polarized scan patterns at the highest frequency (f_g) are provided in Fig. 3.15. It is clear from observation that the Vivaldi suffers from high cross-polarization degradation in the non-principal scanning planes and most heavily in the D-plane. This is different than single-polarized arrays where the worst scanning plane changes depending upon the element depth and unit cell asymmetries (a comparison of single-/dual-polarized Vivaldi array polarization patterns for various element depths and frequencies can be found in Appendix C). On the other hand, the SNA array offers significant improvement throughout the entire $\theta \leq 60^\circ$ scan cone. Outside of this region, dips in the SNA co-polarized pattern can be seen in the E-plane and H-plane that correspond to surface wave excitations as a consequence of the periodic structure embodying the Type 2 SNA fin. With proper design of the slices, these effects are pushed outside of the scanning volume.

These patterns are further extended to CPR and intrinsic cross-polarization ratio (IXR) in Fig. 3.16. The CPR pattern describes the polarization performance assuming that the antenna and wave polarization are known a priori to both correspond to Ludwig's 3rd definition (as has been done throughout this dissertation). It

¹By [26], "for planar apertures in which $A \gg \lambda^2$, the value of the standard directivity is $4\pi A/\lambda^2$, with λ the wavelength and with radiation confined to a half space." For finite wideband array prototypes like the proposed 19×19 SNA arrays, $A \approx \lambda^2$ in the lower frequency band for broadside and $\theta_0 = 60^\circ$ scans at 1.26 GHz and 1.79 GHz, respectively, and the directivity begins to become underestimated before those points. However, the comparative performance between the Vivaldi and SNA array is generally sufficient for our purposes.

Figure 3.15: Co-polarized and cross-polarized patterns for dual-polarized PCB Vivaldi arrays (left) and split SNA array (right) at the highest frequency (f_g).

can be seen in these charts that the CPR accentuates the poor non-principal plane cross-polarization of the Vivaldi array, in which it becomes highly cross-polarized up towards ≈ 10 dB in the D-plane around $\theta = 55^\circ$, and compliments the low cross-polarization of the SNA array to remain < -10 dB out to $\theta = 60^\circ$. For example, comparing the use of each array in a typical on-the-move communication systems link in a free-space medium, the system using a Vivaldi array could provide catastrophic polarization mismatch losses in the non-principal scan planes without correction.

For the case of polarimeters or radars in which the incoming wave polarization is not known a priori and is of arbitrary polarization basis (i.e. natural source of emission, radar scatterers, etc.), IXR is a more suitable figure of merit [30] (c.f. IXR in Section 1.2). To note, lower values of IXR indicate less total relative error in the post-calibration measurement. In the IXR charts (note a scale change as $-\infty < \text{IXR [dB]} < 0$), the SNA array IXR is nearly entirely the same as its CPR, albeit differences outside the area of interest near grazing angles. However, it is visible that the Vivaldi array is even worse in the majority of the scan volume except for the principal planes and, interestingly enough, the D-plane around $\theta = 55^\circ$ which had exhibited high CPR. Upon closer analysis of the definition of IXR, this makes sense as IXR is a measure of the calibratability of the system and is closely related to the orthogonality between dual-polarized channels. Thus, IXR is better for when one polarization (i.e. channel) dominates, and the worst for CPR [dB] = 0 when it would be most vulnerable due to small noise or system fluctuations causing large calibration errors. Ultimately, the Vivaldi array exhibits problematic CPR and IXR that could severely inhibit its use in both communication and sensing systems, while the SNA array exhibits better than 10 dB of both FoMs out to $\theta = 60^\circ$ and provides significant improvements.

Figure 3.16: CPR and IXR for traditional dual-polarized PCB Vivaldi arrays (left) and split SNA array (right) at the highest frequency (f_g).

3.5.2.1 Calibration Study

This section will follow more of an investigate approach regarding the calibration of the SNA and Vivaldi arrays, and will serve to motivate future research work. The previous section touched upon the IXR of each array, which is a cross-polarization FoM for radar and sensing systems c.f. Eq. (1.22), and concluded that the Vivaldi array suffers significantly more than that of the SNA array. Now those same arrays will be calibrated at a frequency near that of f_g at $f = 10.5$ GHz (about 12% less), and assume a calibration (or correction) of the polarization at that frequency and scan angle is desired to achieve approximately -30 dB CPR according to Ludwig's 3rd definition of cross-polarization (i.e. one port is excited, the other is terminated; resulting in linear polarization).

This example correction at a D-plane $\theta_0 = 45^\circ$ infinite array scan is presented in Fig. 3.17 in a CPR vs. frequency chart. Before correction (i.e. uncorrected), one port is originally excited with 1W and the other is terminated. The CPR follows familiar trends for both arrays—the SNA array CPR is already fairly low and constant across the 10:1 operating band (non-dotted lines), while the Vivaldi array CPR steadily deteriorates with increasing frequency. Upon applying the necessary beamforming weights for calibration across both orthogonal ports for each array, both arrays can be seen to have achieved the calibration goal of -30 dB at the target frequency. This is expected as these are ideal beamforming weights correcting a noiseless system model², but are there other expenses as a consequence of the good CPR performance at the target correction? Some observations are outlined here:

- *Frequency dependence:* It is clearly seen in Fig. 3.17 that the CPR of the SNA array away from the correction frequency is nearly independent of calibration,

²In reality, system noise or measurement errors will influence the calibration process; these errors are magnified by poor IXR via (1.23) and would result in worse CPR post-calibration for the Vivaldi array had noise been present in this case.

Figure 3.17: Effects of calibration on cross-polarization ratio versus frequency in dual-polarized Vivaldi and SNA arrays for infinite array D-Plane $\theta_0 = 45^\circ$ scans.

while the calibration actually makes the Vivaldi array worse than it was before calibration below mid-band. This is because, originally, the polarization properties of the SNA array at a certain scan angle are nearly constant with scan angle, while those of the Vivaldi array are not.

- *Side lobe level:* In Chapter 1, it was stated that correctional procedures in a Vivaldi array corrected around the main beam, but the cross-polarization of the side lobe levels highly deteriorated and essentially traded places with the uncorrected co-polarization side lobes (c.f. Fig. A.3 in Appendix A). It is known that the co-polarized side lobes of a UWB-ESA with low CPR are much less sensitive to calibration.

This section merely scratched the surface of an open-ended research opportunity regarding side effects and trade-offs in calibrating dual-polarized phased arrays. What truly comes at the expense of calibration? Some of those questions were addressed

and theorized here, and these inquiries will be re-postulated in the Chapter 6 future work.

3.5.3 A Glimpse into Dual-Polarized Hybrid Stock-Metal Builds

To provide some examples of SNA array design flexibility, hybrid stock-metal SNA array builds are shown alongside a traditional 10:1 all-metal Vivaldi array (e.g. [56]) counterpart in Figs. 3.18(a) and (b). Each array consists of 8×8 dual-polarized elements on slant-linear lattices. In regards to (a), uniformly-spaced foam layers are partitioned in the original Vivaldi flare to form an SNA structure (i.e. a Type 1 structure) [120], while in (b) vertically-integrated PCB cards are inserted into the stock-metal base (i.e. forming a Type 2 structure) [121]. For all cases, array assemblies require no soldering and utilize press-fit RF components and fastening machine screws; the stock-metal is machined with wire electrical discharge machining (EDM) for precision cuts. Both topologies originated as baseline designs from the University of Massachusetts Amherst alongside guidance and support from the Naval Research Laboratory (NRL) Radar Division; NRL facilitated final design work and proof-of-concept demonstrations that determined the SNA arrays offer excellent impedance performance that is near-equivalent or better than their Vivaldi counterparts while delivering significant polarization improvements [120, 121].

3.6 Conclusions

This chapter served to introduce and familiarize the reader with the topology and performance benefits of a new class of UWB-ESAs with low cross-polarization, termed the Sliced Notch Antenna (SNA) array. Infinite and 19×19 finite array simulations of a both single-/dual-polarized SNA arrays and their Vivaldi counterparts were presented to highlight performance similarities and differences; both arrays showed good VSWR but the proposed SNA array achieved >25 dB and >15 dB CPR improve-

(a)

(b)

Figure 3.18: A traditional 10:1 all-metal Vivaldi array from the Naval Research Laboratory Radar Division (left, e.g. [56]) compared with 10:1 SNA array builds (right); both arrays are dual-polarized with 8×8 elements on slant-linear lattices. (a) A hybrid stock-metal and foam SNA build (**SOURCE:** R. Kindt, J. Logan, and M. Vouvakis, “Dual-polarized Sliced Notch Array,” presented at Antenna Applications Symposium, Allerton Park, Monticello, IL, 2016); (b) A hybrid stock-metal and vertically-integrated PCB SNA array build (**SOURCE:** R. Kindt, J. Logan, R. Mital, and M. Vouvakis, “Dual-polarized Sliced Notch Array - Ultra-wideband flares with exceptional polarization control,” presented at IEEE Phased Array Tech. Int. Symp. 2016). The SNA array design concepts and initial simulations were developed at the University of Massachusetts Amherst alongside guidance and support from the Naval Research Laboratory (NRL) Radar Division; NRL facilitated final design work, development, and measurement of the fabricated prototypes.

ment for non-principal plane scans in the $\theta \leq \{45, 60\}^\circ$ cones, respectively. Topological variations in Type 1 and Type 2 SNA arrays were reviewed; both array types offer similar performance and their different structural utility was addressed in single and dual-polarized arrangements.

Chapter 4 will present the theoretical groundwork and enabling operating principles of SNA arrays. After a grasp on the physical nature is developed, design guidelines will be laid out that served as a basis for the models in this chapter. Chapter 5 will later present the measurements and analysis of the first proof-of-concept SNA array build.

CHAPTER 4

THEORY & DESIGN OF LOW CROSS-POLARIZATION VIVALDI ARRAYS

Chapter 3 introduced the topology and various geometrical embodiments of a new class of low cross-polarization UWB-ESAs, termed the Sliced Notch Antenna (SNA) array. Through full-wave simulations, it was shown that the SNA array can indeed provide the same or even better wideband and wide-scan impedance matching than a Vivaldi array while retaining low cross-polarization.

This chapter will shed light into the SNA array operation principles and design guidelines. The presented theory directly builds upon that presented in Chapter 2. It will begin by analyzing the far-field characteristics of SNA arrays through providing simple modifications of the Vivaldi array far-field formulae and demonstrating the improved SNA polarization performance. As elucidated to in Chapter 3, this hinges upon the control of the element vertical-to-horizontal current ratio through implementing strategic capacitively coupled discontinuities across the upper body of the Vivaldi fin. Next, the effect of this control strategy upon the SNA array impedance performance will be investigated by establishing approximate circuit models that are extensions to those developed for Vivaldi arrays in Chapter 2. Lastly, some design guidelines and studies will be provided that further scrutinize the effect of the control strategy implementation. The operation theory and guidelines yield simplistic design rule-of-thumbs that laid the foundation for the SNA arrays introduced in Chapter 3 in addition to the fabricated array in Chapter 5.

4.1 SNA Far-Field Theory

This section will develop an approximate theory for the far-field radiation of single-polarized SNA infinite arrays, and will theoretically prove the enhanced CPR performance.

4.1.1 Analysis Set-up and Assumptions

The theoretical work makes the same assumptions as those of the Vivaldi array in Chapter 2, and will be re-summarized here. The element is assumed to be perfectly matched at all frequency and scan angles, has no coupling to neighboring elements within the array, and no grating lobes exist in the upper half-space ($D_E = D_H = \lambda_g/2$); this leads to the element pattern variation with (θ, ϕ) as the sole contributing far-field factor. The element is also shown to be embedded in a dielectric of permittivity ϵ_r within the tapered slot surrounding space for completeness, but only free-space scenarios ($\epsilon_r = 1$) will be considered. The currents form some aperture distribution dictated by the physical parameters of the topology which are excited by a delta-gap; the distribution is perfectly balanced and symmetric about the feed location or along the center of the SNA fin (i.e. as if a virtual ground plane or electric wall was present) and is localized to the metallized edges of an infinitely thin sheet. Ludwig's 3rd definition of cross-polarization is used. Essentially, we are looking for the CPR contribution from currents radiating in a single element, in which the currents are produced from an infinite array approximation. Practically, this model is limited as there will be some coupling between neighboring elements in addition to impedance mismatch as a function of scan angle that alters the radiated far-field in some way i.e. ripple or even dips in the gain pattern if not designed and tuned properly.

4.1.2 Unit Cell Model

The SNA element is comprised of $N + 1$ segments where $n = 0$ is the base region and $n = 1 \cdots N$ is the sliced region shown in Fig. 4.1(a). The $n = 0$ region is effectively

a standard Vivaldi element from Chapter 2 that is now electrically disconnected (sliced) at some point along the fin to separate it from the upper sliced region; N represents the total amount of these slices along the original Vivaldi fin. Although all segments are totally disconnected, strong capacitive coupling occurs between the segments that is dictated by the geometrical parameters and topology of the slice, thus energy is coupled upwards along the element fin. Within each segment n , the currents can be decomposed into horizontal and vertical pairs along the fin edges to approximate the taper (like for the ‘sections’ in Chapter 2) as depicted in Fig. 4.1(b), while the currents within the sliced region now form another new horizontal current pair (to close a loop).

A model of the currents within one segment is shown in Fig. 4.1(c). The slicing introduces top and bottom slot currents, referred to respectively as $I_{slot_n}^{top}$ and $I_{slot_n}^{bottom}$. These currents are continuous at all corner locations and are assumed to be at a maximum at the center; this assumption is valid since symmetry conditions enforce a virtual short circuit (i.e. electric virtual ground) at the slot center. The slot electrical current length will also always be $\leq \lambda/2$ within the operating band as it cannot be larger than the unit cell half-spacing in free-space conditions. These currents form their own slot array factor AF_{slot_n} controlled by the gap size, g_n , and their own slot space factors SF_{slot_n} controlled by the electrical length of the slot, $\beta\ell_{slot_n}$. SFs of the top and bottom slot currents are assumed to be equivalent as the difference in physical lengths is negligible for our purposes. Critically, the slicing induces a coupling interaction between segments dictated by the physical slot parameters that can be characterized by a complex coupling coefficient, c_n , where $0 \leq |c_n| \leq 1$ (0 represents no coupling to the top segment, 1 represents all energy coupled upwards i.e. appears electrically connected). In this manner, the slice junction behaves similar to that of a coupled line coupler. The effect of c_n can be expressed in the slot AF as a non-uniform amplitude two-element array with elements separated by distance g_n

in the $\hat{\mathbf{e}}_{\mathbf{z}}$ -direction and a phase difference $\angle c_n$ i.e.

$$\text{AF}_{\text{slot}_n}(g_n, c_n) = e^{jkg_n \cos \theta/2} - |c_n|e^{-jkg_n \cos \theta/2 + \angle c_n}. \quad (4.1)$$

In almost all cases, $\angle c_n$ is negligible as the slice gap is almost always electrically small ($< \lambda_g/20$) for impedance matching considerations as will be discussed later. The implied reference location within a segment is at the center of the slot at some position z_n in the $\hat{\mathbf{e}}_{\mathbf{z}}$ -direction. Furthermore, the horizontal and vertical currents along the fin, denoted respectively by I_{h_n} and I_{v_1n}, I_{v_2n} , form their own respective AF_{h_n} and $\text{AF}_{v_1n}, \text{AF}_{v_2n}$ like in the Chapter 2 analysis.

The coupling parameter c_n is key as it dictates how much energy is longitudinally coupled upwards along the fin as a result of the slice implementation. For instance, if all gaps are small to the point where the segments seem electrically connected, this results in all $c_n \approx 1$ and the model regresses to that of a standard Vivaldi with polarization drawbacks. On the other hand if the very bottom gap g_1 is very large then $c_1 \approx 0$; this is effectively a shorter Vivaldi that is known from Chapter 2 to have lower bandwidth or impedance matching. However, if one could systematically devise a frequency-selective mean of controlling this coupling coefficient c_n , such that (i) low-frequency currents couple upwards and form large loops for improving low-frequency operation, i.e. lowering lower band edge and matching, and (ii) high-frequency currents are blocked or regulated from flowing upwards thus limiting radiation from longitudinal currents, then it would seem plausible that a wideband array with good polarization could be achieved. This method can be accomplished through the tuning of c_n within each segment as will be shown by the end of this section.

The primary goal now is to develop the far-field radiation of the SNA array as modified by the effect of slices. The CPR of a segment n of the SNA array can be written as

Figure 4.1: Modeling in the current flows in an SNA array element. (a) Typical SNA array currents; (b) Approximated currents; (c) Current model of a single section.

$$\text{CPR}_n(\theta, \phi, \rho_{\text{sna}_n}) = \frac{|\cos \theta \sin \phi \cos \phi - \rho_{\text{sna}_n} \sin \theta \cos \phi - \cos \phi \sin \phi|^2}{|\cos \theta \sin^2 \phi - \rho_{\text{sna}_n} \sin \theta \sin \phi + \cos^2 \phi|^2}, \quad (4.2)$$

where ρ_{sna_n} is now the vertical-to-horizontal current ratio of the an SNA segment n . Equation (4.2) has the same form as the Vivaldi in Chapter 2 c.f. (2.5) but replaces ρ_{viv_n} with ρ_{sna_n} . For the case of Fig. 4.1(c), ρ_{sna_n} can be approximated by

$$\rho_{\text{sna}_n}(c_n, g_n) = \frac{c_n [I_{v_1n} \ell_{v_1n} \cos(kt_1) \text{AF}_{v_1n} + I_{v_2n} \ell_{v_2n} \cos(kt_3) \text{AF}_{v_2n}]}{c_n I_{h_n} \ell_{h_n} \cos(kt_2) \text{AF}_{h_n} + \text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)} \quad (4.3)$$

in which the segment is assumed to be electrically small such that each current source shares the same z_n location. However, the currents here still include a travelling wave approximation for completeness in which $\cos(kt_{\{1,2,3\}})$ factors incorporate phase shifts induced by the travelling wave nature and $t_{\{1,2,3\}}$ denote progressive locations. In order to produce good CPR, it is desirable for $\rho_{\text{sna}_n} = 0$; this implies that radiation from vertical currents does not exist. Upon observation of (4.3), it is seen that this condition can be met if (i) the vertical length components are heavily outweighed by the horizontal length components i.e. it is low-profile, (ii) the vertical current AFs cancel their contribution, or (iii) $|c_n| = 0$. Critically, this means that for an SNA array element, the vertical-to-horizontal current ratio is not only controlled by the typical vertical and horizontal current-length components, *but also horizontal contributions from the slot in the denominator and coupling effects.*

The effect of slicing can perhaps more clearly be seen by finding ρ_{sna_n} with that of an electrically small Vivaldi array section from Chapter 2 that is modified to incorporate a slice inserted at the bottom of the current model like in Fig. 4.1(c) without the second vertical current segment I_{v_2n} . Recall ρ_{viv_n} to be

$$\rho_{\text{viv}_n} = \frac{I_{v_n} \ell_{v_n} \text{AF}_{v_n}}{I_{h_n} \ell_{h_n} \text{AF}_{h_n}}. \quad (4.4)$$

If a slice is added to the Vivaldi array section (i.e. the slot horizontal current pairs are inserted), the section behaves like that of an SNA array; the current ratio can be written as

$$\rho_{\text{sna}_n} = \frac{c_n I_{v_n} \ell_{v_n} \text{AF}_{v_n}}{c_n I_{h_n} \ell_{h_n} \text{AF}_{h_n} + \text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)} \quad (4.5)$$

where the $\text{SF}_{\text{slot}_n} \text{AF}_{\text{slot}_n}$ factor incorporates the slot horizontal components and c_n is the coupling coefficient affecting the amount of energy coupled to currents above the slot. Equation (4.5) may be further simplified for convenience by

$$\begin{aligned} \rho_{\text{sna}_n} &= \frac{I_{v_n} \ell_{v_n} \text{AF}_{v_n}}{I_{h_n} \ell_{h_n} \text{AF}_{h_n} + \frac{1}{c_n} \text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)} \\ \rho_{\text{sna}_n} &= \rho_{\text{viv}_n} \cdot \frac{I_{h_n} \ell_{h_n} \text{AF}_{h_n}}{I_{h_n} \ell_{h_n} \text{AF}_{h_n} + \frac{1}{c_n} \text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)} \\ |\rho_{\text{sna}_n}| &= |\rho_{\text{viv}_n}| \cdot \left| \frac{I_{h_n} \ell_{h_n} \text{AF}_{h_n}}{I_{h_n} \ell_{h_n} \text{AF}_{h_n} + \frac{1}{c_n} \text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)} \right| \\ |\rho_{\text{sna}_n}| &= |\rho_{\text{viv}_n}| \cdot \left| \frac{1}{1 + \frac{\text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)}{c_n I_{h_n} \ell_{h_n} \text{AF}_{h_n}}} \right| \end{aligned} \quad (4.6)$$

which indicates the original Vivaldi current ratio ρ_{viv_n} is multiplied by a new horizontal current term due to the slice. For cases in which $g_n \neq 0$ (i.e. cases in which the slot gap is not closed, the slot exists, and $|c_n| < 1$), it is seen by (4.6) that

$$\left| \frac{\text{SF}_{\text{slot}_n}(\beta \ell_{\text{slot}_n}) \text{AF}_{\text{slot}_n}(g_n, c_n)}{c_n I_{h_n} \ell_{h_n} \text{AF}_{h_n}} \right| > 0 \quad \text{for } g_n \neq 0 \quad (4.7)$$

since the slot will introduce some horizontal contributions and be non-zero. In the case that the slot gap is closed i.e. a non-sliced Vivaldi element, $g_n = 0$ and $c_n = 1$ and (4.7) goes to zero; resulting in $|\rho_{\text{sna}_n}| = |\rho_{\text{viv}_n}|$ by (4.6) and the section reduces that of a standard Vivaldi element. The ideal case for polarization considerations is when

$c_n \rightarrow 0$ indicating no energy is coupled to vertical components i.e. only the bottom horizontal slot current radiates and $\rho_{\text{sna}_n} \rightarrow 0$ by (4.6). Thus combining (4.6) and the inequality of (4.7) leads to

$$|\rho_{\text{sna}_n}| < |\rho_{\text{viv}_n}|. \quad (4.8)$$

This indicates that the CPR_n c.f. (4.2) of an SNA segment n benefits from lower values of ρ_{sna_n} for non-principal plane scans due to slice insertion; both SNA and Vivaldi arrays theoretically have ideal CPR for principal plane scans. Aggregate SNA element far-zone components are found by the vectorial summation of contributions from all currents within SNA segments. Constructive and destructive interference between segments can also be taken advantage of to further improve the CPR to be better than that of the base Vivaldi i.e. $n=0$. The following subsection will analyze the effects of ρ_{sna_n} on CPR through systematically controlling the slot parameters.

4.1.2.1 Single Section Results

The CPR for a single SNA element segment n is evaluated at a D-plane scan angle $(\theta_0=45^\circ, \phi_0=45^\circ)$ as a function of g_n and c_n in Fig. 4.2 when the vertical and horizontal current lengths take realistic values. Here, the vertical current lengths are four times as large as that of the horizontal current, and the slot is twice as long as the vertical current lengths. The phase induced by the coupling coefficient, $\angle c_n$, is neglected due to the electrically small gap sizes and all wavelengths refer to the grating lobe wavelength, λ_g . The red curve indicates realistic values for $|c_n|$ as a function of g_n and ℓ_{slot_n} and is calculated at f_g using ideal coupled line coupler models in circuit design software (National Instruments AWR Microwave Office® [122] was used in this case). For reference, the CPR of a single Vivaldi section (no slot exists) is shown in the top left corner of the chart where $g_n \rightarrow \lambda/\infty$ and $|c_n| \rightarrow 1$. This yields a CPR of ≈ 0 dB, which is a reasonable value as seen in the findings of Chapter 2, c.f. Fig.

Figure 4.2: CPR of a single SNA element section as a function of c_n and g_n for a D-plane ($\phi_0 = 45^\circ$) $\theta_0 = 45^\circ$ scan. The red curve indicates realistic predicted values for c_n as a function of g_n . All λ values refer to the grating lobe wavelength, λ_g , and $d_n = \lambda/4$.

2.11(a), for cases where $\ell_{v_n} \gg \ell_{h_n}$ for D-plane scans. Conversely, the limiting case of $|c_n| \rightarrow 0$ for $g_n = \lambda/\infty$ (bottom left corner), no energy would be coupled upwards and the CPR results in that of an ideal horizontal (dipole) element in the D-plane (≈ -15 dB).

In reality, c_n is a function of g_n . In Fig. 4.2, the red curve represents the $c_n(g_n)$ red curve obtained through Microwave Office. In this range, the CPR steadily improves from -7.5 dB to -15 dB, and therefore validates what the theory in the previous section suggested—namely that an SNA array region n is capable of offering significant CPR improvements over Vivaldi array through control of the slot parameters.

4.1.2.2 Example Element Results

The full SNA element CPR will now be examined. A sheet SNA element embedded in air is constructed in MATLAB® and ANSYS HFSS® [109] (full-wave infinite array) of the forms shown in Fig. 4.3(a) with $L/D_E \approx 5.2$ ($\approx 10:1$ bandwidth). Several uniform thickness ($D_E/25$) and uniformly-spaced slices ($D_E/2$ or $\ell_{slice_n}/4$ for a total of $N=7$ slices) are introduced to create an SNA element. The geometry is discretized into horizontal and vertical segments (as discussed similarly for a Vivaldi element in Chapter 2) at each blue circle for a total 50 points, and their individual far-field contributions are vectorially summed to yield the far-field components and the resulting CPR. Values of c_n for SNA array CPR are chosen by sampling values of $|c_n|$ across the operating band for an ideal coupled line coupler model in Microwave Office. Note that this must be done for each slice and for several frequencies (different slot lengths in this instance, but gaps could also vary), and this could be implemented in a more practical way by calculating the coupling parameters simultaneously in MATLAB. Nonetheless, it is still a fairly tedious procedure and it is likely to be more convenient to follow design rule-of-thumbs that will be presented later.

Results are shown against infinite array predictions in Fig. 4.3(b) for the SNA element at different scan angles. Overall, strong correlation exists between full-wave and theoretical results for both arrays. The differences are most likely due to the full-wave infinite array incorporating adjacent element coupling effects and surface current radiation that the proposed theory does not account for. Additionally, the coupled line coupler model is only approximate and other factors such as fringing fields were not accounted for. However, the trends largely follow the same pattern and the SNA array results closely track the full-wave predictions (within 3 dB). *Results indicate that a high-profile SNA element is able to retain good polarization performance due to its more well-balanced current ratios along the metallic fin.* However, the introduction of

Figure 4.3: Single-polarized SNA element analysis comparisons between theoretical predictions and full-wave simulations. (a) Models with $H/D_E = 5.2$; (b) CPR vs. frequency for various scan angles.

slices along the Vivaldi fin are also significant topological changes from an impedance perspective, and their effects in that regime will be analyzed in the next section¹.

4.2 SNA Array Impedance Theory

SNA arrays modify the typical Vivaldi array topology to yield improved polarization characteristics by introducing discontinuities (referred to as slices) along the tapered fin that could alter the element input impedance within the operating band. This section will develop an approximate single-mode circuit model for SNA arrays at broadside scan that is based upon the one presented in Section 2.2.1 for Vivaldi arrays that analytically predicts the effect of slicing in input impedance. The goal of this circuit model is to provide a physical explanation to the complex effects taking

¹It should also be noted before proceeding that the slice gaps in this element model are small for reasons that will become more clear in the following section; generally this is critical to maintain *strong capacitive coupling* in order to preserve matching and achieve wider BWs. This is different from antennas in which there is weak coupling to parasitic directors that results in low fractional bandwidths e.g. typically around 2-3% for a Yagi-Uda, and rather the proposed method will be shown to utilize strong coupling (amongst various other design factors discussed in this dissertation) to achieve FBWs greater than 163% (i.e on the order of 10:1 BWR).

Figure 4.4: SNA array views. (a) Infinite two-dimensional single-polarized SNA array; (b) Side view of an SNA element.

place in SNA arrays, and to be used to draw basic design guidelines that can be used as a starting point for optimization.

4.2.1 Approximate Circuit Model Development

A thick all-metal single-polarized stepped notch element of N steps is modified by incorporating several slices as depicted in Figs. 4.4(a-b) for an infinite array and element, respectively. With respect to Fig. 4.4(b), the slices are placed in the center of each notch step with uniform gap thickness, but the number of exclusive slices, their gap thickness, and their location may take on an infinite plurality of design possibilities.

The SNA array impedance analysis will assume that the array is scanned at broad-side only, thus this model may be simplified by exploiting element symmetry about the $\hat{e}_z$ -axis as shown in Fig. 4.5(a) and (b). A virtual ground plane (electric wall) is introduced at the center of the flared slotline to reduce the model to that of a bisected version. A single slice in a certain taper slot step n is highlighted in Fig.

4.5(c) with three key design parameters: the distance from the virtual ground plane, d_n , the slice gap size, g_n , and the thickness of the element, t . Upon ignoring fringing effects in the element thickness, this slice closely resembles an E-plane T-Junction in a parallel-plate waveguide shown in Fig. 4.5(d). The approximate circuit model of the E-plane T-junction is shown in Fig. 4.5(e). This circuit is analagous to that provided in [123] albeit being translated from admittance to impedance parameters for analysis consistency and also allowing the characteristic impedances Z_{slice} and Z_n to differ. The parameters of the equivalent circuit for the E-plane T-Junction in a parallel plate waveguide are given by [123] and adapted to follow the SNA model parameters by

$$\frac{Z_{slice}}{X_C} = \frac{1}{n_{c_n}^2} \left(\frac{2d_n}{\lambda_o} \right) \left[\ln \left(1.43 \frac{d_n}{g_n} \right) + \frac{1}{2} \left(\frac{2d_n}{\lambda_o} \right)^2 \right] + \frac{\pi^2 d_n}{32 \lambda_g} \left(\frac{g_n}{d_n} \right)^2 J_0^2 \left(\frac{\pi g_n}{\lambda_o} \right) \quad (4.9)$$

$$\frac{Z_n}{X_L} = -\frac{\pi^2 d_n}{16 \lambda_o} \left(\frac{g_n}{d_n} \right)^2 J_0^2 \left(\frac{\pi g_n}{\lambda_o} \right) \quad (4.10)$$

$$n_{c_n} = \text{sinc}(\pi g_n / \lambda_o) \quad (4.11)$$

$$n_{turns_n}^2 = n_{c_n}^2 (g_n / d_n) \quad (4.12)$$

where J_0 is the Bessel function of zero order, λ_o refers to the free-space wavelength λ_o , and n_{c_n} is another turns parameter that is nearly always unity for our case as $g_n \ll \lambda_o$. Note that Eqs. (4.9) and (4.10) typically assume $Z_{slice} = Z_n$ (usually represented as normalized admittance values) to solve for susceptance parameters in the equivalent network of E-plane T-junctions in waveguides. The reactive elements of these equations, X_L and X_C , mostly have out-of-band impacts and become secondary effects in the circuit model behavior; this is primarily due to the tapered slot width being much wider than the slot gap ($d_n \gg g_n$) and the operating wavelength being much larger than gap sizes ($\lambda_o \gg g_n$) for typical slicing cases. Consequently, the

Figure 4.5: Progression to slice circuit model analysis. (a) Single-polarized SNA array element; (b) Bisected equivalent model with a virtual ground plane; (c) Zoomed view of a single slice with parameters; (d) Physical relation to an E-Plane T-Junction parallel-plate waveguide; (e) Equivalent circuit model of (d) [123].

reactive network elements will be neglected for our purposes but otherwise may need to be incorporated.

A critical aspect of the equivalent circuit in Fig. 4.5(e) is the transformer turn parameter, n_{turns_n} , which for the E-plane T-junction (or SNA element), represents the ratio of modal voltages in the junction between the collinear throat (or the two Z_n arms of the SNA taper slot step) and the E-plane throat (or Z_{slice} arm corresponding to the SNA slice). This turns parameter is controlled by the slot gap size g_n and the taper slot step half-distance d_n as is apparent from (4.12). Typically, the transformer is a step-up transformer ($n_{turns_n}^2 < 1$) in which more power is coupled through the slot rather than upwards along the fin because $d_n > g_n$ i.e. the tapered slot is wider than the slot gap. However, step-down operation ($n_{turns_n}^2 > 1$) cases can occur when more power remains in the tapered slot either by (i) coupling longitudinally along the fin or (ii) reflecting at the slice interface due to high Z_{slice} . Thicker elements can induce (i) more easily due to occupying more lateral area in the slice (i.e. embodying wide parallel plate capacitors) and increasing the coupling coefficient within the slot; thicker elements consequently require larger slot gap sizes as opposed to thin elements to maintain similar coupling. Thus, insight into coupling effects within the slot can be understood from the transformer n_{turns_n} parameter at each slice junction (which were shown to also be critical for polarization concerns in the preceding far-field discussion).

The approximate circuit model for the SNA array element at broadside scan can be completed as shown in Fig. 4.6. Here, the circuit equivalent of a segment n is given in the left side of the figure, and then inserted midway into each taper slot section n along the cascaded series. A key part of this model is the virtual short-circuit termination of the slice transmission line section. This arises from the symmetry of the model at broadside scan about the center of the fin (the electric field must go to

zero at the center of the slice)². The physical length of the shorted transmission line is thus equivalent to half the physical length of the full slice, $\ell_{slice_n}/2$, where

$$\ell_{slice_n} = D_E - 2d_n. \quad (4.13)$$

It can be easily seen that as d_n increases (the taper slot step width becomes wider), ℓ_{slice_n} decreases³. Furthermore, the input impedance of the shorted stub⁴, Z'_{slice} , can be written as

$$Z'_{slice_n} = j2Z_{slice_n}n_{turns_n}^2 \tan(\beta\ell_{slice_n}/2) \quad (4.14)$$

where $\beta = k_0\sqrt{\epsilon_r}$ and ϵ_r is the relative permittivity of the medium within and surrounding the slice. Z_{slice_n} is multiplied by a factor of 2 that arises from the fact that the half-slot length shorted stub is inserted in series on both paths of the Z_n transmission line due to symmetry (forming the full length slot); both stub lines are tied to ground and are thus in parallel to simplify to a single line with double the characteristic impedance. Z_{slice_n} can be approximated with closed-form expressions for isolated slotlines obtained from conformal mapping [124], or conversely calculated using infinite array slotline simulations, to follow expected design trends i.e. smaller element thickness t , wider slot gap size g_n , and lower ϵ_r lead to increased characteristic impedance due to smaller capacitance per unit length.

The developed circuit model enables us to develop simple preliminary physical insights from basic transmission line theory before proceeding to simulation results.

²For off-broadside scans, this symmetry is broken (particularly in the E-plane) and this short-circuit termination does not hold. It is left as future work to model these effects in a circuit. Broadside scan results remain a strong indication of initial benchmark performance.

³Note that ℓ_{slice_n} and ℓ_{slot_n} (from the previous far-field theory section) refer to the same quantity. The terms ‘slot’ and ‘slice’ will be used interchangeably.

⁴This refers to the shorted stub arising from the slice, and is not associated with stubs used in typical stripline Vivaldi feeding.

Figure 4.6: Single-mode circuit approximation of a single-polarized infinite SNA array element at broadside scan.

- The shorted stubs will become resonant at quarter-wavelength i.e. $\beta\ell_{slice_n}/2 = \pi/2$. For the typical half-wavelength unit cell spacing, this occurs when $\ell_{slice_n} = D_E$ (in free-space). Thus, the slices may cause resonance problems when their electrical length approaches the unit cell size.
- Higher Z_{slice_n} makes the stub response less sharp (damped Q factor). This means larger gap sizes may pose problems if the stub response becomes more broadband and enters the array operating band to disrupt impedance performance; another interpretation is more power is reflected at the slot interface due to higher n_{turns_n} leading to higher Z_{slice_n} .

These observations will be discussed in more detail when addressing design guidelines later in this chapter.

4.2.2 Input Impedance Model Results

This section will act as a direct extension to the procedures followed for the Vivaldi array in Section 2.2.2.2. The developed SNA circuit model will be compared with full-wave results; the models are identical to those in Section 2.2.2.2 except for the introduction of a slice at the end of each tapered step (values of the Vivaldi array parameters can be found in Tables 2.3, B.1, and B.2). SNA full-wave models of $N + 1 = 10$ taper steps for varying thickness $t/D_H = \{0, 0.165, 0.5\}$ are shown in Fig. 4.7(a-c) where a total of $N = 9$ slices are introduced (note that N refers to the total number of slices and is one less than the total number of taper steps). The slices were chosen to be uniform gap size g_n for each respective model, and g_n changes between each different thickness model to realize a constant characteristic impedance of $\approx 85\Omega$ (calculated through an infinite array full-wave analysis like the other slotline values). This impedance was chosen as it (i) sets values that push the minimum threshold for standard PCB fabrication methods in the sheet PEC model i.e. $g_n = 2$ mils and (ii) provides moderate impedance disruption (effects of slot impedance will be touched

Figure 4.7: Full-wave simulation models in Ansys HFSS of three single-polarized infinite all-metal SNA arrays with $N = 10$ taper slot steps and different thicknesses. These are equivalent to those in Chapter 2 c.f. Fig. 2.7 albeit the introduced slices. (a) $t/D_H = 0$; (b) $t/D_H = 0.165$; (c) $t/D_H = 0.5$.

upon in the design guidelines section). Also, as visible in Figs. 4.7(a-c), the chosen slot impedance provides a wide variety of gap sizes when realizing similar Z_{slice} across different thickness models. Thicker elements require larger slices to realize the same characteristic impedance of the slot; this is an important design consideration.

The SNA element circuit model slice parameters n_{turns_n} and ℓ_{slice_n} in each model were computed using (4.13) and (4.12) and tabulated for thicknesses $t/D_H = \{0, 0.165, 0.5\}$ respectively in Tables 4.1, 4.2, and 4.3; the results for each model comparing the circuit and full-wave infinite array simulations of the input impedance (Z_{in}) at broadside scan are shown in Figs. 4.8, 4.9, and 4.10. Each chart corresponds to when only one

Table 4.1: Parameters for $t/D_H = 0$ model of Fig. 4.7(a) and fixed $Z_{slice_n} = 88\Omega$ ($g_n = 2$ mils (0.051 mm)).

n	1	2	3	4	5	6	7	8	9
$g_n[mm]$	0.051	0.051	0.051	0.051	0.051	0.051	0.051	0.051	0.051
$2d_n[mm]$	0.5	1	1.5	2	2.5	3	4.25	5.5	6.75
$n_{turns,n}$	0.451	0.319	0.260	0.225	0.202	0.184	0.155	0.136	0.123
$\ell_{slice,n}/2[mm]$	6	5.75	5.5	5.25	5	4.75	4.125	3.5	2.875

Table 4.2: Parameters for $t/D_H = 0.165$ model of Fig. 4.7(b) and fixed $Z_{slice_n} = 85\Omega$ ($g_n = 1.02$ mm).

n	1	2	3	4	5	6	7	8	9
$g_n[mm]$	1.016	1.016	1.016	1.016	1.016	1.016	1.016	1.016	1.016
$2d_n[mm]$	1.5	1.75	2.25	3	3.75	4.5	5.5	6.5	8.5
$n_{turns,n}$	1.161	1.075	0.945	0.821	0.734	0.670	0.606	0.558	0.488
$\ell_{slice,n}/2[mm]$	5.5	5.375	5.125	4.75	4.375	4	3.5	3	2

Table 4.3: Parameters for $t/D_H = 0.5$ model of Fig. 4.7(c) and fixed $Z_{slice_n} = 84\Omega$ ($g_n = 2.25$ mm).

n	1	2	3	4	5	6	7	8	9
$g_n[mm]$	2.25	2.25	2.25	2.25	2.25	2.25	2.25	2.25	2.25
$2d_n[mm]$	3.25	3.75	4.25	5.25	6	6.75	7.5	8.75	10
$n_{turns,n}$	1.161	1.081	1.015	0.914	0.855	0.806	0.764	0.708	0.662
$\ell_{slice,n}/2[mm]$	4.625	4.375	4.125	3.625	3.25	2.875	2.5	1.875	1

slice is present at a time for a given location. From first glance, the trends in all three cases are in good agreement and successfully emulate the effect of the slice implementation. It is clear that lower slices closer to the feed location (i.e. n nearer to 1) disrupt the element input impedance much more wildly than those of higher slices (i.e. n nearer to 9). Furthermore, these adverse effects begin to occur lower in frequency with lower slices. This is due to the fact that $\ell_{slice,n}$ is inherently longer for slices closer to the feed due to the nature of the tapered slot topology; this pushes a harmful slot resonance into the band at lower frequencies where $Re\{Z_{in}\} \rightarrow 0$ as the electrical length of the shorted stub in the circuit model becomes resonant at $\lambda/4$. Slice #2 manifests a destructive parallel resonance as $Re\{Z_{in}\} \rightarrow \infty$ (as opposed to a series resonance like slice #1) around 10 GHz as it is spaced $\approx \lambda/4$ further from the feed. For each case, the high-frequency quarter-wave stub resonance shifts higher in frequency with increasing slice # as $\ell_{slice,n}$ decreases and alternates from series and parallel resonances based upon its electrical distance from the feed. This occurs until these resonant effects are out-of-band for higher slices (i.e. #5-#8) in which the input impedance begins to emulate very nearly the same performance as if no slice was present (i.e. the input impedance of a Vivaldi array). Ultimately, limits are placed upon how low slices can be placed before their resonances enter the band, which can have implications on the polarization performance as will be discussed in the following section.

Furthermore, the thickest element circuit approximations has larger discrepancies (c.f. Fig. 4.10) that are mainly due to the fact that the approximated Vivaldi circuit baseline model of this thickness was the least accurate to begin with. Thicker models in general also appear to have more severe impedance anomalies from a given slice; part of this is attributed to the baseline reference Vivaldi circuit models again being less accurate with full-wave simulations to begin with in comparison with the thin element. However, this is primarily due to the fact that, for a given slice characteristic

Figure 4.8: Comparison of equivalent circuit model prediction with full-wave simulations for the input impedance (Z_{in}) of an $N=9$ sliced stepped single-polarized infinite SNA array for $t/D_H=0$ (c.f. Fig. 4.7(a)). Each result is for one slice at a time.

Figure 4.9: Comparison of equivalent circuit model prediction with full-wave simulations for the input impedance (Z_{in}) of an $N=9$ sliced stepped single-polarized infinite SNA array for $t/D_H=0.165$ (c.f. Fig. 4.7(b)). Each result is for one slice at a time.

Figure 4.10: Comparison of equivalent circuit model prediction with full-wave simulations for the input impedance (Z_{in}) of an $N = 9$ sliced stepped single-polarized infinite SNA array for $t/D_H = 0.5$ (c.f. Fig. 4.7(c)). Each result is for one slice at a time.

impedance, thicker elements require much larger slice gap sizes and result in higher $n_{turns,n}$ parameters. This significantly inflates the input impedance of a slice c.f. (4.14) and as indicated by Tables 4.1, 4.2 and 4.3; this results in more severe effects on impedance. Thus, with respect to very thin element, thick elements generally need to have smaller gaps to compensate for a given slice characteristic impedance.

It is also noteworthy that the input impedance for the slice #1 cases in the low-band exhibit elevated resistance at the first parallel resonance; this is an artifact of the chosen equidistant length of the tapered steps (i.e. commensurate lines) as by Richards' transformation and Kuroda's identities [28]. In this manner, if one were to use these principles to consider the element model from a distributed low-pass filter implementation perspective (as will be done in some capacity in the next section), it can be understood that the physical length of these commensurate transmission lines (i.e. "unit elements" in filter terminology) impact the filter response. However, it is generally difficult to predict the response beforehand without the use of simulation tools as the characteristic impedance of each "unit element" is different in addition to variations in slice length for different n ; design guidelines for optimization and tuning tend to be more useful for practical design as will be presented in the next section. Nonetheless, the length of these commensurate lines becomes resonant at approximately $\lambda/8$ around 5 GHz and contributes to increased resistance in slice #1; this could be more well-matched by slicing several segments at one time as will now be discussed.

The cases where the element is sliced at all nine locations simultaneously like in Fig. 4.11(a) are shown in 4.11(b-d). Compared to some of the previous individual lower slice cases, the impedance with all slices behaves much more controlled below the high-band cut-off; this can be reasoned to stem from the input impedance with all slices present being proportional to the cascaded responses with each individual slice present (similar to a cascade of filter sections) and this leads to a smoother response

in the pass-band since the majority of individual slice cases do not exhibit poor mid-band matching. However, in the high-band, the input impedance still suffers from a resonance as $Re\{Z_{in}\} \rightarrow 0$ very near the same location as it did for slice #1 previously in Figs. 4.8, 4.9, and 4.10; this indicates the lowest slice dictates the high-end disruptive impedance behavior and is critical to design properly. Thus, if the bottom slice doesn't result in catastrophic resonances in-band, then all higher slices tuned similarly (i.e. same gap size) will inherently also have no resonant behavior in-band since their slice lengths are shorter due to the tapered slot geometry. As mentioned, this may further be understood as a cascade of low-pass filter sections between a source impedance (i.e. element feed impedance) and a load impedance (i.e. free-space) in which the first filter section dictates the earliest cut-off frequency; ensuring the design of the first filter (slice) in the chain does not impede performance then by extension means the proceeding filters also will not if designed similarly (i.e. characteristic impedances of the slots are similar but having higher cut-off frequencies).

Lastly, despite vast differences in element thickness, all three models operate very similarly with good agreement; this implies that this circuit model can satisfy different embodiments ranging from thin metallizations in PCB fabrication or thick metallizations in stock-metal embodiments. Critically, the overall good agreement between full-wave and circuit simulations validates the proposed circuit model. The established trust enables the circuit model for use in setting SNA element design guidelines in the next section.

4.3 Design Guidelines

The preceding SNA element circuit model showed good agreement with full-wave infinite array simulations at broadside scan and thus it can be used to shed greater physical insight into the SNA impedance operation. The goal of this section is to use this model to develop simple guidelines with respect to physical slice parameters that

Figure 4.11: Comparison of equivalent circuit model prediction with full-wave simulations for the input impedance (Z_{in}) of an $N = 9$ sliced stepped single-polarized infinite SNA array for $t/D_H = \{0, 0.165, 0.5\}$ (c.f. Fig. 4.7(c)). Each result incorporates all slices present. (a) SNA diagram of results for (b-d); (b) $t/D_H = 0$ case; (c) $t/D_H = 0.165$ case; (d) $t/D_H = 0.5$ case.

can be used to help facilitate the design of SNA arrays (i.e. for those designed in this dissertation). This will be accomplished using three studies: (i) circuit simulations for impedance tuning studies, (ii) a parametric study between matching and cross-polarization in SNA elements, and (iii) revisiting bandwidth vs. cross-polarization trends in SNA and Vivaldi arrays.

4.3.1 Model

Impedance design guidelines stemming from the validated circuit model will be established to shed light into tuning SNA elements to avoid any catastrophic resonances due to slicing and also maintain the well-matched wideband performance native to Vivaldi arrays. This analysis purely investigates the effects of slicing on impedance behavior by reducing the circuit model to that in Fig. 4.12 where the circuit can essentially be treated as a two-port cascaded series of distributed low-pass filter prototypes with equivalent resistive load terminations⁵. Equivalent loads, in conjunction with removal of the backing cavity and inductive components, simplify the analysis by restricting changes in physical behavior to be purely due to effects manifesting from slices. Additionally, all transmission lines providing physical separation between slices are matched to the load impedances i.e. $Z_{feed} = R_A = Z_0 = 100\Omega$ (this value was chosen for consistency with previous circuit models). After conclusions are drawn about phenomena taking place due to slices, typical Vivaldi element tuning procedures (i.e. tapered slotlines, backing cavity, etc.) can be employed to provide a smooth match to free-space as done for the proposed SNA arrays in Chapter 3 and 5.

⁵Treatment with unequal resistive terminations (i.e. $Z_{feed}=100\Omega$, $R_a=377\Omega$) is possible using LC ladder network synthesis techniques for desired response types i.e. Butterworth, Chebyshev, etc. [125–127]. This approach is based upon solving for low-pass prototype LC sections along the ladder that are treated as incremental lengths of transmission line; this is generally not the case for our concerns here because (i) the transmission lines are of considerable length and (ii) the design goal typically aims plainly for good matching rather than a generalized response type.

Figure 4.12: Simplified circuit for establishment of design guidelines. The circuit can be viewed as a cascaded series of distributed low-pass filter prototypes with resistive load terminations where $Z_{\text{feed}} = R_A$.

The primary objective will be to maintain minimal losses (maximize the filter transmission coefficient) within the operating band up to the array high frequency (i.e. the pass-band of the low-pass filter with cut-off at f_g) by reasoning simulation trends and establishing design guidelines. Assuming lossless materials, the amount of power transferred from Port 1 (Z_{feed}) to Port 2 (R_A) in Fig. 4.12 can be interpreted as the total radiated power per element unit cell normalized to the incident power at the input port; thus only impedance mismatch losses are considered resulting from slices. The discussed approach enables simple and intuitive design guidelines to be drawn from (i) typical distributed element low-pass filter synthesis and (ii) performance trends from variation of the geometrical slice parameters. Such parameters of interest for slices include the

1. *Slice Length*: electrical length, $\beta\ell_{slice_n}$
2. *Slice Gap Size*: g_n
3. *Number of Slices*: N
4. *Periodicity*: the electrical length between adjacent slices, $\beta_0\ell_0$

An analysis of these parameters will be conducted followed by a summary of design guidelines at the end of this section.

4.3.1.1 Slice Electrical Length and Gap Size

The electrical length and gap size of a slice directly affect the slice input impedance Z'_{slice_n} . Recalling (4.14), Z'_{slice_n} can be approximated as a function of slice geometrical parameters as

$$Z'_{slice_n} = j2\sqrt{\frac{L_n}{C_n}} \left(\frac{g_n}{d_n} \right) \tan(\beta_0\sqrt{\epsilon_r}\ell_{slice_n}/2) \quad (4.15)$$

where L_n and C_n are the slot inductance and capacitance per unit length. From an impedance matching standpoint, it is of course desirable for Z'_{slice_n} to be zero or

close to it, such that the slice junctions cause no reflections. However, several factors instead result in heightened Z'_{slice_n} as seen from (4.15) for instance increased electrical slice length (specifically $\beta_0\sqrt{\epsilon_r}\ell_{slice_n}/2 = m\pi/2$ where m is an odd integer), and wider gap size or even decreased element thickness (both due to increased L_n , decreased C_n).

The effect of increasing the input impedance of the slice, Z'_{slice_n} , is shown in Fig. 4.13(a) for the case where the slice length is $\lambda_g/2$. In this case, only one slice junction is present ($N = 1$), $\epsilon_r = 1$, and its inserted location does not matter since $Z_{feed} = R_A = Z_0$. The y-axes correspond to $|S_{21}|$ in dB (c.f. Fig. 4.12 for port definitions) and the x-axes indicates normalized frequency with respect to the grating lobe frequency f_g (assumed to be the half-spacing grating lobe frequency, $f_g = \frac{c}{2D_E}$; $D_E = \lambda_g/2$; $D_E = D_H$). When ℓ_{slice_n} or Z'_{slice_n} become zero, $|S_{21}| = 1$ and results in the ideal lossless all-pass response which would correspond to the case of a contiguous Vivaldi fin. A clear trend is apparent in Fig. 4.13(a) where increased Z'_{slice_n} results in more broadband impedance deterioration around the resonant frequency. This results from a dampened Q factor as the slice input impedance increases. *Thus, it is critical to minimize Z'_{slice_n} for impedance matching considerations; this is usually most practically accomplished (for a given slice length) by decreasing the gap size g_n c.f. Eq. (4.15).* It is worthy to note that increasing the element thickness t or ϵ_r also reduces Z'_{slice_n} due to increased capacitance per unit length C_n (granted the resulting electrical length remains the same). Conversely smaller gap sizes in thinner (or lower ϵ_r) elements can be made larger with higher t or ϵ_r while maintaining the same Z'_{slice_n} . This was demonstrated in Section 4.2.2 for increased t .

For Fig. 4.13(b), Z'_{slice_n} is now fixed to 100Ω with $N = 1$ and $\epsilon_r = 1$, and ℓ_{slice_n} is varied to yield another trend: as ℓ_{slice_n} becomes larger, harmful resonances drops lower in frequency. Each resonance corresponds to $Z'_{slice_n} \rightarrow \infty$ as the shorted stub becomes a quarter-wavelength (recall the actual stub length is $\ell_{slice_n}/2$) or any odd integer

Figure 4.13: Effect of slice input impedances (Z'_{slice_n}) and electrical lengths ($\beta\ell_{slice_n}$) versus normalized frequency when $N = 1$, $\epsilon_r = 1$, and $Z_{feed} = R_A = Z_0 = 100\Omega$. (a) Input impedance trends for $\ell_{slice_n} = \lambda_g/2$; (b) Electrical length trends for $Z'_{slice_n} = 100\Omega$.

multiple due to Eq. (4.15) and the periodic nature of distributed element filters. Thus, if the slice spans the entire element spacing D_E , then the resonance will be located at the high frequency $f = f_g$ and provide a complete impedance mismatch. This results in a low-pass response where the -3 dB point occurs around $f/f_g = 0.75$ for this case. Furthermore, the resonance moves to midband at $f = 0.5$ when $\ell_{slice_n} = \lambda_g$. Although this is not physically possible in free-space since the unit cell is restricted to $D_E = \lambda_g/2$, it provides critical insight if $\epsilon_r > 1$ and the electrical length is increased. *Thus, the electrical slice length needs to be kept short ($\beta\ell_{slice_n} < \lambda_g/2$) to mitigate problematic resonances within the band; this is increasingly more challenging for structures with high dielectric constants since they force the slice length to be physically small (slicing at higher points along the fin).*

4.3.1.2 Number of Slices and Spacing Between Slices

From a filter perspective, the number of slices N is related to filter order. Thus, for a higher number of slices (higher filter order), it is logical that the response should be steeper around cut-off and exhibit more ripple in the passband. Additionally, the physical spacing between the slices also plays a significant role in tuning. For simplicity, we assume the slices are separated periodically by an electrical distance $\beta_0\ell_0$ (which corresponds to the electrical length of the typical taper slot step) but in reality could be aperiodic. The impact of slice periodicity can be rudimentarily understood by considering Richards' transformation and Kurodas' identities [28]. By these principles, shorter and longer ℓ_0 respectively shift f_c upwards and downwards; shorter lines reach a high-end limit by the shorted stub quarter-wave resonance. Passband matching below f_c is furthermore affected by the length of these lines—shorter lines exhibit worse matching at lower frequencies, longer lines ($\ell_0 > \lambda_g/2$) can cause in-band resonances and impedance aberrations at frequencies below f_g .

The results for varying N but constant $\ell_0 = \lambda_g/8$, $Z'_{slice_n} = 100\Omega$, and $\ell_{slice_n} = \lambda_g/2$ are charted in Fig. 4.14(a). This corresponds to a filter that has the same unit cell, but becomes physically larger and higher order as N increases. Trends expectedly indicate that increasing N yields steeper cut-off and also produces more rapid in-band ripple around the original $N = 1$ response (the black curve). *It is also clear that in order to operate up to $f = f_g$ with several slices, ℓ_{slice_n} must be less than $\lambda_g/2$ to compensate for the transition slope and provide a cushion for a sharper high-end cut-off.* In fact, this sharp cut-off was previously observed in the active VSWR charts of the single-polarized SNA array in Chapter 3 in which it was tuned to be above f_g .

The physical spacing between slices ℓ_0 is presented for several cases in Fig. 4.14(b) for $Z'_{slice_n} = 100\Omega$, $\epsilon_r = 1$, $\ell_{slice_n} = \lambda_g/2$, and $N = 9$. This corresponds to having the same order filters, but with larger physical size as ℓ_0 increases. We again assume the physical length of the lines between slices is commensurate, leading to a periodic structure. For the largest periodicity case $\ell_0 = \lambda/2$, the high-end cut-off drops to slightly above midband whereas decreasing ℓ_0 pushes the high-end upwards towards $f/f_g = 1$ (the quarter-wave stub resonant frequency). Despite providing better high-end potential, decreasing ℓ_0 also deteriorates matching within the operating band. *A physical separation between $\lambda_g/16 - \lambda_g/4$ is a reasonable compromise between matching and high-frequency cut-off.* Although not shown explicitly, it can be reasoned that slices with out-of-band resonant effects i.e. $\ell_{slice_n} \ll \lambda_g/2$ do not affect passband input impedance and the spacing is less critical for impedance considerations; the following section will show it is desirable to space slices more closely as to fit more slices along the fin for polarization considerations. Lastly, some spacing cases in Fig. 4.14(b) are different from all prior curves in the sense they are not symmetric about $f/f_g = 1$; the responses are in fact periodic every $4f_c$ due to the definition of Richards' transformation [28] although most of our cases happened to be being

(a)

(b)

Figure 4.14: Effect of total slice number of slices (N) and physical distance between slices (ℓ_0) versus normalized frequency when $Z_{\text{feed}} = R_A = Z_0 = Z'_{\text{slice}_n} = 100\Omega$, $\ell_{\text{slice}_n} = \lambda_g/2$, and $\epsilon_r = 1$. (a) Number of slices trends for $\ell_0 = \lambda_g/4$; (b) Spacing between slices trends for $N=9$.

innately symmetric about the first stub resonance (nonetheless, we are not interested in frequencies above f_g as they are not part of the intended operating band).

4.3.2 Matching vs. Cross-Polarization Trade-Offs

A parametric study of a single-polarized SNA array is provided to analyze design trade-offs between impedance matching and CPR. Full-wave infinite array simulations are carried out for a stripline PEC sheet of SNA arrays that are fed by a standard Knorr balun unbalanced 50Ω feed transition in the middle layer as shown in Fig. 4.15(a). For this analysis, the total number of inserted slices progressively increases in a top-down fashion from $N = 1 \dots 10$. Each slice is physically spaced apart equal distances ($\ell_0 \approx \lambda_g/6.6$) with uniform slice gaps ($g_n = \lambda_g/25$). From our investigation in the previous sections, it is expected that lower slices should perturb the impedance performance more heavily than higher slices, and eventually a harmful resonance proportional to the slot length should come into the high-band (when the half slot length $\ell_{slot_n}/2 \approx \lambda_g/4$ or total slot length $\ell_{slot_n} \approx \lambda/2 = D_E$). It is also expected based upon the polarization theory and design principles throughout Chapter 3 and earlier in this chapter that it is desirable to insert more slices for better polarization considerations and for more well-balanced vertical-to-horizontal current ratios (i.e. more small loops).

Resulting D-plane $\theta_0 = 45^\circ$ scanning CPR and broadside scan VSWR versus frequency are charted in Fig. 4.15(b). As N increases, the CPR steadily improves throughout the entire frequency band and for near $N = \{8, 9\}$ the CPR is nearly flat at -10 dB. This is a dramatic improvement compared to when $N = 1$ (the black reference curve), where the CPR is $+10$ dB at the highest frequency ($f_{high} = f_g$) and steadily deteriorates with increasing frequency like typical Vivaldi arrays. However, in the bottom portion of the chart, increasing N (slice more and lower) has adverse effects on the matching levels. Additionally, a resonance near f_g creeps into the band for high

Figure 4.15: Full-wave design trade-off simulations for increasing the amount of slices. (a) Simulated HFSS unit cell; (b) Simulated D-plane $\theta_0 = 45^\circ$ scanning polarization ratio (top) and broadside VSWR (bottom) versus frequency.

values of N and disrupts performance. This is due to the longer slot lengths at lower slices; for this case $\ell_{slice_n}/2 \approx \lambda_g/2.4$ and the skirt of the resonance comes into the operating band. The effect of this impedance mismatch is also felt in the CPR portion above, where a rise in the CPR occurs for $N = 10$ in the high-band. Critically, the slices need to be carefully tuned when strategically placing slices in the lower portions of the SNA element. One such solution involves making lower slices have smaller gap sizes to decrease the slot characteristic impedance while higher slices can have larger gap sizes as their negative effects are out-of-band. Several other solutions can be implemented with designer creativity to enforce shorter slot lengths and decreased slot characteristic impedance when necessary. Nonetheless, a trade-off exists between matching and cross-polarization and, with careful design and optimization, a 'sweet spot' can be found where the slices can have a small impact on impedance while providing significant CPR improvements.

4.3.3 Performance Trends vs. Element Depth

In Chapter 2, it was shown that Vivaldi array bandwidth increases while CPR degrades as element aspect ratio increases, resulting in an undesirable performance trade-off. These trends are now revisited to include single- and dual-polarized SNA arrays in Figs. 4.16(a) and (b), respectively. The red lines correspond to the right y-axis of CPR (note negative values are on top), while the blue lines correspond to the left y-axis of bandwidth ratio (BWR). The plotted data was acquired from several well-optimized SNA and Vivaldi infinite array PCB-based designs (like those in Chapter 3) of different depths H . SNA array BWR remains very close to that of the Vivaldi array and with further optimization can most likely become equivalent. In terms of polarization performance, the SNA array offers significant CPR improvements approaching 20 dB. Moreover, the SNA CPR is virtually independent of the element aspect ratio. It is also interesting to note that the dual-polarized arrange-

Figure 4.16: Bandwidth ratio (BWR) and cross-polarization ratio (CPR) trends on PCB-based SNA and Vivaldi arrays. (a) Single-polarized arrangements; (b) Dual-polarized arrangements.

ments inherently exhibit better CPR than their single-polarized array counterparts which is fairly well-known and pointed out by [66]. *Conclusively, the SNA array achieves low cross-polarization like that of a low-profile array regardless of element depth while retaining Vivaldi-like bandwidths and matching levels.*

Furthermore, an interesting case study can be found in Appendix F to further quantify the effects of the sliced SNA region on both impedance and polarization when it is completely removed (but the contiguous base remains). The study concludes with the SNA array not only having the same desirable wide-scan impedance performance as its Vivaldi array counterpart (the tall Vivaldi array with the same profile as the original SNA array), but also offers an enhanced polarization performance over a lower profile (truncated) Vivaldi element.

4.3.4 Summary of Design Guidelines and Design Procedure

Learnings from the preceding subsections and the theoretical foundation will now be summarized to compose a basic list of design guidelines and procedures for SNA element tuning. The guidelines assume well-matched impedance operation up to $f = f_g$ (with unit cell dimensions $D_E = D_H = \lambda_g/2 = \frac{c}{2f_g}$) and the lowest possible cross-polarization are desired.

Design Guidelines:

- The electrical length of slices ($\beta\ell_{slice_n}$) should be less than $\lambda_g/2$ to avoid catastrophic resonances within the operating band
- The slice gap (g_n) should be maximized to reduce the slot coupling coefficient for polarization considerations *without causing unacceptable impedance mismatch*
- Slicing should begin as low as possible along the fin to allow more area above to be sliced for polarization considerations *without causing unacceptable impedance mismatch*

- The slices should typically be spaced approximately $\ell_0 \approx \lambda_g/16 - \lambda_g/4$ apart

Using these guidelines, the following basic design procedure was developed and used as a foundation to design the proposed SNA arrays through infinite array simulations in this work:

1. Design a well-matched Vivaldi array element of a particular embodiment
2. Calculate the lowest point along the Vivaldi array fin where $\beta\ell_{slice_n} = \lambda_g/2$. This indicates where the slot resonance will occur at f_g .
3. Insert a slice above this location ($\approx \lambda_g/2$ up the fin). The gap size of the slice should be as large as possible without causing unacceptable impedance mismatch within the operating band. The gap size depends upon the element thickness and various examples for different thicknesses were demonstrated in Section 4.2.2. Typical values can range from $Z_{slice_n} = 50 - 100\Omega$. Slotline impedance simulators (or even approximate expressions [124]) may be employed for rough starting points.
4. Slice repeatedly up the tapered fin with the same gap size periodically spaced at approximately $\lambda_g/4$. Check cross-polarization and impedance performance.
5. If cross-polarization goal is not met, then (i) increase the slice gap sizes where the tapered slot is wider towards the top of the fin for slices that will not have large in-band resonant effects, (ii) decrease the periodic spacing between slices to increase the amount of slices, or (iii) creatively slice lower without hurting impedance performance.

A benefit of the proposed design approach is that it is built upon the foundation of first designing a well-matched Vivaldi array for which robust design guidelines are well-known and available. Following this, the remaining procedure is fairly simple

and revolves mainly around inserting as many slices as possible while eliminating or minimizing the effects of slot resonances.

4.4 Conclusions

This chapter took a deeper look into the core operating principles of SNA arrays. Building upon the groundwork laid in Chapter 2, theoretical models for the SNA far-field were developed that incorporate the effect of slicing on the far-field polarization. It was shown that the CPR can be favorably controlled through slicing implementation and its coupling effects. An approximate single-mode circuit model was developed that relates an inserted slice to that of a simple E-plane T-Junction parallel plate waveguide; good agreement was demonstrated between full-wave and circuit simulations. Design guidelines were then provided based upon this circuit model to assist in impedance tuning; it was shown slices must be kept small in the amount, short in length, and small in gap size to prevent disruption of impedance; this becomes harder to achieve when slicing closer towards the feed. Conversely, it was also shown that it is desirable for polarization considerations to introduce a high amount of slices with large gap sizes at lower points towards the feed; an inherent trade-off between matching and polarization exists that requires careful tuning and optimization. A critical finding was that a high-profile element could achieve well-matched 10:1 impedance performance without lossy materials while maintaining good polarization behavior like that of a low-profile array. The theoretical foundation and principles laid in this chapter formed the basis for the design of the SNA arrays in Chapter 3 across various embodiments, and also the fabricated proof-of-concept in Chapter 5 to be discussed next.

CHAPTER 5

MEASURED PERFORMANCES OF 1.2-12 GHZ SNA AND VIVALDI ARRAYS

Using the theoretical operation principles and design guidelines developed in Chapter 4, this chapter will present the design, fabrication, and measurement of a 19×19 single-polarized Sliced Notch Antenna (SNA) array. The array was designed to operate from 1.2-12 GHz (10:1) to cover near the entirety of four major RF bands (L, S, C, and X). The intent of this chapter is to provide a proof-of-concept of the first fully constructed and tested SNA array, and comparatively evaluate its performance against a counterpart Vivaldi array that was also designed and fabricated. Both arrays were manufactured as multilayer microwave PCB cards consisting of 19 elements as shown in Fig. 5.1(a) and then assembled into 5×19 and 19×19 configurations like those of Fig. 5.1(b) and (c). The arrays were configured and measured in different sizes to (i) verify proper design/fabrication and gain trust in the utilized numerical solvers and (ii) analyze finite array effects on performance. The SNA array and its Vivaldi counterpart are identical in all aspects (depth-to-periodicity dimension, feeding, materials, etc.) except for the reconfiguration of the Vivaldi fin into that of the SNA topology, which introduces no additional mechanical complexity.

This chapter will begin with an overview of the SNA and Vivaldi array element designs followed by a synopsis of the array assembly and measurement configurations. Next, impedance and far-field polarization performance results of the arrays in 1×19 (linear), 5×19 , and 19×19 configurations will be presented using three sources of analysis:

(a)

(b)

(c)

Figure 5.1: Single-polarized 1.2-12 GHz (10:1) SNA array and its Vivaldi array counterpart. (a) Linear (1×19) array PCB cards; (b) Assembled 5×19 arrays; (c) Assembled 19×19 SNA array.

1. Infinite array full-wave simulations
2. In-house Domain Decomposition Finite Element Method (DD-FEM) simulations [118]
3. Center embedded element measurements

In this manner, infinite array simulations will provide the desired performance benchmarks to be achieved by a central finite array element, while in-house DD-FEM simulations will fully incorporate the truncation effects due to finite array sizes and provide an accurate prediction of the actual measured array performance. Altogether, the SNA array prototype will demonstrate 10:1 bandwidth operation and excellent agreement with simulations while providing enhanced impedance performance for E-plane scans and >15 dB CPR improvement over its counterpart Vivaldi array in the D-plane at $\theta = 45^\circ$.

5.1 Element Designs

This section will incorporate a brief synopsis of the element design, fabrication, and assembly process. The design choices were guided using theory and design methods discussed in Chapters 3 & 4.

5.1.1 Topology Overviews

The prototypes are fully fabricated as multilayer stripline PCB boards using Rogers Duroid PTFE substrates ($\epsilon_r = 2.2$). Both designs share several notable design features as can be seen throughout Figs. 5.2(a-d), specifically the incorporation of stripline Knorr balun feeds, unbalanced 50Ω end-launch SMA feed launches, plated vias in the contiguous metal regions to suppress excitation of parallel plate modes while scanning, and tapered fin regions. The arrays also share the same element depth (70 mm) and periodicity (12.5 mm) dimensions, resulting in a half-wavelength

grating lobe frequency of 12 GHz. The feed line transition lengths from the end-launch SMA to the Knorr balun are not drawn to scale and are approximately 30 mm in actuality to allow room for metal blocks to be inserted and embody the ground plane (shown later).

Comparing the two arrays of Figs. 5.2(a) and (b), the SNA array is shown to be dissected into four separate sections that critically differentiate itself from its Vivaldi array counterpart. Aside from having the same feed transition, the SNA array reconfigures the contiguous metal tapered fin of the Vivaldi array into a disconnected series of vertical strips following a tapered envelope in Figs. 5.2(b) and (d). As discussed previously, this topology is critical for achieving good polarization performance while retaining the high-profile necessary for wideband impedance matching. The vertical strips in the disconnected taper are separated by 10 mil gaps, tapered exponentially for a smoother impedance transition to free-space, and do not require the typical vias used to eliminate parallel plate modes in Vivaldi arrays. A transition region between contiguous base and disconnected taper is also designed to be pyramidal in shape to provide a less abrupt impedance transition between the contiguous tapered base and the disconnected SNA fin for improved matching.

Another key aspect in this SNA array prototype is the introduction of a vertical cut (2 mm wide) inserted along the center of the disconnected taper fin down to the disconnected transition region. This assists in high-end impedance tuning by keeping resonant slice lengths in the disconnected region shorter. Perhaps more importantly, *this modification also enables the plausibility of an orthogonal card being inserted for the extension into a dual-polarization arrangement without requiring adjacent electrical connection.* As shown in dual-polarized design examples in Chapter 3, this can dramatically reduce the soldering requirements typically required along the high-profile contiguous Vivaldi fin, and could altogether eliminate soldering if the base did not require soldering (i.e. a stock-metal embodiment). Although the modified

Figure 5.2: Fabricated element design illustrations and pictures. (a) Depiction of the Vivaldi array; (b) Depiction of the SNA array; (c) Highlighted portion of the fabricated Vivaldi array; (d) Highlighted portion of the fabricated SNA array.

impedance theory incorporating this vertical cut is not presented, it can be justified using effective dielectric properties of anisotropic periodic structures that are not near resonance (e.g. capacitive strip gratings or mesh grids [128–131]); this leads to design considerations incorporating small slice gap sizes and small distances between adjacent slices (or slice periodicities).

5.1.2 Array Fabrication and Assembly

As mentioned previously, the arrays are fully fabricated as multilayer PCBs and post-assembled with fastened metal bars. This technique has been used previously [132] and was adapted here for its convenience as a proof-of-concept demonstration. The boards consist of three printed layers and two 20 mil Rogers Duroid cores as shown in the PCB stackup (not drawn to scale) of Fig. 5.3(a). The cores are bonded together by GenClad CuClad 6250 ($\epsilon_r = 2.32$) and non-conductive fill plated vias of 20 mil diameter are drilled through the entire stackup. A portion of the printed artwork of the three layers (L_1, L_2, L_3) are respectively depicted in Figs. 5.3(b) and (c) for both Vivaldi (left) and SNA (right) arrays. As can be seen, the mechanical complexity of both arrays is identical—they utilize the same number of layers, number and material of PTFE cores, and occupied volume.

The arrays were fabricated and connectorized by Modular Components National (MCN). Two 1×19 element rows before connectorization can be seen in Figs. 5.4(a-b), courtesy of MCN. Flaps on the end of the cards with drilled holes were incorporated in the design for assembling the array and ground plane. This process is repeated to create 19 cards and eventually assemble a 19×19 array. This array size was chosen as it is convenient for designating a center element with an odd number of rows and columns, and 19 elements was calculated to be the largest amount of elements that could fit on a standard panel size for these designs. Ideally, the array is as large

Figure 5.3: PCB stackup and respective layer artwork for the fabricated Vivaldi and SNA arrays. (a) PCB stackup; (b) Vivaldi array artwork example; (c) SNA array artwork example.

Figure 5.4: Photos of the proposed Vivaldi and SNA 1×19 array cards in preliminary manufacturing stages (courtesy of Modular Components National) before connectorization. (a) Vivaldi array; (b) SNA array

as possible to mitigate finite array truncation effects and more closely approximate infinite array performance in central elements.

After receiving the cards, the array and ground plane needed to be assembled. The ground plane was formed by machining $5/8'' \times 7/16''$ (0.625 mm x 0.4375 mm) aluminum bars and drilling holes to align with those in the card flaps. These needed to be carefully machined to meet the designed grating lobe spacing in the H-plane. An aluminum rod was inserted through the ground plane blocks and cards and then fastened on both ends. A back-shot of the array with the metal blocks, rods, and array cards are shown in Fig. 5.5 for a 5×19 configurations with connectorized matched loads.

5.2 Measurement Configurations

A brief overview will be provided of the measurement configurations utilized for upcoming results.

Figure 5.5: Back shot of the 5×19 SNA array connectorized with matched loads for center embedded element measurements.

Center element active impedance performance was evaluated using an Agilent Network Analyzer to collect two-port S-parameters (S_{ij} 's) in an anechoic chamber as shown in Figs. 5.6(a) for an example measurement of the 19×19 proposed SNA array. The center element active reflection coefficient can be approximated when all other ports are matched by [24]

$$\Gamma_{mn}(\theta_0, \phi_0) = \sum_{p=1}^M \sum_{q=1}^N S_{mn,pq} e^{-j([p-m]D_E u_0 + [q-n]D_H v_0)} \quad (5.1)$$

where (θ_0, ϕ_0) is the array scan direction, $u_0 = k \sin \theta_0 \cos \phi_0$ and $v_0 = k \sin \theta_0 \sin \phi_0$ are the spatial u-v coordinates, k is the free-space wavenumber, $S_{mn,pq}$ are the measured S-parameters between elements mn and pq , M and N are respectively the number of elements in the $\hat{e}_E$ -/ $\hat{e}_H$ -directions, and D_E and D_H are the physical unit cell dimensions. For the case of the 19×19 array, this procedure required 361 unique measurements to compute the active reflection coefficient for the center element.

Far-field performance was analyzed by measuring the center embedded element far-field patterns in an anechoic far-field chamber. Measurements were collected for principal-plane (E-/H-plane) and diagonal plane (D-Plane) elevation cuts in the forward half-space ($-90^\circ \leq \theta \leq 90^\circ$) at 1° intervals. An exemplary configuration for a

(a)

(b)

Figure 5.6: Examples of impedance and far-field measurement configurations. (a) Anechoic chamber and network analyzer measurement set-up of the 19×19 proposed SNA array; (b) D-plane far-field measurement of a 19×19 SNA array mounted upon blue foam cut at 45° angles.

D-plane measurement is shown in Fig. 5.6(b) in the UMass far-field anechoic chamber. In order to measure the D-plane performance in this chamber, the array was rotated in its azimuthal plane at a 45° angle by utilizing an angled piece of cut foam. This may have some effects on D-plane measurements at wide angles and also makes precise alignment difficult.

5.3 Results: Organization & Precursor

The results sections will begin by presenting infinite array design simulations. The measured results will follow and be split into their own exclusive sections: (1) impedance performance and (2) far-field performance. Each section will simultaneously analyze the proposed SNA array and its counterpart Vivaldi array in this order of array configurations: linear (1×19), 5×19 , and 19×19 . In each case, the measured results will be evaluated with both full-wave finite and infinite array numerical predictions.

Although the primary point of concern will be the performance evaluation of the 19×19 arrays, the electrical behavior of the linear and 5×19 arrays will first be characterized to gain trust in our measurement procedure and simulations. These results are also useful by providing insight into the progressive impact of finite array size truncation on electrical performance. In all cases it will be shown that there is an overall excellent agreement between measurement/simulations and that the proposed SNA array provides the same bandwidth as its Vivaldi counterpart while delivering significant polarization improvements. Logically, it will also be shown that larger array size measurements are closer to infinite array VSWR results and exhibit more stable impedance while scanning in all planes¹. Perhaps less obvious, the far-field polarization performance will also improve with increasing array size.

Lastly, it has been shown repeatedly that cross-polarization issues are more devastating in the mid-/high-band while efforts towards improved impedance matching and wider bandwidth are typically more difficult to achieve in the low-band. Thus, the far-field analysis will focus more upon results in the mid-/high-band while the impedance analysis will concentrate more upon the scrutiny of low-frequency results.

5.4 Results: Infinite Array Performance

5.4.1 Impedance

The infinite array performance comparing the SNA array and its Vivaldi counterpart are presented in Fig. 5.7. Active VSWR levels are generally within ± 0.2 for each scan with the SNA array exhibiting slightly higher values for broadside and

¹As a general rule-of-thumb, the central elements in a planar array will not begin to emulate infinite array performance until the area of the array, A , is proportional to the operating wavelength by $A \approx 25\lambda^2$ [133]. For the measured 19×19 array near the low-band edge, $A \approx \lambda^2$ and thus the array will be electrically small and suffer from low-frequency finite size truncation effects; the central elements will begin to better approximate infinite array performance for frequencies above 6.3 GHz with this array size. Nonetheless, measured performance can still be accurately verified using finite array tools that encapsulate the truncation effects.

Figure 5.7: Infinite array VSWR simulations of single-polarized SNA array with vertical dissections and its Vivaldi counterpart. (a) E-plane scans; (b) H-plane scans.

H-plane scans but improved values for E-plane scans. As compared to the single-polarized SNA array in Chapter 3, this SNA array does not produce a sharp cut-off at the high-frequency for broadside but resonance does appear near f_g for E-plane $\theta_0 = 60^\circ$ scan in this case. Although the theory behind this is not presented fully in this work, it can be justified using effective dielectric properties of anisotropic periodic structures that are not near resonance. It should also be noted that the Vivaldi reference models are slightly different as the current designs utilize Rogers 5880 Duroid ($\epsilon_r = 2.2$) as opposed to Rogers 5880LZ ($\epsilon_r = 1.96$). Both arrays yield overall very good scan impedance performance over a 10:1 bandwidth while operating with optimal aperture sampling.

5.4.2 Polarization

The infinite array co-/cross-polarization vs. frequency for D-plane $\theta = 45^\circ$ scans between the SNA and Vivaldi array are shown in Fig. 5.8. The SNA array achieves ≥ 10 dB CPR improvement above mid-band over the Vivaldi array. For $\phi = 30^\circ, \theta = 60^\circ$ scans in the bottom chart, even larger improvements ≥ 25 dB are achieved in the high-

band. Furthermore, the extent of improvement can best be seen in the infinite array scanning polar patterns within Fig. 5.9. As seen previously, the worst scanning plane for a high-profile single-polarized Vivaldi element veers from the D-plane towards the H-plane, landing around $\phi_0 = 30^\circ$ for this case. The SNA array achieves >15 dB improvement at the worst angle and clears up all of the orange-/red-level CPR within a $\theta \leq 60^\circ$ scan cone. Additionally, the Vivaldi array can be seen to suffer from dips in its co-polarized pattern in the E-plane and extending towards the D-plane near broadside. These phenomena are in fact a class of scan blindnesses first recognized in [134] that are formed by deep cavities in single-polarized Vivaldi arrays. SNA arrays inherently alleviate this as they form a shorter cavity due to their smaller contiguous metal base. Lastly, scan resonances in the SNA array can be observed in the co-polarized pattern in Fig. 5.9(a) at wide angles in the E-plane due to the increased effective permittivity in the anisotropic dielectric medium; the design is on the edge of exciting a surface wave at the $\theta \leq 60^\circ$ scan cone in the E-plane.

5.5 Results: Measured Impedance Performance

This section will present results for the measured impedance performance at broadside and E-/H-plane $\theta_0 = \{45, 60\}^\circ$ scans. The D-plane scans will be omitted as they are approximately the average of the E-/H-plane scans.

5.5.1 Linear 1×19 Arrays

Results for the linear (1×19) arrays in Fig. 5.10 will be provided in this section. Vivaldi and SNA array results will be formatted in left and right columns, respectively.

5.5.1.1 Center Element Active VSWR vs. Frequency

The resulting center element active VSWR of the SNA array and Vivaldi array for broadside and E-/H-plane $\theta = 45/60^\circ$ scans are shown side-by-side in Figs. 5.11(a) and (b) against infinite linear array predictions (dashed purple lines) and 1×19 in-house

Figure 5.8: Infinite array scanning co-/cross-polarization levels versus frequency for a unit cell with 1W incident power at two scan angles for the proposed SNA array and its Vivaldi array counterpart. (a) D-plane ($\phi = 45^\circ$) $\theta = 45^\circ$ scan; (b) $\phi = 30^\circ$ $\theta = 60^\circ$ scan.

Figure 5.9: Polarization patterns for the single-polarized Vivaldi array (left) and SNA array (right) at the highest frequency (f_g).

DD-FEM finite array simulations. The high-end vertical marker signifies the half-wavelength grating lobe frequency (f_g) which is equivalent to the high-frequency ($f_H = f_g$) for optimal Nyquist sampling of the aperture, while the low-end vertical marker indicates the targeted 10:1 bandwidth low-frequency marker for the infinite planar array design. Despite not optimizing the initial design for performance in a linear array, both arrays yield decent active scanning VSWR and bandwidths (naturally, the VSWR values are higher and the bandwidths are lower as compared to an infinite planar array due to the lack of H-plane coupling).

The SNA array operates over a 6.3:1 (1.85-12 GHz) bandwidth with $\text{VSWR} < 2$ at broadside, whereas the Vivaldi array operates over the same bandwidth but with VSWR increasing to 2.75 around the low-end. As comparing the scanning, the SNA array achieves $\text{VSWR} < 2$ and $\text{VSWR} < 3$ out to $\theta = 45^\circ$ and $\theta = 60^\circ$, respectively, for the majority of the mid-band and high-band while the Vivaldi array yields $\text{VSWR} < 3.25$ and $\text{VSWR} < 4$. The Vivaldi array provides a more stable low-end throughout the scans. Aside from the VSWR values, a more critical aspect is the promising accuracy observed when compared to the infinite linear array and 1×19 array simulations at broadside and out to wide scan angles. *These results suggest proper design and fabrication of the array, and trust can be placed in the analysis tools.* It should also be noted that 19 elements was large enough for the center element to accurately approximate performance across the operating band for a *one-dimensional infinite linear array*, but this does not imply that 19 elements is large enough to characterize a *two-dimensional infinite planar array* across the operating band. The following section will further validate the analysis tools using an 5×19 planar array configuration.

Figure 5.10: Side-by-side views of the measured Vivaldi array (left) and SNA array (right).

Figure 5.11: Measured center element active VSWR of linear (1x19) array scans and finite/infinite array simulations. (a) Vivaldi array; (b) SNA array.

Figure 5.12: Side-by-side views of the measured 5×19 Vivaldi array (left) and SNA array (right).

5.5.2 5×19 Arrays

5.5.2.1 Center Element Active VSWR vs. Frequency

For the 5×19 arrays in Fig. 5.12, the measured and simulated results for broadside, $\theta_0 = 45^\circ$ and $\theta_0 = 60^\circ$ scans are shown in Figs. 5.13(a) and (b) in side-by-side fashion for ease of comparison with the Vivaldi and SNA results in the left and right columns, and the E-plane and H-plane scans in the top and bottom rows, respectively.

Examining the E-plane results in the top row first, the finite array simulations and measurements are in good agreement for both array cases. Infinite array results also show good correlation at midband and higher frequencies as the array seems electrically large and is less affected by truncation effects. Additionally, the low-frequency band-edge has shifted further downwards as compared to the linear array case to approximately 1.5 GHz (8:1) due to the introduction of coupling in the H-plane. However, at lower frequencies and wider scan angles, finite array simulations and measurements deviate greatly from infinite array predictions due to array guided surface waves (AGSW) [135] and truncation. The SNA array exhibits an impedance spike ($\text{VSWR} \approx 5$) around 1.8 GHz in the broadside performance that diminishes with scan angle (this is shown to be a truncation effect in Appendix D), while the Vivaldi array E-plane scanning degrades immensely with $\text{VSWR} > 6$ at the low-band for $\theta_0 =$

60°. Nonetheless, the arrays are electrically small at lower frequencies and it is difficult to fully gauge their performance.

The H-plane results in the bottom row display also agree well finite array simulations. In this case, both arrays are stable with scan angle and nearly track the infinite array case except at low frequencies. The SNA array broadside spike diminishes with increasing scan angle again. Interestingly, the Vivaldi array H-plane scans have better performance than the E-plane scans even though the array is electrically larger in the E-plane (19 elements) and smaller in the H-plane (5 elements). This is further testament to the importance of finite array simulation and measurement in wideband arrays.

Although truncation effects distort the low-frequency performance, finite array simulations and measurements have repeatedly shown excellent agreement and the results begin to more closely approach infinite array simulations due to the larger array size. In areas not affected significantly by truncation, the VSWR is well-matched with broadside $VSWR < 2$ and $VSWR < 3$ out to 60° for both arrays. Moreover, it was shown that the truncated array sizes impact the Vivaldi and SNA arrays in different ways that are not easily predictable. Due to this, it is sometimes difficult to differentiate resonances caused by finite array effects from those manifesting from element design. A brief synopsis classifying the impedance resonances in these measured results is provided in Appendix D for the interested reader. Extension to 19×19 array configurations will begin to alleviate several truncation effects in the next section.

5.5.3 19×19 Arrays

This section will encompass active impedance results of the 19×19 Vivaldi and SNA arrays shown in Fig. 5.14.

Figure 5.13: Measured center element active VSWR of 5×19 array E-/H-plane scans and finite/infinite array simulations. (a) Vivaldi array; (b) SNA array.

Figure 5.14: Views of the measured 19×19 Vivaldi array (left) and SNA array (right).

5.5.3.1 Center Element Active VSWR vs. Frequency

The broadside active VSWR charts are presented in Figs. 5.15(a) and (b). All results have good agreement between measurements and finite array simulations; better agreement with infinite array simulations is observed down to mid-band out to 45° due to the larger array. Interestingly, the increase in size to 19×19 from 5×19 actually hurt the center element low-frequency behavior of the Vivaldi array at broadside scan as it now has a ripple effect below 3 GHz, while the SNA low-end is relatively flat down to its band-edge. This indicates that the low-frequency broadside spike in the 5×19 SNA array center active VSWR measurements of Fig. 5.13 was indeed a truncation effect.

At the lower frequencies, the E-plane scanning of the SNA array is better than the Vivaldi while for the H-plane the SNA array has worse VSWR than the Vivaldi. In fact, the center element H-plane low-end is very poor for both arrays and is heavily afflicted by truncation effects with this array size. At 3.8 GHz for broadside scan in the SNA array, there is a 0.4 dB difference between finite array simulation and measurement; there is also some disagreement at the high-end of the SNA array at E-plane $\theta_0 = 60^\circ$ where the measurement is better than the simulation. Possible sources of disagreement can stem from measurement faults, effects not accounted for during simulation (i.e. perturbations from the environment, assembly faults, etc.), or

Figure 5.15: Measured center element active VSWR of 19×19 array E-/H-plane scans and finite/infinite array simulations. (a) Vivaldi array; (b) SNA array.

Figure 5.16: Measured passive center element VSWR/phase of 19x19 SNA array. (a) VSWR; (b) Phase Response.

a slightly defected element(s). A spot check is performed on the passive characteristics of the center element (10,10) to yield the results in Figs 5.16(a) and (b). The passive VSWR and phase response closely track that of the simulated DD-FEM finite array solver (discrete points across the operating band as the black asterisks), confirming the center element itself is behaving as expected. To sum, both arrays operate over a 10:1 bandwidth while they are both afflicted differently by truncation effects; a larger array would be necessary in practice to help alleviate these effects. Nonetheless, the overall agreement between measurement/finite array simulation is good and suggests proper operation.

5.5.3.2 Center Element Active VSWR Scan Volume Map

Results for measured E-/H-plane scan center element active VSWR as a function of scan angle (θ_0) and frequency for both Vivaldi and SNA arrays are presented in Fig. 5.17. The Vivaldi and SNA array E-plane scan remains $\text{VSWR} < 1.5$ over much of the scan volume across the operating band, while at some areas for both it rises up to 2.3. At $\theta_0 = 60^\circ$ in the E-plane of the SNA array, a high VSWR blip occurs due to a surface wave excitation induced by the increased effective permittivity as a consequence of

the SNA topology (this can be suppressed by lowering the effective permittivity of the SNA fin region i.e. using a lower DK dielectric or a more well-designed slicing implementation). For off-broadside E-plane scans near 12 GHz, the Vivaldi array suffers from very high VSWR due to the aforementioned scan blindnesses caused by deep cavities in single-polarized Vivaldi arrays [134]. The SNA array H-plane scanning has some arbitrary VSWR bumps of approximately 3.5 at ($\theta_0=35^\circ$, $f=4.5$ GHz) and ($\theta_0=50^\circ$, $f=11.25$ GHz), and the Vivaldi array has some VSWR=3 points around 3-4 GHz at wide scan angles. The Vivaldi array has poor low-frequency VSWR at 1.5 GHz for broadside and H-Plane scans, while the SNA array does not have a broadside issue but instead only suffers after $\theta_0=25^\circ$ in H-plane scans below 2 GHz.

5.5.3.3 Active VSWR Element Distribution

Broadside, H-plane $\theta_0 = 45^\circ$, and E-plane $\theta_0 = 45^\circ$ scan active VSWR element distributions are respectively charted in Figs. 5.18, 5.19, and 5.20 for $f = \{1.2, 4, 12\}$ GHz. These frequencies were chosen to demonstrate a variety of low-/mid-/high-band performance. A better perspective of the standing waves on the array faces can be observed that manifest from surface waves launched along the array aperture (the AGSW) due to reflection from the edges and subsequent modification of the element currents, leading to harmful variations in the input impedance [135].

At the low-frequency ($f = 1.2$ GHz) in the broadside scan cases for both arrays of Fig. 5.18, the AGSW form symmetric standing wave patterns across the aperture and also manifest themselves differently in each array due to the differences in topologies. As frequency increases, the severity of the truncation effects lessens as can be observed at $f = 4$ GHz, while some periodic oscillation in the VSWR is still seen across the element distribution. The high-band element VSWR levels approximate an infinite array as the array becomes electrically large (each card is close to 20λ long, or $A \approx 400\lambda^2$) and the standing waves oscillate fast and less severely, having almost a

Figure 5.17: Measured center element active VSWR of 19×19 Vivaldi and SNA arrays versus elevation scan angle and frequency. (a) Vivaldi E-Plane; (b) SNA E-plane; (c) Vivaldi H-Plane; (d) SNA H-plane.

negligible effect on element input impedance. Similar conclusions are drawn from the scanning cases in Figs. 5.20 and 5.19 except now the off-broadside scanning causes an asymmetric disruption in impedance performance across the array face. As repeatedly seen, the array size needs to be much larger (which is a realistic occurrence for many UWB-ESA applications) to suppress the severity of truncation effects at lower frequencies and provide more well-behaved element impedance. Nonetheless, an understanding of the behavior of the 19×19 arrays has been provided, and the measured center element results track closely with the finite array simulations to verify proper design.

5.6 Results: Measured Far-Field Performance

The far-field performance results of the SNA and Vivaldi arrays will now be presented. It will again begin with the simplest case (a 1×19 card) and evaluate performance to gain trust in our measurement procedure and simulations, and then follow with measurements of the 5×19 and 19×19 array configurations. Measurements were only taken in the high-band (8-12 GHz) for the linear and 5×19 arrays as the worst cross-polarization issues are in the high-band. The 19×19 array far-field results cover more of the operating band down to 2 GHz.

5.6.1 Linear 1×19 Arrays

5.6.1.1 Center Embedded Element Far-Field Pattern

Center embedded element far-field measurements at 8 GHz, 10 GHz, and 12 GHz were conducted for both linear arrays against equivalent finite array DD-FEM simulations. The results of the Vivaldi array and proposed SNA array are presented respectively in columns in Figs. 5.21(a) and (b). Both measurements have good correlation with the finite array predictions; some discrepancies in amplitude around the pattern dips are most likely due to measurement alignment issues and slight card

Figure 5.18: Simulated broadside scan active VSWR element distribution maps for the 19x19 Vivaldi (left column) and SNA (right column) arrays.

Figure 5.19: Simulated H-plane $\theta=45^\circ$ scan active VSWR element distribution maps for the 19×19 Vivaldi (left column) and SNA (right column) arrays.

Figure 5.20: Simulated E-plane $\theta=45^\circ$ scan active VSWR element distribution maps for the 19 $\times$ 19 Vivaldi (left column) and SNA (right column) arrays.

bending causing asymmetries. Aside from this, the SNA array is seen to provide superior polarization performance out to wider angles due to its broader beamwidth. As a measure of reference at 8 GHz, the SNA array co-polarization (co-pol.) becomes equivalent to the cross-polarization (cx-pol.) at approximately $\theta = 50^\circ$, whereas the Vivaldi array experiences this at $\theta = 35^\circ$. With increasing frequency at 10 GHz and 12 GHz, the Vivaldi array patterns become more directive and the cross-polarization increases more rapidly with elevation scan angle to result in the co-pol. and cx-pol. being equivalent at approximately $\theta = 30^\circ$ and $\theta = 25^\circ$, respectively. The SNA array, however, retains a broad beamwidth and pushes this characteristic out to $\theta \geq 45^\circ$ for both cases.

E-plane measurements were also taken at the same frequencies in Figs. 5.22(a) and (b). Results suggest very good agreement between measurement and simulation of the co-pol. patterns. However, there are large variations in the cx-pol. patterns. This is due to the fact that the principal plane cx-pol. is > 35 dB beneath the co-pol. and the measured signal becomes buried in the chamber noise floor around -30 dB. Some alignment error may also have an impact. Moreover, the co-pol. patterns show good agreement with low principal plane cx-pol. for both arrays. Both arrays overall behave as expected and the SNA array provides a broader pattern beamwidth to improve the polarization performance at wider angles in the D-plane. Good agreement in measurements and simulations suggests proper operation of both arrays and confidence in simulation analysis tools.

5.6.2 5×19 Arrays

5.6.2.1 Center Embedded Element Far-Field Pattern

The 5×19 array D-plane patterns are shown in Figs. 5.23(a) and (b), where both patterns are broadened due to the H-plane coupling and the polarization performances are improved. For $f = 8$ GHz at $\theta = 45^\circ$, the Vivaldi array co-pol. and cx-pol. become

(a)

(b)

(a)

(b)

(a)

(b)

Figure 5.21: Center embedded element far-field co-/cross-polarization D-plane measurements and simulated finite array DD-FEM solver results at various frequencies for linear array (1×19) configurations. (a) Vivaldi array; (b) SNA array.

(a)

(b)

(a)

(b)

(a)

(b)

Figure 5.22: Center embedded element far-field co-/cross-polarization E-plane measurements and simulated finite array DD-FEM solver results at various frequencies for linear array (1×19) configurations. (a) Vivaldi array; (b) SNA array.

equal (0 dB CPR) while the SNA provides a -15 dB CPR. In fact, the SNA array co-pol. remains above the cx-pol. across the entire band at this frequency. In the case of $f = 10$ GHz, the Vivaldi array co-/cx-pol becomes equal at around $\theta = 35^\circ$ in this instance while the SNA array hits this mark at $\theta = 60^\circ$, further cementing the enhanced polarization performance at wide angles. At $\theta = -45^\circ$, the SNA array and Vivaldi CPRs are respectively -7.5 dB and $+3.5$ dB, netting an improvement of -11 dB in the SNA array. Lastly, similar improvement is carried over for the last row of $f = 12$ GHz results, where the Vivaldi array actually exhibits the worst deterioration around $\theta = 37.5^\circ$ with a $+4$ dB CPR and the SNA array offers a -8 dB CPR at the same angle for a 12 dB improvement.

A few points of interest can be observed from these charts. Firstly, the SNA array co-pol. broadside gains exhibit a 2.5 dB dip at 8 GHz and 2 dB dip at 10 GHz, while the Vivaldi array broadside gains remain the maximum. Although this may appear alarming, this is a truncation effect due to the non-square array dimensions that will be validated in the next section for 19×19 array measurements. Secondly, there are discrepancies at wide-angles in the 12 GHz SNA array results that are expected to be caused from (i) card bending having more pronounced effects at higher frequencies and/or (i) the supporting blue foam that rotates the array for the D-plane measurement perturbing results at wide angles. Nonetheless, the SNA array demonstrates significant improvement in polarization performance over its Vivaldi array counterpart in 5×19 array configurations.

Measured patterns of the principal E-/H-planes are also provided in Figs. 5.24(a) and (b) and Figs. 5.25(a) and (b), respectively, for both arrays. The general conclusion from analyzing both sets of charts is there is once again good agreement in all co-polarized gain patterns, while the cross-polarized gains are below the chamber noise floor to acquire; the measured cx-pol. remains well-below that of the co-pol curves (>25 dB in most cases). An interesting phenomenon occurs in the co-polarized

Figure 5.23: Center embedded element far-field co-/cross-polarization D-plane measurements and simulated finite array DD-FEM solver results at various frequencies for 5×19 configurations. (a) Vivaldi array; (b) SNA array.

Figure 5.24: Center embedded element far-field co-/cross-polarization E-plane measurements and simulated finite array DD-FEM solver results at various frequencies for 5×19 configurations. (a) Vivaldi array; (b) SNA array.

Figure 5.25: Center embedded element far-field co-/cross-polarization H-plane measurements and simulated finite array DD-FEM solver results at various frequencies for 5×19 configurations. (a) Vivaldi array; (b) SNA array.

pattern of the Vivaldi array for $f = 12$ GHz between $15^\circ \leq \theta \leq 20^\circ$, where significant dips in the pattern are visible. As mentioned beforehand, these dips result from a class of E-plane scan blindnesses in single-polarized Vivaldi arrays that are associated with waveguide cavity-like modes causing harmful impedance resonances off-broadside [134].

5.6.3 19×19 Arrays

5.6.3.1 Center Embedded Element Far-Field Pattern

Measurements were taken of the 19×19 arrays across a 2-12 GHz operating band. As cross-polarization issues are most severe in the high-band, patterns for only one low-band frequency ($f = 3$ GHz) are provided for brevity in Fig. 5.26. Good correlation between co-polarized measurements is observed with the similar noise floor issue experienced for principal-plane cross-polarization. Even at such a low frequency, the Vivaldi array D-plane cross-polarization is equivalent to the co-polarization at $\theta \approx 60^\circ$, whereas the SNA array attains about -11 dB CPR at $\theta = 60^\circ$, -15 dB CPR at $\theta = 45^\circ$, and, plainly, the CPR remains less than 0 dB across all observation angles. Overall, the results agree well with measurements. A complete set of measurements across the operating band is provided in Appendix E.

Finite array simulations of the 19×19 center embedded element patterns for the D-/H-/E-planes are shown in Figs. [5.27, 5.28, 5.29](a-c) for 8 GHz, 10 GHz, and 12 GHz. The agreement in the co-polarized patterns for each plane and frequency is good except for some minor deviation at wide angles towards grazing. At wider angles in the D-plane, good accuracy is observed overall out to 90° with about 3-4 dB of deviation between $45^\circ \leq \theta \leq 67.5^\circ$ at $f = 12$ GHz. Critically, the SNA array offers significant improvement out to at least $\theta = 60^\circ$ at these frequencies over its Vivaldi counterpart. For instance, the co-polarization becomes equivalent to the cross-polarization (CPR = 0 dB) for the Vivaldi array at $\theta \approx \{40, 35, 30\}^\circ$ respectively for $f = \{8, 10, 12\}$ GHz,

Figure 5.26: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 3$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

whereas the SNA array respectively offers a $\text{CPR} = \{-10, -10.5, -16\}$ dB for those same observation angles and frequencies. Additionally, the 19×19 array eliminates the broadside co-pol. gain dip previously observed in the 5×19 SNA array configuration (specifically in the D-plane).

5.6.3.2 Center Embedded Element D-Plane vs. Frequency

The simulated center element D-plane CPR as a function of frequency vs. observation angle is presented in Fig. 5.30(a) and (b) for the 19×19 Vivaldi and SNA arrays, respectively. Horizontal dashed markers at $\theta = \pm 45^\circ$ are included to provide a common benchmark to evaluate trends. The most glaring trend can be seen in the Vivaldi array (top) as the problematic CPR (high values of red) continually creeps inwards towards angles nearer to broadside with increasing frequency, and eventually crosses that $\theta = \pm 45^\circ$ threshold with $\text{CPR} = 0$ dB as early as around 5 GHz (below midband). On the other hand, the SNA array (bottom) alleviates much of the high CPR throughout the operating band out to $\theta = \pm 45^\circ$ and into $\theta = \pm 60^\circ$. The D-plane CPR takes on a more constant trend as a function of frequency and scan angle, similar to that of a low-profile array, and never crosses the $\theta = \pm 45^\circ$ threshold with $\text{CPR} = 0$ dB. The low-frequency (1.2 GHz) of the SNA array exhibits high CPR at wide angles due to high impedance mismatch in the central elements stemming from AGSW (as seen previously in Fig. 5.18, where the Vivaldi array central elements avoid this issue for the 19×19 array size).

The same charts were produced for center element D-plane measurements across a 2-12 GHz operating band for the 19×19 Vivaldi and SNA arrays respectively in Fig. 5.31 and Fig. 5.32. In each plot, the top is the same simulated data from Fig. 5.30 and the bottom is the measured data. Overall, there is very good agreement between measurement and simulation for each case that signify the superior polarization behavior of the SNA array.

Figure 5.27: Center embedded element far-field co-/cross-polarization D-plane measurements and simulated finite array DD-FEM solver results at various frequencies for 19×19 configurations. (a) Vivaldi array; (b) SNA array.

Figure 5.28: Center embedded element far-field co-/cross-polarization H-plane measurements and simulated finite array DD-FEM solver results at various frequencies for 19×19 configurations. (a) Vivaldi array; (b) SNA array.

Figure 5.29: Center embedded element far-field co-/cross-polarization E-plane measurements and simulated finite array DD-FEM solver results at various frequencies for 19 $\times$ 19 configurations. (a) Vivaldi array; (b) SNA array.

(a)

(b)

Figure 5.30: Simulated 19×19 array center embedded element D-Plane CPR as a function of frequency and observation angle over the 10:1 operating band. (a) Vivaldi array; (b) SNA array.

Figure 5.31: 19×19 SNA array center embedded element D-Plane CPR as a function of frequency and observation angle over a 2-12 GHz band. (a) Simulated (b) Measured.

Figure 5.32: 19×19 Vivaldi array center embedded element D-Plane CPR as a function of frequency and observation angle over a 2-12 GHz band. (a) Simulated (b) Measured.

5.6.3.3 Simulated Scan Gain vs. Frequency

The preceding measurements involved normalized far-field data of the center embedded element pattern and provided no indication of the true gain of the actively scanned array. Active scan measurements are outside the reach of this work due to a limited resource pool, so only finite array simulations were conducted. This is the same solver used throughout this chapter that has shown to be robust and reliable in terms of impedance and far-field performance.

When a single-polarized array scans at (θ_0, ϕ_0) , the realized co-polarized scan gain (G_{co}) is approximately

$$G_{co}(f, \theta_0, \phi_0) \approx \frac{4\pi A}{\lambda^2} \cos(\theta_0) \left(1 - |\Gamma_{avg,11}^a(f, \theta_0, \phi_0)|^2\right) p(f, \theta_0, \phi_0) \epsilon_{cd} \quad (5.2)$$

where A is the aperture area, λ is the operating wavelength, ϵ_{cd} are conductor/dielectric losses, p is polarization efficiency, and $\Gamma_{avg,11}^a$ is the average active reflection coefficient (VSWR) of all elements in M rows and N columns (effectively the return loss seen by the entire radiating aperture) written as

$$\Gamma_{avg,11}^a = \frac{1}{MN} \sum_{m=1}^M \sum_{n=1}^N \Gamma_{m,n,11}^a(f, \theta_0, \phi_0). \quad (5.3)$$

For the following results, ϵ_{cd} is assumed to be unity (PEC conductors, lossless dielectric), but the associated loss was estimated through simulation to be 0.058 dB and 0.048 dB for the SNA and Vivaldi arrays, respectively; the elements use only low-loss dielectric material (Rogers Duroid, $\tan \delta = 0.0009$) and high-conductivity copper traces.

The simulated 19×19 array broadside co-/cross-polarized main beam scan gains are first charted in Figs. 5.33(a) and (b), where (b) is zoomed version with different trace colors of (a) to provide a better perspective on the co-polarized gain. The dotted black line indicates the directivity achievable by a planar array, and f_L indicates the

Figure 5.33: Simulated 19x19 Vivaldi and SNA array co-/cross-polarized gains for broadside scan. (a) Broadside scan; (b) Broadside scan with zoomed view on the co-polarization.

10:1 low-frequency band-edge and f_g is the high-frequency. Both arrays exhibit very low cross-pol. (>50 dB down) across the band and minimal losses. The SNA array experiences a maximum of 0.45 dB drop in co-polarized gain from the directivity while the Vivaldi has a worst-case drop of 0.48 dB at f_L . The Vivaldi array has 0.24 dB higher gain than the SNA array at f_g . The critical points are each actively scanned array exhibits very low loss due to mismatch, and the co-polarized gain curves are within about 0.25 dB of one another throughout the operating band. It is also interesting to note that despite the high losses exhibited by truncation effects in certain elements seen beforehand, the scan gain still meets the 10:1 low-frequency band-edge indicated by f_L . This is due to the average aperture return-loss, $\Gamma_{avg,11}^a$, still being relatively low. Naturally, practical implementations would need to account for such high return losses exhibited by certain elements to prevent damage to backing hardware.

Scan gains for E-/H-plane $\theta_0 = \{45, 60\}^\circ$ scans are presented in Figs. 5.34(a-d) with explicitly co-polarized gain focused charts in Figs. 5.35(a-d). Similar conclusions are drawn in this case—the cross-polarization remains very low out $\theta_0 = 60^\circ$ (better

than 35 dB down) and there are very low losses throughout the majority of the operating band. The largest drop in gain occurs for both arrays at f_g for E-plane $\theta_0 = 60^\circ$ scan cases where the Vivaldi and SNA arrays respectively drop by 2.66 and 5.57 dB. This is a combination of truncation effects, worse impedance matching at wide scan angles, and, in the case of the SNA array, an aforementioned surface wave excitation. The SNA array is generally better matched for E-plane scans aside for at f_g , and the arrays are approximately equivalent in the H-plane scans. Oscillations about the directivity are once again caused by the finite array size, ground plane, and asymmetries enforced by the progressive phase excitations across the aperture. In general, both arrays exhibit similar performance throughout the operating band and are still fairly efficient given their electrically small array size.

Non-principal plane scans towards the $\phi_0 = \{30, 45\}^\circ$, $\theta_0 = \{45, 60\}^\circ$ directions are presented in 5.36(a-d). The $\phi = 30^\circ$ case is considered as single-polarized arrays supporting substantial longitudinal currents force the worst cross-polarization properties to begin to shift from the D-plane ($\phi_0 = 45^\circ$) towards the H-plane ($\phi_0 = 0^\circ$) as discussed in Chapter 2. All blue curves indicate co-polarized gains, red curves indicate cross-polarized gains, and the SNA/Vivaldi arrays are respectively indicated by the solid/dashed curves. For every case, the solid blue curve of the SNA array stays well-above that of the dashed blue curve of the Vivaldi array due to its improved polarization properties, especially in the mid-/high-bands. In the top-left case of the D-plane $\theta_0 = 45^\circ$ scan, cross-polarization levels of the Vivaldi array dramatically rise with increasing frequency and become equal to the co-polarized gain at 6.4 GHz—resulting in about a 3 dB drop from the directivity—and continue to drop for the rest of the band. Similar trends are observed for $\phi_0 = 30^\circ$, $\theta_0 = 45^\circ$ and D-plane $\theta_0 = 60^\circ$ scans in Fig. 5.36(b) and (c), respectively. Special notice is given to the $\phi_0 = 30^\circ$, $\theta_0 = 60^\circ$ scan case of Fig. 5.36(d) where the absolute worst polarization properties of the Vivaldi array are observed. At 6.8 and 9.2 GHz, the co-polarized

Figure 5.34: Simulated 19×19 Vivaldi and SNA array co-/cross-polarized gains for various principal plane scans. (a) E-plane $\theta = 45^\circ$; (b) H-plane $\theta = 45^\circ$; (c) E-plane $\theta = 60^\circ$; (d) H-plane $\theta = 60^\circ$.

Figure 5.35: Simulated 19 × 19 Vivaldi and SNA array co-polarized gain for various principal plane scans. (a) E-plane $\theta = 45^\circ$; (b) H-plane $\theta = 45^\circ$; (c) E-plane $\theta = 60^\circ$; (d) H-plane $\theta = 60^\circ$.

Vivaldi array scan gain drops by 19.4 and 28 dB, respectively, due to significantly deteriorated cross-polarization levels. At these same frequencies, the SNA array experiences 1.74 and 1.05 dB of loss—an improvement of 17.66 and 26.95 dB—and a maximum of 2 dB of loss across the entire operating band. All in all, the SNA array provides significant polarization improvements in the non-principal scan planes over its Vivaldi array counterpart and comparatively sacrifices little-to-no scan loss (sometimes improved) in the principal scan planes.

5.7 Conclusions

This chapter presented the design, fabrication, and measurement of the 10:1 SNA array and Vivaldi array prototypes in 1×19 , 5×19 , and 19×19 single-polarized configurations. Finite array simulations yielded excellent agreement for both measured center embedded element VSWR and far-field patterns in all cases. The SNA array was shown to generally operate as good as its Vivaldi array counterpart with low VSWR levels over a 10:1 band out to $\theta = 60^\circ$ above mid-band. Truncation effects distorted the low-frequency behavior of both arrays in different manners due to the electrically small sizes of the arrays and their different topologies; a larger array would be needed for the central elements to approximate infinite array performance over the entire 10:1 operating band. In an electrically large array, the central elements are capable of achieving $\text{VSWR} < 2.5$ out to $\theta_0 = 60^\circ$ scans.

In regards to far-field polarization behavior, the SNA array was shown to yield significant improvements over the Vivaldi array in the non-principal planes. The SNA array offers >15 dB improvements in D-plane CPR out to wide angles at the high frequency and the CPR remains fairly constant versus frequency; this is in contrast to the Vivaldi array where D-plane CPR levels upwards of 3 dB manifest in the high-band. Active scan simulations suggest the SNA array yields a maximum of 2 dB of co-polarized gain loss in the worst scanning plane within a $\theta_0 \leq 60^\circ$ cone

Figure 5.36: Simulated 19×19 Vivaldi and SNA array co-/cross-polarized gains for various non-principal plane scans. (a) $\theta_0 = 45^\circ$, $\phi_0 = 45^\circ$; (b) $\theta_0 = 45^\circ$, $\phi_0 = 30^\circ$; (c) $\theta_0 = 60^\circ$, $\phi_0 = 45^\circ$; (d) $\theta_0 = 60^\circ$, $\phi_0 = 30^\circ$.

while the Vivaldi array exhibits up to 28 dB of loss at its worst operating point. In comparison with its Vivaldi counterpart, the proposed SNA array retains the same fabrication complexity and sacrifices little-to-no scan loss (sometimes improved) in the principal scan planes over the entire 10:1 operating band. Ultimately, the desirable UWB wide-scan impedance performance of Vivaldi arrays was retained over a 10:1 bandwidth while simultaneously delivering significant polarization improvements in the non-principal scan planes.

CHAPTER 6

EPILOGUE

This dissertation presented a new class of ultra-wideband (UWB) electronically scanned arrays (ESAs), the Sliced Notch Antenna (SNA) array, that offers very wide bandwidth (upwards of 10:1), excellent wide-scan impedance matching and close to optimal polarization radiation characteristics, while maintaining high efficiency (low-loss) and optimal aperture sampling ($f_H = f_g$). Moreover, the developed methods, models, theoretical results and insights propose new directions in UWB-ESA design that challenge the widely-held view in the area that good polarization properties are exclusively a byproduct of low profile; provided theory and design guidelines enable the complete and efficient design methodology of such arrays. Theoretical, simulation and measurement results on exemplary SNA arrays are all in good agreement, validating the proposed array concept and theoretical developments as well as showcasing its benefits over a state-of-the-art Vivaldi array.

6.1 Conclusions

The core of this research hinges upon breaking the common misconception that the profile of an array element is directly responsible for producing high cross-polarization levels in the non-principal scan planes. After deeper scrutiny of the radiation by models that support vertical and horizontal currents in an infinite array unit cell, it was shown that the vectorial superposition of these vertical and horizontal currents propagating along the Vivaldi fin is responsible for those high cross-polarized fields, and not merely the high-profile of the elements. Simple near-to-far field transformations

followed by the co- and cross-polarization transforms show that even a high-profile array is capable of producing low cross-polarization as long as the ratio of vertical-to-horizontal currents is appropriately controlled to effectively produce an element with a broader co-polarized far-field pattern. This requirement can be elegantly satisfied by disconnecting (slicing) the tapered fin of a Vivaldi element. It was found through both a newly developed theoretical model and via numerical simulations that these slices must maintain a very strong capacitive coupling between one another in order to maintain the good impedance behavior of the original (un-sliced) Vivaldi array. This very requirement was found to be in conflict with the polarization findings, as maintaining good polarization requires larger gaps between slices. A compromise for this trade-off was found to be an SNA structure with a large number of slices separated by small gaps.

A very surprising finding was that the sliced regions between orthogonal or coplanar fins *do not* require electrical connection in dual-polarized configurations. The theory behind this development was not presented, but it can be justified using effective dielectric properties of periodic structures that are not near resonance. The practical benefit of this finding was that the proposed SNA array helps alleviate some of these issues in conventional dual-polarized, dual-offset Vivaldi array configurations by providing electrical disconnection between adjacent elements in the modified fin region.

6.2 Contributions

The technological contributions of this dissertation are summarized here:

- To the best of our knowledge, the proposed SNA array is the first 10:1, low-loss array with low VSWR and low cross-polarization over the operating band out to $\theta_0 = 60^\circ$ scans.

- The proposed SNA array can achieve cross-polarization levels equivalent to those of a low-profile array (that are up to $5\times$ shorter).
- The proposed SNA array can be designed easily by first using well-known guidelines from Vivaldi arrays and then adapting the design with the work here to tune for lower cross-polarization.
- Since the SNA concept is based on the Vivaldi array, it too can be implemented in many different embodiments (i.e. PCB, planar, stack-metal, hybrid, etc.) to meet diverse mechanical and system design requirements.
- The proposed SNA arrays optimally sample the array aperture, i.e. $f_H = f_g$, and thus no extra T/R modules are needed.
- More importantly the SNA array is backwards-compatible with Vivaldi UWB-ESA systems. Vivaldi arrays remain the most popular wideband element in UWB-ESA systems but could be coming to their end-of-life cycle in some systems. The proposed SNA array is comprised of the core of Vivaldi arrays with a modified fin section, and could be a suitable low cross-polarization candidate.

Additionally, the theoretical contributions in this dissertation are summarized below:

- This work shed light into the true nature of cross-polarization in Vivaldi arrays.
- A theoretical model was proposed to explain the polarization behavior of both Vivaldi and SNA arrays.
- An approximate circuit model for SNA array impedance behavior was developed.
- Design guidelines for SNA array element tuning were given.

6.3 Future Work

This dissertation has merely opened the door to the theory and design of SNA arrays and various research opportunities are available; a brief summary of some of those topics are discussed here.

Intrinsic Cross-Polarization Ratio Impacts: The concept of intrinsic cross-polarization ratio (IXR) in dual-polarized phased arrays is fairly new and its influence has yet to be fully applied in depth by the majority of the phased array design community. IXR is known to be proportional to the orthogonality between channels and the resulting measurement precision post-calibration in sensing systems, but the physical connection to the radiating aperture is still not well understood. For instance, it was observed in Chapter 3 that dual-polarized Vivaldi arrays at some wide angles e.g. $\phi = 45^\circ$, $\theta > 45^\circ$ have high CPR (poor polarization), but they can still show satisfactory IXR in certain scan areas; this implies that at these scan directions the poor CPR can be effectively calibrated with less error. The question that naturally arises from these two seemingly contradicting FOM results is: what are the implications on system FOMs, e.g. sensitivity, side-lobe levels, and other effects that IXR and CPR ignores. Does calibrating an initially highly cross-polarized beam with some good IXR have any other effects on radiation as compared to a calibration of an initially highly co-polarized beam with that same good IXR?

Scanning Equivalent Circuit Models and Design Guidelines: In comparison to full-wave simulations, circuit models offer intuitive physical insight and faster optimization. The single-mode circuit theory developed here accomplished that for broadside scan, but did not incorporate off-broadside scanning. It would be beneficial to develop a scan model for deeper insight into the SNA array operation. Additionally, the circuit model needs to be adapted for SNA arrays with vertical ‘cuts’ and dual-polarized configurations. The circuit model can also be extended to incorporate realistic feeding structures such as an integrated Knorr balun for a more direct comparison with

conventional PCB-based Vivaldi elements. Furthermore, the SNA array topology can induce the onset of surface wave excitations at wide scan angles, but the full theory and design guidelines characterizing these modes was not fully developed; developing more strict design guidelines for wide angle scanning would be very useful.

Parametric Studies: Only a few parametric studies were conducted in this research to touch upon the trade-offs between impedance and polarization performance as a function of a basic slicing implementation i.e. periodically-spaced uniform gap slices. There are several implementations in which the slicing could be applied, and there may be some optimized periodicities and gap sizes for tuning. More rigorous parametric studies and design guidelines would have useful impact for optimal tuning.

Dual-Polarized Array Theory: The developed theory dealt exclusively with single-polarized arrays. It is noted in this research and in several other works that dual-polarized arrays intrinsically exhibit better polarization properties than single-polarized arrays. It would be useful to develop the theory as to the precise reasoning of this.

Frequency Scalability: The importance of frequency scalability is growing with the emergence of *mm*-wave band applications. Tightly-coupled arrays rely upon small gaps to synthesize high amounts of inter-coupling for low-frequency bandwidth that make scaling to higher frequencies particularly challenging. Relative to these arrays, SNA arrays are less reliant upon gap sizes since the primary mechanic for low-frequency bandwidth is the element profile, and furthermore can be designed in certain embodiments with thicker elements (e.g. stock-metal hybrids) to enable even wider gap sizes. Investigations into SNA array designs pushing high-frequency limits with practical fabrication techniques would provide interesting studies.

Other Design Embodiments: The SNA arrays presented here were implemented in PCB-based designs and hybrid fabrication with stock-metal. The SNA topology could perhaps lend itself to more suitable manufacturing methods with other fabrication techniques or creative implementations.

APPENDIX A

POLARIZATION CORRECTION ON DUAL-POLARIZED VIVALDI ARRAYS

The high off-axis non-principal plane cross-polarization scan levels in Vivaldi arrays has inspired research into external correctional efforts. A correctional procedure incorporating appropriate calibration and excitation re-weighted feeding schemes has been investigated to create cross-polarization nulls at desired scan angles and frequencies for dual-polarized Vivaldi array configurations [69]. This is a less general case of the calibration procedures that have been used in passive sensing systems (and optics) employing dual-polarized polarimeters [29, 30] (as discussed in Section 1.3.2.2). Consider for a given scan angle and frequency, the field components of each orthogonal and independent horizontal and vertical port ($E_{x,H}^0, E_{y,H}^0$ and $E_{x,V}^0, E_{y,V}^0$) can be obtained through toggling the initial complex voltage excitations (a and b). Through this, a transfer matrix (Jones matrix, $\mathbf{J}$) is found and inverted that can determine the desired radiated fields (E_x^{goal}, E_y^{goal}) and the necessary calibrated excitations (a^{goal}, b^{goal}) by

$$\begin{bmatrix} E_{x,H}^0 & E_{x,V}^0 \\ E_{y,H}^0 & E_{y,V}^0 \end{bmatrix}^{-1} \begin{Bmatrix} E_x^{goal} \\ E_y^{goal} \end{Bmatrix} = \begin{Bmatrix} a^{goal} \\ b^{goal} \end{Bmatrix}. \quad (\text{A.1})$$

The goal of the correctional procedures is to transform an initial pair of beamforming weights (a and b) suffering from polarization purity to a calibrated beam pair (a^{goal} and b^{goal}) at the same scan direction. As discussed in Sections 1.2 and 1.3.2.2, inaccuracies in the final calibration can arise from system noise or tolerancing issues due to poor CPR and channels with a poor degree of orthogonality.

A top-level representation of this procedure is shown in Fig. A.1(a), where attenuators and phasors are used to generate the appropriate correction/calibration coefficients independently for a dual-polarized Vivaldi element. An example correction of a D-plane $\theta_0 = 45^\circ$ degree scan at 8.4 GHz is demonstrated in Fig. A.1(b) for infinite array configurations of a Vivaldi element operating over a 12:1 (1-12 GHz) bandwidth. The desired polarization in this instance is horizontal linear and Ludwig's 3rd definition of cross-polarization is used [31]. The y-axis indicates the CPR ($|\mathbf{E}_{\text{cx}}|^2/|\mathbf{E}_{\text{co}}|^2$) where more negative values in dB indicate lower cross-polarization. It is apparent that without correction, the CPR steadily deteriorates with increasing frequency to points where the cross-polarization is greater than or equal to the co-polarization. When the correction is targeted at 8.4 GHz, the CPR improves significantly to achieve near 25 dB improvement. However, since the calibration procedure and uncorrected element CPR are frequency-dependent, the correction gradually becomes less beneficial away from the targeted frequency and eventually has adverse effects. For instance, if -10 dB CPR is required, the correction only meets requirements over 6-10.5 GHz (1.65:1 bandwidth) and, for lower frequencies, the CPR is worse from when there is no correction applied.

A chart of the CPR sensitivity versus D-plane scan angle is shown in Fig. A.2 for a fixed frequency ($f = 8.4$ GHz) in a corrected and uncorrected 9×9 dual-polarized Vivaldi array. The correction is localized at the target correction angle at $\theta_0 = 45^\circ$, and scan angles near broadside have been heavily degraded. Uncorrected and corrected far-field patterns are respectively shown in Fig. A.3(a) and (b), highlighting that correction is narrowband and strictly localized around the main beam. In fact, the cross-polarization levels in virtually all side lobes in the corrected case have become equal to, or even punched through, the co-polarization side lobes. This can pose serious implications like those discussed in Section 1.3.2 including false target indication, jamming vulnerabilities, channel cross-talk, etc.

Figure A.1: Polarization correctional procedures with external weighting feeding schemes in dual-polarized arrays. (a) Top-level diagram of the correction scheme; (b) Frequency response of a corrected and uncorrected infinite Vivaldi array scanned at D-plane $\theta_0 = 45^\circ$. The CPR is significantly improved locally around the 8.4 GHz correction frequency.

Figure A.2: Effect of correctional procedures on cross-polarization ratio versus elevation scan angle for a 9×9 10:1 dual-polarized Vivaldi array fixed frequency (8.4 GHz).

Figure A.3: Array far-field pattern comparison of a 9×9 dual-polarized Vivaldi array with and without correctional procedures for D-plane $\theta_0 = 45^\circ$ scans at various frequencies. (a) Uncorrected; (b) Correction at $f = 8.4$ GHz.

To sum, calibration feeding schemes provide successful correction of cross-polarization at desired scan angles. However, these correctional procedures often do not meet all Vivaldi UWB-ESA system needs as they are:

1. Inherently narrowband
2. Tolerance prone or inaccurate
3. Applicable only to dual-polarized configurations
4. Detrimental to side lobe cross-polarization levels

Additionally, it is time-consuming to create look-up-table (LUT) calibration weights for designated frequencies and scan angles over a wide band and scan volume.

In some applications requiring extremely high polarization isolation, calibration may even be necessary on a UWB-ESA that already has low cross-polarization (i.e. $\text{CPR} = -15$ dB) to further improve CPR levels to better than -30 dB (one such example is for radio telescopes i.e. SKA [136]). Still, a UWB-ESA with intrinsically low cross-polarization throughout its scanning volume offers significant correctional benefits across the operating band. Some of these advantages include:

1. Less sensitive to tolerancing issues or errors
2. Non-destructive to the CPR as a function of frequency
3. Non-destructive to side lobe co-polarization levels

A brief example touching upon this with a proposed dual-polarized SNA array is provided in Section 3.5.2.1.

APPENDIX B

CONTINUATION OF VIVALDI ARRAY ELEMENT CIRCUIT PARAMETER TABLES

This appendix provides the physical dimensions and approximate circuit parameters for thicker element models in Section 2.2.2.2.

Table B.1: Dimensions of an infinite Vivaldi array element with ten taper slot steps and $t/D_H = 0.165$ c.f. Fig. 2.7(b) and its Circuit Approximation Parameters at broadside scan

Physical Dimensions	Approximate Circuit
$Z_0 = 100\Omega$	$Z_0 = 100\Omega$
$D_E, D_H = 12.5\text{mm}$	$R_A = 377\Omega$
$2d_{cav} = 8.9\text{mm}$	$Z_{cav} = 340\Omega$
$2d_1 = 1.25\text{mm}$	$Z_1 = 96\Omega$
$2d_2 = 1.5\text{mm}$	$Z_2 = 106\Omega$
$2d_3 = 1.75\text{mm}$	$Z_3 = 114\Omega$
$2d_4 = 2.25\text{mm}$	$Z_4 = 130\Omega$
$2d_5 = 3\text{mm}$	$Z_5 = 150\Omega$
$2d_6 = 3.75\text{mm}$	$Z_6 = 170\Omega$
$2d_7 = 4.5\text{mm}$	$Z_7 = 188\Omega$
$2d_8 = 5.5\text{mm}$	$Z_8 = 213\Omega$
$2d_9 = 6.5\text{mm}$	$Z_9 = 244\Omega$
$2d_{10} = 8.5\text{mm}$	$Z_{10} = 321\Omega$
$\ell_{cav} = 9.5\text{mm}$	$\beta_{cav}\ell_{cav} = 68.4^\circ$ at 6 GHz
$\ell_n = 7.5\text{mm}$	$\beta_n\ell_n = 54^\circ$ at 6 GHz
	$L_{feed} = 0.7 \text{ nH}$
	$L_{cav} = 2 \text{ nH}$

Table B.2: Dimensions of an infinite Vivaldi array element with ten taper slot steps and $t/D_H = 0.5$ c.f. Fig. 2.7(c) and its Circuit Approximation Parameters at broad-side scan

Physical Dimensions	Approximate Circuit
$Z_0 = 100\Omega$	$Z_0 = 100\Omega$
$D_E, D_H = 12.5\text{mm}$	$R_A = 377\Omega$
$2d_{cav} = 12\text{mm}$	$Z_{cav} = 384\Omega$
$2d_1 = 3\text{mm}$	$Z_1 = 78\Omega$
$2d_2 = 3.25\text{mm}$	$Z_2 = 82\Omega$
$2d_3 = 3.75\text{mm}$	$Z_3 = 90\Omega$
$2d_4 = 4.25\text{mm}$	$Z_4 = 98\Omega$
$2d_5 = 5.25\text{mm}$	$Z_5 = 116\Omega$
$2d_6 = 6\text{mm}$	$Z_6 = 132\Omega$
$2d_7 = 6.75\text{mm}$	$Z_7 = 152\Omega$
$2d_8 = 7.5\text{mm}$	$Z_8 = 177\Omega$
$2d_9 = 8.75\text{mm}$	$Z_9 = 235\Omega$
$2d_{10} = 10\text{mm}$	$Z_{10} = 314\Omega$
$\ell_{cav} = 9.5\text{mm}$	$\beta_{cav}\ell_{cav} = 68.4^\circ$ at 6 GHz
$\ell_n = 7.5\text{mm}$	$\beta_n\ell_n = 54^\circ$ at 6 GHz
	$L_{feed} = 1 \text{ nH}$
	$L_{cav} = 0 \text{ nH}$

APPENDIX C

POLAR PATTERNS OF SINGLE-/DUAL-POL. VIVALDI ARRAYS VS. ELEMENT DEPTH & FREQUENCY

This appendix provides a collection of polarization pattern comparisons between single-/dual-polarized Vivaldi arrays at various frequencies and element aspect ratios (H/D , where H is the element profile and $D = D_E = D_H = \lambda_g/2$). Results were generated through infinite array scan simulations throughout the forward half-space at 5° steps in θ_0 and 7.5° steps in ϕ_0 . The arrays are PCB-based embodiments as presented in the structures of Chapter 3.

Figure C.1: Polarization patterns for single-polarized (left) and dual-polarized (right) PCB-based Vivaldi arrays of $H/D=1$.

Figure C.2: Polarization patterns for single-polarized (left) and dual-polarized (right) PCB-based Vivaldi arrays of $H/D=1.5$.

Figure C.3: Polarization patterns for single-polarized (left) and dual-polarized (right) PCB-based Vivaldi arrays of $H/D=2$.

Figure C.4: Polarization patterns for single-polarized (left) and dual-polarized (right) PCB-based Vivaldi arrays of $H/D=3$.

Figure C.5: Polarization patterns for single-polarized (left) and dual-polarized (right) PCB-based Vivaldi arrays of $H/D=4$.

Figure C.6: Polarization patterns for a 10:1 dual-polarized (right) PCB-based Vivaldi array at a low frequency (4 GHz).

Figure C.7: Polarization patterns for a 10:1 dual-polarized (right) PCB-based Vivaldi array above midband (8 GHz).

Figure C.8: Polarization patterns for a 10:1 dual-polarized (right) PCB-based Vivaldi array at the high frequency (12 GHz).

APPENDIX D

TRUNCATION EFFECT CLASSIFICATION IN SINGLE-POL. 5×19 VIVALDI AND SNA ARRAYS

To provide a better scope of the impedance performance throughout the scan volume and also classify causes of impedance resonances, the measured E-/H-plane scan center element VSWR as a function of scan angle (θ_0) and frequency for both Vivaldi and SNA arrays in Figs. D.1(a,c) and (b,d), respectively. Similar conclusions are drawn from Section 5.5.2.1, but several high VSWR hot spots can better be observed. The in-band problematic resonances will be addressed to classify their cause to better make sense of the measured data.

Classification of some in-band resonances from Fig. D.1:

1. **Vivaldi Array - E-plane; $f > 11.75$ GHz; Off-broadside:** - The Vivaldi array can also be seen to have high VSWR for the E-Plane scan case of Fig. D.1(a) beginning around 11.75 GHz around $\theta_0 \approx 20^\circ$ and being pushed higher in frequency with increasing scan angle. This is a class of E-plane scan blindnesses in single-polarized Vivaldi arrays first recognized by Schaubert [134] that is controlled by the size of the cavity formed between adjacent rows and the ground plane. The SNA array does not exhibit this problematic resonance within the operating band due to inherently having a smaller cavity due to its shortened contiguous metal base.
2. **Vivaldi Array - E-Plane; $1.2 < f < 7$ GHz; $\theta_0 > 30^\circ$:** This high E-plane scan VSWR was seen in the results previously in Fig. 5.13(a). This will be

Figure D.1: Measured center element active VSWR of 5×19 Vivaldi/SNA array versus elevation scan angle and frequency. (a) Vivaldi E-Plane; (b) SNA E-plane; (c) Vivaldi H-Plane; (d) SNA H-plane.

shown to be due to truncation effects upon the absence of this resonant activity in the upcoming 19×19 array results.

3. **SNA Array - E-plane; $f = 12$ GHz; $\theta_0 = 60^\circ$:** The SNA array suffers from high VSWR at the high frequency for E-plane scan at $\theta_0 = 60^\circ$ in Fig. D.1(b) due to a surface wave excitation. This was discussed in the design methodology of Chapter 4, and can be treated by reducing the effective permittivity of the substrate.
4. **SNA Array - E-/H-plane; $f = 1.8$ GHz; Broadside:** It is apparent that the SNA array has a resonance in the low-band. This was seen previously in 5.13(b,d). It will be shown in the upcoming section that this is due to truncation effects as this resonance disappears with a 19×19 array size. Additionally, a closer look can be seen of the standing wave patterns on the finite 5×19 array in the active VSWR element distribution map of Fig. D.2 for broadside/H-plane scans. A strong resonance develops around the central elements in this case, but the elements between the center/edges remain well-matched.

A more complete picture of the effect of truncation on VSWR can be observed from the simulated active VSWR element distribution maps of Figs. D.3(a) and (b). Each square represents the active VSWR level of an element in the 5×19 array. Above 6 GHz, most central elements are well-matched, while severe array guided surface waves (AGSW) begin to form detrimental standing waves at lower frequencies due to the current distribution on the truncated aperture. It is interesting to note that both arrays behave differently due to their innately different topologies, especially at the lower frequencies.

Simulated 5×19 SNA Array VSWR Distribution at $f = 1.8$ GHz

Figure D.2: Simulated 5×19 SNA array active VSWR element distribution map for broadside and H-plane scans at $f = 1.8$ GHz.

(a)

(b)

Figure D.3: Simulated active VSWR element distribution maps of 5×19 Vivaldi/SNA arrays for broadside and E-/H-plane $\theta_0 = 45^\circ$ scans and various frequencies. (a) Vivaldi array; (b) SNA array.

APPENDIX E

MEASURED 19×19 SINGLE-POL. SNA & VIVALDI ARRAY EMBEDDED ELEMENT PATTERNS

This appendix provides additional measured embedded element results for the 19×19 single-polarized SNA and Vivaldi arrays in Chapter 5.

Figure E.1: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 4$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

Figure E.2: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 5$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

Figure E.3: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 6$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

Figure E.4: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 7$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

Figure E.5: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 9$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

Figure E.6: Center embedded element far-field co-/cross-polarization E-/D-/H-plane measurements and simulated finite array DD-FEM solver results at $f = 11$ GHz for the 19×19 array configurations. (a) Vivaldi array; (b) SNA array.

APPENDIX F

CASE STUDY OF SNA FIN PERFORMANCE IMPACT ON DUAL-POLARIZED TAPERED SLOT ARRAYS

The purpose of this study is to quantify the effects of the sliced SNA region on both impedance and polarization. This is done through comparing performance of dual-polarized SNA arrays with and without the SNA fin along with a typical Vivaldi array as a reference. Infinite array 10:1 unit cell models of the reference Vivaldi array, the proposed dual-polarized SNA array from Chapter 3, and deconstructed versions of the same SNA array are shown in Fig. F.1. The maximum active scanning VSWR levels out to $\theta_0 = 60^\circ$ scans for these models are presented in Fig. F.2. Vivaldi and SNA array scanning levels remain low as presented earlier in Fig. 3.11. Removal of the SNA fin region (model #3) and its dielectric (model #4) significantly expectedly degrade impedance performance due to the lack of matching ($\text{VSWR} < 4$) granted from increased profile; inclusion of the dielectric in model #3 provides some impedance matching through the dielectric cover.

Effects on polarization are presented in Fig. F.3(a) for D-plane $\theta_0 = 45^\circ$ scan CPR levels. The Vivaldi array exhibits typical degraded behavior. The deconstructed SNA array models (#3 and #4) yield vastly improved CPR due to their inherent lower profile; their performance is relatively equivalent as the dielectric does not support additional radiating longitudinal currents. Despite a higher profile, the original SNA array (model #1) has the best polarization performance due to reasons discussed throughout Chapters 3 & 4. The same conclusion is drawn from Fig. F.3(b) for a $\theta_0 = 60^\circ$ scan. These are interesting observations as the original SNA array is nearly twice as deep as models #3 and #4 but still improves upon the polarization plainly

Figure F.1: Dual-polarized infinite array unit cell models to be analyzed. (From left to right) Vivaldi array, SNA array, SNA array with slices removed, SNA array with slices and dielectric removed.

Figure F.2: Infinite array maximum active scanning VSWR within the $\theta_0 = 60^\circ$ scan cone for Vivaldi, SNA, and dissected SNA array models of Fig. F.1.

due to its topology enabling more well-balanced vertical-to-horizontal current-length ratios. In doing so, the SNA array not only achieves essentially the same desirable impedance performance out to wide scan angles of the Vivaldi array, but also *enhances the polarization performance with a deeper profile.*

(a)

(b)

Figure F.3: Infinite array D-plane scan polarization ratio (dB) results versus frequency for Vivaldi, SNA, and dissected SNA array models of Fig. F.1. (a) $\theta_0 = 45^\circ$ scan ; (b) $\theta_0 = 60^\circ$ scan. The SNA array (model #2) achieves excellent impedance performance while improving the polarization despite an increased profile (relative to Models #3 and #4).

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