ANALYSIS AND DESIGN OF SLOTTED SUBSTRATE INTEGRATED WAVEGUIDE ANTENNA ON CYLINDRICAL SURFACE

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iv
ABSTRACT

ANALYSIS AND DESIGN OF SLOTTED SUBSTRATE INTEGRATED WAVEGUIDE ANTENNA ON CYLINDRICAL SURFACE

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This thesis presents novel traveling wave conformal shunt slot arrays on cylindrical substrate integrated waveguide (CSIW) placed on a cylindrical surface in the circumferential direction at K-band. Mutual couplings of longitudinal slots on broad wall of CSIW are formulated and active admittance formulas of the slots are derived which are used in Elliott’s design procedure to design slotted CSIW arrays. The designs are carried out at 25 GHz using non-uniform and uniform slot separations. The non-uniform slot array allocates 16 slot elements near resonance where element separations are determined to compensate different spatial phase delays of each element to increase directivity. The slot excitation amplitudes are determined by -20 dB side lobe level (SLL) Taylor distribution using aperture projection method. However, -20 dB SLL is not achieved because of large slot separations. Hence, slot lengths are optimized further to increase bandwidth for beam steering applications at the expense of increasing SLL. Three different uniform slot arrays are designed with
25 slot elements to obtain pencil beam pattern with low SLL using aperture projection method and genetic algorithm. The required slot excitation coefficients are achieved by using slots with large deviations from resonance. Both uniform and non-uniform slot arrays are fabricated using standard printed circuit board (PCB) process. The antennas are measured and about 10° beam steering in H-plane is obtained when frequency is swept between 24-26 GHz in the non-uniform slot array. Pencil beam pattern with desired -20 dB SLL is achieved at 25 GHz in the uniform slot array.

Keywords: Substrate Integrated Waveguide Antenna Array, Conformal Arrays, Slotted Waveguide Array, Traveling Wave Antenna, Slot Antenna Array.
ÖZ

SİLİNDİRİK YÜZEY ÜZERİNDE YARIKLI DALGA KILAVUZU
ANTENİNİN ANALİZ VE TASARIMI

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# TABLE OF CONTENTS

ABSTRACT .......................................................................................................................... v  
ÖZ ................................................................................................................................. vii  
ACKNOWLEDGEMENTS ............................................................................................... x  
TABLE OF CONTENTS .............................................................................................. xii  
LIST OF TABLES ........................................................................................................ xiv  
LIST OF FIGURES ...................................................................................................... xv  

## CHAPTERS

1. **INTRODUCTION** ................................................................................................. 1  
   1.1. Literature Survey on Slotted SIW Antennas .................................................... 3  
   1.2. The Traveling Wave Slot Array on CSIW Developed in This Thesis .......... 7  
   1.3. Thesis Objectives and Organisation ............................................................... 8  

2. **CHARACTERIZATION OF CSIW AND SINGLE LONGITUDINAL SLOT ON CSIW** ........................................................................................................... 11  
   2.1. Characterization of CSIW .............................................................................. 11  
   2.2. Microstrip to SIW Transition Structure.......................................................... 14  
   2.3. Characterization of Single Slot on CSIW ....................................................... 17  

3. **DESIGN PROCEDURE OF SHUNT SLOT ARRAY ON CSIW** ...................... 27  
   3.1. Derivation of the Active Admittance Formulas for the Longitudinal Slots on CSIW ................................................................................................................. 27  
   3.1.1. Modes in the CSIW ..................................................................................... 28  
   3.1.2. Active Admittance Formulation of Longitudinal Slot on the CSIW ...... 31  
      3.1.2.1. Waveguide and Slot Dimensions ...................................................... 32  
      3.1.2.2. Derivation of the First Active Admittance Formula ........................ 34  
      3.1.2.3. Derivation of the Second Active Admittance Formula ................. 43  
   3.2. Design Procedure of Slotted CSIW Array ...................................................... 59
3.3. Numerical Results ........................................................................................................ 64
  3.3.1. External Couplings of Two Slots on PEC Cylinder ................................ 66
  3.3.2. External and Internal Couplings of Two Slots on CSIW ...................... 70
4. DESIGN OF TRAVELING WAVE CSIW SLOT ARRAY ........................................ 79
  4.1. Pattern Synthesis Methods .................................................................................... 80
    4.1.1. Aperture Projection Method ........................................................................ 81
      4.1.1.1. Taylor Line Source Sampling for Planar Arrays .............................. 81
      4.1.1.2. Taylor Line Source Sampling for Slotted CSIW Array .................... 87
  4.2. Slot Array Design on CSIW with Non-Uniform Slot Separation ............... 93
    4.2.1. Fabrication and Measurements.................................................................. 101
      4.2.1.1. 1 x 16 Optimized Slot Array on CSIW .................................. 101
      4.2.1.2. 1 x 4 SIW Power Divider ................................................................ 106
      4.2.1.3. 4 x 16 Slot Array on CSIW ....................................................... 108
  4.3. Slot Array Design on CSIW with Uniform Slot Separation ..................... 113
    4.3.1. Fabrication and Measurements............................................................... 132
      4.3.1.1. 1 x 25 Slot Array on CSIW .................................................. 132
5. CONCLUSIONS AND FUTURE WORK ................................................................. 139
REFERENCES ................................................................................................................... 141
APPENDICES
  A. SYSTEMATIC CONSTRUCTION OF FIELDS IN BEND REGULAR
     WAVEGUIDES ....................................................................................................... 147
  B. FIELDS RADIATED BY APERTURES ON PEC CYLINDER ......................... 151
  C. ASYMPTOTIC FORMS OF HANKEL FUNCTION RATIOS FOR LARGE
     ORDER AND ARGUMENT .................................................................................. 155
CURRICULUM VITAE .................................................................................................... 157
LIST OF TABLES

TABLES

Table 3.1: Eigenvalues, $\varepsilon_{mn}$ and $\mu_{mn}$, of $TE_{mn}$ and $TM_{mn}$ modes for various $m$ and $n$ values. (NA: Not Applicable) .......................................................................................................................... 30

Table 3.2: Comparison of the $Y_{pq}$ values for $z_0 = 0^\circ$ with the literature [45] ............... 67

Table 3.3: Comparison of the $Y_{pq}$ values for $\phi_0 = 0^\circ$ with the literature [45]. ........... 68

Table 3.4: Comparison of the $Y_{pq}$ values for various $z_0$ and $\phi_0$ values with the literature [45] ................................................................. 69

Table 4.1: Slot offsets and lengths of 1 x 16 slot array on CSIW determined by Elliott’s design procedure ................................................................. 96

Table 4.2: Slot offsets and lengths determined after the optimization of 1 x 16 slot array on CSIW designed by Elliott’s design procedure .................. 100

Table 4.3: Comparison of the simulated efficiency values of optimized 1 x 16 slot array on CSIW .................................................................................................................. 100

Table 4.4: Simulated and measured HPBW and cross-polarization values of 1 x 16 slot array on CSIW .............................................................................. 105

Table 4.5: Simulated and measured HPBW and cross-polarization values of 4 x 16 slot array on CSIW .............................................................................. 112

Table 4.6: Slot offsets and lengths for 1 x 25 slot array on CSIW (1st design) ................. 118

Table 4.7: Slot offsets and lengths for 1 x 25 slot array on CSIW (3rd design) ............... 130

Table 4.8: Slot offsets and lengths for 1 x 25 slot array on CSIW after 6th iteration (3rd design) .............................................................................................. 131

Table 4.9: Simulated and measured HPBW and cross-polarization values of 1 x 25 slot array on CSIW .............................................................................. 136
LIST OF FIGURES

FIGURES

Figure 1.1: Slotted CSIW Array................................................................. 2
Figure 1.2: (a) Cross-sectional and (b) top views of the traveling wave slot array on CSIW................................................................. 7
Figure 2.1: (a) SIW structure. (b) Equivalent of SIW.............................. 12
Figure 2.2: Comparison of the propagation constants found by simulations and closed form expressions. ......................................................... 14
Figure 2.3: A schematic view of microstrip line to SIW transition. .......... 15
Figure 2.4: (a) Fabricated microstrip to SIW transition (L = 2 cm). (b) S_{11} and (c) S_{21} results........................................................................................................ 15
Figure 2.5: (a) Fabricated microstrip to SIW transition (L = 10 cm). (b) S_{11} and (c) S_{21} results.................................................................................................................................................. 16
Figure 2.6: TRL calibration standards: (a) Thru (b) Reflect, and (c) Line Standards. ............................................................................................................................................... 18
Figure 2.7: Circuit model of longitudinal slot on CSIW.......................... 19
Figure 2.8: A and D-parameters of the single slot on CSIW for R = 10 cm: (a) real and (b) imaginary parts......................................................... 20
Figure 2.9: B-parameters of the single slot on CSIW for R = 10 cm: (a) real and (b) imaginary parts................................................................. 21
Figure 2.10: C-parameters of the single slot on CSIW for R = 10 cm: (a) Real and (b) imaginary part of the shunt admittance........................................ 22
Figure 2.11: Characterization polynomials of the shunt slot on CSIW: (a) Ratios of the admittance to resonant conductance. (b) Normalized Resonant Conductance. (c) Normalized Resonant Length. ................................................................................. 23
Figure 2.12: (a) Fabricated single slot on microstrip to SIW transition ( \( L = 2 \, cm \)).  
(b) \( S_{11} \) and (c) \( S_{21} \) results. .................................................................................................................. 24

Figure 3.1: E-plane bend regular waveguide. .................................................................................. 29
Figure 3.2: Longitudinal slot array on bended regular waveguide. ........................................... 32
Figure 3.3: Longitudinal slot \((p^{th} \, slot)\) on bended regular waveguide. .............................. 33
Figure 3.4: Bended regular waveguide with no slot \((p^{th} \, slot \, is \, covered \, by \, conducting \, tape)\). .................................................................................................................. 37
Figure 3.5: Circuit model of the longitudinal slot on E-plane bend regular waveguide. 
........................................................................................................................................... 41
Figure 3.6: Longitudinal slot array composed of \(p^{th} \, and \, q^{th} \, slots\) on bended regular waveguide.................................................................................................................. 46
Figure 3.7: (a) Longitudinal slot array with \(q^{th} \, slot \, is \, covered \, by \, conducting \, tape \)  
(case “\(a\)”). (b) Equivalent problem. ......................................................................................... 47
Figure 3.8: (a) Longitudinal slot array composed of \(p^{th} \, and \, q^{th} \, slots\) on bended regular waveguide with \(q^{th} \, slot \, is \, excited \) (case “\(b\)”). (b) Equivalent problem. 50
Figure 3.9: Circuit model representation of a traveling wave shunt slot array. ........................... 60
Figure 3.10: Deformed integration path. .......................................................................................... 64
Figure 3.11: Two slots on cylindrical PEC. .................................................................................... 66
Figure 3.12: (a) Longitudinal slots on the same CSIW. (b) Longitudinal slots on two parallel CSIW.................................................................................................................................. 70
Figure 3.13: Circuit model of two longitudinal slots on the same CSIW. .................................... 71
Figure 3.14: (a) Magnitude and (b) phase of \(S_{11} \) \((R = 10 \, cm, D_{pq} = 5 \, mm, \, and \, z_{q} =  
0.5 \, mm)\). ................................................................................................................................. 72
Figure 3.15: (a) Magnitude and (b) phase of \(S_{11} \) \((R = 10 \, cm, z_{p} = 0.2 \, mm, \, and \, z_{q} =  
0.5 \, mm)\). ..................................................................................................................................... 73
Figure 3.16: (a) Magnitude and (b) phase of \(S_{11} \) \((R = 10 \, cm, z_{p} = -0.2 \, mm, \, and \, z_{q} =  
0.5 \, mm)\).................................................................................................................................... 74
Figure 3.17: (a) Magnitude and (b) phase of \(S_{11} \) \((R = 0.6 \, cm, z_{p} = -0.2 \, mm, \, and \, z_{q} =  
0.5 \, mm)\).................................................................................................................................... 74
Figure 3.18: Circuit model of the longitudinal slot on CSIW....................................................... 75
Figure 3.19: (a) Magnitude and (b) phase of active $S_{11}$ ($R = 10$ cm, $z_q = 0.5$ mm and $D_w = 6$ mm). ................................................................. 76

Figure 3.20: (a) Magnitude and (b) phase of active $S_{11}$ ($R = 10$ cm, $z_p = 0.2$ mm and $z_q = 0.5$ mm). ................................................................. 77

Figure 4.1: Sampling of the Taylor distribution using aperture projection method... 80

Figure 4.2: (a) Sampling Taylor line source distribution randomly with 100 elements. (b) Corresponding radiation patterns. .................................................. 82

Figure 4.3: (a) Sampling half side of the Taylor line source distribution randomly and allocating symmetric weights for the other half (Total 200 elements). (b) Corresponding radiation patterns. .......................................................... 83

Figure 4.4: (a) Sampling Taylor line source distribution randomly with 1000 elements. (b) Corresponding radiation patterns. .................................................. 83

Figure 4.5: (a) Sampling Taylor line source distribution uniformly with 11 elements. (b) Corresponding radiation patterns. .......................................................... 84

Figure 4.6: (a) Sampling Taylor line source distribution randomly with 100 elements and adjusting the sampled weights using element density. (b) Corresponding radiation patterns. .......................................................... 85

Figure 4.7: (a) Sampling Taylor line source distribution randomly with 20 elements and adjusting the sampled weights using element density. (b) Corresponding radiation patterns. .......................................................... 86

Figure 4.8: (a) Element positions of the slotted CSIW array and element positions at the projected aperture (40 element array). (b) Sampling Taylor line source distribution using projected aperture positions. (c) Corresponding radiation patterns for linear arrays. (d) Corresponding radiation patterns for cylindrical slot arrays. .......................................................... 88

Figure 4.9: (a) Radiation patterns for the slotted CSIW array using (a) -30 dB and (b) -40 dB SLL Taylor weights. .......................................................... 90

Figure 4.10: (a) Element positions of the slotted CSIW array and element positions at the projected aperture (25 element array). (b) Sampling Taylor line source distribution using projected aperture positions. (c) Corresponding radiation
patterns for linear array. (d) Corresponding radiation patterns for cylindrical slot array.

Figure 4.11: (a) Element spacing and (b) calculated patterns of $1 \times 16$ slot array on CSIW ($R = 10 \text{ cm}$) for three different slot separation schemes.

Figure 4.12: (a) $S_{11}$ and (b) $S_{21}$ results of $1 \times 16$ SIW slot array on CSIW designed by Elliott’s design procedure.

Figure 4.13: Simulated radiation patterns of $1 \times 16$ arrays (a) designed using Elliott’s design procedure and (b) optimized array.

Figure 4.14: Comparison of the accepted powers and efficiencies of two $1 \times 16$ array on CSIW ($R = 10 \text{ cm}$).

Figure 4.15: (a) Fabricated traveling wave optimized $1 \times 16$ slot array on SIW. (b) $S_{11}$ and (c) $S_{21}$ results of the optimized $1 \times 16$ slotted SIW array.

Figure 4.16: Measurement setup of $1 \times 16$ slot array on CSIW ($R = 10 \text{ cm}$).

Figure 4.17: Radiation patterns of $1 \times 16$ slot array on CSIW.

Figure 4.18: (a) Gain and (b) efficiency results of $1 \times 16$ SIW slot array.

Figure 4.19: (a) Fabricated two back to back $1 \times 4$ power dividers. (b) $S_{11}$ and (c) $S_{21}$ results of the back to back power divider structure.

Figure 4.20: Fabricated traveling wave $4 \times 16$ slot array on SIW.

Figure 4.21: (a) $S_{11}$ and (b) $S_{21}$ results of $4 \times 16$ SIW slot array.

Figure 4.22: Measurement setup of $4 \times 16$ slot array on CSIW ($R = 10 \text{ cm}$).

Figure 4.23: Radiation patterns of $4 \times 16$ slot array on CSIW.

Figure 4.24: (a) Gain and (b) efficiency results of $4 \times 16$ SIW slot array.

Figure 4.25: The function of the Elliott’s design.

Figure 4.26: Slot array on bended regular waveguide.

Figure 4.27. Phase and error graphs for $1 \times 25$ slot array on CSIW ($1^{\text{st}}$ design): (a) Deviation of slot phases from mode voltage phases ($\angle D^1_p$). (b) Deviation of slot phases from mode voltage phases in Elliott’s design ($\angle D^2_p$). (c) Error in the Elliot’s design.

Figure 4.28. Slot voltage (a) amplitudes and (b) phases for $1 \times 25$ slot array on CSIW. (c) H-plane radiation patterns ($1^{\text{st}}$ design).
Figure 4.29. Slot voltage (a) amplitudes and (b) phases for $1 \times 25$ slot array on CSIW. (c) H-plane radiation patterns ($2^{nd}$ design). (d) Error in the Elliot’s design. ................................................................. 125

Figure 4.30. Slot voltage (a) amplitudes and (b) phases for $1 \times 25$ slot array on CSIW. (c) H-plane radiation patterns ($3^{rd}$ design)...................................................... 128

Figure 4.31. Phase and error graphs for $1 \times 25$ slot array on CSIW ($3^{rd}$ design): (a) Deviation of slot phases from mode voltage phases ($\angle D_p^1$). (b) Deviation of slot phases from mode voltage phases in Elliot’s design ($\angle D_p^2$). (c) Error in the Elliot’s design. ................................................................. 129

Figure 4.32: (a) Fabricated traveling wave $1 \times 25$ slot array on SIW. (b) $S_{11}$ and (c) $S_{21}$ results of the $1 \times 25$ slotted SIW array. ................................................................. 133

Figure 4.33: Measurement setup of $1 \times 25$ slot array on CSIW ($R = 10 \text{ cm}$). ......... 133

Figure 4.34: Radiation patterns of $1 \times 25$ slot array on CSIW........................................ 135

Figure 4.35: (a) Gain and (b) efficiency results of $1 \times 25$ SIW slot array.............. 137
Conformal arrays have a specific shape determined by the parameters other than radiation pattern and input match requirements and they can easily be implemented using microstrip technology using a flexible substrate or multifaceted surfaces [1]. However, microstrip arrays can have complex feed networks such as probes feeding patch elements. Series or corporate microstrip feed networks can be used to eliminate this problem but microstrip feed networks suffer from radiation problems. Strip line can be used to eliminate the radiation losses, however, a cross coupling problems are encountered in the feed network with strip line structure. Slotted waveguide arrays (SWA) have some advantages over microstrip antennas such as being low loss, high isolation, and high power handling. Due to these advantages, SWA antennas are widely used in communication and radar systems particularly at microwave wave frequencies. Regarding conformal array applications, microstrip antennas are easier to implement compared to SWA antennas, however, excitation of antenna elements by a waveguide feed network is advantageous compared to microstrip feed network in terms of eliminating radiation losses and cross coupling problems. Moreover, complex feed network structure is not needed in SWA to excite the slots in the same waveguide. Furthermore, the array of waveguides can be formed without cross coupling problems. Recently, a new planar wave guiding structure, namely post wall waveguide or substrate integrated waveguide (SIW) is proposed [2], [3]. Although, dielectric filled SIW suffers from dielectric losses which do not exist in hollow waveguides, SIW takes most of the advantages of regular rectangular waveguides as well as the advantages of microstrip technology such as being low cost and small
size. Moreover, SIW can be easily manufactured using simple printed circuit board (PCB) process which allows monolithic integration of SIW with the other planar circuits. The integration of regular rectangular waveguides with planar feed networks requires complex and bulky integration schemes which require precise manufacturing process difficult to achieve at millimeter wave frequencies. Monolithic integration capability in SIW technology eliminates transition losses due to these complex integration of regular rectangular waveguides with planar feed networks. Due to these advantages, the slotted SIW antennas are good candidates for the conformal array applications especially when SIW is implemented using a flexible substrate.

There is a limited study in the literature on the conformal array applications with slotted waveguide arrays. The aim of this thesis is to analyze, design and implement a traveling wave slot array on a novel cylindrical substrate integrated waveguide (CSIW) structure at K-band where the CSIW lies in the circumferential direction of the cylinder as shown in Figure 1.1.

Figure 1.1: Slotted CSIW Array.
The contents of this chapter, in which the traveling wave slot array structure on CSIW is introduced, can be summarized as follows. Section 1.1 gives the overview of the slotted SIW antennas for both planar and conformal array applications in the literature while Section 1.2 presents the traveling wave slot array on CSIW developed in this thesis. Finally, Section 1.3 explains the detailed objectives and organization of the thesis.

### 1.1. Literature Survey on Slotted SIW Antennas

There has been a growing interest in SIW based antennas in the last few years. Most of the studies regarding SIW slot antennas in the literature employ planar SIW structures [4] - [23]. The first slot waveguide antenna is presented in [4] where a 4 x 4 longitudinal standing wave slot array is formed at X-band. The array is fed by a 1 x 4 microstrip power divider. The antenna is successfully implemented and measured. Using a SIW based power divider with tapered power division, a low side lobe level (SLL) longitudinal slot array on SIW is implemented in [5] and about -30 dB SLL is obtained at X-band. Slotted SIW antennas usually allocate constant spacing between slot elements. In [6], an optimization on the spacing between each element is carried out by using a combination of genetic algorithm and conjugate gradient method for a given pattern and efficiency requirements, and hence, a traveling wave SIW slot array with 10 slots is obtained with about 8% bandwidth at X-band. A series slot array is implemented with transverse slots in [7] using half mode SIW to miniaturize the SIW antennas where the SIW is cut from its centerline such that only one wall of the half mode SIW contains periodic vias while the radiation into free space takes place at the other wall. A narrow 10 dB return loss bandwidth, which is about 275 MHz, is obtained at X-band because of the standing wave array design. Another standing wave series slot array on SIW with improved bandwidth performance is presented in [8] where 16 element 45°-inclined alternating reactance slot pairs are allocated to achieve both impedance matching and uniform field excitation. About
990 MHz 10 dB return loss bandwidth is obtained with 24.3 dBi gain and 53.7% efficiency at Ka-band. There are other slotted SIW studies in the literature where beam forming networks are implemented by SIW technology and they are monolithically integrated with SIW slot arrays [9] - [12]. In [9], a standing wave planar SIW slot array is monolithically integrated with SIW 7 input rotman lens for multibeam applications at 28.5 GHz. The scan coverage is increased by conformal placement of 5 planar multibeam antennas. A 24 beam slot array antenna with 4 x 8 butler matrix is implemented with SIW technology in [10] to achieve 360° coverage over azimuth and -14.5 dB SLL is obtained in all the beam states. The antenna is constructed for multiple input multiple output (MIMO) wireless communications and 3.75% bandwidth is obtained at 16 GHz. The other slot array with 4 x 4 butler matrix implemented at 77 GHz with SIW technology combines 4 broadband couplers to avoid crossovers [11]. Another example is a compact single layer monopulse SIW slot array presented in [12] where 7.39% bandwidth with maximum gain of 18.75 dB at Ka-band is obtained. The antenna has four ports where each port is responsible for exciting the sum and difference patterns in azimuth, elevation, and diagonal planes.

There are also studies to enhance the efficiency and bandwidth of the SIW slot arrays. Dielectric resonators are allocated on top of slots in [13] to increase bandwidth and efficiency. About 11 dB gain is obtained with 1 x 4 elements over 4.7% bandwidth. In [14], the patch elements are located on top of 4 x 4 slots fed by SIW. The dielectric cubes placed on top of patch elements in [13] are replaced by Yagi-like parasitic director in [14], and hence, difficulties in the fabrication of dielectric cubes are eliminated. In [14], about 7.5 GHz impedance bandwidth is obtained at W-band with about 18 dB antenna gain.

There are some SIW leaky wave antenna examples in the literature [15] - [23]. A leaky wave antenna using continuous slot on broad wall of SIW is implemented in [15]. Position of the slot with respect to the center line of the SIW is varied along the
SIW to obtain amplitude taper, and hence, a low SLL about -30 dB is obtained at 35 GHz. The scan ranges of the continuous leaky wave antennas are limited to the forward quadrant only. In order to increase the scan range by radiating into both forward and backward quadrants, periodic and metamaterial based leaky wave antennas are employed [16] - [23]. SIW based periodic leaky wave antenna is first proposed in [16] where the leakage is obtained by increasing the spacing of the vias at one side of SIW. The same type of leaky wave antenna is implemented in [17] by rarely spacing the vias at both sides of the SIW and first higher order mode is used in the SIW structure to obtain better leakage constant compared to fundamental mode of the SIW at Ka-band. A miniaturized leaky wave antenna using half mode SIW is proposed in [18]. In [18], a quasi-omnidirectional pattern at 28 GHz is obtained with an input match below -14 dB between 24.5 GHz and 29.5 GHz. Another half mode SIW leaky wave antenna is developed in [19] to obtain 4 different polarization states at Ka-band. The antenna provides an input match better than -20 dB for all the polarization states and also it achieves about 30° beam steering by sweeping the frequency between 33 GHz and 39 GHz. A very large frequency scanning about 130° is obtained in [20] by using composite right/left handed metamaterial on broad wall of SIW when the frequency is swept between 8.5 GHz and 12 GHz where -10 dB input match is also satisfied in the band. In [20], the same antenna is also implemented using half mode SIW, and hence, the efficiency is increased to 87%. Another SIW leaky wave antenna which contains right/left handed metamaterial is presented in [21] to obtain 6 different polarization states with increased scan range compared to the one presented in [19]. A gain reduction problem near the broadside of the metamaterial based leaky wave antennas is studied in [22] and gain values above 9 dBi is obtained between 8 GHz and 13 GHz by using patch elements inserted below radiating slots in SIW. Moreover, the same structure is also used to improve the boresight bandwidth in [23] by slowing down the rapid transition from backward quadrant to forward quadrant in the metamaterial based leaky wave antennas.
In the literature, most of the studies related to conformal waveguide arrays employ regular rectangular waveguides [24] - [27] and there are limited studies with the SIW conformal arrays [9], [28], [29]. An analytic design procedure is developed in [24] for dielectric filled waveguide small slot arrays and 6 x 9 conformal waveguide slot array is implemented at 9.34 GHz. Another waveguide with edge wall slot array is designed and fabricated in [25]. A large beam steering about 84° from broadside is obtained with -13 dB side lobe level (SLL). The radar cross section of the conformal slot array in [26] is computed using method of moments (MoM) but the slot array is not implemented. All of these slotted waveguide antennas have sectoral cross section extends parallel to the cylinder axis. A conformal waveguide slot array where planar waveguides are placed around the cylinder tangent to the circumferential direction is formed at Ku band using narrow wall slots for 360° azimuth tracking in [27]; however, this antenna is multifaceted, i.e., planar slotted waveguides are used to form the conformal array. There are also conformal slot array examples in the literature using SIW, however, none of them employs SIW fully conformed around the cylinder in the circumferential direction. The conformal SIW array presented in [9] is multifaceted and planar SIW slot antennas are used to increase the scan coverage. In the other conformal SIW array application, a 4 x 4 slotted SIW array is formed using a flexible substrate and 11 dB gain is obtained for planar case at 79 GHz [28]. However, no effort has been made in the design of this antenna to place it on a prescribed conformal surface; it is fabricated only on a flexible substrate. Another SIW conformal array is presented in [29]. The array shows the design and implementation of a shaped beam slotted SIW antenna on cylindrical surface. A flat topped fan beam is obtained with about 38° half power beam width (HPBW) in E-plane at 35 GHz. However, the conformal slotted SIW array is formed by using a SIW having sectoral cross-section which extends in the axial direction of the cylinder.
1.2. The Traveling Wave Slot Array on CSIW Developed in This Thesis

The traveling wave slot array on CSIW structure examined in this thesis is shown in Figure 1.2. The array is designed to operate at K-band. The CSIW is formed by conformal placement of the planar SIW on a cylindrical surface in the circumferential direction. The periodic placement of the vias mimics the side walls of the regular rectangular waveguide. The longitudinal slot antennas are placed on the broad wall of the CSIW to radiate into free space. The array is excited from Port 1 and Port 2 is terminated by a match load to obtain a traveling wave array. When the frequency is varied, steering in the main beam direction in H-plane of the slot array

Figure 1.2: (a) Cross-sectional and (b) top views of the traveling wave slot array on CSIW.
can be obtained which eliminates allocating phase shifters used in phased arrays for beam steering. The steering in the E-plane can be achieved by constructing the arrays of CSIW structures and using phase shifters for each CSIW in the array.

In this thesis, different designs of slotted CSIW arrays are carried out at $25 \, GHz$ using non-uniform and uniform slot separations on CSIW. In both of the slot separation cases, the directivity is increased by using proper slot phase values. The non-uniform slot array is designed for beam steering applications while the uniform slot array is examined for pattern synthesis problem. In the non-uniform slot array, the slot elements near resonance are used and the slot separations are designed to adjust slot phases to increase directivity while required slot phases are obtained by using slots with large deviation from resonance in the uniform slot array. Both non-uniform and uniform slot arrays are fabricated on CSIW using standard printed circuit board (PCB) process. The measurement results of the fabricated antennas show that the non-uniform slot array has about $10^\circ$ frequency steering in H-plane when the frequency is swept between $24 \, GHz$ and $26 \, GHz$. In the uniform slot array, a pencil beam pattern is synthesized with a SLL below $-20 \, dB$. These are the first slotted cylindrical SIW arrays presented in the literature.

1.3. Thesis Objectives and Organisation

The main goal of this thesis is to design, analyze, and implement a traveling wave slot array on a cylindrical substrate integrated waveguide (CSIW) where the SIW lies in the circumferential direction of the cylinder. The specific objectives in the frame of this thesis are summarized as follows:

- Characterization of CSIW and single slot on CSIW.
A propagation constant of the fundamental mode excited in CSIW is determined and the effect of radius of curvature on the propagation constant is examined. Single slot is the basic antenna structure used in the array. The circuit model of a single slot is extracted and self-slot admittance values to be used in the design of the array are calculated for various slot lengths and offsets.

- Mutual coupling and active admittance formulations.

The active admittance formulas are needed in the Elliott’s design procedure to determine the slot offsets and lengths in the slotted CSIW array. The external and internal mutual coupling formulations of the slots on CSIW are carried out and the active admittance formulas are derived.

- Design of traveling wave slot arrays on CSIW

Slotted CSIW arrays are designed for beam steering and pattern synthesis applications using non-uniform and uniform slot separations, respectively. Designed arrays are fabricated using conventional PCB manufacturing techniques.

Beam steering is achieved by varying the frequency in the non-uniform slot array. Pattern synthesis procedure is developed for slotted CSIW arrays with uniform slot separation. A few arrays are designed and fabricated to realize low side lobe level, narrow beam width radiation patterns.

This thesis consists of four chapters explaining the accomplishments of the mentioned objectives. After this introduction chapter, the effect of radius on the propagation constant of the fundamental mode in CSIW is examined in Chapter 2.
Moreover, the characterization of the single slot is carried out using a shunt admittance model.

Chapter 3 presents the derivations of active admittance formulas by taking both internal and external couplings into account. Then, the design steps of the slotted CSIW array using derived active admittance formulas are listed. Finally, the case studies related to the mutual couplings of the slots on perfect electric conductor (PEC) cylinder and the mutual couplings of the slots on CSIW are examined. The mutual couplings are calculated using derived coupling formulas for different cases and they are compared with the simulations and literature to validate them.

Chapter 4 describes the design of slotted CSIW arrays. The slotted CSIW arrays are designed for beam steering and pattern synthesis applications. One slotted CSIW array is designed for frequency steering and the other designs are carried out to obtain pencil beam pattern with low SLL. The fabrication and measurement results of the designed slotted CSIW arrays are also presented in the chapter.

Chapter 5 concludes the results of this study with a discussion on possible suggestions of the future improvements.
CHAPTER 2

CHARACTERIZATION OF CSIW AND SINGLE LONGITUDINAL SLOT ON CSIW

This chapter introduces the SIW structures used in the final design of the traveling wave slot array on CSIW. Section 2.1 gives the CSIW parameters used in this thesis and examines the propagation constant of the fundamental mode in the CSIW for different radius of curvature values while Section 2.2 presents the microstrip to SIW transition structure and loss calculation in the SIW. Finally, Section 2.3 introduces the circuit model of the longitudinal slot and characterization of the single slot element on the CSIW.

2.1. Characterization of CSIW

SIW is a waveguide structure the walls of which are formed by periodically located metallic vias with period, $p$ and diameter, $d$, as shown in Figure 2.1 (a). The width of the SIW, $W$, which is usually greater than the substrate height, $h$, determines the cutoff frequency. A SIW can be modelled by a regular rectangular waveguide as shown in Figure 2.1 (b) where the waveguide have the same dielectric filling and waveguide height, $h$, as in the SIW. The width of the waveguide, $W_{\text{eff}}$, is equal to the equivalent width of the SIW determined by the period and diameter of the vias as well as the width of the SIW, [30]. Using this equivalence, the propagation constant
of the modes in the SIW can be calculated using closed form expressions derived for the waveguides.

The SIW structure used in the slotted CSIW array developed in this thesis is built on Rogers 3003 substrate ($\varepsilon_r = 3$, $\tan\delta = 0.0013$ at 10 GHz) with thickness, $h = 0.5$ mm. The width of the SIW, $W$, is measured from the centers of the vias and it is chosen as 6 mm which is half a free space wavelength ($\lambda_0/2$) at 25 GHz to avoid grating lobes in the E-plane when the array of slotted CSIW is formed. Using 6 mm SIW width, only the fundamental mode propagates at 25 GHz in the SIW which is required in the design of the CSIW array. The period and the diameters of the vias are important parameters responsible from the leakage of the electromagnetic field. The dimensions of vias are determined by considering the fabrication tolerances as well as obtaining a good electrical performance. Via diameters should not be too small so that the vias can be successfully drilled and electroplated in PCB process. In order to use the space on the substrate efficiently, large via diameter values should be avoided. The period of the vias should be small enough to have low leakage loss. However, the distance between two vias should be large enough to have enough strength of the structure to avoid any defects in the fabrication. Considering these
fabrication and electrical performance criteria, the period, \( p \), and diameter, \( d \), of the vias are determined as 1 mm and 0.8 mm, respectively.

The propagation constant determines the excitation phases of the slot elements and it is an important parameter used in the design of slotted CSIW array. Using the closed form expressions given in [4], the equivalent width of the SIW shown in Figure 2.1 (a) is calculated. In the calculation of the equivalent width, the SIW parameters mentioned for the slotted CSIW array is used. By choosing the width of the SIW, \( W = 6 \text{ mm} \), only the fundamental mode is allowed to propagate between 16 GHz and 32 GHz. Hence, the propagation constant is obtained for the fundamental mode in planar SIW using the equivalent width in the standard closed form formulas for the waveguides, [31]. The attenuation constant is ignored in the calculations. The propagation constant formulations for the fundamental mode of the rectangular waveguide bent in its E- or H-plane is presented in [32]. After finding the equivalent width of the CSIW bent over its E-plane as shown in Figure 1.2, the propagation constant of the fundamental mode in CSIW is calculated for different values of radius, \( R \), between 5 cm and 15 cm using the equivalent width of the CSIW and the formulations derived for E-plane bend regular waveguides given in [32]. By considering about 4 mm resonant length of the slots at 25 GHz and the element spacing, the radius values above 5 cm is useful for cylindrical array applications at 25 GHz to have enough number of array elements to obtain good matching as well as satisfying some pattern requirements such as high directivity. The propagation constant values calculated using closed form expressions for the planar case and for different radius of curvature values are compared in Figure 2.2. As it can be seen in Figure 2.2, there is almost no difference between the results for the radius values above 5 cm and planar case.

The simulations of the propagation constant of CSIW are carried out in Ansys High Frequency Structure Simulator (HFSS) to compare with the propagation constants obtained from closed form expressions. The CSIW structure shown in Figure 1.2 is
simulated without slots and microstrip parts in HFSS. The CSIW is excited by the wave port in HFSS representing a regular rectangular waveguide coming from infinity having width equal to the equivalent width of the SIW to reduce reflections. The propagation constants are derived from simulated phase of $S_{21}$ results of the structure for various radius values as shown in Figure 2.2. As it is seen in Figure 2.2, it can be concluded that the simulated propagation constants are nearly the same as the ones calculated by closed form expressions.

### 2.2. Microstrip to SIW Transition Structure

A transition from microstrip line to SIW structure is needed to plug connectors to the microstrip lines and then to be able to measure the SIW devices by the network analyzer. A transition structure designed for the traveling wave antenna study is composed of a microstrip line linearly tapered till the one side of the SIW as shown in Figure 2.3. A $50 \, \Omega$ microstrip line have width, $W_f = 1.25 \, mm$ and length,
$L_f = 4 \, \text{mm}$. The tapered section with dimensions, $W_t = L_t = 2.1 \, \text{mm}$ is connected to SIW with length, $L$. Both microstrip line and SIW is built on $0.5 \, \text{mm}$-thick Rogers 3003 substrate. The same SIW parameters as mentioned in Section 2.1 are used.

The designed transition structure with length, $L = 2 \, \text{cm}$ is fabricated using standard PCB process and it is shown in Figure 2.4 (a). The structure is measured using 2.92

![Figure 2.3: A schematic view of microstrip line to SIW transition.](image)

![Figure 2.4: (a) Fabricated microstrip to SIW transition ($L = 2 \, \text{cm}$). (b) $S_{11}$ and (c) $S_{21}$ results.](image)
end launch connectors operating till 40 GHz. The measured results are compared with the simulations in Figure 2.4 (b) and (c). As it can be seen in Figure 2.4 (b) and (c), there is a good agreement between measured and simulated results where the effect of the connectors is taken into account in the simulations. The measured return loss is better than 10 dB between 20 GHz and 30 GHz. Measured insertion loss is about 0.91 dB at 25 GHz and the simulations of the structure with connectors shows about 0.57 dB insertion loss which is quite close to the measured one.

The other transition structure with length, $L = 10 \text{ cm}$ is also fabricated in which the other parameters are the same with the transition structure shown in Figure 2.4 (a). The fabricated transition is presented in Figure 2.5 (a). The simulations of the structure with connectors are carried out and they are compared with the measurements as shown in Figure 2.5 (b) and (c). A good agreement between the

Figure 2.5: (a) Fabricated microstrip to SIW transition $(L = 10 \text{ cm})$. (b) $S_{11}$ and (c) $S_{21}$ results.
measured and simulated s-parameters of the structure is obtained especially in the $S_{11}$ parameters shown in Figure 2.5 (b). Measured and simulated insertion losses of the structure at 25 GHz are found as 2.13 dB and 1.48 dB, respectively. In order to understand the small differences between the simulations and measurements, many fabrications are carried out for the devices and highly non-uniform manufacturing is observed between the same samples in the same fabrication cycle, i.e., yield of the fabrication is low. Devices that give the best results are chosen, and hence, fabrication process might need to be improved further for better results.

Using the measured insertion loss values for the transition structures with 2 cm and 10 cm SIW lengths, the loss inside the SIW is calculated as 0.153 dB/cm at 25 GHz. Hence, the measured losses due to the microstrip feed line with length, $L_f = 4 \text{ mm}$, and the transition part with length, $L_t = 2.1 \text{ mm}$ are calculated as about 0.3 dB for one sided transition at 25 GHz. The loss inside the SIW includes dielectric and conductor losses while the losses at the transition part include dielectric, conductor, and radiation losses due to the discontinuity encountered in the microstrip transition.

### 2.3. Characterization of Single Slot on CSIW

Characterization of the single slot is one of the important steps in designing the slotted waveguide array. It is known that longitudinal slot on broad face of planar rectangular waveguide is modeled as shunt admittance [33]. When the other slots are absent in the array, the shunt admittance of single slot does not contain the effects of mutual couplings, hence the admittance of single slot is called self-slot admittance. Since the cross sections of the planar SIW and CSIW are the same and the length of the slot is small compared to the radius of curvature, the shunt admittance model is not affected much by the radius. Hence, a longitudinal slot on CSIW can also be modeled as shunt admittance. In the planar regular rectangular waveguide case, the circuit model of the slot can be extracted by feeding the structure using a wave port.
and embedding the ports till the slot reference plane in the simulator. However, in the simulation of CSIW, there is no port parameter calculation in HFSS for both cylindrical waveguide structures and planar SIW. Hence, Thru-Reflect-Line (TRL) calibration standards shown in Figure 2.6 are used to extract the circuit model of single slot on CSIW.

The circuit model of a longitudinal slot on CSIW is extracted using the parameters of

Figure 2.6: TRL calibration standards: (a) Thru (b) Reflect, and (c) Line Standards.
the SIW given in Section 2.1. The CSIW is placed on a cylindrical surface with radius $R = 10 \text{ cm}$ as shown in Figure 1.2 and single slot is assumed to exist on the CSIW. Moreover, only $\hat{a}_z$-directed electrical field is assumed in the slot region [34]. TRL calibration standards shown in Figure 2.6 are also placed on cylindrical surface with radius $R = 10 \text{ cm}$ to eliminate the effect of the transition from microstrip line to CSIW and to extract the slot admittance. The slot width should be small in order not to disturb the fundamental mode in CSIW. Moreover, narrow slot assumption which assumes only $\hat{a}_z$-directed electrical field in the slot region shown in Figure 1.2 can be achieved better by reducing the width of the slot [34]. By considering the fabrication tolerances and mentioned electrical criteria, the width of the slot, $w^s$, is chosen as $0.25 \text{ mm}$. A single slot having width of $w^s = 0.25 \text{ mm}$ is placed on a Thru standard with length $L$ as indicated in Figure 2.6 (a) where the slot center is located at the reference plane which cuts the CSIW two equal pieces with CSIW length $L/2$. The length and offset of the $p^\text{th}$ slot are abbreviated as $L_p$ and $z_p$, respectively, as shown in Figure 1.2 (b) and Figure 2.6 (a). The effects of curved edges, which are encountered in PCB milling process in the fabrication, are taken into account by properly rounding the edges of the slots in the simulations. The slot length, $L_p$ extends till the rounded edges of the slots.

The circuit model of single longitudinal slot placed on the CSIW is assumed to be a shunt admittance as shown in Figure 2.7 where $Y_s/G_0$ is the normalized self-admittance of the slot, where $G_0$ and $\gamma$ are the characteristic impedance and the

![Figure 2.7: Circuit model of longitudinal slot on CSIW.](image-url)
propagation constant of the CSIW, respectively. The CSIW with single slot simulations are carried out for different slot offsets, \( z_p \) and lengths, \( L_p \) at 25 GHz for \( R = 10 \text{ cm} \). The simulations of the TRL standards without the slot are also carried out at 25 GHz for \( R = 10 \text{ cm} \) and the \( ABCD \) parameters of the slot are extracted for different slot offsets and lengths. Extracted \( A, D \) parameters are shown in Figure 2.8 while \( B \)-parameters are presented in Figure 2.9. As it is seen in Figure 2.8 and Figure 2.9, \( A \)- and \( D \)-parameters are close to unity and \( B \)-parameters are close to zero for slot offsets till 0.5 mm and slot lengths around resonance, which is the case for shunt admittance. Hence, it is proved that the longitudinal slot on CSIW can also be modeled as shunt admittance.

![Graph](image)

**Figure 2.8:** \( A \) and \( D \)-parameters of the single slot on CSIW for \( R = 10 \text{ cm} \): (a) real and (b) imaginary parts.
The shunt admittance values of the slot are equal to the $C_m$ parameters. $C_m$ parameters, i.e., real ($G_s/G_0$) and imaginary ($B_s/G_0$) parts of the normalized shunt admittance, $Y_s/G_0$, are given in Figure 2.10. It can be concluded from Figure 2.10 that the resonant conductance and the resonant length increase with the slot offset [35]. In the design of a slotted CSIW array, the desired complex slot voltage distribution is determined by the pattern requirements. Then the desired active admittance values corresponding to the desired slot voltage distribution are determined. After that the slot offsets and lengths which satisfy desired active admittance values are found. Hence, the slot admittance values calculated between discrete slot offset and length values might be needed to sustain desired excitation voltage. So, it is useful to derive the characterization polynomials using Figure 2.10 to be able to extract any slot admittance that corresponds to any slot offset till 0.5 mm and any slot length between 3.5 mm and 5 mm. The accuracy of the characterization polynomials increases as the number of slot offset and length values to carry out simulations is increased. For this reason, the simulations are carried out with 0.1 mm slot offset steps and 0.025 mm

Figure 2.9: $B$-parameters of the single slot on CSIW for $R = 10 \, cm$: (a) real and (b) imaginary parts.
slot length steps which are dense enough for good interpolation between discretely calculated values.

The characterization polynomials are derived by using the curves for different slot offsets shown in Figure 2.10. The first characterization polynomial curves are determined as follows. First of all, the slot lengths are normalized with respect to the resonant length, \( L_r \). After that, the conductance, \( G_s \) and susceptance, \( B_s \) values are normalized with respect to the resonant conductance, \( G_r \). These calculations are carried out for different slot offset values and the first characterization polynomial curve shown in Figure 2.11 (a) is found by taking the average of the curves. The second and third characterization polynomial curves are the normalized resonant conductance, \( G_r/G_0 \) and the normalized resonant lengths, \( kL_r \) values as a function of the slot offset as shown in Figure 2.11 (b) and (c), respectively, where \( k \) is the wavenumber in the CSIW [33]. The characterization polynomials are determined by finding the polynomial coefficients of the curves shown in Figure 2.11.
The longitudinal slot characterization is carried out for $R = 5 \text{ cm}$, $R = 15 \text{ cm}$, and planar case. The characterization polynomials are determined for these cases and they are presented in Figure 2.11. As it is seen in Figure 2.11, there is almost no difference between different radius values. The radius values affect the resonant length of the slot, especially for the small slot offset values.

Figure 2.11: Characterization polynomials of the shunt slot on CSIW: (a) Ratios of the admittance to resonant conductance. (b) Normalized Resonant Conductance. (c) Normalized Resonant Length.
A single slot is designed and fabricated on the SIW transition structure with length, $L = 2 \, cm$ shown in Figure 2.4 (a). The fabricated single slot on the SIW transition structure is shown in Figure 2.12 (a). The single slot has slot offset, $z_p = 0.5 \, mm$ and slot width, $w_s = 0.25 \, mm$. The length of the slot is $L_p = 4.3423 \, mm$ which is the resonant length of the slot for $0.5 \, mm$ slot offset determined from Figure 2.10. The measured s-parameters of the single slot is compared with the simulations in Figure 2.12 (b) and (c). Good agreement between the simulated and measured values is obtained. In Figure 2.12 (b), the $S_{11}$-parameters are slightly above $-10 \, dB$ at some frequency points. Compared to the $S_{11}$-parameters of the transition structure with no slots shown in Figure 2.4 (b), the input match is disturbed by the existence of the slot in Figure 2.12 (b) which is an expected result because of the shunt admittance value introduced by the slot. In the slot array design, slot offsets and spacings should be

Figure 2.12: (a) Fabricated single slot on microstrip to SIW transition ( $L = 2 \, cm$). (b) $S_{11}$ and (c) $S_{21}$ results.
properly designed so that total admittance seen by excitation side of the array should be close to characteristic impedance of the waveguide to avoid mismatch caused by the slots. The effect of the other slots on shunt admittance model should also be taken into account in the design as described in Chapter 3.
CHAPTER 3

DESIGN PROCEDURE OF SHUNT SLOT ARRAY ON CSIW

This chapter presents the derivation of the active admittance formulas of the longitudinal slots on broad wall of E-plane bend regular waveguide by taking the external and internal couplings into account. Section 3.1 presents the description and formulation of the problem and the derivation of the active admittance formulas while Section 3.2 explains the design procedure for the slotted CSIW arrays using derived active admittances. Section 3.3 gives the numerical results for the coupling of the slots on cylindrical surface and on CSIW.

3.1. Derivation of the Active Admittance Formulas for the Longitudinal Slots on CSIW

The characterization of the single slot on CSIW is carried out in Section 2.3. As it is stated in Section 2.3, the admittance of the single slot on CSIW is called self-slot admittance since the other slots are absent in CSIW. When the other slots are present in the array, the self-slot admittance values used to present the longitudinal slots are disturbed by the mutual couplings and the slots are presented by the active slot admittance values. In other words, the active slot admittance is the admittance of the slot when the other slots are present in the array. The active admittance is also assumed to be shunt admittance like in the self-admittance circuit model. The active admittance is found by considering both the external and internal couplings in the
slotted CSIW array. The slot couplings through free space outside the SIW are considered as external couplings while the slot couplings through the path inside the SIW are called internal couplings.

The active admittance formulas containing the external and internal mutual couplings of the slots are important parameters used in the Elliott’s design procedure [34], [36]. The active admittance formulas derived in [34] and [36] contains the self-slot admittance terms and the mutual coupling terms for the planar slotted waveguide antennas. Similar to the planar slotted waveguides, the active admittance formulas for slotted CSIW antennas can be derived by the help of reciprocity theorem. The fundamental mode and higher order modes are used in the derivation of the active admittance formulas. So, it is proper to examine the mode characteristics in the CSIW before the derivation of the active admittance formulas.

### 3.1.1. Modes in the CSIW

The CSIW used in this study resembles E-plane bend regular waveguides. The E-plane bend regular waveguides differ only in containing the continuous walls compared to the CSIW. The schematic view of the E-plane bend regular waveguide is presented in Figure 3.1. The bended regular waveguide has PEC boundaries at \( \rho = \rho_1, \rho_2 \) and \( z = 0, a \). Since traveling wave slot array is examined in this thesis, the E-plane waveguide is assumed to be excited from one end and it is matched at \( \phi = \phi_1, \phi_2 \). The E-plane bend regular waveguide is formed by bending the regular rectangular waveguide on cylindrical surface such that its broad wall with length \( a \) lies parallel to the z-axis as shown in Figure 3.1, [32]. Since the CSIW presented in Figure 1.2 is bended in a similar manner, the modes inside the CSIW can be examined using the E-plane bend regular waveguides. Because the waveguide is bended, the hybrid mode fields, i.e., transverse magnetic to \( \hat{a}_z \)-direction (\( TM_{mn}^{z} \)) and
transverse electric to $\hat{a}_z$ -direction ($TE^{z}_{mn}$), exist in the bend regular waveguides [37]. Using the systematic construction of fields [38], $TM^{z}_{mn}$ and $TE^{z}_{mn}$ mode field components are derived for the E-plane bend regular waveguide in Appendix A. These field components are used in the application of reciprocity theorem to find the active admittance formulas.

The eigenvalues, $\epsilon_{mn}$ and $\mu_{mn}$, for $TM^{z}_{mn}$ and $TE^{z}_{mn}$ modes, respectively can be found from roots of (A.20) and (A.33), respectively [37]. The CSIW is formed by using the substrate and SIW parameters given in Section 2.1. In Figure 3.1, the radius, $R = (\rho_1 + \rho_2)/2$ is taken as 10 cm and the broad wall dimension, $a$ is determined by the equivalent width of the CSIW, [4]. The height of the bended regular waveguide, $(\rho_2 - \rho_1)$, is equal to the thickness of the substrate used in CSIW. Using these parameters the eigenvalues, $\epsilon_{mn}$ and $\mu_{mn}$ for $TM^{z}_{mn}$ and $TE^{z}_{mn}$ fields, respectively, are calculated at 25 GHz and presented in Table 3.1. In Table 3.1, the $m$
values explicitly used in (A.20) and (A.33) are listed. However, different $n$ values are assigned in order to discriminate two cases with same $m$ value. The $n$ values are started from 0. As it is seen in Table 3.1, $TE_{10}^z$ mode is the fundamental mode as in regular rectangular waveguides. The other bended regular waveguide modes are found to be similar to the $TE$ and $TM$ fields with respect to the direction of propagation in planar waveguides with the same transverse cross-section shown in Figure 3.1 except the case where $m = 0$ and $n = 1$. For $m = 0$ and $n = 1$, $TE_{01}^z$ field components vanishes which can be seen in Appendix A and $TM_{01}^z$ modes are excited unlike $TE$ and $TM$ modes in planar waveguides. For this reason, the value for the eigenvalue $\mu_{01}$ is assigned as not applicable (NA). The other cases where the eigenvalues assigned as NA in Table 3.1 corresponds to the cases such that there is no root of (A.20) or (A.33).

Table 3.1: Eigenvalues, $\epsilon_{mn}$ and $\mu_{mn}$, of $TE_{mn}^z$ and $TM_{mn}^z$ modes for various $m$ and $n$ values. (NA: Not Applicable).

<table>
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<td>-j63.30</td>
</tr>
<tr>
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<td>-j21.70</td>
<td>NA</td>
<td>3</td>
<td>1</td>
<td>-j64.60</td>
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<tr>
<td>5</td>
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<tr>
<td>6</td>
<td>0</td>
<td>-j34.10</td>
<td>NA</td>
<td>5</td>
<td>1</td>
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<tr>
<td>7</td>
<td>0</td>
<td>-j40.10</td>
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<td>1</td>
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<td>-j46.12</td>
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<td>NA</td>
<td>9</td>
<td>1</td>
<td>-j81.60</td>
<td>-j81.60</td>
</tr>
</tbody>
</table>
The propagation constant values in the E-plane bend regular waveguide can be found by dividing the eigenvalues listed in Table 3.1 by the radius, $R$. If this procedure is carried out for $TE_{10}^{z}$ mode, it can be seen that the resulting propagation constant value is so close to the propagation constant values calculated in Figure 2.2 at 25 GHz in Section 2.1.

3.1.2. Active Admittance Formulation of Longitudinal Slot on the CSIW

The active admittance formulations of the longitudinal slot on the CSIW are carried out using the procedure similar to one explained in [34]. The formulations are carried out for the longitudinal slots on E-plane bend regular waveguide and the equivalent width of SIW is used for the longitudinal slots on CSIW. Two formulas are derived for the active admittance values which are used as first and second design equations in the design of slotted CSIW array. In the first active admittance formula, it will be assumed that the total slot voltage is known after all couplings take place. Then the relation between the active admittance and circuit model parameters, such as, slot voltage, mode voltage, and characteristic impedance, etc., is found. The first design formula is used to find the relation between the active admittances of successive slots in the traveling wave array while the value of the active admittance is calculated explicitly in the second active admittance formula. In the second active admittance formula, the sources of the slot voltages are examined and the self-admittance is modified by mutual coupling terms. The accuracy of the active admittance value increases as the number of coupling terms considered in the second design equation increases. Before the active admittance formulations, the variables which define the bended regular waveguide and slot dimensions used in the active admittance formulations are introduced.
3.1.2.1. Waveguide and Slot Dimensions

The schematic view of longitudinal slot array on bended regular waveguide is shown in Figure 3.2. The bend regular waveguide is formed between cylindrical PEC surfaces located at $\rho = \rho_1$ and $\rho = \rho_2$. Then the longitudinal slots are cut on the surface at $\rho = \rho_2$. The waveguide has PEC boundaries on the waveguide walls located between $z = z_1$ and $z = z_2$ with $z_2 > z_1$ in $\hat{z}$-direction. Both the region defined by $\phi_L < \phi < \phi_F$ for $\rho < \rho_2$ and the other region defined by $\phi_F < \phi < \phi_L$ for $\rho < \rho_1$ are filled with PEC. Moreover, the structure extends to infinity in $\pm \hat{z}$-direction so that in the regions outside the waveguide, i.e., the region for $z < z_1$ and $z > z_2$, the structure becomes a cylindrical PEC volume with radius $\rho_2$. The waveguide region is covered by fictitious closed surface $S$ which is composed of the surfaces $S_F$, $S_L$ and $S_3$. The surface $S_3$ is located on the waveguide boundaries at

![Figure 3.2: Longitudinal slot array on bended regular waveguide.](image-url)
\( \rho = \rho_1, \ \rho = \rho_2, \ z = z_1, \) and \( z = z_2. \) The waveguide is fed from Port1 located at \( \phi = \phi_F \) on the surface \( S_F \) and Port2 located at \( \phi = \phi_L \) on the surface \( S_L \) is terminated by a matched load to obtain traveling wave slot array.

In order to describe the slot dimensions, the \( p^{th} \) slot in the slot array shown in Figure 3.2 is chosen. The \( p^{th} \) slot is presented in Figure 3.3. The start and end positions of the \( p^{th} \) slot in \( \hat{a}_\phi \)-direction are \( \phi = \phi_1^p \) and \( \phi = \phi_2^p \), respectively. The total angular displacement of the slot is \( 2\phi_0^p \) while the start and end positions of the slot in \( \hat{a}_z \)-direction are \( z = z_1^p \) and \( z = z_2^p \), respectively. The slot width, \( w^p \), is the same for all the slots in the slot array shown in Figure 3.2 and it is assumed to be very thin so that only \( \hat{a}_z \)-directed electric field exist in the slot. Considering these slot and waveguide dimensions following equations can be written.

Figure 3.3: Longitudinal slot (\( p^{th} \) slot) on bended regular waveguide.
where $\phi_p$ and $z_p$ are the center positions of the slot in the global coordinate system. $z_p$ is the slot offset measured with respect to the centerline of the waveguide as shown in Figure 1.2. $z^{p,0}$ is the center position of the slot in $\hat{a}_z$-direction measured from the wall of the waveguide. The derivation of the active admittance formulas are carried out using the field expressions given in Appendix A where the formulas are derived for waveguide located at $z_1 = 0$ and $z_2 = a$.

### 3.1.2.2. Derivation of the First Active Admittance Formula

The derivation of the active admittances is based on finding the backward scattered mode amplitude for each slot in the slot array shown in Figure 3.2 as explained in [33], [34], and [36]. Then using the relation between the backward scattered mode amplitude and the transmission line model, the first active admittance formula is derived. The backward scattered mode amplitude, $B_p$, for the $p^{th}$ slot in the array can be found and the result can be generalized to all the slots in the array. First of all, the slotted bended regular waveguide array shown in Figure 3.2 is excited by the
fundamental mode, \((TE_{10}^z)\) mode, incident from Port1 located at the surface \(S_F\) and Port2 at the surface \(S_L\) is terminated by a matched load. As a result of this excitation, the slot voltage \(V_{p,1}^s\) appears between the terminals of the \(p^{th}\) slot in the waveguide. Because of the reflections of the fundamental mode and higher order modes from the other slots in the waveguide and also due to the external couplings, the total slot voltage becomes \(V_p^s\) for the \(p^{th}\) slot. The sources of \(V_p^s\) will be examined in detail in Section 3.1.2.3 while deriving the second active admittance formula. In deriving the first active admittance formula, it will be assumed that total slot voltage \(V_p^s\) is known in the array. The first active admittance formula gives the relation between the slot voltage, \(V_p^s\), and the active admittance. The slot voltage, \(V_p^s\), and hence, the active admittance causes backward and forward scattering waves propagating in \(-\hat{a}_\phi\) and \(+\hat{a}_\phi\)-directions respectively. The backward and forward scattered mode amplitudes for the \(p^{th}\) slot will be derived only for \(TE_{mn}^z\) modes because 9 higher order modes appear as \(z_{mn}TE\) modes before \(z_{mn}TM\) higher order modes are excited as shown in Table 3.1. The internal mutual couplings of the slots are considered only for the first higher order mode in the second design equation for planar waveguides, [36]. The effect of the higher order mode couplings increases as the height of the waveguide decreases, [36]. By considering the small thickness of the substrate, which determines the height of the bended regular waveguide with equivalent width of the CSIW used in this thesis, considering the effect of only the first higher order mode is enough and a few higher order modes can also be taken into account if desired. For this reason, the backward and forward scattered coefficients are not calculated for \(TM_{mn}^z\) modes in slotted bended regular waveguide array.

The backward and forward scattered mode amplitudes for \(TE_{mn}^z\) modes which are abbreviated as \(B_{p}^{TE_{mn}}\) and \(C_{p}^{TE_{mn}}\), respectively, caused by the slot voltage \(V_p^s\) will be derived by the help of the reciprocity theorem. Before applying the reciprocity
Theorem, it is assumed in the derivations that the slot width, $w'$ is small enough so that only $a_z$-directed electric field appears in the slot region. Moreover, it is also assumed that the scattering is not affected by the propagation direction of the field used to excite the waveguide, i.e., the slot mode is symmetrical and its shape is not affected by the direction of propagation of the waveguide mode [33]. Since the slot is longitudinally placed on the broad wall of the waveguide, the circuit model of the slot can be assumed as a shunt admittance as stated in Section 2.3.

The reciprocity theorem can be applied to the fields around the $p^{th}$ slot in the slot array shown in Figure 3.2. The zoomed view of the $p^{th}$ slot in the slot array while the array is excited is shown in Figure 3.3. The closed fictitious surface $S$ is defined around the $p^{th}$ slot to apply the reciprocity theorem as shown in Figure 3.3. The surface, $S$ is composed of three parts. The first and second parts are $S_1$ and $S_2$ lying before and after the $p^{th}$ slot region, respectively. The third part, $S_3$ lies skin tight against the waveguide walls and it contains the surface $S_p$, defined just below the slot region. The fields $E_1$ and $H_1$ inside the surface $S$ are $TE_{mn}$ and $TM_{mn}$ mode fields scattered by the slot to the region $\phi < \phi^p$ and $\phi > \phi^p$ and they are equal to the slot mode in the slot region. Similar to finding the scattering coefficients of the slots for $TE$ modes in planar waveguides described in [33], the fields $E_2$ and $H_2$ inside the closed surface $S$ are chosen as $TE_{mn'}$ mode fields exist in the absence of the slot, i.e., when the slots in the waveguide is covered by conducting tape as shown in Figure 3.4. In Figure 3.4, the surface $S_3$ contains the fictitious surface, $S_p$ to represent the $p^{th}$ slot region. However, the slot surface $S_p$ is shorted in Figure 3.4. Similar to [33], $E_2$ and $H_2$ are chosen as field components used to find the forward and backward scattering coefficients of the slot, they are not the modes exciting the waveguide. As stated before, the waveguide is excited by $TE_{10}$ mode, and hence, the slot voltage $V^p_s$ appears between the slot terminals. The usage of $E_2$ and $H_2$ is just an artifice to obtain scattering coefficients for $TE_{mn'}$ mode, [33]. Since the closed surfaces defined
in Figure 3.3 and Figure 3.4 are geometrically the same, the reciprocity theorem presented in (3.6) can be applied between the fields \( (E, H) \) and \( (E, H) \) as, [33].

\[
\int_S \left( \mathbf{E}_1 \times \mathbf{H}_2 - \mathbf{E}_2 \times \mathbf{H}_1 \right) d\mathbf{S} = 0, \quad S = S_1 \cup S_2 \cup S_3
\]

(3.6)

Since the tangential component of \( \mathbf{E}_1, \mathbf{E}_1^{\text{tan}} = 0 \) on the surface \( S_3 - S_p \) in Figure 3.3 and the tangential component of \( \mathbf{E}_2, \mathbf{E}_2^{\text{tan}} = 0 \) over the surface \( S_3 \) in Figure 3.4, it is obtained that

\[
\int_{S_1} \left( \mathbf{E}_1 \times \mathbf{H}_2 - \mathbf{E}_2 \times \mathbf{H}_1 \right)d\mathbf{S} = \int_{S_p} \left( \mathbf{E}_{\text{slot}} \times \mathbf{H}_2 \right)d\mathbf{S}.
\]

Hence, (3.6) becomes

\[
\int_{S_p} \left( \mathbf{E}_{\text{slot}} \times \mathbf{H}_2 \right)d\mathbf{S} = \int_{S_1} \left( \mathbf{E}_2 \times \mathbf{H}_1 - \mathbf{E}_1 \times \mathbf{H}_2 \right)d\mathbf{S} + \int_{S_2} \left( \mathbf{E}_2 \times \mathbf{H}_1 - \mathbf{E}_1 \times \mathbf{H}_2 \right)d\mathbf{S}
\]

(3.7)
where \( \overline{E}_{\text{slot}} \) is the slot field found by considering the assumptions on the slots mentioned before and \( \overline{E}_1 = \overline{E}_{\text{slot}} \) on \( S_p \) as shown in Figure 3.3.

The fields \( \overline{E}_1 \) and \( \overline{H}_1 \) on the surfaces \( S_1 \) and \( S_2 \) are composed of summation of \( TE_z^{\pm} \) and summation of \( TM_z^{\pm} \) modes derived in Appendix A and they are given by

\[
\overline{E}_1 = \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} E_1^{TE_{m\omega}} + \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} E_1^{TM_{m\omega}} , \quad \phi < \phi_1^p \text{ or } \phi > \phi_2^p
\]  

(3.8)

\[
\overline{H}_1 = \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} H_1^{TE_{m\omega}} + \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} H_1^{TM_{m\omega}} , \quad \phi < \phi_1^p \text{ or } \phi > \phi_2^p
\]  

(3.9)

The fields \( \overline{E}_2 \) and \( \overline{H}_2 \) are composed of only \( TE_z^{\pm} \) mode components, i.e., \( \overline{E}_2 = E_2^{TE_{\omega_2'}} \) and \( \overline{H}_2 = H_2^{TE_{\omega_2'}} \). Using mode orthogonality [39], the integrals over \( S_1 \) and \( S_2 \) on the right hand side (RHS) of (3.7) becomes

\[
\int \left( \overline{E}_2 \times \overline{H}_1 - \overline{E}_1 \times \overline{H}_2 \right) d\overline{S} = \int \left( E_2^{TE_{\omega_2'}} \times \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} H_1^{TE_{m\omega}} - \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} E_1^{TM_{m\omega}} \times H_2^{TE_{\omega_2'}} \right) d\overline{S}
\]  

(3.10)

If \( E_2^{TE_{\omega_2'}} \) and \( H_2^{TE_{\omega_2'}} \) are chosen as \( TE_{m\omega_2'} \) mode fields propagating in \( +\hat{\alpha}_\theta \)-direction, the backward scattered mode amplitudes, \( B_p^{TE_{\omega_2'}} \) are derived. Since \( d\overline{S} \) is in \( \pm\hat{\alpha}_\theta \)-direction in (3.10), only transverse field components (\( \hat{\alpha}_\rho \) and \( \hat{\alpha}_z \)-components) of \( E_1^{TE_{m\omega}} , H_1^{TE_{m\omega}} \) and \( E_2^{TE_{\omega_2'}} , H_2^{TE_{\omega_2'}} \) are required for the integrals taken over \( S_1 \) and \( S_2 \).

The transverse field components of \( E_1^{TE_{m\omega}} , H_1^{TE_{m\omega}} \) in the regions of \( S_1 \) and \( S_2 \) can be written using the fields derived in Appendix A as
\[ \vec{E}_{1,t}^{T\varepsilon_m} = \begin{align*} &\sum_{m=1}^{\infty} \sum_{n=0}^{\infty} B_p^{T\varepsilon_m} \sin \left( \frac{m \pi}{a} \right) B_{m\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) \mu_{mn} e^{j \mu_{mn}(\phi - \phi_p)} \hat{a}_p, \quad \phi < \phi_p^p \\ &\sum_{m=1}^{\infty} \sum_{n=0}^{\infty} C_p^{T\varepsilon_m} \sin \left( \frac{m \pi}{a} \right) B_{m\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) \mu_{mn} e^{-j \mu_{mn}(\phi - \phi_p^p)} \hat{a}_p, \quad \phi > \phi_p^p \end{align*} \] (3.11)

\[ \vec{H}_{1,t}^{T\varepsilon_m} = \begin{align*} &\frac{1}{2} \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} B_p^{T\varepsilon_m} \frac{m \pi}{a} \cos \left( \frac{m \pi}{a} \right) B_{m\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) k_{m\varepsilon_m}^{T\varepsilon_m} e^{j \mu_{mn}(\phi - \phi_p)} \hat{a}_p \\
&+ \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} C_p^{T\varepsilon_m} \frac{m \pi}{a} \sin \left( \frac{m \pi}{a} \right) B_{m\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) k_{m\varepsilon_m}^{T\varepsilon_m} e^{-j \mu_{mn}(\phi - \phi_p)} \hat{a}_z \\
&\frac{1}{2} \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} C_p^{T\varepsilon_m} \frac{m \pi}{a} \cos \left( \frac{m \pi}{a} \right) B_{m\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) k_{m\varepsilon_m}^{T\varepsilon_m} e^{-j \mu_{mn}(\phi - \phi_p)} \hat{a}_p \\
&+ \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} C_p^{T\varepsilon_m} \frac{m \pi}{a} \sin \left( \frac{m \pi}{a} \right) B_{m\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) k_{m\varepsilon_m}^{T\varepsilon_m} e^{j \mu_{mn}(\phi - \phi_p)} \hat{a}_z \end{align*} \] (3.12)

Let \( \vec{E}_{1,t}^{T\varepsilon_m} \) and \( \vec{H}_{1,t}^{T\varepsilon_m} \) be \( T\varepsilon_m \) mode field propagating in +\( \hat{a}_\phi \)-direction. The transverse field components of \( \vec{E}_{1,t}^{T\varepsilon_m} \) and \( \vec{H}_{1,t}^{T\varepsilon_m} \) are given by

\[ \vec{E}_{2,t}^{T\varepsilon_m} = \frac{i}{\rho} \sin \left( \frac{m' \pi}{a} \right) B_{m'\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) \mu_{m'n'} e^{j \mu_{m'n'}(\phi - \phi_p)} \hat{a}_p \] (3.13)

\[ \vec{H}_{2,t}^{T\varepsilon_m} = \frac{1}{2} \cos \left( \frac{m' \pi}{a} \right) B_{m'\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) \mu_{m'n'} e^{-j \mu_{m'n'}(\phi - \phi_p)} \hat{a}_p + \] \[ \frac{1}{2} \sin \left( \frac{m' \pi}{a} \right) B_{m'\varepsilon_m}^{T\varepsilon_m} \left( k_p^{T\varepsilon_m} \rho \right) \mu_{m'n'} e^{j \mu_{m'n'}(\phi - \phi_p)} \hat{a}_z \] (3.14)

It can be shown that \( \int \left( \vec{E}_{2,t}^{T\varepsilon_m} \times \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} \vec{H}_{1,t}^{T\varepsilon_m} - \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} \vec{E}_{1,t}^{T\varepsilon_m} \times \vec{H}_{2,t}^{T\varepsilon_m} \right) dS = 0 \) in (3.10). Then using (3.7) - (3.14) and mode orthogonality the back scattering mode amplitudes are derived as
\[ B_p^{TE_{mn}^z} = \frac{1}{2} \int_{S_{mn}} \left( \overline{E}_{slot} \times \overline{H}_2^{TE_{mn}^z} \right) d\overline{S} \]  
(3.15)

where

\[ \overline{E}_{mn}^{TE_{mn}^z} = j \frac{\rho}{\rho} \sin \left( \frac{m \pi z}{a} \right) B_{\rho \mu} \left( k_{\rho \mu} \rho \right) \mu_{mn} \hat{a}_{\rho} \]  
(3.16)

\[ \overline{H}_t^{TE_{mn}^z} = \frac{1}{z} \frac{m \pi}{a} \cos \left( \frac{m \pi z}{a} \right) B_{\rho \mu} \left( k_{\rho \mu} \rho \right) \mu_{mn} \hat{a}_{\rho} + \frac{k_{mn}^{TE_{mn}^z}}{z} \sin \left( \frac{m \pi z}{a} \right) B_{\rho \mu} \left( k_{\rho \mu} \rho \right) \hat{a}_z \]  
(3.17)

\[ \overline{H}_2^{TE_{mn}^z} = \begin{bmatrix} \frac{1}{z} \frac{m \pi}{a} \cos \left( \frac{m \pi z}{a} \right) & B_{\rho \mu} \left( k_{\rho \mu} \rho \right) e^{-j\mu_{mn}(\phi - \phi')} \hat{a}_z \\ 0 & 0 \end{bmatrix} \]  
(3.18)

and where the slot field is given by

\[ \overline{E}_{slot} = \frac{V_p}{w} \cos \left( \frac{\pi (\phi - \phi')}{2\phi_0} \right) \hat{a}_z \]  
(3.19)

By assuming \( \overline{E}_{2^{TE_{mn}^z}} \) and \( \overline{H}_{2^{TE_{mn}^z}} \) mode fields propagating in -\( \hat{a}_\phi \)-direction, the forward scattering \( TE_{mn}^z \) mode amplitudes \( C_p^{TE_{mn}^z} \) can be found and it can be shown that

\[ C_p^{TE_{mn}^z} = -B_p^{TE_{mn}^z} \]  
(3.20)
The general expression of $B^{TE_p}$ can be obtained from (3.15) as

$$B^{TE_p} = K^{TE_p} f^{TE_p} V^x_p$$

(3.21)

where

$$K^{TE_p} = \frac{m\pi}{a^2} \frac{B_{mn} \left( k_p^{TE_p} / \rho_2 \right)}{N_{\rho_{mn}}} \sin c \left( \frac{mw}{2a} \right)$$

(3.22)

$$f_p^{TE_p} = \frac{\left( \pi / \phi_0^p \right)}{\left( \pi / 2\phi_0^p \right)} \cos \left( \mu \phi_0^p \right) \cos \left( \frac{m\pi z_{p,0}}{a} \right)$$

(3.23)

and where

$$N_{\rho_{mn}} = \int_{\rho_{mn}} \frac{1}{\rho} \left[ B_{mn} \left( k_p^{TE_p} / \rho \right) \right] d\rho$$

(3.24)

In (3.22), since $w^x \ll 2a$, the term $\sin c \left( \frac{mw}{2a} \right)$ ≈ 1 for $m = 1$.

In order to relate the active admittance to the slot voltage, the circuit model presented in Figure 3.5 can be used. In Figure 3.5, the active admittance of the $p^{th}$ slot is

![Figure 3.5: Circuit model of the longitudinal slot on E-plane bend regular waveguide.](image)

41
assumed to be placed between transmission lines. The transmission lines with characteristic impedance, \( G_0 \), represent the E-plane bend regular waveguide before and after the slot. The incident and scattered fields are assumed to be \( TE_{10} \) waveguide modes. \( A_{10} \) is due to the excitation of the waveguide plus the reflections coming from the slots located before the \( p^{th} \) slot and \( D_{10} \) is due to the reflections coming from the succeeding slots. \( B_{10} \) and \( C_{10} \) are the back and forward scattered voltage coefficients. The multi-reflection effects due to \( D_{10} \) are not taken into account. The mode voltage, \( V_p \), can be defined between the upper and lower walls of the waveguide. Hence, by considering the sign change in the \( \hat{a}_\rho \)-component of the electrical field in \( TE_{10} \) mode with respect to the direction of propagation in Appendix A, the following equations can be derived [33],

\[
\frac{Y_p^a}{G_0} = \frac{2B_{10}}{A_{10} - D_{10} - B_{10}} \quad (3.25)
\]

\[
B_{10} = -C_{10} = \frac{1}{2} \frac{Y_p^a}{G_0} V_p \quad (3.26)
\]

Hence, the back scattered mode amplitudes can be equated as

\[
B_{10}^{TE_{10}} = B_{10} = \frac{1}{2} \frac{Y_p^a}{G_0} V_p \quad (3.27)
\]

In order (3.27) to be valid the backward scattered power levels must be the same. The backward scattered power levels can be equated as follows.

\[
\frac{1}{2} |B_{10}|^2 G_0 = \frac{1}{2} \text{Re} \left\{ \int_{S_i} \left( B_p^{TE_{10}}^\ast \overline{E_i}^{TE_{10}} \times \left( B_p^{TE_{10}} \overline{H_i}^{TE_{10}} \right) \right) d\overline{S} \right\} \quad (3.28)
\]
The first active admittance formula due to $TE_{10}^z$ mode scattering can be derived using (3.21), (3.27), and (3.28) as

$$\frac{Y_p}{G_0} = K_1 f_p^{TE_{10}^z} \frac{V_p^s}{V_p}$$ \hspace{1cm} (3.29)

where

$$K_1 = \frac{j\pi}{a} B_{\rho_\rho} \left( k_{\rho_\rho}^{TE_{10}^z} \rho_2 \right) \overline{2\mu_{10}} \left( \alpha \mu G_0 a \right) \left( k_{\rho_\rho}^{TE_{10}^z} \right)^2$$ \hspace{1cm} (3.30)

$$f_p^{TE_{10}^z} = -\left( \frac{\pi}{2\phi_0^z} \right) \cos \left( \mu_{10} \phi_0^z \right) \sin \left( \frac{\pi p}{a} \right)$$ \hspace{1cm} (3.31)

3.1.2.3. Derivation of the Second Active Admittance Formula

The second active admittance formula contains the self-slot admittance and the related mutual coupling terms used to modify the self-slot admittance. In order to obtain the second design equation, the sources of the slot voltage, $V_p^s$ are examined for the $p^{th}$ slot in the waveguide. The slot voltage can be considered as composed of four parts as follows

$$V_p^s = V_{p,1}^s + V_{p,2}^s + V_{p,3}^s + V_{p,4}^s$$ \hspace{1cm} (3.32)

where $V_{p,1}^s$ is the slot voltage due to the $TE_{10}^z$ mode excitation of the waveguide and $V_{p,2}^s$ is the slot voltage excited by the $TE_{10}^z$ mode reflected from the succeeding slots.
$V_{p,3}^s$ is the slot voltage caused by the external couplings of the slot with the other slots while $V_{p,4}^s$ is due to the internal higher order mode couplings in the waveguide.

First of all, the $p^{th}$ slot is assumed to be the only slot in the E-plane bend regular waveguide. Hence, the active admittance in Figure 3.5 becomes a self-slot admittance, $Y_p^s/G_0$, which corresponds to the slot voltage, $V_{p,1}^s$. This case corresponds to a transmission line model with a shunt self-admittance which is excited by the incident voltage with amplitude $A_{i0}$ and scattering coefficients $B_{i0}$ and $C_{i0}$ appear as a result of this excitation. Since there are no other slots in the waveguide $D_{i0} = 0$. Hence using (3.25), the self-admittance is given by

$$\frac{Y_p^s}{G_0} = \frac{2(B_{i0}/A_{i0})}{1-(B_{i0}/A_{i0})} \quad (3.33)$$

The term $(B_p^{TE_{i0}}/A_p^{TE_{i0}})$ is equivalent to $(B_{i0}/A_{i0})$ and if it is used in (3.33), following equation is obtained.

$$B_p^{TE_{i0}} = \frac{Y_p^s}{G_0} A_p^{TE_{i0}} \quad (3.34)$$

Using (3.21) for the relation between $B_p^{TE_{i0}}$ and $V_{p,1}^s$, for the single slot case, it is obtained that

$$V_{p,1}^s = \frac{1}{f_p^{TE_{i0}}} K^{TE_{i0}} \frac{Y_p^s}{G_0} A_p^{TE_{i0}} \quad (3.35)$$
If the other slots exist in the waveguide, there is $TE_{10}^z$ mode incident from $-\hat{a}_\phi$ direction with mode amplitude, $D_{p}^{TE_{10}^z}$ which causes the slot voltage, $V_{p,z}^s$. The relation between $D_{p}^{TE_{10}^z}$ and $V_{p,z}^s$ is given by

$$V_{p,z}^s = -\frac{1}{f_p} \frac{Y_p}{K_{TE_{10}^z}} \frac{G_0}{2} D_{p}^{TE_{10}^z}$$

(3.36)

If all the calculations are carried out till (3.21) to find the back scattering coefficient for $D_{p}^{TE_{10}^z}$ excitation, it will be clear that the back scattering coefficient for $D_{p}^{TE_{10}^z}$ excitation is similar to finding the forward scattering coefficient, $C_{p}^{TE_{10}^z}$, for $A_{p}^{TE_{10}^z}$ excitation. Since $C_{p}^{TE_{10}^z} = -B_{p}^{TE_{10}^z}$, there is a sign difference between (3.35) and (3.36).

Up to now, the slot voltage terms due to $TE_{10}^z$ mode in the waveguide are found. The slot voltages $V_{p,3}^s$ and $V_{p,4}^s$ due to external and internal couplings, respectively, will be derived using reciprocity theorem. In the derivations, the similar procedure presented in [34] and [36] will be used. First of all, only the $p^{th}$ and $q^{th}$ slots are assumed to be present in the slot array as shown in Figure 3.6. In Figure 3.6, the bended regular waveguide is excited by $TE_{10}^z$ mode with amplitude $A_{p}^{TE_{10}^z}$ from Port1. Port2 is terminated by a matched load in order to obtain traveling wave array. As a result of this excitation, the voltages $V_{p}^s$ and $V_{q}^s$ appears between the terminals of the $p^{th}$ and $q^{th}$ slots, respectively. $V_{p}^s$ and $V_{q}^s$ are the voltages which takes all the internal and external couplings into account. If the backward scattered mode amplitude in the $p^{th}$ slot due to the slot voltage $V_{q}^s$ is found then it can be related to the slot voltages $V_{p,3}^s$ and $V_{p,4}^s$ using (3.21).
Two different cases, namely, case “a” and case “b” are defined to apply the reciprocity theorem. In case “a”, the E-plane bend regular waveguide is excited from Port1 when only the $p^{th}$ slot exist in the array while in case “b”, $q^{th}$ slot is assumed to exist in addition to $p^{th}$ slot in the waveguide where only the $q^{th}$ slot is excited which causes the slot voltages $V_{p,3}^s$ and $V_{p,A}^s$ in the $p^{th}$ slot. The reciprocity theorem applied between two cases is given by, [33]

$$\int_V \left( \mathbf{E}^b \cdot \mathbf{J}^a - \mathbf{H}^b \cdot \mathbf{M}^a \right) dV = \int_V \left( \mathbf{E}^a \cdot \mathbf{J}^b - \mathbf{H}^a \cdot \mathbf{M}^b \right) dV$$  \hspace{1cm} (3.37)

where the superscripts “a” and “b” are used to describe case “a” and case “b”, respectively. $\mathbf{J}^a$ and $\mathbf{M}^a$ are the equivalent electric and magnetic currents,
respectively, which are the sources of the fields $\overline{E^a}$ and $\overline{H^a}$ in volume $V$ in case “$a$”. Similarly, $\overline{J^b}$ and $\overline{M^b}$ are the equivalent electric and magnetic currents, respectively, which are the sources of the fields $\overline{E^b}$ and $\overline{H^b}$ in volume $V$ in case “$b$”. The volume $V$ is the same for “$a$” and “$b$” cases. The detailed description of volume $V$, case “$a$” and case “$b$” will be given in the following steps of the derivation.

The sources in “$a$” and “$b$” cases are the incident field at Port1 and the slot voltage, $V^s_q$, respectively, as shown in Figure 3.6. In order to obtain reaction of the incident field at Port1 on the $q^{th}$ slot, the slot voltage $V^s_q$ should be killed in case “$a$”. This is accomplished by covering the $q^{th}$ slot by a conducting tape such that $V^s_q = 0$ as shown in Figure 3.7 (a). In Figure 3.7 (a), case “$a$” is described such that $TE_{10}^a$ mode with amplitude $A^a$ is incident from Port1 while Port2 is terminated by a matched load.

![Figure 3.7](image)

Figure 3.7: (a) Longitudinal slot array with $q^{th}$ slot is covered by conducting tape (case “$a$”). (b) Equivalent problem.
load and \( q^{th} \) slot is shorted. As a result of this excitation, the voltage \( V^{q,a}_p \) is excited at the center of the slot and \( TE_{10}^z \) modes with amplitude \( B^a \) and \( C^a \) are scattered in \(-\hat{\varphi}\) and \( \hat{\varphi}\)-directions, respectively.

The equivalent problem of case “\( a \)” is shown in Figure 3.7 (b) where the equivalent currents on the fictitious surface \( S \) are defined using equivalence principle. The volume \( V \) in case “\( a \)” is defined by the fictitious surface \( S \) as shown in Figure 3.7 (b). The surface \( S \) is composed of three parts which are \( S_1, S_2, \) and \( S_3. \) \( S_1 \) and \( S_2 \) are the surfaces used to describe the cross-section of the bended regular waveguide in front of Port1 and Port2, respectively. The surface \( S_3 \) is located at \( \rho = \rho_1 \) and \( \rho = \rho_2. \) The surface \( S_3 \) extends outside the waveguide and covers the full circumference of the cylinder except the \( p^{th} \) slot surface \( S_p. \) \( S_2 \) contains the \( q^{th} \) slot surface \( S_q \) at both inside and outside the waveguide. Moreover, \( S_3 \) is also defined on the waveguide walls at \( z = z_1 \) and \( z = z_2 \) inside the waveguide and the surface \( S_3 \) extends to infinity in \( \pm \hat{z} \)-directions outside the waveguide. The fields \( \vec{E}^a \) and \( \vec{H}^a \) in case “\( a \)” shown in Figure 3.7 (a) are supported by the equivalent currents \( \vec{J}^a \) and \( \vec{M}^a \) on surface \( S \) in the equivalent problem shown in Figure 3.7 (b). Using the field expressions derived in Appendix A, the equivalent currents \( \vec{J}^a \) and \( \vec{M}^a \) on \( S_1 \) and \( S_2 \) are given by

\[
\vec{J}^a = \begin{cases}
-\frac{1}{z} \cos\left(\frac{\pi}{a}\right) B^{l}_{\mu_0} k_{l_{\rho} \rho} k_{l_{\varphi} \varphi} \hat{a}_z \\
\frac{k_{l_{\rho} \rho}^2}{z} \sin\left(\frac{\pi}{a}\right) B^{l}_{\mu_0} k_{l_{\rho} \rho} \hat{a}_\rho
\end{cases}
\begin{pmatrix}
A^a e^{-j\mu_0 (\varphi - \varphi')} \\
B^a e^{j\mu_0 (\varphi - \varphi')}
\end{pmatrix}
(\rho, \mu, z) \in S_1
(3.38)
\]

\[
\vec{M}^a = \begin{cases}
\frac{j}{\rho} \sin\left(\frac{\pi}{a}\right) B^{l}_{\mu_0} k_{l_{\rho} \rho} \mu_{l_{\varphi} \varphi} \hat{a}_z \\
0
\end{cases}
\begin{pmatrix}
A^a e^{-j\mu_0 (\varphi - \varphi')} \\
B^a e^{j\mu_0 (\varphi - \varphi')}
\end{pmatrix}
(\rho, \mu, z) \in S_1
(3.39)
\]

48
Since the fields are set to zero between Port1 and the surface $S_1$, both $\mathbf{J}^a$ and $\mathbf{M}^a$ are found on $S_1$. In [34], the surface $S_2$ is let to go to infinity in planar waveguides and no obstacles are assumed which causes reflections. Hence, no sources are assumed from the matched side of the waveguide. In Figure 3.7 (b), matched load is used at Port2 which avoids the reflections, and hence, there are no sources incident from $-\hat{\phi}$-direction. Since the fields between the surface $S_2$ and Port2 exist both in Figure 3.7 (a) and (b) due to the matched load at Port2, the equivalent currents $\mathbf{J}^a$ and $\mathbf{M}^a$ on $S_2$ are found as zero. Regarding the equivalent currents on $S_3$, it can be stated that only equivalent electric currents ($\mathbf{J}^a$) exist on $S_3$ because the tangential electric field in case “$a$”, $E_{\text{tan}}^a = 0$ on $S_3$. It will be understood that $\mathbf{J}^a$ on $S_3$ need not to be determined in applying the reciprocity theorem. Moreover, only the field $\mathbf{H}^a$ is required to be determined on the surface $S_q$.

In case “$b$”, Port1 in the slot array shown in Figure 3.6 is not excited and it is terminated by a matched load. Case “$b$” is described as in Figure 3.8 (a). In Figure 3.8 (a), the $q^{th}$ slot is excited by slot voltage, $V_q^{1,b}$ while Port1 and Port2 are terminated by a matched load. As a result of this excitation, $TE_{10}^q$ modes with amplitude $B_{\text{all}}^b$ and $C_{\text{all}}^b$ are scattered in $-\hat{\phi}$ and $\hat{\phi}$-directions, respectively.

The equivalent problem for case “$b$” is shown in Figure 3.8 (b). The surface $S$ shown in Figure 3.8 (b) is exactly the same as the surface presented in Figure 3.7 (b). Hence, the volumes $V$ in case “$a$” and “$b$” defined by the surface $S$ are the same and the reciprocity theorem given in (3.37) can be applied between “$a$” and “$b$” cases. In the equivalent problem of case “$b$” shown in Figure 3.8 (b), since the fields $E^b$ and $H^b$ exist between Port1 and surface $S_1$ and also between Port2 and surface $S_2$, $\mathbf{J}^b = \mathbf{M}^b = 0$ on $S_1$ and $S_2$. Moreover, only the equivalent electric currents ($\mathbf{J}^b$) exist on the surface $S_3$ because the tangential electric field in case “$b$”, $E_{\text{tan}}^b = 0$ on $S_3$. Equivalent magnetic sources are placed external and internal to the waveguide on
the surface of the $q^{th}$ slot, $S_q$, and the $q^{th}$ slot is covered by PEC. The equivalent magnetic currents located exterior and interior of the bended regular waveguide are indicated as $\mathcal{M}_{q,\text{ext}}^b$ and $\mathcal{M}_{q,\text{int}}^b$, respectively, in Figure 3.8 (b).

The field components of $\mathbf{E}^b$ and $\mathbf{H}^b$ are assumed to be single mode which is $TE_{10}^z$ mode in the bended regular waveguide. The values of $\mathbf{E}^b$ and $\mathbf{H}^b$ on $S_2$ and $S_3$ are not needed in applying the reciprocity theorem, and hence, $\mathbf{E}^b$ and $\mathbf{H}^b$ fields on $S_i$ are given by

$$E^b_\rho = -\frac{j}{\rho} \sin \left( \frac{\pi z}{a} \right) B_{\rho_0} \left( k_{\rho_0} \rho \right) \mu_{10} B_{\text{eff}}^b e^{j\rho_0(\phi-\varphi)} \right), \ (\phi, \rho, z) \in S_i \tag{3.40}$$
\[ E^b_\phi = \sin \left( \frac{\pi}{a} \right) B'_{\rho_0} \begin{pmatrix} k_{\rho}^{\text{TE}_{10}} \\ \rho \end{pmatrix} k_{\rho}^{\text{TE}_{10}} B^b_{\text{all}} e^{j\mu_0 (\theta - \phi')} \), (\phi, \rho, z) \in S_i \] (3.41)

\[ E^b_z = 0 \), (\phi, \rho, z) \in S_i \] (3.42)

\[ H^b_\rho = \frac{1}{2} \frac{\pi}{a} \cos \left( \frac{\pi}{a} \right) B'_{\rho_0} \begin{pmatrix} k_{\rho}^{\text{TE}_{10}} \\ \rho \end{pmatrix} k_{\rho}^{\text{TE}_{10}} B^b_{\text{all}} e^{j\mu_0 (\theta - \phi')} \), (\phi, \rho, z) \in S_i \] (3.43)

\[ H^b_\phi = j \frac{1}{2} \rho \frac{\pi}{a} \cos \left( \frac{\pi}{a} \right) B'_{\rho_0} \begin{pmatrix} k_{\rho}^{\text{TE}_{10}} \\ \rho \end{pmatrix} B^b_{\text{all}} e^{j\mu_0 (\theta - \phi')} \), (\phi, \rho, z) \in S_i \] (3.44)

\[ H^b_z = \frac{(k_{\rho}^{\text{TE}_{10}})^2}{4} \sin \left( \frac{\pi}{a} \right) B'_{\rho_0} \begin{pmatrix} k_{\rho}^{\text{TE}_{10}} \\ \rho \end{pmatrix} B^b_{\text{all}} e^{j\mu_0 (\theta - \phi')} \), (\phi, \rho, z) \in S_i \] (3.45)

The reciprocity theorem given in (3.37) is applied between case “a” and case “b”. If it is considered that all the equivalent currents are defined on the surface \( S \), only the tangential field components on the surface \( S \) are required to be determined. Moreover, (3.37) reduces to the surface integral in the regions where the current sources are located. Hence, (3.37) becomes

\[ \int_S (\overline{E}^b \cdot \overline{J}^a - \overline{H}^b \cdot \overline{M}^a) dS = \int_S (\overline{E}^a \cdot \overline{J}^b - \overline{H}^a \cdot \overline{M}^b) dS , \quad S = S_1 \cup S_2 \cup S_3 \] (3.46)

Since \( \overline{J}^a = \overline{M}^a = 0 \) on \( S_2 \), \( \overline{M}^a = 0 \) on \( S_3 \), and \( \overline{E}^b_{\text{un.g}} = 0 \) on \( S_1 \), the left hand side (LHS) of (3.46) becomes

\[ \int_S (\overline{E}^b \cdot \overline{J}^a - \overline{H}^b \cdot \overline{M}^a) dS = \int_{S_1} (\overline{E}^b \cdot \overline{J}^a - \overline{H}^b \cdot \overline{M}^a) dS \] (3.47)
In (3.46), \( \int_{S} \left( E^a \mathcal{J}^b \right) dS = 0 \) because \( E^a_{\text{tan},g} = 0 \) on \( S_3 \) and \( \mathcal{J}^b = 0 \) on \( S_1 \) and \( S_2 \).

Moreover, \( \int_{S} \left( H^a \cdot \mathcal{M}^b \right) dS = \int_{S_q} \left( \mathcal{H}^a_{\text{ext}} \cdot \mathcal{M}^b_{q,\text{ext}} + \mathcal{H}^a_{\text{int}} \cdot \mathcal{M}^b_{q,\text{int}} \right) dS_q \) because \( \mathcal{M}^b_{q,\text{ext}} = \mathcal{M}^b_{q,\text{int}} = 0 \) on \( S_1, S_2 \), and \( S_3-S_q \) where \( \mathcal{H}^a_{\text{int}} \) and \( \mathcal{H}^a_{\text{ext}} \) are the magnetic fields inside and outside the bended regular waveguide in case "a". Hence, RHS of (3.46) becomes

\[
\int_{S} \left( E^a \mathcal{J}^b - H^a \cdot \mathcal{M}^b \right) dS = -\int_{S_q} \left( \mathcal{H}^a_{\text{ext}} \cdot \mathcal{M}^b_{q,\text{ext}} + \mathcal{H}^a_{\text{int}} \cdot \mathcal{M}^b_{q,\text{int}} \right) dS_q \quad (3.48)
\]

Equating RHS of (3.47) and RHS of (3.48), (3.37) becomes

\[
\int_{S_q} \left( E^b \mathcal{J}^a - H^a \cdot \mathcal{M}^b \right) dS_1 = -\int_{S_q} \left( \mathcal{H}^a_{\text{ext}} \cdot \mathcal{M}^b_{q,\text{ext}} + \mathcal{H}^a_{\text{int}} \cdot \mathcal{M}^b_{q,\text{int}} \right) dS_q \quad (3.49)
\]

In (3.49), only the \( q^{th} \) slot is assumed in addition to the \( p^{th} \) slot in the waveguide. If the other slots are also assumed to exist in the waveguide, (3.49) can be generalized as, [34], [36],

\[
\int_{S_q} \left( E^b \mathcal{J}^a - H^a \cdot \mathcal{M}^b \right) dS_1 = -\sum_{q=1}^{N} \int_{S_q} \left( \mathcal{H}^a_{\text{ext}} \cdot \mathcal{M}^b_{q,\text{ext}} \right) dS_q - \sum_{q=1}^{M} \int_{S_q} \left( \mathcal{H}^a_{\text{int}} \cdot \mathcal{M}^b_{q,\text{int}} \right) dS_q \quad (3.50)
\]

where \( M \) is the total number of the slots in the same waveguide and \( N \) is the total number of the slots in the array. The prime signs on the summations in (3.50) indicates that the case \( q = p \) is excluded from the summations.

Inserting (3.38), (3.39), and (3.40)-(3.45) into (3.50), the expression for the backward scattered mode amplitude, \( B_{all}^b \) is derived for the \( p^{th} \) slot as...
\[ B_{all}^b = \frac{1}{jaN_{\mu_0}\mu_0} \left( \left( k_{TE_{10}}^2 \right) / 2 \right) \left[ \sum_{q=1}^{N} \int_{S_q} \left( \frac{H_{\text{ext}}^a}{A^a} M_q^b \right) dS_q + \sum_{q=1}^{M} \int_{S_q} \left( \frac{H_{\text{int}}^a}{A^a} M_q^b \right) dS_q \right] \]  

(3.51)

where \( H_{\text{int}}^a \) is composed of two parts for the scattered fields in case “a” given by

\[ H_{\text{int}}^a = H_{\text{int}}^{a1} + H_{\text{int}}^{a2} \]  

(3.52)

where \( H_{\text{int}}^{a1} \) and \( H_{\text{int}}^{a2} \) are \( TE_{10}^z \) and \( TE_{m0}^z \) mode components of \( H_{\text{int}}^a \), respectively.

\( B_{all}^b \) can be divided into four parts as follows

\[ B_1^b = \frac{1}{jaN_{\mu_0}\mu_0} \left( \left( k_{TE_{10}}^2 \right) / 2 \right) \sum_{q=1}^{p-1} \int_{S_q} \left( \frac{H_{\text{int}}^{a1}}{A^a} M_q^b \right) dS_q \]  

(3.53)

\[ B_2^b = \frac{1}{jaN_{\mu_0}\mu_0} \left( \left( k_{TE_{10}}^2 \right) / 2 \right) \sum_{q=p+1}^{M} \int_{S_q} \left( \frac{H_{\text{int}}^{a1}}{A^a} M_q^b \right) dS_q \]  

(3.54)

\[ B_3^b = \frac{1}{jaN_{\mu_0}\mu_0} \left( \left( k_{TE_{10}}^2 \right) / 2 \right) \sum_{q=1}^{N} \int_{S_q} \left( \frac{H_{\text{ext}}^a}{A^a} M_q^b \right) dS_q \]  

(3.55)

\[ B_4^b = \frac{1}{jaN_{\mu_0}\mu_0} \left( \left( k_{TE_{10}}^2 \right) / 2 \right) \left[ \int_{S_{p-1}} \frac{H_{\text{ext}}^a}{A^a} M_{p-1,int}^b dS_{p-1} + \int_{S_{p-1}} \frac{H_{\text{int}}^{a2}}{A^a} M_{p-1,int}^b dS_{p-1} \right] \]  

(3.56)

where \( B_1^b \) is the internal \( TE_{10}^z \) backward scattering mode amplitudes due to the \( TE_{10}^z \) mode excitation of the slot array and also due to the \( TE_{10}^z \) mode reflections of all the slots before the \( p^{th} \) slot. \( B_2^b \) is the internal \( TE_{10}^z \) backward scattering mode amplitudes
due to the \( TE_{10}^z \) mode reflections of the slots located after the \( p^{th} \) slot. The effect of \( B_1^b \) and \( B_2^b \) has already been treated in (3.35) and (3.36). \( B_1^b \) and \( B_4^b \) are the backward scattering \( TE_{10}^z \) mode amplitudes due to the external and internal couplings, respectively. \( B_3^b \) is caused by the slot voltage, \( V_{p,3}^s \). Using (3.55), \( V_{p,3}^s \) can be derived.

In (3.55), \( \overline{H}_{ext}^a \) is the magnetic field radiated by the slot with slot voltage \( V_p^{s,a} \). The equivalent magnetic source for the \( q^{th} \) slot is given by

\[
\overline{M}_{q,ext}^b = \frac{V_q^{s,b}}{w^z} \cos \left( \frac{\pi (\phi - \phi_q^a)}{2\phi_q^a} \right) \hat{a}_\phi
\]

where \( V_q^{s,b} \) is the slot voltage of the \( q^{th} \) slot in case “b”. Since \( \overline{M}_{q,ext}^b \) has only \( \hat{a}_\phi \) component, only \( \hat{a}_\phi \) component of \( \overline{H}_{ext}^a \) is needed in (3.55). The \( \hat{a}_\phi \) component of \( \overline{H}_{ext}^a \) is given using the derivations presented in Appendix B, [38] as

\[
H_{ext,a}(\rho_2,\phi,z) = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{in\phi} \int_{-\infty}^{\infty} \left\{ \frac{f}{\sqrt{k_0^2 - k_z^2}} \right\} \left\{ \begin{array}{c} \left( \frac{n k_z}{\rho_2} \right)^2 \frac{1}{\omega \mu_0 (k_0^2 - k_z^2) W(n,k_z)} \\ -\omega \epsilon_0 W(n,k_z) \end{array} \right\} e^{i k_z z} dk_z
\]

where

\[
W(n,k_z) = \frac{H_n^{(2)}(\rho_2 \sqrt{k_0^2 - k_z^2})}{H_n^{(2)}(\rho_2 \sqrt{k_0^2 - k_z^2})}
\]
and where

\[
\tilde{E}_z^a(n, k_z) = \frac{1}{\pi} \frac{V_{p}^{x,a} e^{-j\kappa z}}{w^r} \sin \left( \frac{k_z w^r}{2} \right) e^{-j\phi^p} \cos(n\phi_0^p) \left[ \frac{\pi}{\phi_0^p} \left( \frac{\pi}{2\phi_0^p} - n^2 \right) \right]
\]  
(3.60)

Hence inserting (3.57) and (3.58) into (3.55), and carrying out the calculations, (3.55) becomes

\[
B^b_j = \frac{1}{j a N_{\mu_0} \mu_0 \left( k_{z_{\infty}} \right)^2 \frac{1}{2}} \sum_{q=1}^{N} \frac{V_{q}^{x,b} V_{p}^{x,a}}{A^a} T_{pq}
\]  
(3.61)

where

\[
T_{pq} = \sum_{n=-\infty}^{\infty} F_{pq}(n) G_{z}^{pq}(n)
\]  
(3.62)

and where

\[
F_{pq}(n) = e^{j(\phi^q-\phi^p)} \cos(n\phi_0^q) \cos(n\phi_0^p) \left[ \frac{1}{\phi_0^q} \left( \frac{\pi}{2\phi_0^q} - n^2 \right) \right] \left[ \frac{1}{\phi_0^p} \left( \frac{\pi}{2\phi_0^p} - n^2 \right) \right]
\]  
(3.63)

\[
G_{z}^{pq}(n) = \int_{0}^{\infty} \frac{j \cos(k_z (z^q - z^r))}{2\sqrt{k_0^2 - k_z^2}} \left[ \sin \left( \frac{k_z w^r}{2\pi} \right) \right]^2 \left[ \frac{(nk_z)^2}{\omega \mu_0 \rho_2 k_0^2 k_z^2 W(n, k_z)} \right] \frac{1}{\omega \varepsilon_0 \rho_2 W(n, k_z)} dk_z
\]  
(3.64)
In (3.61), $V_q^{s,b}$ is the slot voltage of the $q^{th}$ slot in the slot array, i.e., when all the slots exist in the waveguide array. Hence the superscript “$b$” can be dropped in $V_q^{s,b}$.

Moreover, it is clear that $V_{p,1}^{s,a} = V_p^{s}$ and the ratio, $V_{p,1}^{s,a} / A^a$ in (3.61) can be found using (3.35). Note that $A_{p}^{TE_{01}} = A^a$ in (3.61) because $A^a$ is $TE_{10}^z$ mode voltage amplitude used to excite the waveguide in the absence of the other slots as described in case “$a$”. When these changes are made in (3.61) and also using (3.21) for the relation between $V_{p,3}^{s}$ and $B_3^b$, $V_{p,3}^{s}$ is found as

$$V_{p,3}^{s} = \frac{\hat{z}N_{\mu_{10}} \rho_1^3 \left(k_{p}^{TE_{01}}\right)^2}{B_{\mu_{10}} \left(k_{p}^{TE_{01}}\rho_2\right)} \frac{Y_p^{s}}{G_0} \sum_{q=1}^{N} V_q^{s} T_{pq}^{*} \left(3.65\right)$$

The last voltage value to be found is $V_{p,4}^{s}$ which is due to the internal higher order mode couplings. The internal couplings is derived for the general case of $TE_{m0}^z$ couplings because the effect of the internal couplings becomes significant when the height of the waveguide is small, [36] and many $TE_{m0}^z$ modes appear before $TM_{mn}^z$ modes as shown in Table 3.1. Similar procedure described in [36] is used to derive the internal couplings of the slots. The backward scattering $TE_{10}^z$ mode amplitude, $B_4^b$ due to the $TE_{m0}^z$ couplings is derived in (3.56). As a result of the internal couplings, the slot voltage, $V_{p,4}^{s}$ induces across the terminals of the slot. Hence, the backward scattering $TE_{10}^z$ mode amplitude, $B_4^b$ is scattered due to the slot voltage $V_{p,4}^{s}$ . Note that in deriving (3.56) from (3.51), the expression for $B_4^b$ is found only for the neighbors of the $p^{th}$ slot for simplicity. The equivalent magnetic currents, $\overline{M}_{p-1,\text{int}}^{b}$ and $\overline{M}_{p+1,\text{int}}^{b}$, in (3.56) are given by
\[
\overline{M}_{p-1,\text{int}}^b = -\frac{V_p^s}{w^s} \cos \left( \frac{\pi(\phi - \phi^{p-1})}{2\phi_0^{p-1}} \right) \hat{a}_\phi 
\]
(3.66)

\[
\overline{M}_{p+1,\text{int}}^b = -\frac{V_{p+1}^s}{w^s} \cos \left( \frac{\pi(\phi - \phi^{p+1})}{2\phi_0^{p+1}} \right) \hat{a}_\phi 
\]
(3.67)

Note that similar to the external coupling derivations using (3.35), the superscript “\(b\)” is dropped in \(V_{p-1}^s\) and \(V_{p+1}^s\) in (3.66) and (3.67), respectively. Since \(\overline{M}_{p-1,\text{int}}^b\) and \(\overline{M}_{p+1,\text{int}}^b\) are \(\hat{a}_\phi\) - directed, only \(\hat{a}_\phi\) - components of \(\overline{H}^{a,2}\) are needed in (3.56) and they are given by

\[
H_{a,2}^{\text{int,}\phi} = \begin{cases} 
B_p^{\text{TE}_a} \frac{j}{2} \frac{m\pi}{a} \cos \left( \frac{m\pi z}{a} \right) B_{p,a0} \left( k_{p,a0} \rho_2 \right) \mu_{m0} e^{j\mu_{a0}(\phi-\phi')} \chi \in S_{p-1} \\
-C_p^{\text{TE}_a} \frac{j}{2} \frac{m\pi}{a} \cos \left( \frac{m\pi z}{a} \right) B_{p,a0} \left( k_{p,a0} \rho_2 \right) \mu_{m0} e^{-j\mu_{a0}(\phi-\phi')} \chi \in S_{p+1} 
\end{cases} 
\]
(3.68)

Using (3.20) and (3.21) for the calculation of \(C_p^{\text{TE}_a}\) and \(B_p^{\text{TE}_a}\), respectively, and inserting (3.66)-(3.68) into (3.56), it is found after lengthy derivations that

\[
B_4 = \frac{1}{j a N_{\mu_0} \mu_{10} \left( k_{p,a0}^{\text{TE}_a} \right)^2 \chi^2} \frac{V_p^s}{w^s} \sum_{m=2}^{T} h_{p,m}^{*} \left[ h_{p-1,m}^{*} V_{p-1}^s + h_{p+1,m}^{*} V_{p+1}^s \right] 
\]
(3.69)

where

\[
h_{p,m}^{*} = K^{\text{TE}_a} f_p^{\text{TE}_a} 
\]
(3.70)

\[
h_{p-1,m}^{*} = \frac{j}{2} \frac{m\pi}{a} \mu_{m0} B_{p,a0} \left( k_{p,a0}^{\text{TE}_a} \rho_2 \right) \sin d \left( \frac{m w^s}{2a} \right) e^{j\mu_{a0}(\phi^{p-1}-\phi')} f_p^{\text{TE}_a} 
\]
(3.71)
\[ h'_{p+1, m} = \frac{j \pi}{2} \frac{m \pi}{a} \mu_{m0} B_{\mu m0} \left( k_{p}^{TE} \rho_{2} \right) \sin \left( \frac{m \pi}{2a} \right) e^{-j \mu_{m0} (\theta^{p+1} - \phi^v)} f_{p+1}^{TE} \quad (3.72) \]

Using the relation between \( B_{4}^{b} \) and \( V_{p,4}^{s} \) given in (3.21) for \( TE_{10}^{z} \) mode and also using (3.35) for the ratio \( V_{p}^{s,a}/A^{a} \), \( V_{p,4}^{s} \) is derived using (3.69) as

\[ V_{p,4}^{s} = -\frac{z^a}{j \mu_{10} \pi^2} \hat{a} \left( k_{p}^{TE} \right)^2 N_{p}^{TE} \left( k_{p}^{TE} \rho_{2} \right)^2 \left( \frac{1}{G_{0}} \sum_{m=2}^{T} h_{p,m}^{s} \left[ h'_{p-1,m} V_{p-1}^{s} + h'_{p+1,m} V_{p+1}^{s} \right] \right) \quad (3.73) \]

The active admittance formula of the shunt slot in terms of the incident and scattered mode amplitudes in the circuit model can be derived as [33], [34]

\[ \frac{Y_{p}^{a}}{G_{0}} = \frac{2 B_{10}^{p}}{A_{10}^{p} - D_{10}^{p} - B_{10}^{p}} \quad (3.74) \]

where \( A_{10}^{p} \) and \( D_{10}^{p} \) are the incident mode voltage amplitudes while \( B_{10}^{p} \) is the scattered mode voltage amplitude for the \( p^{th} \) slot as described in Figure 3.5. In (3.74), \( A_{10}^{p} \) and \( D_{10}^{p} \) can be replaced by \( A_{p}^{TE} \) and \( D_{p}^{TE} \) while \( B_{10}^{p} \) can be replaced by \( B_{all}^{p} \). The values of \( A_{p}^{TE} \) and \( D_{p}^{TE} \) can be found from (3.35) and (3.36), respectively. Also using (3.21) to obtain \( B_{all}^{b} = K^{TE} f_{p}^{TE} V_{p}^{s} \), where \( V_{p}^{s} = V_{p,1}^{s} + V_{p,2}^{s} + V_{p,3}^{s} + V_{p,4}^{s} \) derived in this section, one can arrive at the second active admittance formula given by
\[
\frac{Y_p^a}{G_0} = \frac{2f_p^{TE_{m0}}}{2N_{\mu_{mn}}a^3 \left( k_p^{TE_{m0}} \right)^2} \left( \sum_{q=1}^{N} \frac{V_q}{V_p} \frac{T_{pq}}{V_p} - \sum_{m=2}^{T} h_{p,m} \left[ h_{p-1,m} \frac{V_{p-1}}{V_p} + h_{p+1,m} \frac{V_{p+1}}{V_p} \right] \right)
\]  

(3.75)

In (3.75), \( T \) is the total number of higher order \( TE_{m0} \) modes before \( TM_{mn} \) modes excite. It is usually enough to consider only the \( TE_{20} \) higher order mode couplings. Hence, \( T \) is taken as 2 in (3.75) for the calculation of the active admittances carried out in this thesis.

### 3.2. Design Procedure of Slotted CSIW Array

The active admittances derived in Section 3.1 can be used to design slotted CSIW arrays. In this study, the design procedure presented by Elliott [34], [40] for the planar slotted waveguides is used. There are two types of planar slotted waveguide arrays, namely, standing wave and traveling wave slot array which can be designed using the active admittance formulations. Since a traveling wave type of array is considered to be implemented on CSIW, the design procedure for the traveling wave type of slot waveguide array is outlined which can be used for CSIW case. The traveling wave slot array is excited from one side and the other side of the array is terminated by a match load. The slots are usually placed in an alternating manner as shown in Figure 1.2 which causes a phase shift of \( \pi \) in addition to the phase delay between the succeeding slots. The spacing between the slot elements in the design is slightly greater or smaller than \( \frac{\lambda_g}{2} \) in order to prevent the reflections at each slot to add up, and hence, to prevent the mismatch. Since the slots are not separated by \( \frac{\lambda_g}{2} \), allocating alternating slots does not help to have the same phase for all the slots. Hence beam is tilted from broadside of the array. If the slots are placed in an
alternating manner, the element spacing lower than $\lambda_g/2$ and greater than $\lambda_g/2$ corresponds to backward firing and forward firing radiation cases, respectively [33].

The admittance of the single longitudinal slot is the self-admittance of the slot characterized in Section 2.3. For the $p^{th}$ slot in the waveguide, the self-admittance is abbreviated as $Y_p^a/G_0$ in (3.75). The self-admittance is usually found by full wave simulators such as HFSS and then characterization polynomials are derived. The characterization polynomials are used in the design of the slotted waveguide array. When the other slots are present in the array, the slot admittance is presented by the active shunt admittances that include the effects of mutual couplings with the other slots as in (3.75). The circuit model of the traveling wave array composed of longitudinal slots on the broad wall of the waveguide is shown in Figure 3.9. The circuit model is constructed by using the normalized active slot admittances, $Y_p^a$ for $p = 1, 2, \ldots, M$ where $M$ is the total number of the slots in the same waveguide with spacing, $d$, in the array. The slot separations need not necessarily to be taken as uniform and the design procedure can be modified for non-uniform slot separation values. The normalization of the active slot admittances are carried out with respect to the characteristic impedance of the waveguide. The power-voltage definition of the characteristic impedance of the waveguide is used in normalizing the active slot admittances [41]. Since the reflections from the end of the circuit are not desired in the traveling wave array, the circuit model is terminated by a match load at the end.

Figure 3.9: Circuit model representation of a traveling wave shunt slot array.
The design of the slotted waveguide array is based on two equations derived in Section 3.1.2. Using (3.29), the ratio of the active admittances of two slots in the waveguide can be written as

\[
\frac{Y_p^a(z_p, l_p)}{Y_q^a(z_q, l_q)} = \frac{f_p^{TE}(z_p, l_p)Y_p}{f_q^{TE}(z_q, l_q)Y_q^*}
\]  

(3.76)

where \(z_p\) is the slot offset and \(l_p = \rho_2 \phi_0^p\) is the half-length of the \(p^{th}\) slot, respectively.

Total admittance seen by the \(p^{th}\) slot is given by

\[
Y_p = Y_p^a + \frac{Y_{p-1} \cos(\beta_{10}d) + j \sin(\beta_{10}d)}{\cos(\beta_{10}d) + j Y_{p-1} \sin(\beta_{10}d)}
\]  

(3.77)

where \(Y_p\) is total admittance seen from right in the region of the \(p^{th}\) slot shown in Figure 3.9 and \(\beta_{10}\) is the propagation constant of the fundamental mode in the waveguide.

The complex mode voltages of consecutive slots are related by

\[
V_p = V_{p-1} \left[ \cos(\beta_{10}d) + j Y_{p-1} \sin(\beta_{10}d) \right]
\]  

(3.78)

Using (3.77) and (3.78), (3.76) can be rewritten for two successive slots as

\[
\frac{Y_p^a(z_p, l_p)}{f_p^{TE}(z_p, l_p)} = \frac{Y_{p-1}^a(z_{p-1}, l_{p-1})}{f_{p-1}^{TE}(z_{p-1}, l_{p-1})} \frac{V_{p}^*/V_{p-1}}{\cos(\beta_{10}d) + j Y_{p-1} \sin(\beta_{10}d)}
\]  

(3.79)

The design procedure can be summarized as follows:
• The ratio of the complex slot voltages $V_{p}^{s}/V_{q}^{s}$ are determined by the pattern requirements, such as low SLL or desired beam shape. The phases of the slot voltages should be taken into wherever needed, such as in the mutual coupling calculations and in (3.79).

• The spacing $d$ between the elements is determined by the pattern requirements such as direction of the maximum radiation and avoiding grating lobes. Small $d$ may cause strong mutual coupling that might not allow making a design.

• Mutual coupling terms in (3.75) are calculated assuming zero slot offsets and resonant slot lengths for all the slots in the array.

• An initial value for the last slot offset $z_{1}$ is chosen. It will affect the values of the other slot offsets and determine the power dissipated at the load.

• The mode voltage at the first slot is taken as unity. In case of multiple parallel waveguides fed by coherent modes, the phase of the mode voltage is determined by the phase of the mode at the input port of the waveguide. If the first slot is desired to be at resonance, the phase value for the slot which is the same as the phase of the mode voltage is used. Although there are complex multiplicative terms in (3.29), this convention can still be used which causes a phase shift with the same amount for all the slots in all the waveguides. Slot length $l_{1}$ is determined by diminishing the imaginary part of (3.75) if the first slot is desired to be at resonance. If the first slot is not at resonance, $l_{1}$ is determined such that the required phase value is satisfied in (3.75). The amount of deviation of the slot from resonance is determined by the deviation of the phase of the slot voltage from the phase of the mode voltage. Using the slot offset, $z_{1}$ and length, $l_{1}$, the self-admittance of the slot is determined by the characterization polynomials. Hence, $Y_{i}^{a}(z_{1},l_{1})/f_{i}^{TE_{0}}(z_{1},l_{1})$ is determined. $Y_{i}$ is found using (3.77). The value of $Y_{2}^{a}(z_{2},l_{2})/f_{2}^{TE_{0}}(z_{2},l_{2})$ is determined using (3.79).
• Next step is to search for \((z_2, l_2)\) such that the same \(Y^s_2(z_2, l_2)/f_2^{TEu}(z_2, l_2)\) value is found as the previous step using (3.75). The error function is formed by taking the absolute value of the difference between two \(Y^s_2(z_2, l_2)/f_2^{TEu}(z_2, l_2)\) values found using different equations, i.e., using (3.75) and (3.79). The error function is minimized to find the best slot offset and length pair which satisfies desired active admittance value.

• The procedure is repeated until all \((z_N, l_N)\) pairs are found.

• The mutual coupling terms in (3.75) are recalculated with new set of slot offsets and lengths. The procedure listed above is repeated until final slot offsets and lengths converge.

• If the value of \(Y_m\) does not give good matching, the separation \(d\) between the slots can be altered slightly and the design procedure should be repeated.

After finding all the slot offsets and lengths, power delivered load is calculated using the power dissipated at the \(p^{th}\) slot which is given by

\[
P_p = \frac{1}{2} \text{Re} \left( V_p V_p^* \left(\frac{Y_p}{G_0}\right)^* \right) \quad (3.80)
\]

and power delivered to load can be written as

\[
P_L = \frac{1}{2} \frac{V_1 V_1^*}{V_1 V_1^* + \sum_{p=1}^{M} P_p} \quad (3.81)
\]

If the power delivered to the load is greater than the desired limit, the value of the initial slot offset, \(z_1\) should be increased and the design procedure should be repeated.
3.3. Numerical Results

The numerical results are presented to examine the mutual couplings of the slots using the formulations carried out in Section 3.1. The validity of the Elliott’s design procedure has already been proved in the literature for the slotted planar waveguides and also in Chapter 4 for the slotted CSIW. In this part, the derived mutual coupling formulas are compared with the planar formulations, formulations of previous studies, and HFSS simulations.

There are some difficulties in the internal and external coupling calculations. In the calculation of the internal couplings, the evaluations of the Bessel functions with complex orders are required in some of the terms such as in (3.24) and (A.31). Such kind of calculations is carried out using the Symbolic Math Toolbox in MATLAB. The calculation of the external couplings requires the numerical integral to be taken over complex domain because the term, $G_{pq}^n(\zeta)$ in (3.64) has poles at $\zeta = k_0$. In order to avoid pole and branch point singularities, the integral path is deformed as shown in Figure 3.10. In order to avoid computer underflow and overflow while carrying out the calculations of the Hankel function terms in (3.64) are written in the form of ratios as in (3.59) and the asymptotic forms for the ratios of the Hankel

![Deformed integration path](image)

Figure 3.10: Deformed integration path.
functions of large orders and arguments derived in Appendix C are used wherever needed in the integration.

The path of integration shown in Figure 3.10 is divided into three regions, similar to the ones presented in [42] - [44]. Three regions are represented as

\[
\Gamma_1 : k_z = k_0 \frac{(1 + jT_1)T_1}{T_1}, \quad 0 \leq t_1 \leq T_1
\]

\[
\Gamma_2 : k_z = k_0 \left[ 1 + jT_1 + \left( \sqrt{1 + T_2} - (1 + jT_1) \right) \frac{T_2}{T_2 - T_1} \right], \quad 0 \leq t_2 \leq T_2 - T_1
\]

\[
\Gamma_3 : k_z = k_0 \left[ \sqrt{1 + T_2} + \left( \sqrt{1 + T_3} - \sqrt{1 + T_2} \right) \frac{T_3}{T_3 - T_2} \right], \quad 0 \leq t_3 \leq T_3 - T_2
\]

where \( \Gamma_1, \Gamma_2, \) and \( \Gamma_3 \) are paths of integration shown in Figure 3.10.

Three regions in Figure 3.10 are uniformly sampled. The mutual coupling term presented in (3.62) is computed by assuming resonant slots at 25 GHz. Two different codes are generated to calculate (3.62) at 25 GHz. The number of \( n \) values to be summed at each step is kept constant in both of the codes. In the first code, the value of \( T_3 \) is kept constant and the summation is carried out over \( n \) in (3.62) till convergence is obtained. In the second code, the value of \( T_3 \) is extended at each summation step and the summation is also carried out for previous and new \( n \) values. The first code is preferred to be used in the calculations because it is faster due to the constant value of \( T_3 \). Two codes are compared and the parameters are determined for the first code as \( T_1 = 0.1, T_2 = 1, T_3 = 30 \), and maximum number of terms in (3.62) to be summed over \( n \) at each step is determined as 200 including positive and negative \( n \) values.
3.3.1. External Couplings of Two Slots on PEC Cylinder

The external coupling formulations derived for the slots on cylindrical surface in this thesis are compared with the literature. For this purpose, two slots are assumed to be placed on PEC cylinder as shown in Figure 3.11. Mutual admittance of the slots are calculated and compared with the results presented in [45]. The widths of the slots in [45] shown Figure 3.11 are \( w_p = w_q = 0.4 \) in and lengths of the slots are \( L_p = L_q = 0.9 \) in. The radius of the cylinder is \( \rho_2 = 1.991 \) in. The angular displacement between the centers of the slots is \( \phi_0 \) while the displacement between the centers of the slots in \( \hat{a}_z \)-direction is \( z_0 \). The mutual admittance calculation frequency is 9 GHz. Similar to the mutual admittance definition given in [45], the mutual admittance of the slots on PEC cylinder is defined as

\[
Y_{pq} = -\frac{1}{V_q V_p} \iint_{S_q} \overline{\left( \overline{H}^p \cdot \overline{M}_q \right)} d\mathbf{S}_q \tag{3.85}
\]

![Figure 3.11: Two slots on cylindrical PEC.](image)

66
where $V_p^s$ and $V_q^s$ are the slot voltages while $\overline{H}_p^p$ is the magnetic field radiated by the $p^{th}$ slot. $\overline{M}_q^q$ is the equivalent magnetic current of the $q^{th}$ slot defined on the surface of the $q^{th}$ slot, $S_q$. The magnitude of the mutual admittance in dB scale is given by

$$\left( Y_{pq} \right)_{dB} = 20 \log_{10} \left| Y_{pq} \right|$$

(3.86)

The mutual admittance values presented by (3.85) are derived for cylindrical PEC case by using the formulations similar to the ones derived in Section 3.1.2.3. It is found that the mutual admittance given in (3.85) equals to the minus of the external mutual coupling term, $-T_{pq}$, presented in (3.62). Hence, comparing the mutual admittance calculations with the literature also tests the accuracy of the external coupling formulations derived in this thesis.

Modal solution, asymptotic solution and geometrical theory of diffraction (GTD) solution for the mutual admittance of the slots are derived in [45]. Modal solution is the exact solution. In [45], the mutual admittance of the slots is calculated for various $\phi_0$ and $z_0$ values at 9 GHz using modal solution, asymptotic solution, and GTD solution and they are listed in Table 3.2, Table 3.3, and Table 3.4. In Table 3.2 and Table 3.3, the mutual admittance values presented by (3.85) are listed. The mutual admittance values presented by (3.85) are compared with the literature [45].

Table 3.2: Comparison of the $Y_{pq}$ values for $z_0 = 0''$ with the literature [45].

<table>
<thead>
<tr>
<th>$z_0 = 0''$</th>
<th>Modal Solution [45]</th>
<th>Asymptotic Solutions</th>
<th>This Thesis</th>
</tr>
</thead>
<tbody>
<tr>
<td>30°</td>
<td>-81.33 dB</td>
<td>-83.14 dB</td>
<td>-81.34 dB</td>
</tr>
<tr>
<td></td>
<td>-77°</td>
<td>-60°</td>
<td>-75°</td>
</tr>
<tr>
<td>40°</td>
<td>-89.87 dB</td>
<td>-91.11 dB</td>
<td>-90.02 dB</td>
</tr>
<tr>
<td></td>
<td>168°</td>
<td>180°</td>
<td>170°</td>
</tr>
<tr>
<td>50°</td>
<td>-96.37 dB</td>
<td>-97.43 dB</td>
<td>-96.72 dB</td>
</tr>
<tr>
<td></td>
<td>58°</td>
<td>69°</td>
<td>61°</td>
</tr>
<tr>
<td>60°</td>
<td>-101.97 dB</td>
<td>-102.93 dB</td>
<td>-102.48 dB</td>
</tr>
<tr>
<td></td>
<td>-49°</td>
<td>-39°</td>
<td>-47°</td>
</tr>
</tbody>
</table>
Table 3.3, the modal solution and asymptotic solution derived in [45] are compared with the asymptotic solutions given in [46] and [47] while the modal solution and GTD solution derived in [45] are compared in Table 3.4.

Table 3.3: Comparison of the $Y_{pq}$ values for $\phi_0 = 0^\circ$ with the literature [45].

<table>
<thead>
<tr>
<th>$\phi_0 = 0^\circ$</th>
<th>Modal Solution [45]</th>
<th>Asymptotic Solutions</th>
<th>This Thesis</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5&quot;</td>
<td>-62.62 dB -72°</td>
<td>-61.7 dB -68°</td>
<td>-62.54 dB -72°</td>
</tr>
<tr>
<td>2&quot;</td>
<td>-71.78 dB -117°</td>
<td>-70.96 dB -118°</td>
<td>-71.66 dB -116°</td>
</tr>
<tr>
<td>8&quot;</td>
<td>-81.84 dB -34°</td>
<td>-80.80 dB -34°</td>
<td>-81.83 dB -37°</td>
</tr>
<tr>
<td>16&quot;</td>
<td>-86.48 dB -4°</td>
<td>-85.26 dB -4°</td>
<td>-86.6 dB -1°</td>
</tr>
<tr>
<td>40&quot;</td>
<td>-91.95 dB -115°</td>
<td>-90.83 dB -112°</td>
<td>-92.46 dB -110°</td>
</tr>
</tbody>
</table>

The mutual admittance calculations of the slots are carried out using the formulations derived in this thesis and they are compared with the results presented in [45], [46], and [47] as shown in Table 3.2, Table 3.3, and Table 3.4. As it is seen in Table 3.2, Table 3.3, and Table 3.4, the calculations carried out using the formulations derived in this thesis is in a good agreement with the results presented in [45], [46], and [47].
Table 3.4: Comparison of the $Y_{pq}$ values for various $z_0$ and $\phi_0$ values with the literature [45].

<table>
<thead>
<tr>
<th>$\phi_0$</th>
<th>$z_0$</th>
<th>Modal Solution [45]</th>
<th>GTD Solution [45]</th>
<th>This Thesis</th>
</tr>
</thead>
<tbody>
<tr>
<td>0°</td>
<td>0.5&quot;</td>
<td>-62.62 dB -72°</td>
<td>-62.54 dB -72°</td>
<td>-61.53 dB -72.07°</td>
</tr>
<tr>
<td></td>
<td>2&quot;</td>
<td>-71.78 dB -117°</td>
<td>-71.66 dB -116°</td>
<td>-70.76 dB -116.09°</td>
</tr>
<tr>
<td></td>
<td>8&quot;</td>
<td>-81.84 dB 34°</td>
<td>-81.83 dB 37°</td>
<td>-80.61 dB 31.95°</td>
</tr>
<tr>
<td></td>
<td>40&quot;</td>
<td>-91.95 dB -115°</td>
<td>-92.46 dB -110°</td>
<td>-91.05 dB -113.20°</td>
</tr>
<tr>
<td>30°</td>
<td>2&quot;</td>
<td>-77.42 dB 175°</td>
<td>-77.69 dB 177°</td>
<td>-76.46 dB 175.49°</td>
</tr>
<tr>
<td>60°</td>
<td>2&quot;</td>
<td>-90.00 dB -3°</td>
<td>-90.17 dB -1°</td>
<td>-88.74 dB -4.11°</td>
</tr>
<tr>
<td>90°</td>
<td></td>
<td>-102.52 dB 120°</td>
<td>-103.10 dB 116°</td>
<td>-101.99 dB 114.00°</td>
</tr>
<tr>
<td>30°</td>
<td>0&quot;</td>
<td>-81.33 dB -77°</td>
<td>-81.34 dB -75°</td>
<td>-80.18 dB -77.28°</td>
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<td>40°</td>
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<td>60°</td>
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<td>-101.97 dB -49°</td>
<td>-102.48 dB -47°</td>
<td>-101.03 dB -49.57°</td>
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3.3.2. External and Internal Couplings of Two Slots on CSIW

The formulations derived for external and internal couplings of the slots on E-plane bend regular waveguides are compared with the simulations to see the validity of them. For this purpose the CSIW with longitudinal slot structure is chosen with the same CSIW parameters presented in Section 2.1. The CSIW with radius $R$ is used in the calculations at 25 GHz as shown in Figure 1.2. Two different cases are considered as shown in Figure 3.12 (a) and (b), respectively. In the first case shown in Figure 3.12 (a), two slots with slot offsets $z_p$ and $z_q$, respectively, and lengths $L_p$ and $L_q$, respectively, lie in the same CSIW while in the second case shown in Figure 3.12 (b), the slots are located at two separate parallel CSIW structures. The CSIW structure is formed using 0.5 mm-thick Rogers 3003 substrate. The diameter and period of the vias are 0.8 mm and 1 mm, respectively. The widths of the slots are chosen as 0.25 mm. The width of the CSIW, $W$ measured from the centers of the vias, is 6 mm. In Figure 3.12, the wave ports are used to excite the CSIW at 25 GHz.

Figure 3.12: (a) Longitudinal slots on the same CSIW. (b) Longitudinal slots on two parallel CSIW.
instead of the microstrip transitions. Port1 is excited in Figure 3.12 (a) while Port1 and Port3 are excited in Figure 3.12 (b). All the other ports are terminated by a matched load. The width of the wave port is chosen as the equivalent width of the CSIW in order to reduce the reflections at the ports.

In both of the cases shown in Figure 3.12 (a) and (b), the simulation results are compared with the transmission line models composed of shunt active admittances representing the slots. The active admittance calculations are carried out using the active admittance formulation given in (3.75). Note that both external and internal coupling terms are used in (3.75) for the case given in Figure 3.12 (a) while only the external coupling terms are used for the structure given in Figure 3.12 (b). Furthermore, all the s-parameters are also calculated by ignoring the mutual couplings and also by using the planar coupling formulations given in [34] and [36]. The calculations for the planar case are carried out by modifying the self-admittance term in (3.75) by the coupling terms derived for the planar case given in [34] and [36].

First of all, the first case is considered as shown in Figure 3.12 (a). The circuit model of two longitudinal slots on the same CSIW shown in Figure 3.12 (a) is presented in Figure 3.13. In the simulations, both of the slots are excited when Port1 is excited. Hence, the circuit model shown in Figure 3.13 contains active admittance terms. However, s-parameters calculated using the circuit model shown in Figure 3.13 corresponds to the passive s-parameters in the simulator because Port2 is terminated

![Figure 3.13: Circuit model of two longitudinal slots on the same CSIW.](image-url)
by a matched load in the calculation of $S_{11}$ parameter in the simulations. The derived active admittance formulas are used in the transmission line model shown in Figure 3.13 and the related s-parameters are calculated. In the calculations, the effect of the transmission line lengths, $L_l$ are extracted and ports are embedded till the centers of the slots both in the simulations and closed form calculations. In order to see the effect of slot offset in Figure 3.12 (a) for $R = 10 \text{ cm}$, the value of the slot separation distance $D_{pq}$ is kept constant at 5 mm and $z_q$ is set to 0.5 mm. The slot offset $z_p$ is varied between -0.5 mm and 0.5 mm while the slot lengths are $L_p = L_q = 4.2 \text{ mm}$ and the results are compared in Figure 3.14. As it is seen in Figure 3.14, there is almost perfect agreement between the formulations with mutual coupling calculations and simulations. Since the couplings for cylindrical and planar cases almost converge for radius values greater than $2\lambda_0$, there is almost no difference between the calculations using planar and cylindrical formulations.

The effect of the distance between slots, $D_{pq}$ measured from centers of the slots, are examined by setting the slot offsets, $z_p = 0.2 \text{ mm}$, $z_q = 0.5 \text{ mm}$ and by setting the lengths, $L_p = L_q = 4.2 \text{ mm}$ for $R = 10 \text{ cm}$. The value of $D_{pq}$ is varied between 5 mm

Figure 3.14: (a) Magnitude and (b) phase of $S_{11}$ ($R = 10 \text{ cm}$, $D_{pq} = 5 \text{ mm}$, and $z_q = 0.5 \text{ mm}$).
and 25 mm and the simulations are carried out by exciting the structure from Port1 while Port2 is terminated by a matched load. The simulations are compared with the calculations carried out using active admittance formulas and transmission line model in Figure 3.15. As it is seen in Figure 3.15, there is almost no difference between the s-parameters in different cases. Since the effect of the mutual couplings are not significant for \( z_p = 0.2 \) mm in Figure 3.14 (a), there is almost no difference between the calculations carried out by considering and ignoring the mutual coupling effects in Figure 3.15 (a).

![Figure 3.15: (a) Magnitude and (b) phase of \( S_{11} \) (\( R = 10 \) cm, \( z_p = 0.2 \) mm, and \( z_q = 0.5 \) mm).](image)

The calculations and simulations shown in Figure 3.15 are repeated for the case where \( z_p = -0.2 \) mm for \( R = 10 \) cm to see the effect of the mutual coupling. The simulated and calculated results are compared in Figure 3.16. As it is seen in Figure 3.16, the mutual couplings are significant for the \( D_{pq} \) values between 5 mm and 6 mm. The simulated response is in a good agreement with the values obtained by taking the mutual coupling into account; however, there is almost no difference between the planar and cylindrical equations.
In order to see the difference between the cylindrical and planar equations, the value of the radius $R$ is reduced from 10 cm to 0.6 cm for the case presented in Figure 3.16. The simulations and calculations are carried out and presented in Figure 3.17. As it is

Figure 3.16: (a) Magnitude and (b) phase of $S_{11}$ ($R = 10 \text{ cm}$, $z_p = 0.2 \text{ mm}$, and $z_q = 0.5 \text{ mm}$).

In order to see the difference between the cylindrical and planar equations, the value of the radius $R$ is reduced from 10 cm to 0.6 cm for the case presented in Figure 3.16. The simulations and calculations are carried out and presented in Figure 3.17. As it is

Figure 3.17: (a) Magnitude and (b) phase of $S_{11}$ ($R = 0.6 \text{ cm}$, $z_p = 0.2 \text{ mm}$, and $z_q = 0.5 \text{ mm}$).
seem in Figure 3.17, the results are similar to the ones presented in Figure 3.16. Using such a low radius value does not make any difference between the cylindrical and planar equations.

Secondly, similar calculations are carried out for the case shown in Figure 3.12 (b). The circuit model of the \( p^{th} \) slot shown in Figure 3.12 (b) is presented in Figure 3.18. In Figure 3.18, the active admittance model is used because both Port1 and Port3 are excited with the same amplitude and phase in Figure 3.12 (b). In the calculations, the effect of the transmission line lengths \( L_1 \) are extracted and ports are embedded till the centers of the slots both in the simulations and closed form calculations. In the circuit model, the s-parameters calculated using active admittance values corresponds to calculating the active s-parameters in the simulator. The active s-parameters calculated in the simulator is given by

\[
ActiveS_{mn} = \sum_{m=1}^{M'} a_m S_{mn} \tag{3.87}
\]

where \( n \) stands for the port number and \( a_m \) is the excitation coefficient at Port \( m \). \( M' \) is the total number of ports in the structure and \( S_{mn} \) is the s-parameter representing the coupling between Port \( m \) and Port \( n \).

---

Figure 3.18: Circuit model of the longitudinal slot on CSIW.
First of all, the waveguide separation, $D_w$, is kept constant at 6 mm with $z_q = 0.5$ mm in Figure 3.12 (b) for $R = 10$ cm. Then the value of $z_p$ is varied by setting $L_p = L_q = 4.2$ mm. The calculated results are compared with the simulations in Figure 3.19. As it is seen in Figure 3.19, there is good agreement between different cases. The other simulations are carried out by sweeping $D_w$ between 6 mm and 24 mm for $R = 10$ cm, $z_p = 0.2$ mm, $z_q = 0.5$ mm, and $L_p = L_q = 4.2$ mm. The closed form calculations are also carried out for the same $D_w$ values and they are compared with the simulations in Figure 3.20. Again good agreement between the closed form calculations carried out by transmission line model and simulations is obtained and mutual couplings are nicely expected. Almost no difference between the planar and cylindrical formulations is obtained because of electrically large radius as stated before.

![Figure 3.19: (a) Magnitude and (b) phase of active $S_{11}$ ($R = 10$ cm, $z_q = 0.5$ mm and $D_w = 6$ mm).](image)

In Figure 3.14 and Figure 3.19, symmetrical s-parameter graphs are obtained when the mutual couplings are not taken into account. This is because; the slots are coupled in different manner when the signs of the slot offsets are changed since the
distance between the slots change. When the mutual couplings are ignored, the active admittance values in the circuit model become self-slot admittances which can be seen in (3.75). Since the self-slot admittances are not affected by the sign change in the slot offset, the s-parameter graphs in Figure 3.14 and Figure 3.19 becomes symmetrical. The phases are also not affected by the sign change in the slot offset because both the slot offset \( V_p \) and the term \( f_{p}^{TE_{10}} \) in (3.31) experience 180° phase shift as a result of the change in the sign of the slot offset. Hence, the phase value of the admittance does not change in (3.29) with respect to the sign of the slot offset.

Figure 3.20: (a) Magnitude and (b) phase of active \( S_{11} (R = 10 \text{ cm}, z_p = 0.2 \text{ mm and } z_q = 0.5 \text{ mm}) \).
CHAPTER 4

DESIGN OF TRAVELING WAVE CSIW SLOT ARRAY

This chapter presents different designs of traveling wave longitudinal slot array on CSIW at 25 GHz. The designs of slot array on CSIW are carried out using non-uniformly distributed slot elements near resonance and uniformly spaced slot elements with large deviations from resonance. In both of the methods applied in slot separations, the aperture projection method is used to reduce SLL. The non-uniform slot array is designed on CSIW for beam steering applications while the uniform slot array is designed for pattern synthesis applications. The slotted CSIW arrays presented in this chapter look like periodic leaky wave antennas in the sense that the fundamental mode propagating in the CSIW is a slow wave for frequencies above 20 GHz due to dielectric filling in the CSIW. The slow waves do not radiate and the radiation into free space is obtained by periodic placement of the slots as in the periodic leaky wave antennas. Hence, the radiation into both backward and forward quadrants, i.e., before and after the broadside direction, can be obtained by frequency change [48].

Section 4.1 examines the pattern synthesis methods, specifically the aperture projection method for the slotted CSIW arrays. Section 4.2 presents the slotted CSIW array with non-uniformly distributed slot elements near resonance while Section 4.3 shows a pattern synthesis application with uniformly distributed slot array on CSIW where the slot elements are largely deviated from resonance. The fabrication and measurements of the traveling wave slot arrays on CSIW are given in each section.
4.1. Pattern Synthesis Methods

Pattern synthesis of an array can be carried out by considering different objectives, such as pencil beam and low SLL, shaped beam synthesis, and some performance indices such as directivity, cross-polarization, and bandwidth. For conformal array applications, some of the classical methods such as Woodward-Lawson, Dolph-Chebyshev synthesis, Fourier synthesis and the methods using Taylor line source synthesis or optimization tools such as genetic algorithm and particle swarm optimization can be used [1], [49]. In this thesis, the aim is to obtain low SLL and narrow HPBW using slot array on CSIW. So, aperture projection method as shown in Figure 4.1 is used together with Taylor line source distribution to reduce the SLL. Genetic algorithm optimization is used together with the aperture projection method to design the phases of the elements in the pattern synthesis applied in Section 4.3. The aperture projection method is examined in detail for the slot array on cylindrical surface in the next section. The effect of the uniform and non-uniform sampling of the Taylor line source distribution, i.e., the effect of the element density on the array weights, [50], [51], is searched by considering the isotropic sources on cylindrical surface and then the study is extended by taking the effect of the element pattern into account.

![Figure 4.1: Sampling of the Taylor distribution using aperture projection method.](image-url)
4.1.1. Aperture Projection Method

The aperture projection method is used to determine the excitation of the antenna elements distributed on conformal surface such as cylindrical surface as shown in Figure 4.1. The method is applied by assuming a planar aperture constructed by projecting the antenna element positions onto the planar aperture as shown in Figure 4.1. In the aperture projection, the element weights are determined by the desired amplitude taper such as Taylor distribution, the radiation pattern of which is known. The effect of element pattern and sampling density is taken into account to determine the excitation amplitudes of the slot elements. Before examining the aperture projection method, the pattern synthesis using Taylor line source sampling is examined for planar arrays in terms of uniform and non-uniform sampling.

4.1.1.1. Taylor Line Source Sampling for Planar Arrays

To examine Taylor line source sampling, continuous Taylor line source distribution of length $5\lambda_0$, where $\lambda_0$ is the free space wavelength, is assumed on z-axis. Uniform phase distribution is assumed, and hence, the broadside of the radiating source is $\theta = 90^\circ$. The Taylor line source assumed is -20 dB SLL ($\bar{n} = 3$) distribution for all the cases as shown in Figure 4.2 (a). The radiation pattern corresponding to the continuous Taylor line source is plotted in Figure 4.2 (b). As it is seen in Figure 4.2 (b), the radiation pattern of the Taylor line source satisfies -20 dB SLL as expected. Different sampling cases of this distribution are carried out by assuming isotropic elements.

First of all, the Taylor line source is sampled by 100 isotropic elements the positions of which are randomly chosen within the Taylor line source length. A uniform probability density function is used to determine the positions of 100 elements within
Figure 4.2: (a) Sampling Taylor line source distribution randomly with 100 elements. (b) Corresponding radiation patterns.

the Taylor line source. The weights of the sampled elements, $W_p$, are shown by red crosses in Figure 4.2 (a) where $p$ stands for the $p^{th}$ element in the array. The radiation pattern of the sampled array shown in Figure 4.2 (b) indicates that non-uniform sampling has adverse effects on the radiation pattern such that the desired SLL could not be achieved.

Secondly, half of the Taylor line source is sampled randomly by 100 elements starting from $z' = -2.5 \lambda_0$ till $z' = 0$ as shown in Figure 4.3 (a) and symmetric weights, and hence, symmetric element positions are used for the other half (between $z' = 0$ and $z' = 2.5 \lambda_0$) resulting in total 200 element array. The corresponding radiation patterns are shown in Figure 4.3 (b). Figure 4.3(b) indicates that randomly sampling half side of the Taylor distribution and using symmetric weights for the other half does not help to obtain desired SLL.
In the third case, the Taylor line source is sampled randomly with 1000 elements as shown in Figure 4.4 (a) and the corresponding radiation patterns are plotted in Figure 4.4 (b). As it is seen in Figure 4.4 (b), the radiation pattern of the sampled array gets

Figure 4.4: (a) Sampling Taylor line source distribution randomly with 1000 elements. (b) Corresponding radiation patterns.
closer to the ideal Taylor source pattern but there are still side lobes above -20 dB even if such a large number of sampling is carried out. In Figure 4.4 (b), the sampling of Taylor line source is carried out many times. The SLL values close to the ideal case are usually obtained. Although, the radiation pattern graph shown in Figure 4.4 (b) is rarely encountered among the trials, it is chosen to indicate that using such kind of large number of elements to sample the Taylor line source distribution might not be enough to satisfy desired SLL and radiation pattern.

Finally, the Taylor line source is uniformly sampled by 11 elements with half a free space wavelength, $\lambda_0/2$, spacing as shown in Figure 4.5 (a). As it is seen in the corresponding radiation patterns presented in Figure 4.5 (b), although the nulls of the array does not coincide with the nulls of the desired radiation pattern, -20 dB SLL is satisfied with this uniform sampling. Hence, it can be concluded that non-uniform sampling the Taylor line source has adverse effects on SLL.

In order to examine the effect of the element density, again different cases of sampling the Taylor line distribution are examined. First of all, the Taylor line source

![Figure 4.5](image_url)

**Figure 4.5:** (a) Sampling Taylor line source distribution uniformly with 11 elements. (b) Corresponding radiation patterns.
distribution is sampled again randomly with 100 elements as shown in Figure 4.6 (a). Then the weights of the isotropic antenna elements are determined by altering the sampled the Taylor line source weights with the element density using

$$W^D_p = \frac{W^s_p}{D_p}$$

(4.1)

where $W^s_p$ is the weight determined by directly sampling the Taylor line source indicated by red crosses in Figure 4.6 (a). $D_p$ is the element density of the $p^{th}$ element in the array and $W^D_p$ is the weight of the $p^{th}$ element corrected by the element density. The element density is defined as

$$D_p = \begin{cases} 
\frac{2}{d'_p + d'_{p-1}} & p = 2, 3, ..., M - 1 \\
\frac{1}{d'_p} & p = 1, M 
\end{cases}$$

(4.2)

Figure 4.6: (a) Sampling Taylor line source distribution randomly with 100 elements and adjusting the sampled weights using element density. (b) Corresponding radiation patterns.
where $M$ is the total number of slot elements in the waveguide and $d'_p$ is the element separation of the sampled Taylor line source distribution as shown in Figure 4.1. The weights of the elements corrected by the element density, $W_p^D$ using \((4.1)\), are presented by green crosses in Figure 4.6 (a). As it is seen in Figure 4.6 (a) that the weights are substantially altered by the element density and as a result of this alteration, the expected SLL values can be obtained as shown in Figure 4.6 (b) where the pattern shape of the sampled array is quite close the ideal Taylor line source pattern. Secondly, the Taylor line source is sampled randomly by 20 elements and the element weights are also adjusted using element density as shown in Figure 4.7 (a). The corresponding radiation patterns presented in Figure 4.7 (b) indicates that considering only the element density is not suitable and spacing between some of the elements should not be too large to cause undesired SLL values. When drawing this result, the experiment shown in Figure 4.6 is repeated many times, and each time desired SLL value is reached. This is because the number of elements in Figure 4.6 (a) is high enough so that the element spacing is not too high for some of the elements resulting in high SLL values.

Figure 4.7: (a) Sampling Taylor line source distribution randomly with 20 elements and adjusting the sampled weights using element density. (b) Corresponding radiation patterns.
In this part, the aperture projection method using Taylor line source sampling is examined for the slot arrays placed on cylindrical surface shown in Figure 4.1. Different cases are examined at 25 GHz for \( R = 10\text{cm} \). The far-field coordinates shown Figure 4.1 are used in the radiation pattern graphs presented in this section. Moreover, only the first half of the radiation patterns are plotted for the region \( 0^\circ < \phi < 90^\circ \) to enhance the visibility of the patterns plotted on the same graph. The other half in region \( 90^\circ < \phi < 180^\circ \) can be found by taking the symmetry of the first half with respect to the y-axis. In Figure 4.1, \( \hat{n} \) is the normal vector showing the desired main beam direction, \( \phi_m^n \). \( \vec{r}_p \) is the position vector of the \( p^{th} \) element in the array. The phase of the \( p^{th} \) element in the array is determined as \( -k_0 \hat{n} \cdot \vec{r}_p \) where \( k_0 \) is the wavenumber in free space, and hence, the element phases are compensated and main beam is directed to \( \phi_m^n \). In this section, all the element phases are determined so that \( \phi_m^n = 90^\circ \).

First of all, 40 slot elements are assumed in order to cover large angular area. Using 40 slot elements equally spaced by \( 0.7\lambda_g = 0.531\lambda_0 \), about \( 142.33^\circ \) angular region is covered by the slot elements as shown in Figure 4.8 (a) where \( \lambda_g \) stands for the guided wavelength determined by using the CSIW parameters given in Section 2.1. By using the projected positions of the slot elements, the element positions of the linear array are obtained as shown in Figure 4.8 (a). The element weights for the linear array, \( W_p^S \) are found by sampling -20 dB SLL Taylor (\( \tilde{n}=3 \)) distribution as shown in Figure 4.8 (b). The array factor for the linear array is calculated by assuming isotropic elements and it is compared with the ideal Taylor line source radiation pattern in Figure 4.8 (c). As it is seen in Figure 4.8 (c), the array factor of the linear array deviates from the desired pattern. The weights of the linear array elements are scaled by the element density of the linear array using (4.1) and the
coefficients for the linear array, $W_p^D$ are found. The array factor calculated using the coefficients, $W_p^D$ is seen to successfully mimic the ideal radiation pattern in Figure 4.8 (c). This is an expected result as shown previously in Section 4.1.1.1. Hence, the element density of the linear array is required to be taken into account while finding the element coefficients for the cylindrical slot array.

Figure 4.8: (a) Element positions of the slotted CSIW array and element positions at the projected aperture (40 element array). (b) Sampling Taylor line source distribution using projected aperture positions. (c) Corresponding radiation patterns for linear arrays. (d) Corresponding radiation patterns for cylindrical slot arrays.
The coefficients, \( W_p^D \), are presented by square markers in Figure 4.8 (b) while circle markers in Figure 4.8 (b) shows the element weights for the cylindrical slot array calculated by taking both the element density and element factor into account which is given by

\[
W_p^{DE} = \frac{W_p^s}{D_p E_p}
\]  

(4.3)

where \( E_p \) is the element pattern of the slot toward the broadside of the array and \( W_p^{DE} \) is the excitation coefficient of the \( p^{th} \) element scaled by both element density and element factor.

The radiation patterns for the cylindrical slot array are calculated using closed form expressions by assuming that the slot elements have required phases to compensate the phase delays due to different path lengths measured from the position of the slot element to the planar aperture. In the pattern calculations, the effect of element pattern is taken into account. The radiation patterns of the cylindrical slot array are calculated using both of the coefficients, \( W_p^D \) and \( W_p^{DE} \). The radiation patterns of the cylindrical slot array are compared with the ideal Taylor line source radiation pattern in Figure 4.8 (d). In Figure 4.8 (d), although there are increased side lobes compared to the ideal Taylor line source radiation pattern due to the slot elements located at the regions close to the beginning and the end of the array, both cases satisfy desired -20 dB SLL. To examine the radiation toward the regions close to \( \phi = 0^\circ \) and \( \phi = 180^\circ \), the same Taylor line source distribution with length equal to the length of the cylindrical slot array is conformed to a cylindrical surface and the radiation pattern of Taylor cylindrical line source pattern is computed. As it is seen in Figure 4.8 (d), the Taylor cylindrical line source pattern has larger side lobes compared to the cylindrical slot array patterns because of conforming the Taylor line source distribution on a cylindrical surface. In Figure 4.8 (d), lower side lobes are obtained
by using the coefficients $W_p^D$ compared to using the coefficients $W_p^{DE}$ for the cylindrical slot array because the coefficients $W_p^D$ looks like the Taylor line source distribution with SLL lower than -20 dB. However, the cylindrical slot array pattern with excitation coefficients $W_p^{DE}$ mimics the desired radiation pattern better around the main beam. In Figure 4.8 (b), the increased element density toward edges is compensated by fall off of the element pattern and the element weights, $W_p^E$ does not deviate much from the weights, $W_p^s$ obtained by direct sampling of Taylor line source distribution. So, directly using Taylor weights, $W_p^s$ can be employed for slot array on CSIW with uniform slot separation around the cylinder.

The same analysis using 40 elements and 0.7λg element spacing is carried out using -30 dB SLL Taylor ($\tilde{n}=3$) and -40 dB SLL Taylor ($\tilde{n}=6$) distributions. The resultant radiation patterns are presented in Figure 4.9 (a) and (b), respectively. As it is seen in Figure 4.9 (a) and (b), -35 dB SLL is the limit for 40 element slot array such that

Figure 4.9: (a) Radiation patterns for the slotted CSIW array using (a) -30 dB and (b) -40 dB SLL Taylor weights.
using -40 dB SLL Taylor distribution gives minimum -34.7 dB SLL in the radiation patterns. By employing the number of elements less than 40 on radius $R = 10 \, cm$, the SLL can be reduced below -35 dB using -35 dB SLL or -40 dB SLL Taylor weights.

By considering the directivity requirements, the total angular coverage of the slot elements on CSIW is determined as about $80^\circ$. With this reduced angular coverage compared to the case presented in Figure 4.8, the full wave simulations of the CSIW array is easier to carry out. Moreover, the time in the Elliott’s design procedure to determine the slot offsets and lengths are reduced. Using the aperture projection method, the slotted CSIW arrays are designed with non-uniform and uniform slot separation values. First of all, the element spacings are determined to compensate the phases of the cylindrical slot array to get in phase contributions from each element toward the broadside direction and this configuration is achieved by 16 slot elements with about $76^\circ$ angular coverage. Hence, non-uniform slot separations are used in this design. Although, the slots are required to be close to the resonance which is easier to design, some of the element spacings are quite far away from $0.7\lambda_g$ which increases the SLL. This design is presented in Section 4.2. The other design allocates 25 slot elements uniformly distributed with $0.7\lambda_g$ spacing which has about $87^\circ$ angular coverage as shown in Figure 4.10 (a) and the phase compensation is achieved using slots largely deviating from resonance.

The same analysis as explained in Figure 4.8 is carried out in Figure 4.10 which starts by sampling the -20 dB SLL Taylor ($\tilde{n}=3$) line source distribution as shown Figure 4.10 (b). As it is seen in Figure 4.10 (c) and (d), all of the patterns satisfy -20 dB SLL criteria. Since the element weights, $W_{p}^{DE}$ are close to the Taylor line source distribution as mentioned before, 25 element array with uniform slot separations on CSIW is decided to be designed in the Elliot’s code with element weights, $W_{p}^{DE}$. The design of this array is presented in Section 4.3.
Figure 4.10: (a) Element positions of the slotted CSIW array and element positions at the projected aperture (25 element array). (b) Sampling Taylor line source distribution using projected aperture positions. (c) Corresponding radiation patterns for linear array. (d) Corresponding radiation patterns for cylindrical slot array.
4.2. Slot Array Design on CSIW with Non-Uniform Slot Separation

The design of 1 x 16 element longitudinal slot array on CSIW is carried out at 25 GHz for radius, \( R = 10 \text{ cm} \). The schematic view of the slotted CSIW array used in the simulations is shown in Figure 1.2. The slotted CSIW array is formed using the CSIW and substrate parameters mentioned in Section 2.1. The slot width is chosen as \( w^s = 0.25 \text{ mm} \). The aim of this design is to develop a proof-of-a-concept prototype to demonstrate a frequency scanning with circumferential slot array on CSIW. Hence, a directivity of the radiation pattern should be high to observe the rotation of the beam clearly. Moreover, the SLL is desired to be low. Hence, the excitation amplitudes of the slot elements, \( W^s_p \) are determined by sampling -20 dB SLL Taylor (\( \tilde{n} = 3 \)) distribution as shown in Figure 4.1. Distances between the slots \( d_p, p = 1, 2, \ldots, 15 \) shown in Figure 4.1 are determined to obtain the slot element phases such that in phase contribution from each element is achieved and directive main beam is obtained in \( \hat{a}_y \)-direction. Without compensating the phases of the elements and using uniform element spacing, larger HPBW values can be obtained by using -40 dB SLL Taylor (\( \tilde{n} = 3 \)) distribution [52], however, the frequency steering might not be clearly observed with such a high HPBW values especially when the narrow bandwidth of the slotted waveguide array is considered. In the schematic view of 1 x 16 slot array on CSIW shown Figure 1.2 which is used in the simulations, the array is excited from Port 1, i.e., the traveling wave propagates in \( -\hat{a}_\phi \)-direction, and Port 2 is terminated by a matched load. The same convention used in Figure 4.1 such that the traveling wave propagates in \( -\hat{a}_\phi \)-direction and the first element having separation \( d_1 \) in Figure 4.1 is the one closest to the matched load. The only difference between Figure 4.1 and 1 x 16 slot array simulations is that the broadside of the array in Figure 4.1 is in \( \hat{a}_y \)-direction while the broadside of 1 x 16 slot array is located at \( \phi = 48^\circ \) in the simulations. This difference is caused by locating the first slot element at \( \phi = 4.57^\circ \) after designing the array using the convention presented in Figure 4.1, and hence, the last slot element position coincides at \( \phi = 80.57^\circ \) in the simulations.
The phase compensation procedure, i.e., determining the slot separation values, is carried out using three different slot alternation schemes. In the first case, all the slots are located at the same half side of the CSIW, i.e., the slots are not alternated. In the second case, slots are alternated sequentially as shown in Figure 1.2 (b). In the third case, some of the slots are alternated and some of them are not alternated with respect the previous slot. In the third case, the alternation of the $p^{th}$ slot is determined by the aim that gives minimum $d_{p,1}$ value, i.e., $d_{p,1}$ values are calculated by altering and not altering the $p^{th}$ slot and alternation is decided if it gives minimum slot separation $d_{p,1}$. Hence, three set of slot separations are obtained and plotted in Figure 4.11 (a). For all three cases, there is minimum allowed slot separation value. If some of the element separations are below the minimum limit, high mutual coupling levels do not allow making the Elliott’s design. The minimum allowed slot separation value is determined by iteratively decreasing the minimum allowed slot separation value and applying the Elliott’s design procedure till the Elliott’s design procedure cannot be achieved due to high mutual couplings. The minimum allowed slot separation value is found to be nearly the same for all three cases.

Figure 4.11: (a) Element spacing and (b) calculated patterns of $1 \times 16$ slot array on CSIW ($R = 10 \text{ cm}$) for three different slot separation schemes.
An in-house code is developed that calculates the far-field pattern of slot array on cylindrical surface. The slot elements are assumed to be tangentially placed on the cylindrical surface and the element factors are also taken into account. The patterns of $1 \times 16$ slot array for three different slot separation cases are calculated and plotted as shown in Figure 4.11 (b). The radiation patterns are plotted at $\theta = 90^\circ$ plane with respect to the coordinate system indicated in Figure 4.1. The patterns show about $5^\circ$ HPBW which is enough to observe the frequency scanning. Although, -20 dB SLL Taylor ($\bar{n} = 3$) distribution is used to determine the weights of the elements, -20 dB SLL could not be obtained in the radiation pattern graphs in Figure 4.11 (b) in all three cases, because the Taylor distribution is not sampled dense enough and because of large slot separation values obtained after compensating the element phases as described in Section 4.1.1. As it can be seen in Figure 4.11 (a), the minimum value for the average of the slot separations is obtained for the third case which provides lowest SLL values in Figure 4.11 (b). Hence, the slot separation values in the third case are decided to be used in the further design steps and the fabrication.

The propagation constant and the characterization polynomials derived for the CSIW is used together with the slot separation values for the third case in the Elliott design procedure described in Section 3.2. The Elliott’s design procedure is modified to be used with different slot separation values. In Chapter 2, it is observed that the shunt admittance model is not affected much by the radius of the curvature and almost no difference between the planar SIW and CSIW is observed for $R = 10 \text{ cm}$. Moreover, almost no difference is observed between the radius $R = 10 \text{ cm}$ ($8.33\lambda_0$ at 25 GHz) and planar case in terms of mutual couplings in Section 3.3 and in [53]. Hence, the active admittance formulations derived for the planar case given in [34] and [36] are used in the Elliott’s design procedure. The Elliott’s design procedure is implemented in MATLAB to determine the slot offsets and lengths of the slot array on CSIW.

In the Elliott’s design, the slot offsets and lengths which satisfy desired complex slot voltages are determined. The Elliott’s design procedure is a complicated routine such
that the pattern requirements, i.e., the complex slot voltages, are required to be satisfied together with a good input match. In the planar case, usually the slot separations are changed if the required input match is not obtained, however, slot separations affect the phase compensation and it is required to obtain directive main beam in slotted CSIW array. The value of the first slot element can be varied to improve the input match, however, the first slot offset determines the power delivered to load as mentioned in Section 3.2. The other parameters which can be changed to improve the input match are the slot separation values or number of elements used in the slotted CSIW array. In the design, the number of elements can be changed or minimum allowed slot separation value might be increased and phase compensation can be reapplied till obtaining a good input match in the Elliott’s design procedure. Increasing the number of elements or slot separations might increase SLL.

In the Elliott’s design procedure, a 1 x 16 slot array on CSIW \((R = 10 \text{ cm})\) with separation values determined by the third case given in Figure 4.11 (a) is considered and the slot offsets and lengths are determined at 25 GHz. The calculated slot offset and length values are tabulated in Table 4.1. Since there is no input match problem encountered in the Elliott’s design procedure, the slot separations or number of elements are not changed to improve input match as mentioned previously. The array designed by Elliott’s design procedure is simulated in HFSS without microstrip

**Table 4.1: Slot offsets and lengths of 1 x 16 slot array on CSIW determined by Elliott’s design procedure.**

<table>
<thead>
<tr>
<th>Element #</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slot Offsets (mm)</td>
<td>-0.250</td>
<td>-0.290</td>
<td>-0.176</td>
<td>-0.164</td>
<td>-0.161</td>
<td>0.223</td>
<td>0.232</td>
<td>0.204</td>
</tr>
<tr>
<td>Slot Lengths (mm)</td>
<td>4.012</td>
<td>4.066</td>
<td>4.006</td>
<td>3.954</td>
<td>3.939</td>
<td>4.009</td>
<td>3.986</td>
<td>3.960</td>
</tr>
<tr>
<td>Element #</td>
<td>9</td>
<td>10</td>
<td>11</td>
<td>12</td>
<td>13</td>
<td>14</td>
<td>15</td>
<td>16</td>
</tr>
<tr>
<td>Slot Offsets (mm)</td>
<td>0.179</td>
<td>0.164</td>
<td>0.153</td>
<td>0.118</td>
<td>0.078</td>
<td>0.058</td>
<td>-0.087</td>
<td>0.055</td>
</tr>
</tbody>
</table>
transitions to compare with the Elliott’s design code implemented in MATLAB. The simulation results are shown in Figure 4.12. In the Elliott’s design, the power delivered to load at 25 GHz is calculated as 9.1 % and $S_{11}$ value at 25 GHz is found as -18.05 dB. The simulated results of the array show about -19.7 dB $S_{11}$ value at 25 GHz which is in a quite good agreement with the Elliott’s design code. The simulated value of $S_{21}$ is -9.59 dB at 25 GHz which indicates that 11 % of the power is delivered to load. In the Elliott’s design code, the power delivered to load is found as 9.1 % which shows the validity of the Elliott’s design procedure implemented in a MATLAB code. Small difference between the results can be caused due to the loss tangent of the substrate not taken into account in the Elliott’s design procedure.

The simulated radiation patterns of the $1 \times 16$ slot array designed by Elliott’s design procedure is plotted between 24 GHz and 26 GHz as shown in Figure 4.13 (a). As it is seen in Figure 4.13 (a), about 10° beam steering is obtained when the frequency is steered between 24-26 GHz band. Note that the main beam is directed to $\phi = 48^\circ$ at 25 GHz which is different from the pattern given in Figure 4.11 (b) as stated before. The SLL around -16 dB is obtained at 25 GHz both in the simulated patterns shown in Figure 4.13 (a) and in the patterns calculated by closed form expressions given in Figure 4.13 (b).

![Graphs](image)

Figure 4.12: (a) $S_{11}$ and (b) $S_{21}$ results of $1 \times 16$ SIW slot array on CSIW designed by Elliott’s design procedure.
Figure 4.11 (b). In Figure 4.13 (a), the SLL increases to about -13 dB as a result of frequency steering. Desired -20 dB SLL is not obtained because of large element spacing as stated before. The simulation results of the accepted power and radiation efficiency of the slotted CSIW array designed by Elliott’s design procedure are presented in Figure 4.14. The accepted power is calculated by subtracting the sum of reflected power and the power absorbed at the matched load from the power incident to the slot array. It can be seen in Figure 4.14, about 88% power is accepted around 25 GHz. Part of the accepted power is radiated and the other part is dissipated as dielectric and conductor losses. Due to these losses, the efficiency of 1 x 16 slot array reduces to 55% at 25 GHz.

In order to increase the bandwidth of the 1 x 16 slot array for beam steering applications, the slot lengths shown in Table 4.1 are optimized in Ansys HFSS to increase the radiated power, and hence, the efficiency. The sequential nonlinear programming optimization in Ansys HFSS is carried out to maximize the accepted power of the array between 24 GHz and 26 GHz in the simulations. The accepted
power is used in the optimizations because fast frequency sweeps can be used to calculate the accepted power in Ansys HFSS which is faster compared to the calculating and saving the fields radiated by the antenna at each frequency step and then calculating the radiated power. As a result of the optimizations, about 200 MHz increase is obtained in the 70% accepted power bandwidth as shown in Figure 4.14. The slot offsets and lengths found after the optimizations are presented Table 4.2. As it is seen in Table 4.2, the slot offsets are not changed compared to the values given in Table 4.1 because only the slot lengths are optimized. After the optimizations, the efficiency values of the array are calculated and presented in Figure 4.14. As it is seen in Figure 4.14, the efficiency bandwidth also increases as a result of increasing the accepted power bandwidth. The simulated radiation patterns of the optimized 1 x 16 slot array are shown in Figure 4.13 (b). The band of the slot array is increased at the expense of increased SLL. As it is seen in Figure 4.13 (b), the SLL is increased to about -10 dB compared to the patterns before the optimization presented in Figure 4.13 (a).
The efficiency calculations are also carried out by setting both dielectric and conductor losses to zero in the optimized $1 \times 16$ slot array antenna simulations. The efficiency results are compared with the results obtained by simulations where all the losses are taken into account in Table 4.3. As it is seen in Table 4.3, the efficiency of the optimized array is below 50% because of high dielectric and conductor losses of large SIW array at K band.

Table 4.2: Slot offsets and lengths determined after the optimization of $1 \times 16$ slot array on CSIW designed by Elliott’s design procedure.

<table>
<thead>
<tr>
<th>Element #</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slot Offset (mm)</td>
<td>-0.250</td>
<td>-0.290</td>
<td>-0.176</td>
<td>-0.164</td>
<td>-0.161</td>
<td>0.223</td>
<td>0.232</td>
<td>0.204</td>
</tr>
<tr>
<td>Slot Length (mm)</td>
<td>3.941</td>
<td>4.047</td>
<td>3.956</td>
<td>4.154</td>
<td>3.900</td>
<td>4.040</td>
<td>4.122</td>
<td>3.890</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Element #</th>
<th>9</th>
<th>10</th>
<th>11</th>
<th>12</th>
<th>13</th>
<th>14</th>
<th>15</th>
<th>16</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slot Offset (mm)</td>
<td>0.179</td>
<td>0.164</td>
<td>0.153</td>
<td>0.118</td>
<td>0.078</td>
<td>0.058</td>
<td>-0.087</td>
<td>0.055</td>
</tr>
<tr>
<td>Slot Length (mm)</td>
<td>3.997</td>
<td>4.138</td>
<td>3.829</td>
<td>3.909</td>
<td>3.989</td>
<td>4.103</td>
<td>4.119</td>
<td>3.801</td>
</tr>
</tbody>
</table>

The efficiency calculations are also carried out by setting both dielectric and conductor losses to zero in the optimized $1 \times 16$ slot array antenna simulations. The efficiency results are compared with the results obtained by simulations where all the losses are taken into account in Table 4.3. As it is seen in Table 4.3, the efficiency of the optimized array is below 50% because of high dielectric and conductor losses of large SIW array at K band.

Table 4.3: Comparison of the simulated efficiency values of optimized $1 \times 16$ slot array on CSIW

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Lossy</th>
<th>Lossless</th>
</tr>
</thead>
<tbody>
<tr>
<td>24 GHz</td>
<td>15.7%</td>
<td>22.0%</td>
</tr>
<tr>
<td>24.5 GHz</td>
<td>34.9%</td>
<td>53.9%</td>
</tr>
<tr>
<td>25 GHz</td>
<td>49.6%</td>
<td>79.2%</td>
</tr>
<tr>
<td>25.5 GHz</td>
<td>44.3%</td>
<td>70.5%</td>
</tr>
<tr>
<td>26 GHz</td>
<td>21.0%</td>
<td>29.3%</td>
</tr>
</tbody>
</table>
4.2.1. Fabrication and Measurements

The optimized 1 x 16 slot array is selected to be fabricated. Moreover, 4 x 16 slot array on CSIW is formed and fabricated by using 1 x 16 optimized slot arrays connected using a 1 x 4 SIW power divider. The fabrication of the slot arrays are carried out in in-house facilities by using standard PCB process. SIW structures are formed by via hole plating process. The fabrications are carried out using 0.5 mm thick Rogers 3003 substrate. The substrate has 35 μm thick copper cladding on both sides. First of all, via holes are drilled which is followed by electroplating process for about 2 hours to coat the walls of drilled holes with copper. Then microstrip lines and slots are formed by stripping the excess copper using PCB milling.

4.2.1.1. 1 x 16 Optimized Slot Array on CSIW

The optimized 1 x 16 slot array with parameters indicated in Table 4.2 is fabricated and it is shown in Figure 4.15 (a). The structure is measured and the s-parameters are compared with the simulations in Figure 4.15 (b) and (c). In Figure 4.15 (b) and (c), very good agreement between the measured and simulated results is obtained. The matching of the structure is below -10 dB between 20 GHz and 30 GHz. The measured and simulated $S_{11}$ value at 25 GHz is about -20 dB. $S_{21}$ values at 25 GHz are -8.15 dB and -10.3 dB for the simulations and measurements, respectively. Hence, the power delivered to load is calculated as about 15.3 % for the simulations and 9.3 % for the measurements at 25 GHz. Using the loss values calculated for the transition and SIW parts without slots in Section 2.2 and also using the measured s-parameters shown in Figure 4.15 (b) and (c), the radiated power of the antenna can be roughly calculated by ignoring the effect of slots at 25 GHz. The radiated power is calculated as around 40 % of the input power, i.e., the efficiency of the antenna is expected to be around 40 % at 25 GHz.
The radiation patterns and the gain of the antenna are measured in an anechoic chamber using the measurement setup shown in Figure 4.16. The radiation pattern measurements are taken with 1° resolution in the H-plane. The H-plane pattern

Figure 4.15: (a) Fabricated traveling wave optimized $1 \times 16$ slot array on SIW. (b) $S_{11}$ and (c) $S_{21}$ results of the optimized $1 \times 16$ slotted SIW array.

The radiation patterns and the gain of the antenna are measured in an anechoic chamber using the measurement setup shown in Figure 4.16. The radiation pattern measurements are taken with 1° resolution in the H-plane. The H-plane pattern

Figure 4.16: Measurement setup of $1 \times 16$ slot array on CSIW ($R = 10 \text{ cm}$).
measurements are taken between 24 GHz and 26 GHz with 0.5 GHz steps and the measured patterns are compared with the simulated results in Figure 4.17. In Figure 4.17 (a) - (e), the measured radiation patterns between 24 GHz and 26 GHz are compared with the simulation results. Good agreement between the simulated and measured data is obtained around the main beam.

In Figure 4.17 (a)-(e), some of the measured side lobes are seen to be deviating from the simulations. This deviation might be due to some fabrication imperfections. For example, the difference between the slot offset values required to be satisfied is too small as shown in Table 4.2 which might not accurately be satisfied in PCB milling process. Moreover, the substrate is heated in the electroplated step which causes substrate to expand which causes a stress on the substrate. In addition to the substrate stress, the stress level of the copper changes after the electroplating process which cause misalignment problems encountered between drill and the milling layers in the fabrication. This misalignment can easily affect values of the slot offsets required to be satisfied.

In Figure 4.17 (f), the main beam positions of the measured radiation patterns are compared with the simulations to observe the frequency steering. As it is seen in Figure 4.17 (f), about 10° beam steering is observed as a result of frequency scanning between 24 GHz and 26 GHz. The maximum deviation of the measured beam positions from the simulated data is about 1° for each measurement frequency. 1° error in the main beam positions is an expected result when it is considered that the measurements are taken with 1° angular resolution.
Figure 4.17: Radiation patterns of $l \times l6$ slot array on CSIW.
The HPBW values are computed by interpolating the simulated and measured H-plane patterns originally generated by 1° angular resolution and then calculating the HPBW numerically. The measured HPBW and cross-polarization values are compared with the simulations in Table 4.4. As it is seen in Table 4.4, there is a very good agreement between the simulated and measured results.

Table 4.4: Simulated and measured HPBW and cross-polarization values of 1 x 16 slot array on CSIW.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>HPBW</th>
<th>Cross-pol.</th>
</tr>
</thead>
<tbody>
<tr>
<td>24 GHz</td>
<td>5°</td>
<td>5°</td>
</tr>
<tr>
<td>24.5 GHz</td>
<td>5.36°</td>
<td>5.18°</td>
</tr>
<tr>
<td>25 GHz</td>
<td>5.19°</td>
<td>5.59°</td>
</tr>
<tr>
<td>25.5 GHz</td>
<td>4.85°</td>
<td>5.36°</td>
</tr>
<tr>
<td>26 GHz</td>
<td>5.15°</td>
<td>5.47°</td>
</tr>
</tbody>
</table>

The gain of the array is measured using the standard gain horn antenna. The measured gain and efficiency values are compared with the simulations in Figure 4.18. Since there is a good agreement between the measured radiation patterns and the simulations in Figure 4.17, the directivity values in the simulations are used to calculate the measured efficiency values. About 11.8 dB gain is obtained at 25 GHz in the simulations while the measured gain is about 11.58 dB which is quite close to the simulated gain. The behaviors of the simulations and measurements are the same and there is maximum about 2.5 dB difference between measured and simulated gain values at 24 GHz. The efficiency of the antenna is expected as 40 % at 25 GHz using measured s-parameters at the beginning of Section 4.2.1.1, and it is found around 45 % in the simulations and the measurements which is quite close to the expected value.
A 1 x 4 SIW power divider structure is designed and fabricated to feed 4 x 16 CSIW slot array. Two power dividers are connected back to back and fabricated. The picture of the fabricated power divider is shown in Figure 4.19 (a). T-junction structure presented in [5] is used to obtain 1 x 2 power division. Then H-plane Y-junctions are implemented to obtain 1 x 4 power division. Additional vias are implemented at two sides of the Y-junction to have equal power division in the simulations. The power divider is fed by optimized microstrip line to SIW transition with $L_f = 7.5 \text{ mm}$. The other transition, substrate and via parameters are the same as the ones listed in Section 2.1 and Section 2.2. The positions and the diameters of three vias at each side of the power divider shown in Figure 4.19 (a) are optimized in Ansys HFSS to have a good transition between 24 GHz and 26 GHz.

### Figure 4.18: (a) Gain and (b) efficiency results of 1 x 16 SIW slot array.
The measured s-parameters of the power divider structure are compared with the simulated results in Figure 4.19 (b) and (c). A good agreement between simulation and measurement results is obtained. Input matching of the power divider is better than $-10 \, dB$ and insertion loss is about $2 \, dB$ between $24 \, GHz$ and $26 \, GHz$ band. The transition structures as shown in Figure 2.3 with $L_f = 7.5 \, mm$ are also manufactured for different SIW lengths and the transition loss in the power divider is measured as $0.56 \, dB$ at $25 \, GHz$ for one sided transition. Considering $1.92 \, dB$ insertion loss of the power divider structure at $25 \, GHz$ shown Figure 4.19 (a) and using $0.56 \, dB$ loss due to the transition, the insertion loss of only the SIW part of back to back power divider structure can be calculated as $0.8 \, dB$. Hence, the insertion loss in the SIW part of one sided power divider structure is $0.4 \, dB$. Although $1.92 \, dB$ insertion loss is good enough for most of the applications, the losses due to the transition to SIW can be

Figure 4.19: (a) Fabricated two back to back $1 \times 4$ power dividers. (b) $S_{11}$ and (c) $S_{21}$ results of the back to back power divider structure.
reduced by using shorter feed lines and/or by using CPW feed line which has lower radiation losses or by using aperture coupled feeding of the SIW by another waveguide which does not have radiation losses. In 4 x 16 slot array structure, a shorter transition with \( L_f = 4 \text{ mm} \) is used to reduce the loss and also to avoid the microstrip cracks which occur while bending the antenna.

### 4.2.1.3. 4 x 16 Slot Array on CSIW

The fabricated traveling wave 4 x 16 slot array is shown in Figure 4.20. The array is composed of microstrip line to SIW transition mentioned in Section 2.2, 1 x 16 slot array optimized as described in Section 4.2, and 1 x 4 SIW power divider structure mentioned in Section 4.2.1.2. Total length of the array is 19.82 cm. If the transition parts with lengths \( L_f = 4 \text{ mm} \) and \( L_t = 2.1 \text{ mm} \) are excluded from the structure the length of SIW region in the 4 x 16 array is found as 18.6 cm.

![Figure 4.20: Fabricated traveling wave 4 x 16 slot array on SIW.](image)

The s-parameters of the fabricated 4 x 16 slot array are measured and compared with the simulations in Figure 4.21. The simulations of 4 x 16 CSIW array are carried out
by taking the effect of connectors into account. Good agreement between the simulations and measurements are obtained except about 5 dB difference in $S_{21}$ values between 24 GHz and 25.5 GHz. The reason for the discrepancy between the $S_{21}$ values are searched by conducting some experiments on SIW (without slots) pieces and microstrip lines fabricated on the same material. During the measurements, non-uniform manufacturing of the devices are observed between the same types of samples from the same production cycle. It is observed that there is a good agreement between the measured and simulated $S_{21}$ values of microstrip lines. On the other hand, 1-2 dB differences are observed in the $S_{21}$ values of SIW transitions. It is predicted that these additional losses observed in the measurements are due to the fabrication imperfections.

The radiation patterns of the fabricated array are measured in an anechoic in the H-plane using the measurement setup shown in Figure 4.22. The radiation patterns of the array are measured by $1^\circ$ angular resolution between 24 GHz and 26 GHz band with 0.5 GHz steps. The measured patterns are compared with the simulations in Figure 4.23. In Figure 4.23 (a)-(e), the simulated and measured patterns are compared between 24 GHz and 26 GHz with 0.5 GHz steps. It can be seen in

Figure 4.21: (a) $S_{11}$ and (b) $S_{21}$ results of 4 x 16 SIW slot array.
Figure 4.23 (a)-(e) that measured patterns are in a very good agreement with the simulations. Better agreement between the measured and simulated SLL values is obtained compared to 1 x 16 slot array patterns shown in Figure 4.17. Small discrepancies in SLL values might be caused by small errors in the placement of the array on foam cylinder which is quite flexible material. Moreover, it is inevitable to have rough surface while cutting the foam to make a cylindrical surface.

In order to observe the amount of frequency steering, the measured main beam positions are compared with the simulations in Figure 4.23 (f). As it is seen in Figure 4.23 (f), about 10° beam steering is observed as a result of frequency scanning between 24 GHz and 26 GHz. There is about 1° maximum error observed at 24 GHz. This is a good result when the measurements taken with 1° angular steps are concerned. The cross polarization levels and HPBW values of the patterns are also measured and compared with the simulations in Table 4.5. Good agreement between the measured and simulated cross polarization levels is obtained. The HPBW values are calculated by interpolating the measured and simulated radiation patterns and they are also in a good agreement with each other as shown in Table 4.5.
Figure 4.23: Radiation patterns of 4 x 16 slot array on CSIW.
The gain of the array is measured for different frequencies using the standard gain horn antenna. The measured gain and efficiency values of $4 \times 16$ slot array on CSIW are compared with the simulations in Figure 4.24. Since there is a good agreement between the measured radiation patterns and the simulations in Figure 4.23, the directivity values in the simulations are used to calculate the measured efficiency values. About $16 \text{ dB}$ gain is obtained at $25 \text{ GHz}$ in the simulations while the measured gain is about $14.5 \text{ dB}$. There is about maximum $2.5 \text{ dB}$ deviation between measured and simulated gain values at $26 \text{ GHz}$. As it is seen in Figure 4.24, the

<table>
<thead>
<tr>
<th></th>
<th></th>
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<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>24 GHz</td>
<td>5.92°</td>
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<td>-21.56 dB</td>
<td>-18.48 dB</td>
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<tr>
<td>24.5 GHz</td>
<td>5.39°</td>
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<tr>
<td>25 GHz</td>
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<td>5.75°</td>
<td>-24.23 dB</td>
<td>-23.90 dB</td>
</tr>
<tr>
<td>25.5 GHz</td>
<td>4.91°</td>
<td>5.36°</td>
<td>-23.36 dB</td>
<td>-23.35 dB</td>
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<tr>
<td>26 GHz</td>
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<td>4.96°</td>
<td>-17.51 dB</td>
<td>-20.57 dB</td>
</tr>
</tbody>
</table>

Table 4.5: Simulated and measured HPBW and cross-polarization values of $4 \times 16$ slot array on CSIW.

Figure 4.24: (a) Gain and (b) efficiency results of $4 \times 16$ SIW slot array.
measured and simulated data is consistent. The dielectric and conductor losses cause degradation in the efficiency values which is proved in the results presented in Table 4.3. The differences between the simulations and measurements in the gain and efficiency values are due to the fabrication imperfections. The fabrication imperfections may cause a considerable degradation in the performance of the structure designed at such high frequency applications which require high precision in the fabrication.

4.3. Slot Array Design on CSIW with Uniform Slot Separation

In this section, different \( l \times 25 \) slotted CSIW arrays are designed at 25 GHz to satisfy predefined radiation pattern criteria, i.e. the slotted CSIW array designs are examined in terms of pattern synthesis problem. For this purpose, a SLL value below -20 dB with pencil beam pattern is determined to be achieved by a traveling wave slot array on CSIW. Three different designs are carried out using the substrate and microstrip to SIW transition parameters mentioned in Section 2.1 and Section 2.2. 25 slot elements are used where the slot elements are uniformly distributed by \( 0.7\lambda_g \) spacing on CSIW with \( R = 10 \text{ cm} \) as shown in Figure 4.1. In the designs, first of all the required slot voltages are determined which gives the desired radiation pattern. The expression of the required slot voltage to be satisfied by the \( p^{th} \) slot in the array is given by

\[
V_p = a_p e^{j\psi_p} \tag{4.4}
\]

where \( a_p \) and \( \psi_p \) are the amplitude and phase of the required \( p^{th} \) slot voltage, respectively. Then using these required slot voltages, the proper slot offset and length values are found by applying Elliott’s design procedure as shown in Figure 4.25.
In the 1st 1 x 25 uniform slot array design, the aperture projection method is used to determine the excitation amplitudes, $W_p^{DE}$ shown in Figure 4.10 (b) where these weights are determined for one of the design cases presented in Section 4.1.1.2. Hence, the required slot amplitudes are chosen as $a_p = W_p^{DE}$ for $p = 1, 2, \ldots, 25$. In order to increase the directivity in the desired direction, $\hat{n}$, which is not necessarily in $\hat{a}_y$-direction as shown in Figure 4.1, the proper slot phase values, $\psi_p$ are determined which compensate phase delays due to different path lengths between the slots and planar aperture where the normal vector of the planar aperture, $\hat{n}$ shows the direction of the main beam. In the 1st design, the direction of the main beam is determined as $\phi_m = 92^\circ$ and the slot phases are determined as $\psi_p = -k_0 \hat{n} \cdot \hat{r}_p$. The reason of choosing the main beam direction as $92^\circ$ will be explained later. The slot phases are satisfied by allowing large deviations in the slot lengths from their resonance values in the design. In order to expect the deviation of the slot from the resonance, the mode voltages are initially assumed to be traveling only in the excitation direction exciting each slot without encountering any reflections due to the slots as shown in Figure 4.26. Hence, the mode voltage exciting the $p^{th}$ slot is given by

$$V'_p = a_p^v e^{i\phi_p} \quad (4.5)$$

where $a_p^v$ and $\phi_p = -L_p \beta_{10}$ are amplitude and phase of the mode voltage. $\beta_{10}$ is the propagation constant of the fundamental mode used to excite the waveguide and $L_p$
is the distance between the Port1 and the center of the slot as shown in Figure 4.26.

In (4.5), prime sign is used in the mode voltage expression in order to discriminate
the mode voltage calculated in (4.5) from the actual mode voltage calculated in
(3.78). The difference between (3.78) and (4.5) is that the actual mode voltages are
calculated in (3.78) while the Elliott’s design procedure is applied; however, the
approximate mode voltages in (4.5) are used before the Elliott’s design procedure
while making initial design on the slot voltages. The actual mode voltages found
using (3.78) are given by

\[ V_p = a_p^e e^{j\phi_p^e} \quad (4.6) \]

where \( a_p^e \) and \( \phi_p^e \) are the amplitude and phase of the actual mode voltage for the \( p^{th} \)
slot, respectively.

The relation between the slot voltage and mode voltage is previously derived in
(3.29). In order a slot to be at resonance, the active admittance presented in (3.29)
should be purely real positive. Since the term \( K_1 \) in the active admittance formula in
(3.29) is purely imaginary, there should be \( 90^\circ \) phase difference between the slot
phase and mode voltage phase when the slot is at resonance. Hence, the phase value,
\( \phi_p^l \) is defined as \( \phi_p^l = \phi_p - 90^\circ \) such that the slot is assumed to be at resonance if
\[ \phi_p^1 = \psi_p \text{ or } \phi_p^1 = \psi_p + 180^0 \]. Note that the case where \( \phi_p^1 = \psi_p + 180^0 \) corresponds to alternating the slot to the other half side of the waveguide compared to the case where \( \phi_p^1 = \psi_p \). The deviation of the slot from resonance is defined for the \( p^{th} \) slot as

\[ D_p^1 = e^{(j\psi_p - \psi_p)} \] (4.7)

Similarly, the phase value, \( \phi_p^2 \) is defined as \( \phi_p^2 = \phi_p^1 - 90^0 \) such that the slot is actually at resonance if \( \phi_p^2 = \psi_p \) or \( \phi_p^2 = \psi_p + 180^0 \). The actual deviation of the slot from resonance is defined for the \( p^{th} \) slot as

\[ D_p^2 = e^{(j\psi_p - \psi_p)} \] (4.8)

Since the values of \( D_p^2 \) are not known before determining the slot offsets and lengths in the Elliott’s design procedure, the initial design of slot phases, \( \psi_p \) are carried out using (4.7).

In the 1st 1 x 25 uniform slot array design, \( D_p^1 \) is plotted by cross labels in Figure 4.27 (a) on the unit circle. The slots corresponding to the \( D_p^1 \) points located at \( 0^0 \) and \( 180^0 \) in Figure 4.27 (a) are considered to be at resonance for the positive and negative slot offset directions, respectively. The deviation of the slot phase, \( \psi_p \) from the phase, \( \phi_p^1 \) determines the deviation of the slot from the resonance. \( \pm 50^0 \) deviation from the resonance is considered to be safe enough to avoid numerical problems [54]. Moreover, the decrease in the efficiency of the array due to allocating elements out of resonance is limited by preventing the slots to deviate from resonance more than \( 50^0 \).
The graph shown in Figure 4.27 (a) is divided into four regions using the constant phase lines defined by \( \pm 50^\circ \) deviation from 0° and 180°. The regions I and III in Figure 4.27 (a) are considered to be the safe regions in which the slots deviate from resonance by maximum amount of \( \pm 50^\circ \). In Figure 4.27 (a), changing the resonant slot element requires a constant phase shift applied in \( \psi_p \) values for all the elements. Hence, it causes a rotation of the phase deviation values on the unit circle. Changing

Figure 4.27. Phase and error graphs for 1 x 25 slot array on CSIW (1st design): (a) Deviation of slot phases from mode voltage phases (\( \angle D^1_p \)). (b) Deviation of slot phases from mode voltage phases in Elliott’s design (\( \angle D^2_p \)). (c) Error in the Elliot’s design.
the desired main beam direction also affects the required slot phases. Hence, the main beam direction, $\phi_m^a$ is chosen as 92° as shown in Figure 4.1 and also the first element which is closest to the match load in the CSIW is chosen as a resonant element in order to maximize the phase points in regions I and III. The phase of the resonant element is adjusted by applying a constant phase shift on all of the desired slot voltage phases found by $\psi_p = -k_o \hat{n} \tilde{r}_p$ so that $\phi_p = \psi_1$. Hence, the required slot phases are determined.

The Elliott’s design procedure is applied using the slot excitation amplitudes, $a_p$ and phases, $\psi_p$ to determine the slot offsets and lengths for the $1^{st}$ design. In the Elliott’s design, the active admittance formulas derived for longitudinal slot array on cylindrical waveguide in Section 3.1.2 are used. The slot offsets and lengths found for the $1^{st} 1 \times 25$ slot array design on CSIW are listed in Table 4.6. The phase deviations of the slots, $D_p^2$ calculated after the Elliott’s design, are plotted on the unit

Table 4.6: Slot offsets and lengths for $1 \times 25$ slot array on CSIW ($1^{st}$ design).
circle in Figure 4.27 (b). As it is seen Figure 4.27 (b), some of the phase values get closer to the +50° line compared to the phase values shown in Figure 4.27 (a). Although, the slot phases are affected by initially assuming a traveling wave mode voltage, Figure 4.27 (a) and (b) shows that designing the slot phases by assuming $D_p^1$ is a good approximation.

The error function used in the Elliott’s design procedure is defined in Section 3.2. The error function calculated at the final step in the Elliott’s design procedure is plotted for each slot in Figure 4.27 (c). As it is seen in Figure 4.27 (c), there is large error above 50%. The large error observed at the 25th slot in Figure 4.27 (c) is due to the small slot offset value as listed in Table 4.6 for the 25th slot. In the Elliott’s design, numerical errors might be encountered when it is required to find the small slot offset values because small slot offsets have small shunt admittance values which are highly affected by the mutual couplings. Hence, numerical errors might be encountered if the slot offset value required to be determined is small depending on the value of the mutual coupling value for the slot under concern. The error above 10% for the slots other than 25th slot occurs for 9th, 12th, 13th, 15th, and 16th slots which are close to the center of the array. These errors in 5 different slots are caused by 5 slots largely deviating from resonance as shown in Figure 4.27 (b). Since the imaginary part of the self-admittance shown in Figure 2.10 (b) has a limited range, ±90° deviation from the resonance cannot be reached and errors increase as the slots deviate from resonance.

The slot amplitudes, $a_p$ and phases, $\psi_p$ desired to be satisfied in the 1st 1 x 25 slot array design is compared to the HFSS simulations in Figure 4.28 (a) and (b), respectively. The HFSS simulations are carried out by using the slot offsets and lengths given in Table 4.6 determined by the Elliott’s design procedure. In Figure 4.28 (a), the maximum relative deviation of the amplitude coefficients found by using Elliott’s design from the desired taper is about 0.18. This amount of error is
usually encountered in the Elliott’s design. However, large fluctuations observed at the peak positions are unusual if Elliott’s design is successfully applied.

The radiation pattern of the slots on CSIW are calculated in the H-plane by using an in-house code developed for calculating the slot array pattern where the slot element pattern is taken into account in the code. Using the desired slot excitation coefficients ($V_p$), the radiation pattern of the desired slot array on CSIW is calculated. The

![Figure 4.28](image_url)  

Figure 4.28. Slot voltage (a) amplitudes and (b) phases for $1 \times 25$ slot array on CSIW. (c) H-plane radiation patterns ($I^{th}$ design).
calculated pattern is compared with the HFSS simulation in Figure 4.28 (c) where the simulations are carried out using slot offsets and lengths found from the Elliott’s design listed in Table 4.6. Due to the fluctuations in the amplitude terms of the Elliott’s design in Figure 4.28 (a), desired SLL could not be reached in the H-plane radiation pattern of the Elliott’s design shown in Figure 4.28 (c). The difference between SLL values in the radiation pattern of the Elliott’s design and the desired array pattern is about 2.7 dB.

The fluctuations in the amplitude distribution of the Elliott’s design Figure 4.28 (a) is tried to be reduced by using an iterative procedure. The iterative procedure is developed using similar approach presented in [55]. In [55], the electric fields found by Method of Moments (MoM) formulation are scaled to obtain desired pattern while the active admittance formulas are scaled and HFSS simulations are used in this thesis to obtain desired slot voltage distribution in the simulations. The initial values of the slot offsets and lengths can be determined using Elliott’s design procedure, [55]. Once the Elliott’s design is completed, the term \( Y_p^a(z_p, l_p)/f_p^{TE}\) in (3.79) is known for all the slots for \( p = 1, 2, 3, \ldots, 25 \). This term is modified in each iteration as

\[
\frac{Y_p^{a,i}(z_p^i, l_p^i)}{f_p^{TE,i}(z_p^i, l_p^i)} = \alpha_p^i \frac{Y_p^{a,i-1}(z_p^{i-1}, l_p^{i-1})}{f_p^{TE,i-1}(z_p^{i-1}, l_p^{i-1})}
\]  

(4.9)

where \( \alpha_p^i \) is a scaling constant while \((z_p^i, l_p^i)\) is the slot offset and length pair for the \(p^{th}\) slot in the \(i^{th}\) iteration. In each iteration, the scaling constant \( \alpha_p^i \) is determined by comparing the simulated slot voltages with the desired slot voltages and new slot offset and length values are found for the slots using the scaled active admittance terms in (4.9). Then the simulations are repeated with new slot offsets and lengths to find new slot voltages in the simulations for the next iteration. The iterations are repeated until the convergence is obtained in the simulated and desired slot voltages.
The term \( Y_p^{a,0}(z_p^0, l_p^0) / f_p^{TE_{in},0}(z_p^0, l_p^0) \) which appears at the RHS of (4.9) is determined from Elliott’s design procedure for \( i = 1 \) and \( p = 1, \ 2, \ 3, \ \ldots, \ 25 \). The values of \( Y_p^{a,s}(z_p^{l}, l_p^0) / f_p^{TE_{in},0}(z_p^{l}, l_p^0) \) are determined for each slot in each iteration by searching the values of \( (z_p^{l}, l_p^0) \) as described in the 6th step of the Elliott’s design procedure presented in Section 3.2. In (4.9), it is assumed that scaling the active admittance value for the \( \text{p}^{th} \) slot does not affect the other active admittance values, similar to the idea of aperture decoupling as described in [55]. The mutual coupling terms can be recalculated using the slot offset and length values calculated from previous iteration and they can be used in the next iteration. Since the slot offsets and lengths are not expected to deviate much from their initial values, the mutual coupling terms which are calculated in the last step of the Elliott’s design procedure can be used in all the iterations to save the time required for the iterations. If the mutual couplings are ignored in the iterations, convergence can still be obtained. Furthermore, the substrate and metal losses, which are not considered in the Elliott’s design and encountered in the simulations, are also taken into account in this iteration technique.

The scaling constant \( \alpha_p^l \) in the iteration is given by

\[
\alpha_p^l = \sqrt{\frac{V_p^{des}}{V_p^{Sim,i-1}}} \tag{4.10}
\]

where \( V_p^{des} \) is the desired complex slot voltage distribution to be satisfied and they are already plotted in Figure 4.28 (a) and (b) for the \( 1^{st} \) 1 x 25 slot array design. The complex slot voltages, \( V_p^{Sim,i-1} \), are determined for all the slots by using the simulations carried out with the slot offset and length pairs \( (z_p^{l-1}, l_p^{l-1}) \) found at the \( (i - 1)^{st} \) iteration. Note that for \( i = 1 \), the values of \( V_p^{Sim,0} \) have already been plotted in Figure 4.28 (a) and (b) as dashed curves. The square root in (4.10) is used to avoid the oscillations of the slot voltages, \( V_p^{Sim,i-1} \) found at each iteration, and hence, to obtain smooth convergence in the slot voltage distribution. Both the amplitudes and
phases of the slot voltages are corrected in each iteration step. The iterations are carried out by taking the first slot as a reference such that the first slot offset, length, and the first slot voltage values are not changed in the iterations. This is the reason for the desired and simulated slot voltages for the first slot element to be equal in Figure 4.28 (a) and (b) and they are equal in all the iteration steps. Hence, the value of \( \left(z_p^i, l_p^f\right) \) are not changed for \( p = 1 \) in each iteration step. Moreover, the values of \( V_{p, Sim, i}^{Sim, i-1} \) obtained in the simulations for each iteration step are scaled by the same complex constant such that \( V_{p, Sim, i}^{Sim, i-1} \) is equated to \( V_{p, des}^{Sim, i} \) in each iteration step.

The iteration procedure can be used to achieve any slot distribution, however, the excitation weights to be realized are limited by the range of the phase values which can be achieved by the slots around the resonance. The iterations are carried out for the 1st 1 x 25 slotted CSIW array. The excitation coefficients and the radiation pattern diverge from desired patterns after 4 iterations as shown in Figure 4.28 because the phase values near 90° in Figure 4.27 (a) and (b) could not be achieved by the slot.

The 2nd 1 x 25 slot array design on CSIW is carried out by determining the desired slot voltages using a genetic algorithm code. The phase values, \( \Psi_p \) and the slot amplitudes, \( a_p \), are the inputs of the algorithm for \( p = 1, 2, 3, ..., 25 \). Hence, total 50 inputs constitute the genomes of the algorithm. The slot phases, \( \psi_p \), are designed using the input phases, \( \Psi_p \) in the code. The input values of the algorithm changes between 0 and 1. The amplitudes \( a_p \) are set free to vary continuously between 0 and 1, however, restrictions are applied on the input phase values, \( \Psi_p \). Since all the slot phases, \( \psi_p \) usually could not be designed by phase compensation technique so that phases of \( D^i_p \) are positioned in the safe regions I and III as shown in Figure 4.27 (a), and also since this situation causes problems in obtaining desired radiation pattern, the restrictions on slot phases are applied in the genetic algorithm. This is
accomplished as follows. First of all, the input phases, $\Psi_p$ varying between 0 and 0.5 are mapped to the angle values between $-35^\circ$ and $35^\circ$ and also $\Psi_p$ values varying between 0.5 and 1 are mapped to the angle values between $145^\circ$ and $215^\circ$. As a result of this mapping, the phase values, $\Psi_p^1$ are obtained. Then the slot voltage phases, $\psi_p$ are obtained by adding $\phi_p^1$ values to $\Psi_p^1$ values. Hence, the phase of $D_p^1$ in (4.7) becomes $\Psi_p$ which have maximum $\pm 35^\circ$ deviation from $0^\circ$ and $180^\circ$. After determining the slot phases, the H-plane radiation pattern of the slot array on CSIW is calculated using in-house code mentioned before. The SLL of the H-plane radiation pattern is calculated and it is minimized till the desired SLL value is reached in the code. No criterion is applied for HPBW because HPBW value becomes good enough when the SLL is suppressed low enough. The optimization is terminated when the SLL below desired value is reached. The population size in genetic algorithm is chosen as 500 in order to reach the global solution. The genetic algorithm optimizations are carried out and the desired radiation pattern is obtained as shown in Figure 4.29 (c). In Figure 4.29 (c), the main beam direction of the desired radiation pattern is $\phi_m^\circ = 80^\circ$ which is in the backward quadrant defined by $0^\circ < \phi_m^\circ < 90^\circ$ where $90^\circ$ is the broadside angle. This can be explained as follows.

The slotted waveguide antennas resemble to the periodic leaky wave antennas when the waveguide is dielectric filled and the fundamental mode propagating in the waveguide is a slow wave. In this study, the fundamental mode in the CSIW is a slow wave at 25 GHz. The main difference between uniform and periodic leaky wave antennas is that the former can radiate only into the forward quadrant, i.e., $90^\circ < \phi_m^\circ < 180^\circ$ while the latter can radiate into both forward and backward quadrants [48].

The slot amplitudes, $a_p$ and phases, $\psi_p$, found in the optimization are also shown as desired amplitudes and phases in Figure 4.29 (a) and (b), respectively. Note that the desired slot amplitudes, $a_p$ and phases, $\psi_p$ shown in Figure 4.29 differs from the
Figure 4.29. Slot voltage (a) amplitudes and (b) phases for \( l \times 25 \) slot array on CSIW. (c) H-plane radiation patterns (2\(^{nd}\) design). (d) Error in the Elliot’s design.

desired slot amplitudes and phases shown in Figure 4.28 because the slot voltage coefficients are determined by genetic algorithm in the 2\(^{nd}\) design shown in Figure 4.29 while the aperture projection method is used together with phase compensation in the 1\(^{st}\) design shown in Figure 4.28.

The Elliott’s design procedure is applied using the slot amplitudes, \( a_p \) and phases, \( \psi_p \) determined in the genetic optimization for the 2\(^{nd}\) \( l \times 25 \) slot array design on CSIW. Hence, the slot offsets and lengths of the 2\(^{nd}\) \( l \times 25 \) slot array are determined.
Using the slot offsets and lengths determined in the Elliott’s design, the simulation of the 2\textsuperscript{nd} 1 x 25 slot array is carried out. As it is seen in Figure 4.29 (c), the radiation pattern obtained from the simulations of the Elliott’s design does not satisfy desired -20 dB SLL because of the difference between the corresponding amplitude and phase curves obtained from simulations presented in Figure 4.29 (a) and (b), respectively. The maximum value of the error function in the Elliott’s design is found above 90 % only for the 12\textsuperscript{th}, 21\textsuperscript{st}, and 24\textsuperscript{th} slots as shown in Figure 4.29 (d). Although the phases of the elements are properly determined, this error is caused by the small slot offset values tried to be determined as mentioned before. As it is seen in Figure 4.29 (a), the desired amplitude taper reduces abruptly for the 12\textsuperscript{th} slot which makes the slot offset value of the 12\textsuperscript{th} slot so small that causes errors in the Elliott’s design procedure. Moreover, the desired voltage amplitudes for the 21\textsuperscript{st} and 24\textsuperscript{th} slots are also small which requires small slot offset values vulnerable to errors for these slots.

The iterations are carried out to find the slot offsets and length which satisfy desired slot voltages, and hence, to find the desired pattern. The slot voltages found after 7 iterations are plotted in Figure 4.29 (a) and (b). Iterations are terminated at 7 because no significant improvement is observed compared to the slot voltages and SLL value in the 6\textsuperscript{th} iteration. As it is seen in Figure 4.29 (a) and (b), the slot voltages at the 7\textsuperscript{th} iteration are quite close the desired curve. However, there are still some difference between the 7\textsuperscript{th} iteration curves and desired slot voltages. Although the radiation pattern of the 7\textsuperscript{th} iteration shown in Figure 4.29 (c) almost converge to the desired radiation pattern, the desired -20 dB SLL criteria could not be satisfied due to these small differences. The difference between the desired SLL and the SLL obtained at 7\textsuperscript{th} iteration is 2.43 dB.

In the iterations of the 2\textsuperscript{nd} 1 x 25 slot array design, the smooth convergence could not be obtained because of high values of the error function found for the 12\textsuperscript{th} slot which is located at the center of the array. The errors arising from small slot offset values at
two ends of the waveguide might be acceptable because the required amplitude taper reduces toward ends. Since there is an abrupt change in the amplitude taper at the 12th slot, the offset value of the 12th slot is too small which causes an error above 100%. This error at the center slot affects the radiation pattern. Due to such large errors, the number of iterations to be carried out increases and as it is seen in Figure 4.29 (a), the convergence is still not obtained after 7 iterations. There is still some difference between the slot voltages for the 12th slot having small slot offset which makes the iteration procedure difficult. Hence, large oscillations in the desired slot voltage amplitudes are not suitable because they increase the number of small slot offset values to be implemented which increases the errors in the code. Moreover, the iterations could not be implemented due to these small slot offsets. It is more proper to use the slot voltage amplitude distribution which monotonically increases till the center of the array and then decreases toward the end slot.

The 3rd design of 1 x 25 slot array on CSIW is carried out by employing a smooth variation in the desired slot voltage amplitude. For this reason, the slot amplitudes, \( a_p \), are determined by the weights, \( W_p^{DE} \), calculated using (4.3). \( W_p^{DE} \) values are calculated similar to the ones shown in Figure 4.10 using the aperture projection method with -25 dB SLL Taylor (\( \hat{n} = 3 \)) line source distribution. The phase only synthesis is carried out using genetic algorithm to determine the slot phases, \( \psi_p \). Using the procedure similar to the 2nd 1 x 25 slot array design on the input phases, \( \Psi_p \), the genetic algorithm optimization is applied such that the maximum allowed phase deviation in \( D_p \) from 0° and 180° is ±50° which is chosen large enough to help the code to find the desired pattern. As a result of disturbing the compensated slot phases by restricting the maximum deviation of the slot elements from resonance to ±50°, the SLL values are expected to increase. The genetic algorithm code is used with a safety margin of 2.5 dB in the SLL goal such that the desired SLL value is set to -22.5 dB. This safety margin is applied in the code in order to compensate possible increase in the SLL due to the errors encountered in the Elliott’s design procedure.
Similar to the 2\textsuperscript{nd} \(1\times25\) slot array design, the population size in genetic algorithm is chosen as 500 and the optimizations are carried out for the 3\textsuperscript{rd} design. The desired slot phases, \(\psi_p\), are determined after the optimizations. The desired slot amplitudes, \(\alpha_p = W_p^{DE}\), determined from -25 dB SLL Taylor (\(\bar{\eta} = 3\)) distribution and the desired slot phases, \(\psi_p\), determined by the genetic optimization are shown in Figure 4.30 (a) and (b), respectively. The radiation pattern obtained from the desired slot voltages and phases is the pattern synthesized in the genetic algorithm which is plotted as a desired pattern shown in Figure 4.30 (c). In Figure 4.30 (c), the main beam direction

Figure 4.30. Slot voltage (a) amplitudes and (b) phases for \(1\times25\) slot array on CSIW. (c) H-plane radiation patterns (3\textsuperscript{rd} design).
of the desired radiation pattern is $\phi^a = 88^\circ$. In the optimization, the SLL of the desired radiation pattern is found as around -22.4 dB.

The deviation of the slot phases from resonance, $D_p^1$, found after the genetic algorithm code is shown in Figure 4.31 (a). As it is seen in Figure 4.31 (a), all the slots are in the safe regions which are regions I and III. The Elliott’s design

![Figure 4.31](image)

Figure 4.31. Phase and error graphs for 1 x 25 slot array on CSIW (3rd design): (a) Deviation of slot phases from mode voltage phases ($\angle D_p^1$). (b) Deviation of slot phases from mode voltage phases in Elliott’s design ($\angle D_p^2$). (c) Error in the Elliot’s design.
procedure is carried out to find the slot offsets and lengths to satisfy desired slot voltage distribution. The slot offsets and lengths found after the Elliott’s design procedure is listed in Table 4.7. The error obtained in the Elliott’s design code is presented in Figure 4.31 (c). The high error above 50% is calculated for the 24th slot. As it is seen in Table 4.7, the value of the offset is small for the 24th slot and the large error is caused by this small slot offset value. The phase deviation values, $D_p^2$, calculated after the Elliott’s design code are plotted in Figure 4.31 (b). As it is seen in Figure 4.31 (b), the slots do not deviate much from the safe region after the Elliott’s design procedure.

The simulations are carried out using the slot offset and length values listed in Table 4.7 found by Elliott’s design. The slot voltage distribution found from simulations of the Elliott’s design is compared with the desired slot voltages in Figure 4.30 (a) and (b). The maximum difference between the slot amplitudes is found as about 0.3 around the peak position in Figure 4.30 (a). Furthermore, the maximum difference in the phases is 0.486 rad at the 24th slot. These differences between the Elliott’s design

Table 4.7: Slot offsets and lengths for 1 x 25 slot array on CSIW (3rd design).

<table>
<thead>
<tr>
<th>Element #</th>
<th>Slot Offset (mm)</th>
<th>Slot Length (mm)</th>
<th>Element #</th>
<th>Slot Offset (mm)</th>
<th>Slot Length (mm)</th>
</tr>
</thead>
<tbody>
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<td>4.175</td>
<td>14</td>
<td>0.117</td>
<td>3.777</td>
</tr>
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</tr>
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<td>-0.092</td>
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</tr>
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<td>18</td>
<td>0.142</td>
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</tr>
<tr>
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<td>-0.141</td>
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</tr>
<tr>
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<td>20</td>
<td>0.055</td>
<td>3.938</td>
</tr>
<tr>
<td>8</td>
<td>-0.377</td>
<td>3.904</td>
<td>21</td>
<td>-0.076</td>
<td>3.951</td>
</tr>
<tr>
<td>9</td>
<td>0.452</td>
<td>4.422</td>
<td>22</td>
<td>0.061</td>
<td>3.782</td>
</tr>
<tr>
<td>10</td>
<td>0.144</td>
<td>3.790</td>
<td>23</td>
<td>-0.197</td>
<td>3.680</td>
</tr>
<tr>
<td>11</td>
<td>-0.458</td>
<td>4.343</td>
<td>24</td>
<td>-0.057</td>
<td>4.238</td>
</tr>
<tr>
<td>12</td>
<td>-0.079</td>
<td>3.789</td>
<td>25</td>
<td>0.042</td>
<td>3.897</td>
</tr>
<tr>
<td>13</td>
<td>0.275</td>
<td>3.969</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
code and simulations are usually encountered. Due to these differences both in the slot amplitudes and phases, the simulated radiation pattern of the Elliott’s design shown in Figure 4.30 (c) has small deviation from the desired radiation pattern.

The iterations as mentioned before are carried out to obtain desired slot excitation coefficients at the slots. The design found using Elliott’s design procedure is iterated and the slot offsets and lengths listed Table 4.8 are found after 6 iterations. The simulations are carried out using the slot parameters listed in Table 4.8 for the 6th iteration. The slot voltages and corresponding radiation pattern simulated for the 6th iteration are presented in Figure 4.30. As it is seen in Figure 4.30 (a) and (b), the simulated slot voltages for the 6th iteration are quite close to the desired slot voltages. Although, the SLL could not be improved further in the iterations in Figure 4.30 (c), some of the side lobes are seen to get closer to the desired pattern. The difference between the desired SLL and the SLL obtained at 6th iteration is 1.54 dB. Moreover, the simulated $S_{11}$ is reduced from -11.5 dB to -13.5 dB after 6 iterations at 25 GHz.

Table 4.8: Slot offsets and lengths for 1 x 25 slot array on CSIW after 6th iteration (3rd design).

<table>
<thead>
<tr>
<th>Element #</th>
<th>Slot Offset (mm)</th>
<th>Slot Length (mm)</th>
<th>Element #</th>
<th>Slot Offset (mm)</th>
<th>Slot Length (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-0.250</td>
<td>4.175</td>
<td>14</td>
<td>0.110</td>
<td>3.797</td>
</tr>
<tr>
<td>2</td>
<td>-0.128</td>
<td>4.068</td>
<td>15</td>
<td>-0.286</td>
<td>3.756</td>
</tr>
<tr>
<td>3</td>
<td>-0.175</td>
<td>4.096</td>
<td>16</td>
<td>0.174</td>
<td>3.893</td>
</tr>
<tr>
<td>4</td>
<td>-0.194</td>
<td>4.111</td>
<td>17</td>
<td>-0.086</td>
<td>3.925</td>
</tr>
<tr>
<td>5</td>
<td>-0.178</td>
<td>4.022</td>
<td>18</td>
<td>0.126</td>
<td>3.976</td>
</tr>
<tr>
<td>6</td>
<td>-0.211</td>
<td>3.947</td>
<td>19</td>
<td>-0.120</td>
<td>4.092</td>
</tr>
<tr>
<td>7</td>
<td>-0.223</td>
<td>3.915</td>
<td>20</td>
<td>0.060</td>
<td>3.961</td>
</tr>
<tr>
<td>8</td>
<td>-0.341</td>
<td>3.868</td>
<td>21</td>
<td>-0.073</td>
<td>3.945</td>
</tr>
<tr>
<td>9</td>
<td>0.381</td>
<td>4.265</td>
<td>22</td>
<td>0.069</td>
<td>3.810</td>
</tr>
<tr>
<td>10</td>
<td>0.136</td>
<td>3.811</td>
<td>23</td>
<td>-0.163</td>
<td>3.686</td>
</tr>
<tr>
<td>11</td>
<td>-0.409</td>
<td>4.249</td>
<td>24</td>
<td>-0.056</td>
<td>4.104</td>
</tr>
<tr>
<td>12</td>
<td>-0.089</td>
<td>3.820</td>
<td>25</td>
<td>0.047</td>
<td>3.942</td>
</tr>
<tr>
<td>13</td>
<td>0.255</td>
<td>3.957</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
4.3.1. Fabrication and Measurements

The slot array which is found after 6th iteration in the 3rd design of 1 x 25 slot array on CSIW is selected for fabrication. The slot offsets and lengths of the array to be fabricated are listed in Table 4.8. The fabrication of the slot array is carried out by Ilfa company which is founded for the fabrication of various PCB products. The array is fabricated on 0.5 mm thick Rogers 3003 substrate where the substrate has 35 µm thick copper cladding on both sides. Electroless nickel immersion gold (ENIG) surface plating is applied on the copper metallization to protect the copper from oxidation.

4.3.1.1. 1 x 25 Slot Array on CSIW

The fabricated 1 x 25 slot array is shown in Figure 4.32 (a). The s-parameters of the 1 x 25 slot array are measured and compared with the simulations in Figure 4.32 (b) and (c). The simulations are carried out both by using zero metal thickness and 35 µm copper thickness in HFSS. The finite conductivity boundary condition is assigned on the metal surfaces with zero metal thickness using copper conductivity in HFSS. As it is seen in Figure 4.32 (b) and (c), there is quite good agreement between the simulated and measured results. Both simulated and measured $S_{11}$ values shown in Figure 4.32 (b) are below -10 dB between 20 GHz and 30 GHz. The measured $S_{21}$ values successfully mimic the simulated $S_{21}$ results in Figure 4.32 (c). The simulations carried out using 35 µm copper thickness are in better agreement with the measured results compared to the simulations with zero metal thickness.
The radiation patterns of the 1 x 25 slotted CSIW array is measured in an anechoic chamber using the measurement setup shown in Figure 4.33. The patterns are measured and compared to simulation results. The measurement setup is shown in Figure 4.33.

Figure 4.32: (a) Fabricated traveling wave 1 x 25 slot array on SIW. (b) $S_{11}$ and (c) $S_{21}$ results of the 1 x 25 slotted SIW array.

The radiation patterns of the 1 x 25 slotted CSIW array is measured in an anechoic chamber using the measurement setup shown in Figure 4.33. The patterns are measured and compared to simulation results. The measurement setup is shown in Figure 4.33.

Figure 4.33: Measurement setup of 1 x 25 slot array on CSIW ($R = 10 \text{ cm}$).
measured between 24 GHz and 26 GHz with 0.5 GHz steps. The measurements are taken with 1° angular resolution. The measured radiation patterns are compared with the simulations in Figure 4.34. The simulated radiation patterns in Figure 4.34 are obtained using 35 µm copper thickness. Good agreement is observed between the measured and simulated radiation patterns in Figure 4.34. The SLL at 25 GHz is below -20 dB in the simulations of the 25 element slotted CSIW array obtained after 6 iterations shown in Figure 4.30 (c) because this simulation does not take the conductor thickness into account. A waveguide with small length equal to the thickness of the copper is obtained in the slot regions, and hence, the slot voltages are affected which disturbs the SLL values in the initial design. In order to reduce the simulation time, the conductor thickness was ignored in the simulations and it was also not expected that the conductor thickness has such a large influence on the slot excitation coefficients. For this reason, the fine tuning of the design was not applied by considering the conductor thickness. It is concluded that if the design described in Figure 4.30 is carried out by considering the metal thickness in the simulations, desired SLL values can be reached.

The simulated and measured main beam positions are compared in Figure 4.34 (f). Good agreement is observed between the simulated and measured main beam positions. About 10° beam steering is obtained when the frequency is changed between 24 GHz and 26 GHz. Since the measurements are taken with 1° angular resolution, the maximum difference between the measured and simulated main beam positions is about 1°.
Figure 4.34: Radiation patterns of $1 \times 25$ slot array on CSIW.
The HPBW values of the patterns are calculated by interpolating the measured and simulated patterns. Moreover, the cross polarization levels of the patterns are measured. The HPBW values and cross polarization levels measured for different frequencies between 24 GHz and 26 GHz are compared with the simulations which take the effect of conductor thickness into account in Table 4.9. Good agreement is obtained between the measured and simulated results in Table 4.9.

Table 4.9: Simulated and measured HPBW and cross-polarization values of 1 x 25 slot array on CSIW.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>HPBW</th>
<th>Cross-pol.</th>
</tr>
</thead>
<tbody>
<tr>
<td>24 GHz</td>
<td>4.90°</td>
<td>5.00°</td>
</tr>
<tr>
<td>24.5 GHz</td>
<td>4.83°</td>
<td>5.00°</td>
</tr>
<tr>
<td>25 GHz</td>
<td>4.58°</td>
<td>5.00°</td>
</tr>
<tr>
<td>25.5 GHz</td>
<td>4.88°</td>
<td>5.02°</td>
</tr>
<tr>
<td>26 GHz</td>
<td>4.81°</td>
<td>4.91°</td>
</tr>
</tbody>
</table>

The gain values of 1 x 25 slot arrays on CSIW are measured using the standard gain horn antenna for the frequencies between 24 GHz and 26 GHz with 0.5 GHz steps. The measured gain values are compared with the simulations which take the effect of conductor thickness into account in Figure 4.35 (a). Since there is good agreement between the simulated and measured radiation patterns in Figure 4.34, the simulated directivity values are used to find the measured efficiency of the array in Figure 4.35 (b). The measured and simulated data is seen to be consistent with each other in Figure 4.35. The simulated gain at 25 GHz is about 15.3 dB while the measured gain is about 14.5 dB. The maximum deviation between the measured and simulated gain is about 1.47 dB at 25.5 GHz. The small differences between the simulations and measurements in the gain and efficiency values are due to the small fabrication imperfections.
The slotted CSIW arrays with uniform and non-uniform slot separations have both some advantages and disadvantages compared to each other. Using non-uniform slot separations is advantageous compared to uniform slot separations on CSIW because the former has slots near resonance while the latter allocates slots with large deviations from resonance, and hence, less number of active admittance calculations is required around the resonance for the design of slot array on CSIW with non-uniform slot separations. Furthermore, some of the complex slot voltage phases cannot be realized in the slotted CSIW array with uniform slot separation because the imaginary part of the slot admittance has limited range. In order to obtain desired radiation pattern using uniform slot separations, some of the optimization techniques can be used to find proper phase values both satisfying the desired pattern criteria and limiting the deviation of slots from resonance; however, the design complexity is increased. Furthermore, the pattern bandwidth of the slotted CSIW array with uniform slot separations might be adversely affected by the slots largely deviating from resonance because undesired slot modes might appear as a result of small frequency change. On the other hand, using uniform slot separations is advantageous compared to allocating non-uniform slot distribution in the slotted CSIW array design because a pattern with desired SLL can be synthesized by using uniform slot separations. In the slotted CSIW array with non-uniform slot separations, the element

Figure 4.35: (a) Gain and (b) efficiency results of $1 \times 25$ SIW slot array.
separations are properly determined by the phase distribution requirements to obtain desired main beam and large element spacings might be encountered in the design which may cause undesired SLL values as discussed in Section 4.1.1. As a conclusion, based on the experience gained throughout this thesis study, uniform spacing can be preferred to meet high gain and low SLL design requirement of slotted CSIW arrays.
CHAPTER 5

CONCLUSIONS AND FUTURE WORK

The main goal of this thesis is to analyze, design, manufacture, and measure a traveling wave slot array at $25 \, GHz$ on a novel CSIW structure where the CSIW lies in the circumferential direction of the cylinder. Different designs are carried out using non-uniform and uniform slot separations for beam steering and pattern synthesis applications. The slot separations are designed properly to obtain in phase contribution of each slot element to increase directivity in the non-uniform array. The aperture projection method is used to decrease the SLL, however, desired SLL could not be obtained in the non-uniform slot array because of large slot separation values. Hence, the slot lengths of the non-uniform array are optimized further to increase the bandwidth for beam steering applications. Although the SLL is increased to $-10 \, dB$ after the optimizations, the fabricated non-uniform slot array has about $10^\circ$ beam steering when the frequency is swept between $24 \, GHz$ and $26 \, GHz$. Three different designs are carried out using the uniform slot array to obtain pencil beam pattern and $-20 \, dB$ SLL. The desired pattern is obtained at $25 \, GHz$ using both the aperture projection method and phase only synthesis applied by genetic algorithm in the uniform slot array.

Based on the research carried out in the frame of this thesis, following conclusions can be drawn as follows:

- The mutual coupling and active admittance formulations are carried out for the slots on a novel CSIW structure using E-plane bend regular waveguides.
These novel formulations can be used to design slotted bend regular waveguide arrays which is not available in the literature.

- The derived formulations are validated by full wave simulations. The active admittance formulas are used in the Elliott’s design procedure to design slotted CSIW arrays. The measurement results and the full wave simulations show that this novel array structure can successfully be designed using the proposed design formulas.

- The Elliott’s design procedure is applied together with aperture projection method and genetic algorithm to satisfy both the desired radiation pattern and input match criteria. The iteration procedure is proposed for fine tuning of the slot voltages after the Elliott’s design procedure.

- The slot array designs are successfully carried out for frequency steering and beam shaping applications. The antennas are fabricated and good agreement is obtained between the simulated and measured results.

The future works are summarized as follows:

- The active admittance formulations and Elliott’s design procedure can be improved to take dielectric and conductor losses into account in the design.

- The iteration procedure can be carried out by taking the metal thickness into in the design of uniform slot array.

- The band of the slotted CSIW array can be improved by designing group of arrays operating at different frequencies in order to increase the scan coverage for the frequency steering applications.

- The dielectric losses which reduce the efficiency of the slotted CSIW array can be eliminated by improving the fabrication process to remove part of the substrate in CSIW.
REFERENCES


APPENDIX A

SYSTEMATIC CONSTRUCTION OF FIELDS IN BEND REGULAR WAVEGUIDES

In a homogeneous source-free region, the equations satisfied by the electric and magnetic fields are given by [38]

\[-\nabla \times \mathbf{E} = \hat{z} H\] (A.1)

\[\nabla \times \mathbf{H} = \hat{y} E\] (A.2)

\[\nabla \cdot \mathbf{E} = 0\] (A.3)

\[\nabla \cdot \mathbf{H} = 0\] (A.4)

The field components due to magnetic vector potential can be found using

\[\nabla \cdot \mathbf{H}^A = 0 \Rightarrow \mathbf{H}^A = \nabla \times \mathbf{A}\] (A.5)

\[\mathbf{E}^A = \hat{y}^{-1}(\nabla (\nabla \cdot \mathbf{A}) + k^2 \mathbf{A})\] (A.6)

The field components due to electric vector potential can be found using

\[\nabla \cdot \mathbf{E}^F = 0 \Rightarrow \mathbf{E}^F = -\nabla \times \mathbf{E}\] (A.7)

\[\mathbf{H}^F = \hat{z}^{-1}(\nabla (\nabla \cdot \mathbf{E}) + k^2 \mathbf{E})\] (A.8)
where \( \hat{y} = j \omega \varepsilon \) and \( \hat{z} = j \omega \mu \). \( \varepsilon \) and \( \mu \) are the permittivity and permeability of the medium, respectively, and \( \omega \) is the angular frequency.

The field components inside the E-plane bend regular waveguide shown in Figure 3.1 can be found using (A.5)-(A.8). The magnetic vector potential and transverse magnetic to z \( (TM^z_{mn}) \) modes in the E-plane bend regular waveguide shown in Figure 3.1 are given by [37]

\[
\vec{A} = \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} \bar{A}_{mn} \cos\left(\frac{m \pi x}{a}\right) C_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right) e^{z j \varepsilon_{mn} \phi} \hat{z} \tag{A.9}
\]

\[
E_{\rho}^A = -\frac{1}{\hat{y}} \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} \bar{A}_{mn} \frac{m \pi}{a} \sin\left(\frac{m \pi x}{a}\right) C'_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right) e^{z j \varepsilon_{mn} \phi} \tag{A.10}
\]

\[
E_{\phi}^A = \frac{j}{\hat{y} \rho} \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} \bar{A}_{mn} \frac{m \pi}{a} \sin\left(\frac{m \pi x}{a}\right) C'_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right) \varepsilon_{mn} e^{z j \varepsilon_{mn} \phi} \tag{A.11}
\]

\[
E_{z}^A = \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} \bar{A}_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right)^2 \cos\left(\frac{m \pi x}{a}\right) C_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right) e^{z j \varepsilon_{mn} \phi} \tag{A.12}
\]

\[
H_{\rho}^A = \pm \frac{j}{\rho} \sum_{m=0}^{\infty} \sum_{n=1}^{\infty} \bar{A}_{mn} \cos\left(\frac{m \pi x}{a}\right) C_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right) \varepsilon_{mn} e^{z j \varepsilon_{mn} \phi} \tag{A.13}
\]

\[
H_{\phi}^A = -\sum_{m=0}^{\infty} \sum_{n=1}^{\infty} \bar{A}_{mn} \cos\left(\frac{m \pi x}{a}\right) C_{mn} \left(\frac{k_{TM}^{+}}{\rho}\right) \varepsilon_{mn} e^{z j \varepsilon_{mn} \phi} \tag{A.14}
\]

\[
H_z^A = 0 \tag{A.15}
\]

\[
\vec{E}^A = E_\rho^A \hat{\rho} + E_\phi^A \hat{\phi} + E_z^A \hat{z} \tag{A.16}
\]

\[
\vec{H}^A = H_\rho^A \hat{\rho} + H_\phi^A \hat{\phi} + H_z^A \hat{z} \tag{A.17}
\]
where \( J_{\varepsilon_{mn}}(k_{\rho}^{TE_{mn}}\rho) \) and \( Y_{\varepsilon_{mn}}(k_{\rho}^{TM_{mn}}\rho) \) are Bessel functions of first and second kinds, respectively, with order \( \varepsilon_{mn} \) and argument \( k_{\rho}^{TE_{mn}}\rho \). \( \varepsilon_{mn} \) are the roots of (A.20). The prime sign indicates the derivative with respect to the argument.

The electric vector potential and transverse electric to \( z \) (\( TE_{mn}^z \)) modes in the E-plane bend regular waveguide shown in Figure 3.1 are given by [37]

\[
F = \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} F_{mn} \sin \left( \frac{m\pi z}{a} \right) B_{\mu_{mn}}(k_{\rho}^{TE_{mn}}\rho) e^{\pm j\mu_{mn}\phi} \hat{a}_z
\]  
(A.21)

\[
E_\rho^F = \pm \frac{j}{\rho} \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} F_{mn} \sin \left( \frac{m\pi z}{a} \right) B_{\mu_{mn}}(k_{\rho}^{TE_{mn}}\rho) \mu_{mn} e^{\pm j\mu_{mn}\phi}
\]  
(A.22)

\[
E_\phi^F = \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} F_{mn} \sin \left( \frac{m\pi z}{a} \right) B_{\mu_{mn}}'(k_{\rho}^{TE_{mn}}\rho) k_{\rho}^{TE_{mn}}\rho e^{\pm j\mu_{mn}\phi}
\]  
(A.23)

\[
E_z^F = 0
\]  
(A.24)

\[
H_\rho^F = \frac{1}{2} \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} F_{mn} \frac{m\pi}{a} \cos \left( \frac{m\pi z}{a} \right) B_{\mu_{mn}}'(k_{\rho}^{TE_{mn}}\rho) k_{\rho}^{TE_{mn}}\rho e^{\pm j\mu_{mn}\phi}
\]  
(A.25)

\[
H_\phi^F = \pm \frac{j}{2\rho} \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} F_{mn} \frac{m\pi}{a} \cos \left( \frac{m\pi z}{a} \right) B_{\mu_{mn}}(k_{\rho}^{TE_{mn}}\rho) \mu_{mn} e^{\pm j\mu_{mn}\phi}
\]  
(A.26)
\[ H_z^F = \sum_{m=1}^{\infty} \sum_{n=0}^{\infty} F_{mn} \frac{(k_{\rho}^{TE}_{mn})^2}{2} \sin \left( \frac{m \pi z}{a} \right) B_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) e^{z j \mu_{mn} \phi} \]  
(A.27)

\[ E^F = E^F_{\rho} \hat{\alpha}_{\rho} + E^F_{\phi} \hat{\alpha}_{\phi} + E^F_z \hat{\alpha}_z \]  
(A.28)

\[ \overline{H}^F = H^F_{\rho} \hat{\alpha}_{\rho} + H^F_{\phi} \hat{\alpha}_{\phi} + H^F_z \hat{\alpha}_z \]  
(A.29)

\[ k_{\rho}^{TE}_{mn} = \sqrt{k^2 - \left( \frac{m \pi}{a} \right)^2}, \quad \text{Im}(k_{\rho}^{TE}_{mn}) \leq 0 \]  
(A.30)

\[ B_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) = J'_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) Y_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) - J_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) Y'_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) \]  
(A.31)

\[ B'_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) = \frac{\partial B_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right)}{\partial \left( k_{\rho}^{TE}_{mn} \rho \right)} \]  
(A.32)

\[ B'_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) \bigg|_{\rho = \rho_2} = \frac{\partial B_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right)}{\partial \left( k_{\rho}^{TE}_{mn} \rho \right)} \bigg|_{\rho = \rho_2} \]  
(A.33)

where \( J_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) \) and \( Y_{\mu_{mn}} \left( k_{\rho}^{TE}_{mn} \rho \right) \) are Bessel functions of first and second kinds, respectively, with order \( \mu_{mn} \) and argument \( k_{\rho}^{TE}_{mn} \rho \cdot \mu_{mn} \) are the roots of (A.33).
APPENDIX B

FIELDS RADIATED BY APERTURES ON PEC CYLINDER

The cylindrical transform of the tangential electric fields on the PEC cylindrical surface with aperture shown in Figure B1 is given by, [38]

$$\widetilde{E}_z(n, k_z) = \frac{1}{2\pi} \int_0^{2\pi} d\phi \int_{-\infty}^{\infty} dz E_z(\rho, \phi, z) e^{-jm\phi} e^{-jk_z z}$$  \hspace{1cm} (B.1)

$$\widetilde{E}_\phi(n, k_z) = \frac{1}{2\pi} \int_0^{2\pi} d\phi \int_{-\infty}^{\infty} dz E_\phi(\rho, \phi, z) e^{-jm\phi} e^{-jk_z z}$$  \hspace{1cm} (B.2)

Figure B1: PEC cylindrical surface with aperture.
The inverse transformation is

\[ E_z(\rho, \phi, z) = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} \tilde{E}_z(n, k_z) e^{j\rho k_z^2 dk_z} \] (B.3)

\[ E_\phi(\rho, \phi, z) = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} \tilde{E}_\phi(n, k_z) e^{j\rho k_z^2 dk_z} \] (B.4)

Using the wave functions

\[ A_z = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} f_n(k_z) H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2}) e^{j\rho k_z^2 dk_z} \] (B.5)

\[ F_z = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} g_n(k_z) H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2}) e^{j\rho k_z^2 dk_z} \] (B.6)

and using (A.1)-(A.8), the fields radiated by the aperture field can be found as

\[ E_z(\rho, \phi, z) = \frac{1}{j \omega \varepsilon_0} \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} (k_z^2 - k_0^2) f_n(k_z) H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2}) e^{j\rho k_z^2 dk_z} \] (B.7)

\[ E_\phi(\rho, \phi, z) = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} \left[ -\frac{n k_z}{j \omega \varepsilon_0 \rho} f_n(k_z) H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2}) + \frac{g_n(k_z) k_0^2 - k_z^2 H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2})}{g_n(k_z) k_0^2 - k_z^2 H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2})} \right] e^{j\rho k_z^2 dk_z} \] (B.8)

\[ H_z(\rho, \phi, z) = \frac{1}{j \omega \mu_0} \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} (k_z^2 - k_0^2) g_n(k_z) H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2}) e^{j\rho k_z^2 dk_z} \] (B.9)

\[ H_\phi(\rho, \phi, z) = \frac{1}{2\pi} \sum_{n=-\infty}^{\infty} e^{jnkz} \int_{-\infty}^{\infty} \left[ -\frac{n k_z}{j \omega \mu_0 \rho} g_n(k_z) H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2}) + \frac{f_n(k_z) k_0^2 - k_z^2 H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2})}{f_n(k_z) k_0^2 - k_z^2 H_n^{(2)}(\rho \sqrt{k_0^2 - k_z^2})} \right] e^{j\rho k_z^2 dk_z} \] (B.10)
where $H_n^{(2)}$ is the Hankel function of second kind and $H_n^{(2)}(z)$ is the derivative of the Hankel function of second kind with respect to its argument.

By equating (B.3) to (B.7) and (B.4) to (B.8) at $\rho = \rho_2$

$$f_n(k_z) = \frac{j \omega \varepsilon_{0} \bar{E}_z(n, k_z)}{(k_0^2 - k_z^2) H_n^{(2)}(\rho_2 \sqrt{k_0^2 - k_z^2})}$$

(B.11)

$$g_n(k_z) = \frac{1}{\sqrt{k_0^2 - k_z^2} H_n^{(2)}(\rho_2 \sqrt{k_0^2 - k_z^2})} \left[ \bar{E}_\phi(n, k_z) + \frac{n k_z}{\rho_2 (k_0^2 - k_z^2)} \bar{E}_z(n, k_z) \right]$$

(B.12)
APPENDIX C

ASYMPTOTIC FORMS OF HANKEL FUNCTION RATIOS FOR LARGE ORDER AND ARGUMENT

The asymptotic forms for the ratios of the Hankel function of second kind and its derivative, \( H_n^{(2)}(z) / H_n^{(2)}(z) \) is derived using the formulations given in [56]. The Hankel function of the second kind can be written in terms of Bessel functions of first and second kinds as [56]

\[
H_n^{(2)}(z) = J_n(z) - jY_n(z) \tag{C. 1}
\]

The recurrence formulas are given by [56] as

\[
H_n^{(2)}(z) = H_{n-1}^{(2)}(z) - \frac{n}{z} H_n^{(2)}(z) \tag{C. 2}
\]

\[
H_n^{(2)}(z) = -H_{n+1}^{(2)}(z) + \frac{n}{z} H_n^{(2)}(z) \tag{C. 3}
\]

The asymptotic forms of the Bessel functions of first and second kinds are given as, [56]

\[
J_n(z) \approx \frac{1}{\sqrt{2\pi n}} \left( \frac{ez}{2n} \right)^n, \quad n \to \infty \tag{C. 4}
\]
\[ Y_n(z) \approx -\frac{2}{\pi n} \left( \frac{ez}{2n} \right)^{-n}, \quad n \to \infty \] (C. 5)

Using (C. 1), (C. 2), (C. 4), and (C. 5), the asymptotic form of \( H_{n}^{(2)}(z)/H_{n}^{(2)}(z) \) can be found as

\[
\lim_{n \to \infty} \frac{H_{n}^{(2)}(z)}{H_{n}^{(2)}(z)} = -\frac{n}{z} \] (C. 6)

The asymptotic form for large negative orders can be found by the help of the formula given by [56] as

\[ H_{-n}^{(2)}(z) = (-1)^n H_{n}^{(2)}(z) \] (C. 7)

Using (C. 1), (C. 3), (C. 4), (C. 5), and (C. 7), the asymptotic form of \( H_{n}^{(2)}(z)/H_{n}^{(2)}(z) \) can be found for large negative orders as

\[
\lim_{n \to \infty} \frac{H_{n}^{(2)}(z)}{H_{n}^{(2)}(z)} = \frac{n}{z} \] (C. 8)

The asymptotic form expressions of the Hankel function of second kind for large argument can be written for fixed integer order \( n \) and as, [56]

\[ H_{n}^{(2)}(z) \approx e^{-\frac{j \pi n}{4}} \left( \frac{2}{z} \right)^{n} e^{-\frac{j z \pi}{2}}, \quad -2\pi < \arg(z) < \pi, \quad |z| \to \infty \] (C. 9)

Inserting (C. 9) into (C. 2) or (C. 3)

\[
\lim_{|z| \to \infty} \frac{H_{n}^{(2)}(z)}{H_{n}^{(2)}(z)} = -j \] (C. 10)
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INTERNATIONAL JOURNAL PUBLICATIONS

INTERNATIONAL CONFERENCE PUBLICATIONS
1. O. Bayraktar and O. A. Civi, “Cylindrical substrate integrated waveguide (CSIW) slot arrays with uniform and nonuniform spacings,” in Proc. 31st General Assembly and Scientific Symposium (URSI GASS), Beijing, China, Aug. 16-23, 2014.


