THESIS FOR THE DEGREE OF LICENTIATE OF ENGINEERING

Baluns, Hybrids and Power Dividers for Ultra-Wideband Antennas

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To Sana & Kumail

Baluns, Hybrids and Power Dividers for Ultra-Wideband Antennas

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Abstract

UWB (ultra wideband), depending on the application requirement, is usually referred to 2:1 bandwidth for radio/wireless applications. Here in this thesis it is also used to denote extremely wideband antenna and components at several octave bandwidth, particularly for high frequency applications, such as the 1 – 10 GHz antenna system needed for the Square Kilometer Array (SKA) radio telescope and the 2 – 14 GHz antenna system for the VLBI2010 (Very Long Base line Interferometer) project. One of the critical parts of an UWB antenna is its feed circuits, because this network must deliver the signal from the UWB antenna to LNAs without introducing losses that severely degrade performance in low noise systems like SKA and VLBI2010. The first part of this thesis is primarily associated with designing UWB baluns for the decade bandwidth Eleven feed. The second part of the thesis addresses UWB components using the novel gap waveguide technology. The gap waveguide is an UWB technology, because it inherently is more broadband than normal hollow waveguides and can be used over more than octave bandwidth. Also, the gap waveguide is less dispersive than normal hollow waveguide.

The decade bandwidth Eleven antenna can be considered to have either four differential ports or eight single-ended ports, and different ways of combining the eight ports are needed for different purposes and applications. The most feasible combination is when the antenna is fed by four baluns; the balun being a device that transforms a balanced two-wire line to an unbalanced coaxial or microstrip line. The first part of the thesis presents a new passive balun solution for the Eleven antenna and a way to integrate four of these baluns together with the antenna in such a way that the four differential ports transform to four single-ended ports. It is also important to verify and evaluate the radiation efficiency of a multiport antenna before being integrated in the system. This thesis addresses how to measure the radiation efficiency of a multiport antenna excluding the losses in the feeding network used for the measurement, particularly when the impedance match between the antenna and the feeding network is not so perfect.

The gap waveguide makes it possible to realize low loss circuits at millimeter and sub-millimeter waves, without having the problems of metal contact between joining metal pieces present when using normal hollow waveguides. The present work includes a study of the resemblance between the groove gap waveguide and the standard hollow rectangular waveguide, and between the ridge gap waveguide and the normal hollow ridge waveguide. The dispersion diagrams and characteristic impedances have been compared. These results are very useful when designing, simulating and measuring gap waveguide components of different kind, because they show under which conditions and accuracies standard waveguide interfaces can be used.

Finally, the thesis presents an UWB microstrip power divider that was designed and packaged using gap waveguide technology, and a ridge gap waveguide ring hybrid. The power divider has a bandwidth of 2–14 GHz, and it is intended to be used in the feed network for the Eleven feed.

Keywords: Ultra-wideband, Eleven feed, baluns, Multiport feeding network, gap waveguide, power divider, packaging, ring hybrid.

Contents

List of papers

A. H. Raza, J. Yang, "UWB Planar Balun Solutions for the Eleven Antenna", Submitted to IET Microwave, Antennas & Propagation.

B. H. Raza, J. Yang, A. Hussain, "Measurement of Radiation Efficiency of Multiport Antennas with Feeding Network Corrections", IEEE Antennas and Wireless Propagation Letters, Vol. 11, pp. 89 - 92, 2012.

C. H. Raza, J. Yang and M. Pantaleev, "A compact UWB passive balun solution for cryogenic 2 – 13 GHz eleven feed for future wideband radio telescopes", 5th Eur. Conf. On Antennas and Propagation (EuCAP 2011), Rome, Italy, 11 – 15 April 2011.

D. H. Raza, J. Yang, P.-S. Kildal, E. Alfonso, "Resemblance Between Hollow Waveguides and Gap Waveguides", Submitted to IET Microwave, Antennas & Propagation.

E. H. Raza, J. Yang, "A Low Loss Rat Race Balun in Gap Waveguide Technology", 5th Eur. Conf. On Antennas and Propagation (EuCAP 2011), Rome, Italy, 11 – 15 April 2011.

F. H. Raza, J. Yang, "Compact UWB Power Divider Packaged by Using Gap-Waveguide Technology", 6th Eur. Conf. On Antennas and Propagation (EuCAP 2012), Prague, Czech Republic, 26 – 30 March 2012.

Other related publications by the Author not included in this thesis:

- J. Yang, H. Raza, "Empirical Formulas for Designing Gap-Waveguide Hybrid Ring Coupler", Submitted to IEEE Microwave and Wireless Components Letter.

- J. Yang, M. Pantaleev, T. Ekebrand, P.-S. Kildal, H. Raza, J. Yin, J. Jonsson, L. Helldner, A. Emrich, B. Klein, "Development of the Cryogenic 2-14 GHz Eleven Feed System for VLBI2010", 6th Eur. Conf. On Antennas and Propagation (EuCAP 2012), Prague, Czech Republic, 26 – 30 March 2012.

- S. Rahiminejad, A. U. Zaman, E. Pucci, H. Raza, V. Vassilev, S. Haasl, P. Lundgren, P.-S Kildal, P. Enokssona, "Design of Micromachined Ridge Gap Waveguides for Millimeter-Wave Applications", Proc. Eurosensors XXV, September 4-7, 2011, Athens, Greece.

Preface

This report is a thesis for the degree of licentiate of engineering at Chalmers University of Technology. The work is divided into two main parts: UWB balun solutions for the eleven feed and Wideband gap waveguide components. This work has been supported in parts by the Swedish Foundation for Strategic Research (SSF) within the Strategic Research Center CHARMANT, Pakistan's NESCOM scholarship program and partially by Swedish Research Council VR. My principle supervisor is Associate Prof. Jian Yang and Prof. Per-Simon Kildal is my examiner. The work was carried out between March 2010 and August 2012 at antenna group of Chalmers.

Acknowledgement

First and foremost, I would like to thank my supervisor, Associate Prof. Jian Yang for his support and confidence in my work. His patience and encouragement were very useful to me throughout the course of this research. I would also like to thank Prof. Per-Simon Kildal for accepting me as a PhD student in his group and for giving me the opportunity to work on creative projects like Eleven feed and gap waveguide. He is a true inspiration for me.

I thank, from the bottom of my heart, to my office in Pakistan who have supported me throughout these two and a half years at Chalmers with their true commitment.

I would like to acknowledge Dr. Miroslav Pantaleev for making baluns for Eleven feed and Ahmed, for helping me in doing measurements of the Eleven feed in reverberation chamber. Many thanks and gratitude to all the former and present members of the Antenna group, Prof. Jan Carlsson, Associate Prof. Mariana, Dr. Rob Maaskant, Dr. Chen, Ashraf, Oleg, Aidin, Elena, Astrid, Dr. Esperanza and Carlo for making an excellent research environment.

My wife Sana's love and support, my son Kumail's patience and my little daughter Abeeha's smile contributed to my work more than one could ever imagine. Their smiles, love and support encourage me to achieve career goal. Finally, with all my heart, I would like to thank my parents for always supporting me.

Syed Hasan Raza Zaidi
Göteborg, October 2012

Chapter 1

INTRODUCTION

This thesis deals with two research topics: an ultra-wideband (UWB) balun feed network for the Eleven antenna, and gap-waveguide components.

1.1 UWB balun feeding to the Eleven antenna

Ultra-wideband (UWB) radio/wireless systems require that their antenna(s) can receive or/and transmit signals over a wide frequency range [1]. These types of antennas are called as UWB antennas. Generally, a UWB antenna can have a broad frequency range, e.g. $(f_h/f_l) > 1.22$ in order to be 'ultra wide' but actually it depends on the system requirements and the center frequency of the bandwidth [2]. UWB antennas are one of the key devices in many UWB systems, such as broadband wireless communications, Electronic Warfare, medical imaging and radar systems, etc. UWB antennas offer many desirable features, and at the same time often with a simple RF front-end.

There are mainly four types of UWB antennas: the scaled structure (such as Bow-Tie dipoles, biconical dipoles and log-periodic dipole arrays), the self-complementary structure (self-complementary spiral antennas), the traveling wave structure (Vivaldi antennas) and the multiple reflection (or resonance) structure (dielectric resonator antenna) [3], [4].

High gain, high radiation efficiency and good impedance matching simultaneously across a wideband are real challenges for UWB antenna design. The feeding network design is critical issue to meet these challenges, due to the often required compact size, manufacturability with low cost, mechanical stability, low ohmic loss, and good matching performance.

The first topic of this thesis discusses the feeding mechanism for a UWB antenna, the Eleven feed [5]-[12], see Figure 1.1, for high frequency applications, such as 1 – 10 GHz version for the Square Kilometer Array (SKA) project and the 2 – 14 GHz version for the Very Long Base line Interferometer (VLBI2010) projects. The Eleven feed is a decade bandwidth log periodic dipole array antenna which has been developed at Chalmers University of Technology since 2005. It has been shown that the Eleven feed has good features, especially for the applications in radio telescopes. Compared to several other candidate feeds [13]–[15] for the SKA and the VLBI2010 projects, the

Eleven feed has many advantages over them in terms of the size, low profile and simple geometry , constant phase center, constant directivity, low cross-polar side lobe levels and moreover constant beamwidth over decade bandwidth. In addition, the multi-port Eleven antenna has been studied for its versatility, such as for use in monopulse tracking systems [16] and UWB MIMO systems[17], [18]. However, the Eleven feed has more ports (total eight ports) than other candidates, which requires more low-noise amplifiers (LNAs) in the whole antenna system. Therefore, different feeding networks for the antenna have been suggested in along with different combinations, where the passive balun solution is an attractive one [19]-[21].

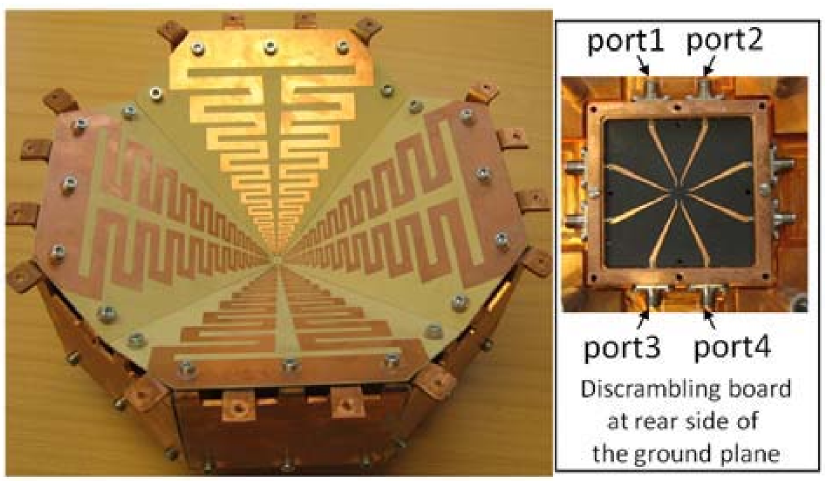


Figure 1.1: Photo of the manufactured Eleven feed, and the descrambling board at the rear side of the ground plane [20].

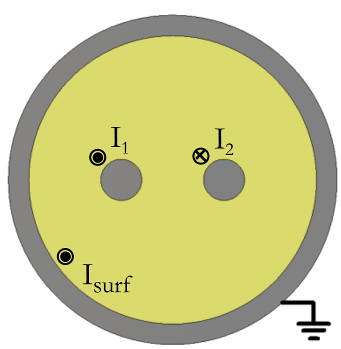


Figure 1.2: Shielded two wire balanced transmission line

A balun, one of the most common feeding devices for antennas, is a transition from a balanced transmission line to an unbalanced transmission line[22]. In balanced line, the two lines have equal potential but 180° phase difference and no current flows through the grounded shield if there is any, i.e. $I_{surf} = 0$, in ideal case, see Figure 1.2. If surface current is not equal to zero, this indicates that common modes are excited, which always needs to be avoided in balun design.

1.2 Gap Waveguide Components

The second topic of this thesis deals with the developments of wideband gap waveguide components.

The gap waveguide is a new planar microwave circuit technology which was originally proposed by P.-S. Kildal [23], see Figure 1.3. This waveguide aims to make low loss circuits at mm and sub-mm wave, particularly above 60 GHz, because it can be made by only metals. By using conventional rectangular waveguide, the tolerance requirements are difficult to meet and the metal contact for the waveguide walls imposes a difficulty for manufacturing. By using microstrip type transmission line, the loss in the dielectrics is often unacceptable. Moreover, the gap waveguide technology can be used for packaging of the microstrip circuits for suppressing the unwanted radiations without having any cavity resonances over a wideband.

One of the basic problems while simulating gap waveguide structure in commercial electromagnetic solvers is how to model the ports for gap waveguide in order to have accurate and fast simulations. A numerical study on the modeling of the gap waveguide port is presented in the thesis, with a method to estimate the characteristic impedance of the gap waveguide from the reflection coefficient.

The development of two compact gap-waveguide devices is also presented in the thesis.

1.3 Outline of the Thesis

A brief overview of various balun techniques for UWB antennas in literature and a basic summary of the new gap-waveguide technology are presented in Chapter 2, offering an understanding of the context of my work in developments of UWB balun and gap-waveguide components. Chapter 3 discusses a new planar UWB balun solution to the Eleven antenna. Numerical study on ports for simulating the gap waveguide circuits is described in Chapter 4. Chapter 5 presents two microwave devices by using gap waveguide. The conclusion ends the first part of the thesis. The second part consists of six papers of mine, which is the basis of this licentiate thesis.

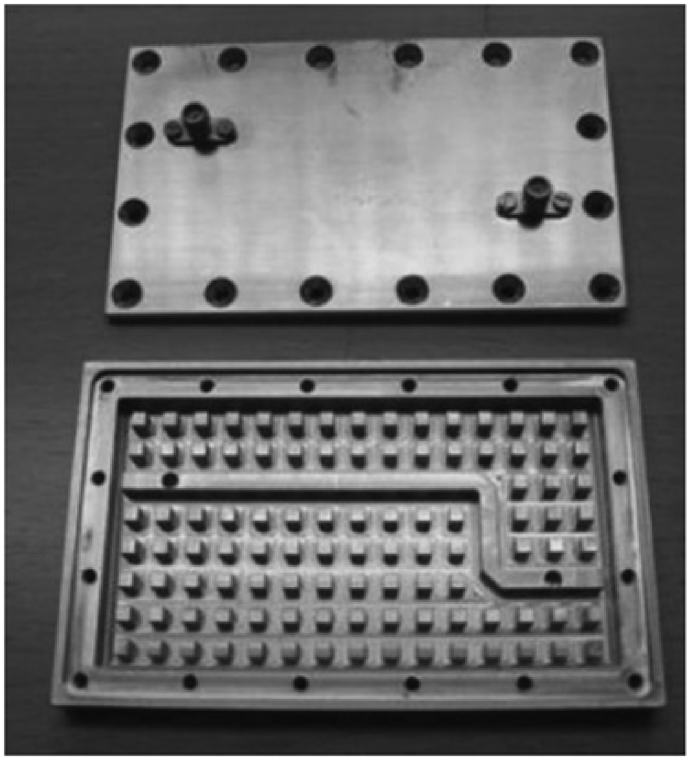


Figure 1.3: Photo of the 12–18 GHz demonstrator designed and measured for verifying the ridge gap waveguide technology [24].

C h a p t e r 2

OVERVIEW OF BALUN TECHNOLOGY AND THE GAP WAVEGUIDE TECHNOLOGY

Antennas, if not fed properly, do not work as predicted by the design. Many antennas require balanced-line feeding, while power/signal transmission is mostly carried out by unbalanced line. Therefore, a balanced-to-unbalanced transformer, a balun, is a key issue in antenna design.

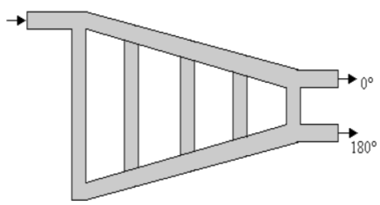
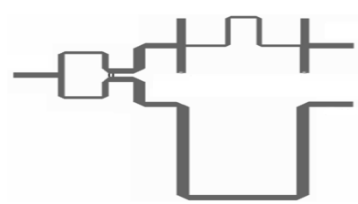
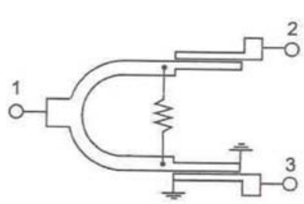
To design a UWB balun is a challenge. A lot of wideband balun designs have been published [25]-[32] and some of them are summarized below and in Table 2.1.

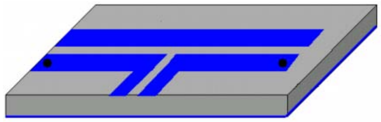
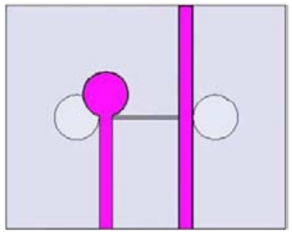
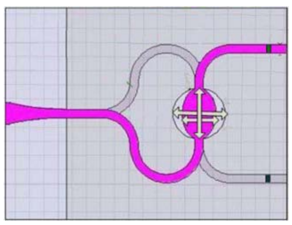
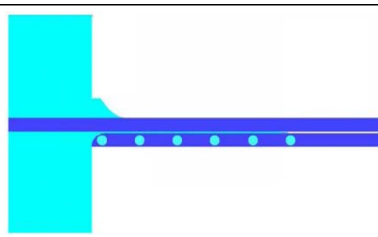
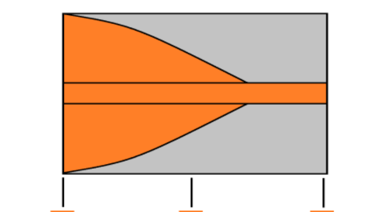
Log-periodic balun is an octave bandwidth balun. The design and structure is based on the log-periodic antenna theory [25]. The bandwidth is depending on the number of sections of resonators. This structure is big in size so it will be difficult to accommodate four of such type of balun on the back of the Eleven feed. There is another wide-band microstrip balun implemented on a single-layer printed circuit board consisting of a wide-band Wilkinson power divider and a non coupled-line broad-band 180^{o} phase shifter [26]. The 10-dB return loss of this balun has been achieved across the bandwidth from 1.7 GHz to 3.3 GHz, or 64%. Due to this only octave bandwidth limitation, this balun is not suitable for decade bandwidth Eleven feed. Broadband phase shifting following the Wilkinson power divider in the above design can also be achieve if use 3 dB 90^{o} hybrid coupler [27] in which the coupled and transmitted port being short circuited or open circuited. It was shown that this balun performed well within 5-11 GHz band. Again the bandwidth, achieved by this balun, is not satisfactory according to the Eleven feed requirement. The only attractive balun on single layer PCB fulfilling requirements of Eleven feed is the marchand balun [28]. This balun has 11:1 bandwidth with less than 3^{o} maximum phase imbalance. However, the results show that it has high losses as well as the S_{11} is not less than -10 dB.

UWB out-of-phase power divider, utilizing both sides of PCB [29], can also be used to feed antenna. This design is based in the transformation from microstrip to slot line and then again slot line to the two arms of microstrip line making out-of-phase power divider. It is shown that the -10 dB reflection coefficient is between 3–11 GHz with about 1 dB transmission loss. Similarly, there is another three layer PCB design for out-of-phase power division based on parallel strip line to microstrip line transition [30] making the output ports on opposite side of middle layer. The output ports are not in the same plane. The performance of this is similar to that of the previous one. The decade bandwidth balun based on microstrip-to-coplanar stripline transition presented in [31] has a good matching performance but with a high transmission loss

The microstrip-to-parallel stripline transition presented in [32] is a simple design and easy to manufacture. Our work is to investigate different tapering techniques to obtain a better performance with a more compact size.

Table 2.1 Wideband Baluns in Literature

		Bandwidth, Phase Imbalance	Size [for 1 GHz]	Comments
Log – Periodic balun [25]		~ 67 % or 2:1 From 1.8–3.8 GHz ± 10°	57 mm x 100 mm	Large structure
Broadband power divider followed by delay line phase shifter [26]		64 % or 2:1 From 1.7–3.3 GHz ± 5°	70 mm x 70 mm	Large structure
Broadband power divider followed by coupled line phase shifter [27]		75% From 5–11 GHz ± 7°	56 mm x 30 mm	Very complex structure.

Marchand balun [28]		~ 167% or 11:1 From 2–22 GHz ± 3°	Size not specified	Multi layers High losses & S_{11} higher than -10 dB
UWB out of phase power divider [29]		115% or ~ 4:1 From 3–11 GHz ±0.25°	40 mm x 20 mm	Large area requires to bend the output ports.
UWB out of phase power divider [30]		115% or ~ 4:1 From 3 – 11 GHz ±0.5°	20 mm x 20 mm	Large area require to bend the output ports.
UWB Microstrip to CPS transition [31]		10:1 From 4 – 40 GHz	63.5 mm length	Large structure and difficult to package it inside.
Microstrip to paralle stripline transition [32]		10:1 From 1.5 – 15 GHz	60 mm length	Large structure and difficult to package it inside.

2.1 Introduction to Gap Waveguide

Gap waveguide technology is a new technique to make low loss circuits at millimeter and sub-millimeter waves [23]. The waveguide can be realized without having any dielectric material inside it, and there is no need to have electrical contact between the top and the bottom metal plates, the latter being provided with pins (or rather posts). These pins create a periodic metal texture that blocks parallel-plate waves to propagate everywhere except along a metal ridge or groove between the pins. The bandwidth of this particular behavior is referred to as a parallel-plate stopband. This stopband makes it also possible to use the gap waveguide technology for packaging of both passive and active microwave circuits realized in other technologies such as microstrip and co-planar waveguides[33][34].

The basic theory of this new type of waveguide is that one of the parallel plates is textured by periodic structures, preferably in the form of rectangular shapes. These periodic structures produce artificially high surface impedance, ideally Perfect Magnetic Conductors (PMC), and together with a smooth metal plate on the top of them they generate a stopband for parallel-plate modes. If there is a metallic ridge or groove within this periodic texture and the distance between the top metal plate and the periodic structure is less than the quarter wavelength, only fundamental mode (quasi TEM mode in ridge gap waveguide and TE_{10} mode in case of groove gap waveguide) will propagate along the ridge or the groove [35], as illustrated in Figure 2.1. For this reason this type of waveguide is called as gap waveguide. The important parameters for defining the stopband are the period of pin structure, the length of air gap between the top of pins and upper smooth plate, and the height and the width of pins [36]. The height of pins is usually set as the quarter wavelength at the center frequency of the stopband. Note that the air gap thickness must be smaller than quarter wavelength [36] and width of the pin plus the distance between two pins must be less than half the wavelength.

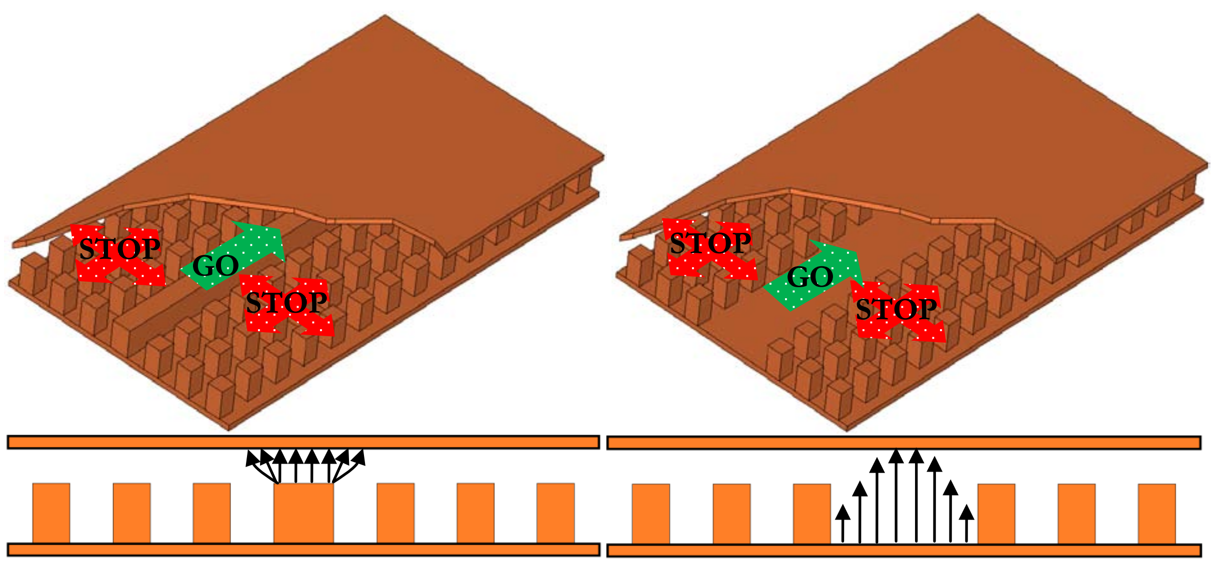


Figure 2.1: Front and top view of basic structures of ridge gap waveguide (left) and groove gap waveguide (right) depicting the propagation region.

With the presence of air gap, i.e. without having any dielectric material and only metal involve everywhere, both the ridge and the groove gap waveguides have lower loss than microstrip lines. The advantage of the gap waveguide compared to normal hollow waveguides is that the latter is difficult to manufacture at high frequency. The normal procedure is that the hollow waveguides and waveguide components are manufactured from two metal pieces and they must be joined with a good metal contact everywhere between them, which is difficult, whereas this is not needed for the gap waveguides.

C h a p t e r 3

UWB BALUN FEEDING SOLUTION TO THE ELEVEN FEED

Orientation of the balun on the back of Eleven feed is shown in Figure 3.1, where baluns are just represented by a symbolic box at this stage. With this configuration, the box of the balun can be replaced by any type of the first four baluns described in Table 2.1. But most of them are of octave bandwidth, whereas requirement of Eleven feed is decade bandwidth balun. So these baluns can't be used as for feeding of the Eleven antenna. However there are some designs available based on Marchand balun [28] having bandwidth of about 11:1 but it has insertion loss of about 2 dB and S_{11} is higher than -10 dB over most of bandwidth. Similarly other designs in the Table 2.1 are available, based on transition from microstrip to other types of transmission line. This involves utilization of both sides of PCBs. So they can't be arrange on the back of Eleven feed, in the orientation shown in Figure 3.1, because there will be ground wall on one side.

Microstrip to stripline transition as shown in Figure 3.2, usually achieves more than decade bandwidth[32]. They are easy to manufacture and several tapering techniques are available to match the low impedance unbalance line with a high impedance balanced lines. Due to low losses and broad bandwidth nature, this balun can be a suitable candidate for the Eleven feed. Orientation of this type of balun is shown in Figure 3.3 where it is connected to Eleven feed petal vertical to the ground plane. This type of arrangement requires solid metallic frames to hold them stably. Those frames may cause resonances at some frequency points which can't be avoided. Similarly, when the balun will pass through the hole of ground plane to connect to the feed petal, this may also generate the common modes.

An ideal balun generating a fully differential mode at the output has an electric wall in the symmetry plane of the balanced port. Dividing the output load in that symmetry plane, we can make a three-port device with the loads of impedance Z_1 at the output ports. The performance of the balun can be described by a three port scattering matrix, as shown by the Figure 3.4, where Port 1 is the unbalance port, and Ports 2 and 3 are the two output balanced ports of the balun. The important parameters for describing the performance of balun are (1) the reflection coefficient, (2) ohmic losses and (3) Common – mode rejection ratio (CMRR) $(S_{31} - S_{21})/(S_{31} + S_{21})$. The common–mode is rejected if S_{31} is equal to S_{21} in magnitude and 180° out of phase with respect to S_{21}. An ideal balun has a fully differential mode at the output, i.e., the common-mode is totally rejected.

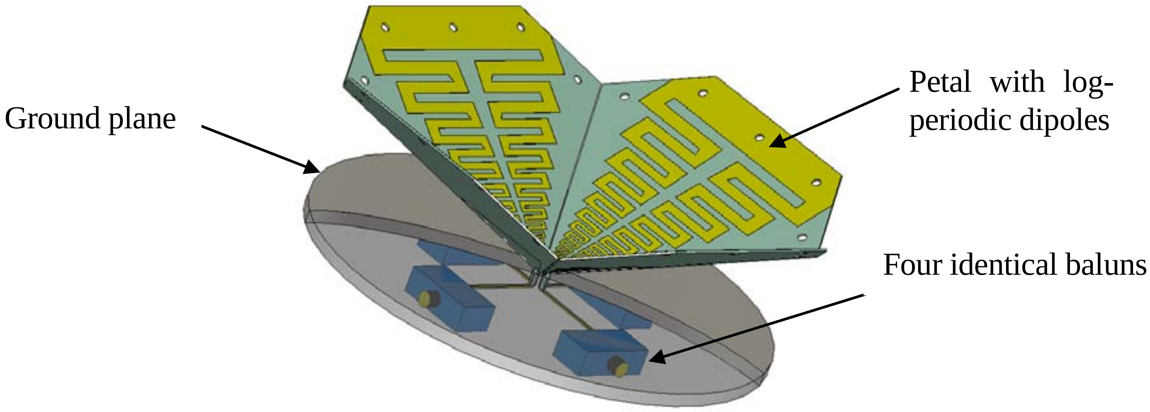


Figure 3.1: Eleven antenna with the baluns

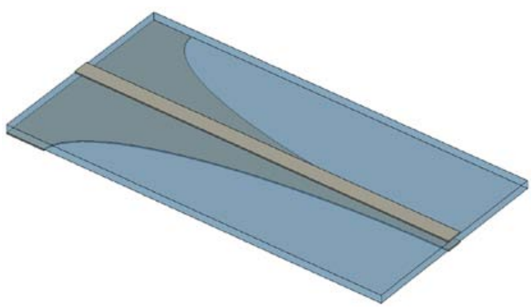


Figure 3.2: Simple microstrip to stripline balun.

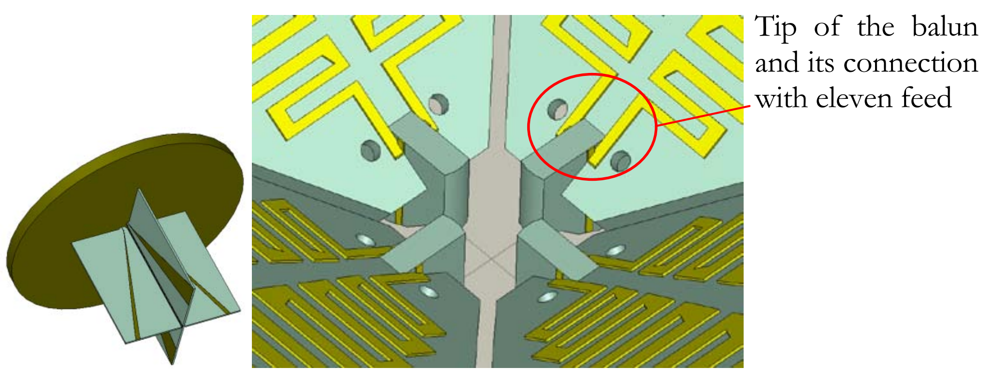


Figure 3.3: Arrangement of four baluns behind the ground plane for two polarization and connection with the Eleven feed.

In literature, there is another terminology describing the balance quality is 'balun ratio' [37] or BR. It is define as the ratio of the current exciting desired balance mode to the current exciting undesired unbalance mode. In term of S-parameters, the balun ratio is $(S_{31} + S_{21})/(S_{31} - S_{21})$.

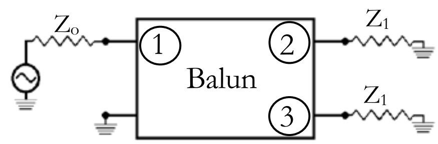


Figure 3.4: Block diagram of a typical balun.

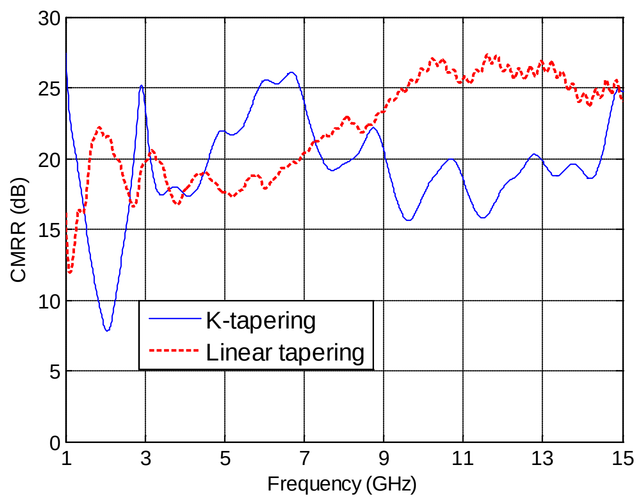


Figure 3.5: Simulated CMRR of the two baluns.

3.1 Design and Modeling

Since a balun can be consider as a matching network for transforming the high impedance balanced load to an unbalanced load. For Eleven feed the differential impedance is very high which require a broadband matching technique. Two versions of the Eleven feed have been made with two different balun designs. The first version of balun was designed using optimization technique available in CST Microwave Studio, which we called 'Linear Tapering'. This design doesn't provide the satisfactory reflection coefficient. In the second version, the tapering of the balun has been done using Klopfenstein technique, which is the optimum technique for broad band matching. Figure 3.5 shows the simulated CMRR, calculated using CST Microwave Studio, of the two versions which is about 17 dB over the desire bandwidth. Balun with linear tapering seems have lower CMRR at about 2 GHz. One method for confirming the performance of the balun is to connect it in the back–to–back configuration at their balanced ports, see Figure 3.6. Simulated and measured reflection and transmission coefficients are shown in Figure 3.7. The reflection coefficient is below -10 dB for 2.5 –

16 GHz, which means that a single balun should have below – 12 dB performance. The measured transmission coefficient agrees well with the simulated one, except for some dips at a few frequency points. This may be caused by some resonances within the metal support frame.

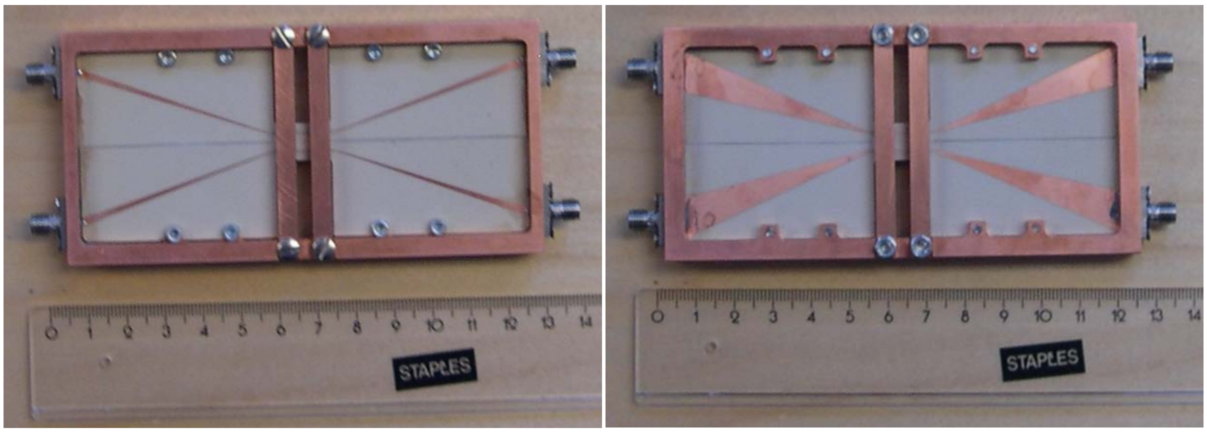


Figure 3.6: Tapered microstrip balun with back-to-back configuration.

Detailed design of linear tapering balun is shown in Figure 3.8. Each twin balun board is 50 mm wide and 50.87mm long. Optimization has been done on the criteria of length and tapering of the balun. Substrate material used in the balun is the same as used for the petal of Eleven feed, i.e. 0.762 mm thick Rogers TMM3. The baluns are then shielded with a metallic box. The metallic box is also partitioned diagonally, so as to reduce the coupling between the baluns. The cavity resonances introduced by the shielded metallic box can be suppress by using 1 mm thick Eccosorb MCS absorbing sheet, which is placed on the inner side of the shielded metal box. This balun transform the 200 ohm differential impedance of the Eleven feed to single ended 50 ohm. This can easily be achieved by gradually decreasing the width of both upper and lower conductors of the balun.

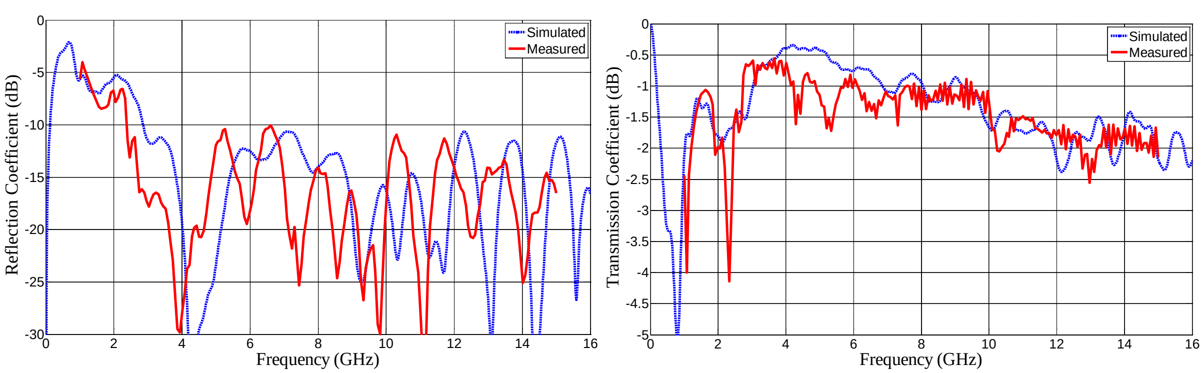


Figure 3.7: Simulated and Measured S-parameters of back-to-back balun.

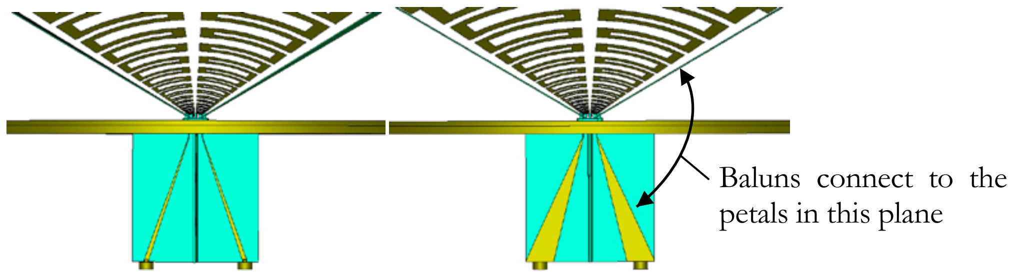


Figure 3.8: Geometry of the version 1 twin tapered microstrip line balun shows the front (left) and rear (right) sides of the circuit board.

Version 2 of the balun is based on Klopfenstein tapering technique, which we called K-Tapering, and is an optimum method in the sense that the reflection coefficient is minimum within the bandwidth. Other changes are that the differential impedance of the Eleven feed is set to have value of 250 ohm. For this high impedance matching, the substrate material has been changed to 0.787 mm thick Rogers 5880. Similarly the size of the board is now increased to 110 x 71.25 [mm], which is twice wider than the linear tapering balun, see Figure 3.9.

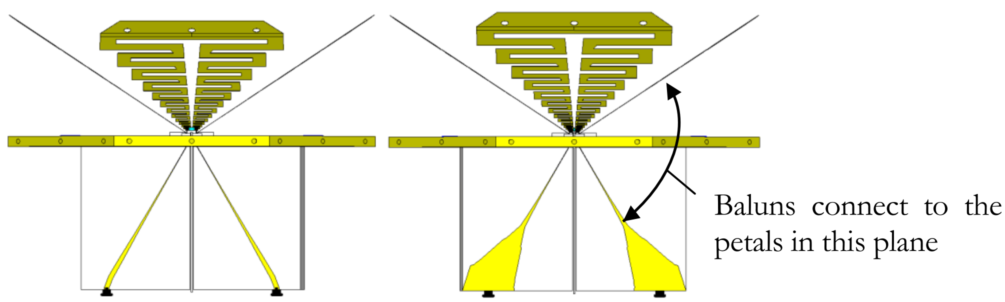


Figure 3.9: Geometry of the version 2 twin tapered microstrip line balun shows the front (left) and rear (right) sides of the circuit board.

3.2 Measurement Setup and Results

The two versions of the balun have been manufactured and the performance of the Eleven feed have also been measured both in anechoic chamber and in the reverberation chamber [38]. Figure 3.10 shows the two prototypes of the Eleven feed together with a compact low-loss decade-bandwidth power combiner using Gap-wavguide technology [39]. The power combiner has a size of 31.58×15.79×7.9 mm^3 and a reflection coefficient below -12 dB over 2-14 GHz.. The dimensions of the Eleven feed remain same in both cases. The balun with linear tapering is shown in left side of Figure 3.10, whereas balun with Klopfenstein tapering technique is shown middle. . The actual performance of the balun is measured by connecting it with the Eleven feed and operate it in a metallic box. With the enclosure in a metallic box, the characteristic impedance of the balun lines changes as well as the common mode start to propagate along the line. This will affect the BOR_1 efficiency [40]-[43]. Numerical studies show that the CMRR of both the baluns are around 17 dB in the desired bandwidth. This also includes the affect of ground plane of the Eleven feed as the balance lines are passing through the holes in the ground plane and this cannot be avoided. Similarly the cavity resonance, which might be excited due to the metallic box, can be suppressed by absorbing sheet.

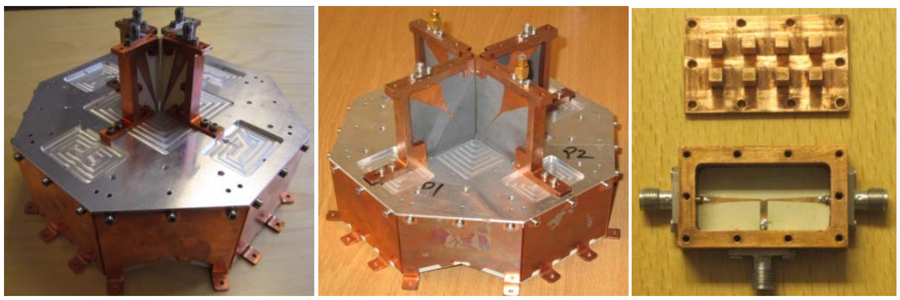


Figure 3.10: Two versions of the Eleven feed with balun, linear tapering balun (Left) and K-tapering balun (middle). The power combiner (right) is used to combine the two output ports of each polarization. All measurements with the Eleven feed include the power combiner.

Figure 3.11 shows the reflection coefficient of the Eleven feed with balun (four linear tapering baluns and two wideband power combiners). The result shows that S_{11} is below -8 dB except few frequency points. This is measured by properly terminating the ports of the baluns of other polarization. It should be noted that the mechanical tolerance has an effect on the reflection coefficient performance of the feed, especially the contact point where the balance twin line connect to the antenna terminal, see Figure 3.3 again.

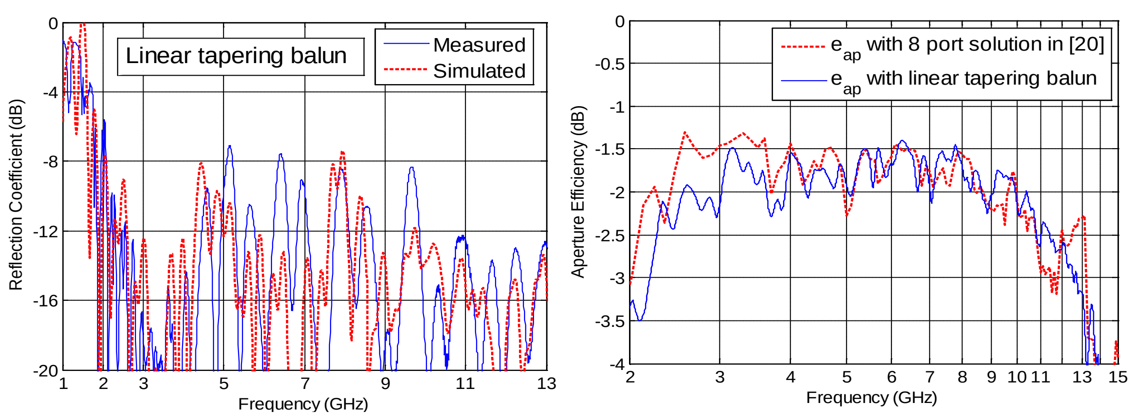


Figure 3.11: Measured reflection coefficient (left) and Efficiencies of the feed (right) with balun feeding.

The radiation patterns of the Eleven feed at various frequencies have been measured and they are satisfying the property of constant radiation characteristics of Eleven antenna as stated in [20], over the whole 2 - 14 GHz bandwidth. The total aperture efficiency and its sub – efficiencies as defined in [40] and [41] are calculated based on radiation patterns measurement. In order to assess the performance of the baluns with the Eleven feed, we compare the aperture efficiency e_{ap} of the Eleven feed with the baluns to that with 8-port solution in [20]. Note that the feed illuminates a symmetrical paraboloid with a subtended angle of 2×60^o and the center blockage loss is neglected in the e_{ap} calculation. From Figure 3.11, we see that the aperture efficiency of the feed with the linear tapering balun feeding fallows that with the 8-port feeding, with 0.5 dB lower from 1 to 4 GHz.

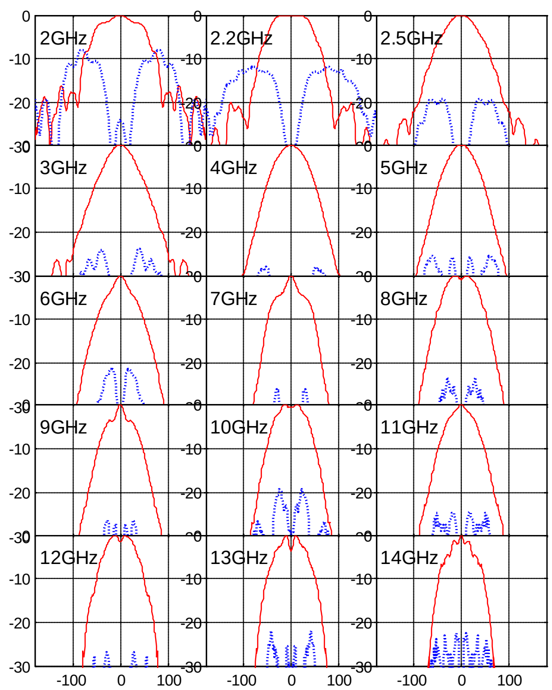


Figure 3.12: Measured co – and cross – polar BOR_1 radiation patterns in φ = 45 plane.

Figure 3.12 shows the measured co and cross – polar radiation patterns of the BOR_1 components in the φ = 45° plane. BOR_1 pattern are measured by the method describe in [43].

The second prototype of the K-tapering balun feeding has been made in order to improve the reflection coefficient. Broadband matching technique were used to improve the matching between the balance port and antenna terminal. Figure 3.13 shows the measured reflection coefficient of the Eleven feed with balun of K-tapering. The result shows that S_{11} is now improved and below -10 dB except for few frequency points where it goes up to -8 dB at around 5 GHz and between 8 – 9.5 GHz, and these are small peaks. Similarly the aperture efficiency, along with BOR_1 efficiency is shown in Figure 3.13. By definition, BOR_1 efficiency is a measure of the power lost in higher order azimuth variations of the far field of the antenna. These variations represent losses, and they can never contribute to the directivity, in fact they contribute to the side lobes [43]. The aperture efficiency follows that with the 8-port feeding up to 8 GHz (0.5 dB lower from 1 GHz to 4 GHz), then however degraded between 8-9.5 GHz. From 10 to 15 GHz, the aperture efficiency of the K-tapering balun feeding is much better than that of the 8-port solution: 1–1.5dB improvement up to 15 GHz. The reason that the e_{ap} for the K-tapering balun feeding is degraded between 8–9.5 GHz, we believe, is that the CMRR performance of the K-tapering balun does not have very robust immunity to strong reflection at the interface between the balun and the Eleven feed, which we did not consider when we designed the balun separately. As it is shown in Figure 3.13, there is a strong reflection (about -8 dB) between the balun and the feed for the K-tapering feeding. This reflection excites common modes which then cause the BOR_1 efficiency, a measure of how symmetric the radiation pattern is, degrading, as shown in Figure 3.13. This phenomenon leads to low aperture efficiency. As a contrast, the linear tapering balun has better immunity to strong reflection, where though the reflection coefficient is high (a bit above -8 dB), the e_{ap} is not degraded. However, the performance of reflection coefficient and aperture efficiency at high frequency end of the linear tapering balun is inferior to that of the K-tapering balun. A study is now being carried out to solve this problem of the BOR_1 efficiency drop between 8 – 9 GHz.

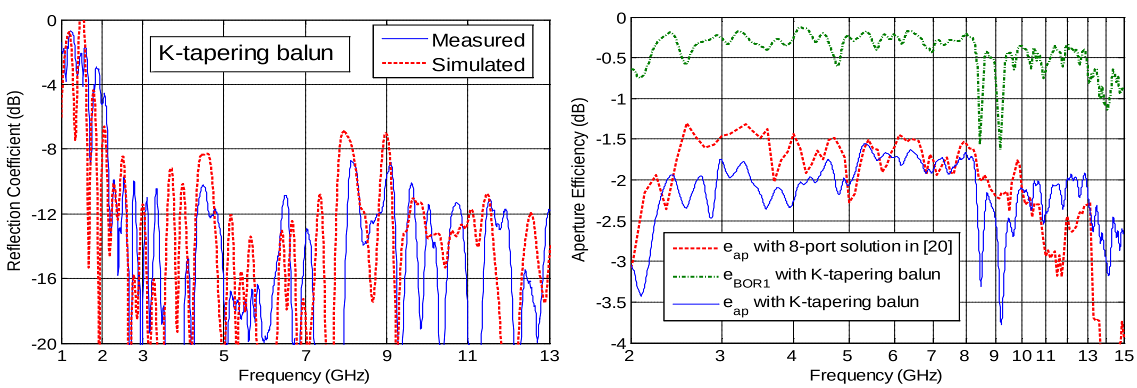


Figure 3.13: Measured reflection coefficient (left) and Efficiencies of the feed with balun (right).

Figure 3.14 shows the measured co and cross – polar radiation patterns of the BOR_1 components in the $\varphi = 45^o$ plane. It can be seen that the radiation pattern is not the same as that in the case of linear tapering at 9 GHz.

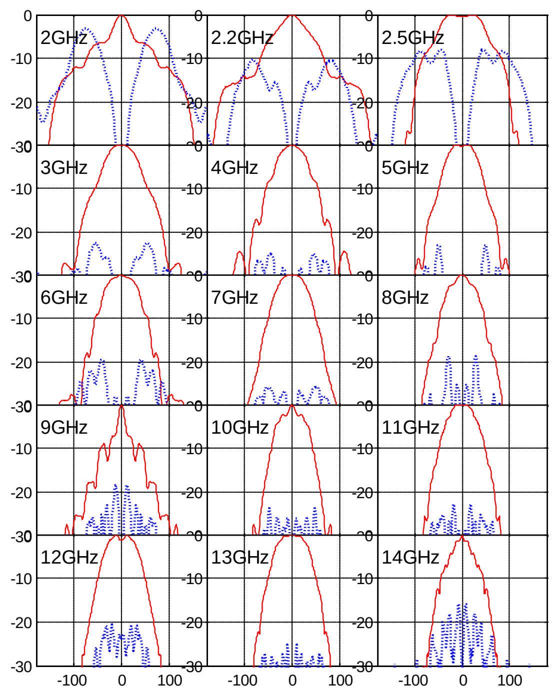


Figure 3.14: Measured co – and cross – polar BOR_1 radiation patterns in φ = 45 plane.

3.3 Measurement of Radiation Efficiency of Multiport Antennas

It is often required, particularly for integrated antenna systems in radio telescopes, that the radiation efficiency of a multiport antenna, excluding the losses in the feeding network, should be measured before being integrated in the system. The radiation efficiency of a multiport antenna e_{rad_ANT} consists of two factors: the ohmic losses in the antenna itself and the so-called decoupling efficiency that accounts for power returned to non-excited ports, as defined in [44] (Note that e_{rad_ANT} here corresponds to ε_{totrad} in [44]). Both factors depend on the excitations provided by feeding networks. However, the test feeding network is often built up of commercially available components in order to reduce the development cost. The losses in the test feeding network could be large, and the impedance match between the antenna and the test feeding network could be far from the perfect case. Thus, it is not a trivial task to obtain the radiation efficiency of a multiport antenna excluding the losses in feeding network. A rigorous feeding network correction approach for obtaining the radiation efficiency of a multiport antenna has been presented. This method uses measurement data of the total radiation efficiency of a multiport antenna with a multiport feeding network, the S-matrices of the feeding network and the antenna, to calibrate out the losses in the feeding network.

Figure 3.15 shows the block diagram of a general multiport antenna with a multiport feeding network. By multiport antenna, we refer it to the multiport antenna without feeding network. The total radiation efficiency of the whole antenna system with the feeding network e_{tot_ANTFN} is defined as

$$e_{tot\_ANTFN} = \frac{P_{rad}}{|a_1|^2} \tag{1}$$

where P_{rad} is the total radiated power from the antenna and $|a_1|^2$ is the input power.

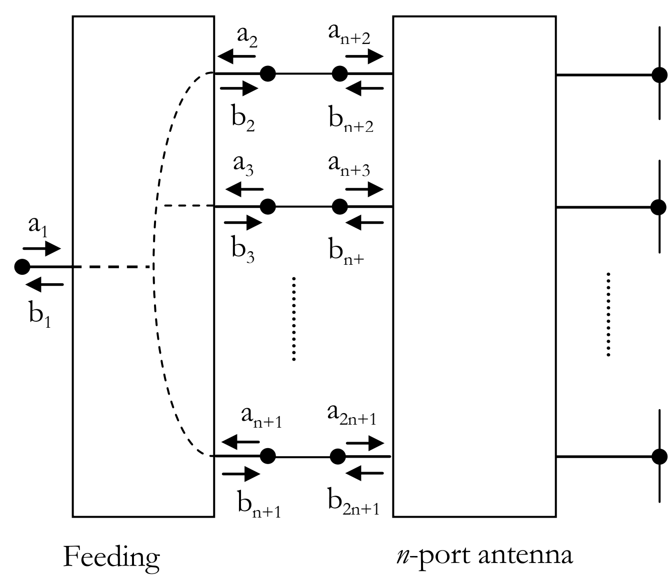


Figure 3.15: Block diagram of the configuration of an n-port antenna with its feeding network.

The radiation efficiency e_{rad_ANT} of the multiport antenna can be expressed as

$$e_{rad\_ANT} = \frac{P_{rad}}{\sum_{k=n+2}^{2n+1}(|a_k|^2 - |b_k|^2)} \tag{2}$$

which measures both the ohmic losses and the decoupling efficiency to non-excited ports in the multiport antenna. The multiport feeding network can be expressed by the following S-matrix

$$\begin{bmatrix} \boldsymbol{b}_I \\ \boldsymbol{b}_{II} \end{bmatrix} = \begin{bmatrix} \boldsymbol{S}_{I,I} & \boldsymbol{S}_{I,II} \\ \boldsymbol{S}_{II,I} & \boldsymbol{S}_{II,II} \end{bmatrix} \begin{bmatrix} \boldsymbol{a}_I \\ \boldsymbol{a}_{II} \end{bmatrix} \tag{3}$$

where

$$\boldsymbol{b}_I = [b_1], \qquad \boldsymbol{a}_I = [a_1], \qquad \boldsymbol{b}_{II} = \begin{bmatrix} b_2 \\ \vdots \\ b_{n+1} \end{bmatrix}, \qquad \boldsymbol{a}_{II} = \begin{bmatrix} a_2 \\ \vdots \\ a_{n+1} \end{bmatrix}$$

$$\boldsymbol{S}_{I,I} = [S_{11}], \qquad \boldsymbol{S}_{I,II} = [S_{12} \quad \cdots \quad S_{1,n+1}]$$

$$\boldsymbol{S}_{II,I} = \begin{bmatrix} S_{21} \\ \vdots \\ S_{n+1,1} \end{bmatrix}, \qquad \boldsymbol{S}_{II,II} = \begin{bmatrix} S_{22} & \cdots & S_{2,n+1} \\ \vdots & \ddots & \vdots \\ S_{n+1,2} & \cdots & S_{n+1,n+1} \end{bmatrix}$$

where S_{ij} is the S parameter of the feeding network from port j to port i. Similarly, the multiport antenna can also be expressed by an S-matrix as

$$\boldsymbol{b}_{III} = \boldsymbol{S}_{III}\boldsymbol{a}_{III} \tag{4}$$

where

$$\boldsymbol{b}_{III} = \begin{bmatrix} b_{n+2} \\ \vdots \\ b_{2n+1} \end{bmatrix}, \qquad \boldsymbol{a}_{III} = \begin{bmatrix} a_{n+2} \\ \vdots \\ a_{2n+1} \end{bmatrix}, \quad \boldsymbol{S}_{III} = \begin{bmatrix} S_{n+2,n+2} & \cdots & S_{n+2,2n+1} \\ \vdots & \ddots & \vdots \\ S_{2n+1,n+2} & \cdots & S_{2n+1,2n+1} \end{bmatrix}$$

From Figure 3.15, we have,

$$\boldsymbol{a}_{II} = \boldsymbol{b}_{III}, \;\; \boldsymbol{b}_{II} = \boldsymbol{a}_{III} \tag{5}$$

Therefore, from (3), (4) and (5), we can obtain

$$\boldsymbol{a}_{III} = \left(I - \boldsymbol{S}_{II,II}\boldsymbol{S}_{III}\right)^{-1}\boldsymbol{S}_{II,I}\boldsymbol{a}_I \tag{6}$$

where I is the identity matrix. Therefore, e_{rad_ANT} can be expressed as

$$e_{rad\_ANT} = \frac{P_{rad}}{|\boldsymbol{a}_{III}|^2 - |\boldsymbol{b}_{III}|^2} \tag{6}$$

$$= \frac{e_{tot\_ANTFN}}{\left|\left(I - \boldsymbol{S}_{II,II}\boldsymbol{S}_{III}\right)^{-1}\boldsymbol{S}_{II,I}\right|^2 - \left|\boldsymbol{S}_{III}\left(I - \boldsymbol{S}_{II,II}\boldsymbol{S}_{III}\right)^{-1}\boldsymbol{S}_{II,I}\right|^2}. \tag{7}$$

Thus, the radiation efficiency e_{rad_ANT} of a multiport antenna can be calculated by measured data of the total radiation efficiency e_{tot_ANTFN}, and S-matrices of the feeding network and the multiport antenna. An approximate method was used in [20] to determine the radiation efficiency, where a perfect impedance matching was assumed between the antenna and the feeding network, i.e. $S_{III} = 0$, which leads from (7) to

$$e_{rad\_ANT\_approx} = \frac{e_{tot\_ANTFN}}{\left|\boldsymbol{S}_{II,I}\right|^2} \tag{8}$$

The above expression corresponds to the case that the radiation efficiency of a multiport antenna is approximately equal to the total radiation efficiency of the antenna with the feeding network subtracted by the insertion loss of the feeding network.

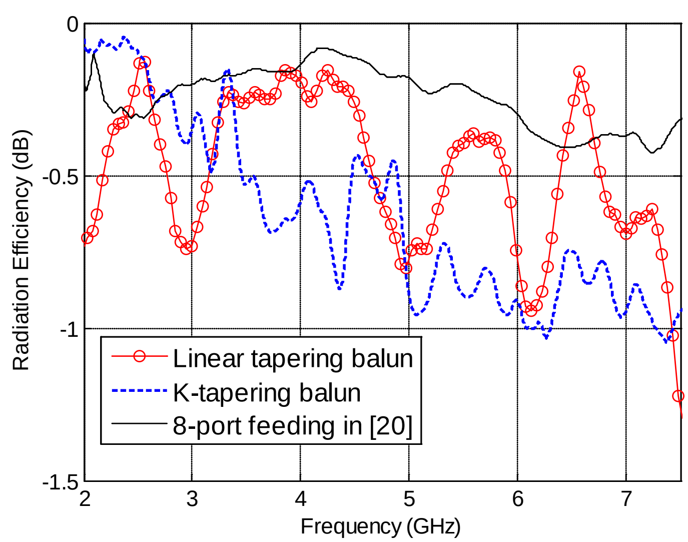


Figure 3.16: Radiation efficiency of the Eleven antenna obtained from the measurement and simulation.

Figure 3.16 shows the measured radiation efficiency by using the reverberation chamber technique over 2–8 GHz, which is the available operating frequency band of the chamber [38]. It is observed that the radiation efficiency of the balun feedings, including baluns and power combiner, is less than -1 dB over 2–8 GHz, a very good performance. But it is still 0.5 dB more loss than that of the 8-port feeding solution in [20], which is a drawback for the passive balun solution.

Chapter 4

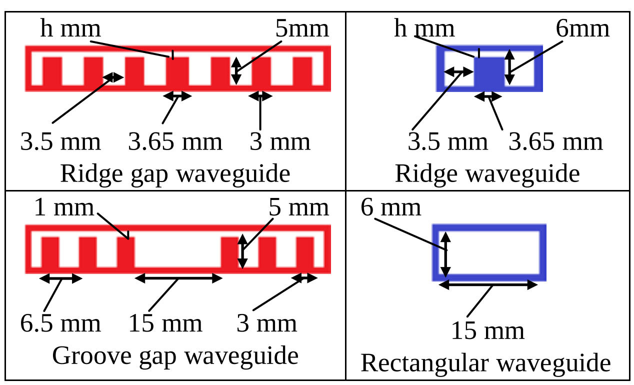
NUMERICAL PORTS AND CHARACTERISTIC IMPEDANCE OF GAP WAVEGUIDE

In order to obtain the propagation characteristics of the gap waveguide, a numerical study on the similarity between the hollow (rectangular and ridge) waveguides and the gap (groove and ridge) waveguides has been carried out in term of dispersion diagram and characteristic impedance.

4.1 The Dispersion Diagram

The dispersion diagrams of the rectangular/ridge waveguides and the groove/ridge gap waveguides are simulated for the infinite periodic structures and infinitely long along the direction of propagation, shown in Figure 4.1, using the Eigenmode solver in CST Microwave Studio. The material used here in the simulation is perfect electric conductor with no surface roughness. Cross sectional views of the geometries with the dimensions used in the simulation are summarized in Table 4.1 , in agreement with [24] in order to get a targeted stopband is from 10 to 20 GHz. Thickness of the pins and the pins' periodicity also play significant roles to determine the stopband.

Table 4.1: Cross-sections with dimensions of the simulated ridge and groove gap waveguides (left), and of the equivalent hollow ridge and rectangular waveguides (right).



The simulated dispersion diagrams of the ridge gap waveguide and classical hollow ridge waveguide are shown in Figure 4.1 for four different gap heights above the ridge, i.e. h = 0.5 mm, 1 mm, 2 mm and 3 mm. The pin heights and the air gap above pins remain the same for all cases. We see that the stopband of the parallel-plate modes for these specific cases is between 11 and 22 GHz. The fundamental mode of the ridge gap waveguide is seen to have a dispersion curve that is very close to the light line if the gap is kept small, as expected. However, it moves away from the light line when the gap between the ridge and the top metal plate is increased. However, the dispersion curves of the ridge gap waveguide are very close to the fundamental mode curves of the equivalent hollow waveguides within the stopband of the parallel-plate modes. Similarly we have in Figure 4.2 compared the dispersion diagrams for the case of groove gap waveguide and normal rectangular waveguide. The groove gap waveguide has exactly the same height as that of the rectangular waveguide, i.e. 6 mm. The stopband for this specific groove gap waveguide is between 11 and 19 GHz. The fundamental mode propagating for both type of waveguides are almost similar within the stopband.

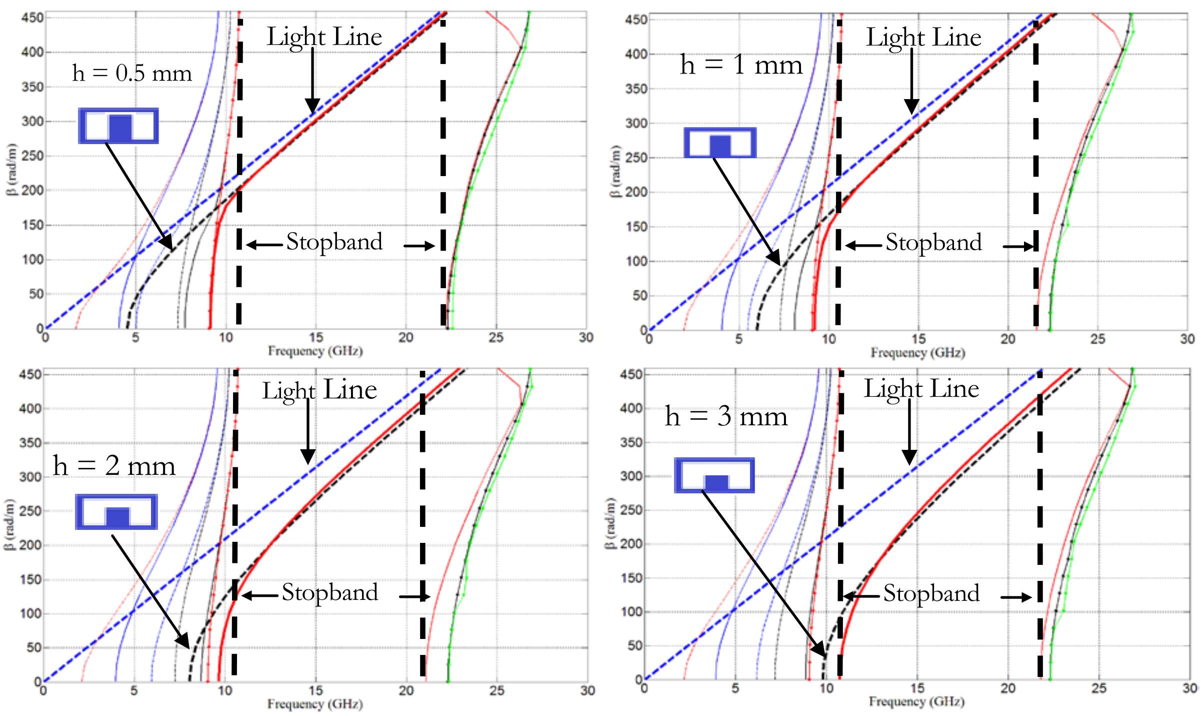


Figure 4.1: Dispersion diagram of ridge gap waveguide with different gaps heights above the ridge. The pins and air gaps are the same for all cases. The two black vertical lines mark the beginning and end of the stopband, and the dashed solid line is the dispersion diagram of the equivalent hollow ridge waveguide.

4.2 Direct Transition Between Hollow Waveguide and Gap Waveguide

The numerical studies in this section have been done in both the two commercial codes CST and HFSS on two gap waveguides with a length of 91.5 mm shown in Figure 4.3. We can easily observe that the results by CST and HFSS are very similar, see Figure 4.4.

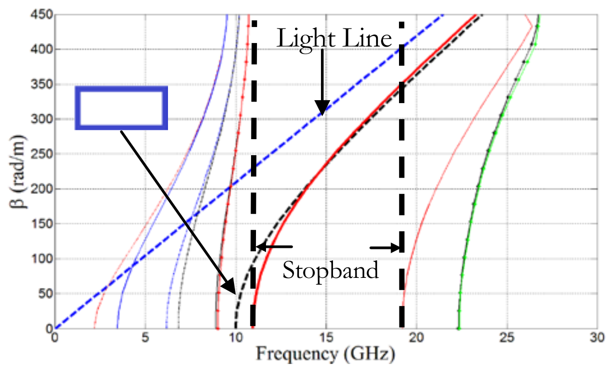


Figure 4.2: Dispersion diagrams of groove gap waveguide and equivalent rectangular waveguide (dashed black line). The two black vertical lines mark the beginning and end of the stopband.

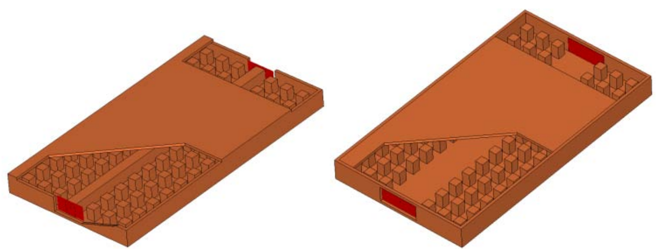


Figure 4.3: Port configuration for simulating the S-parameters of the gap waveguides.

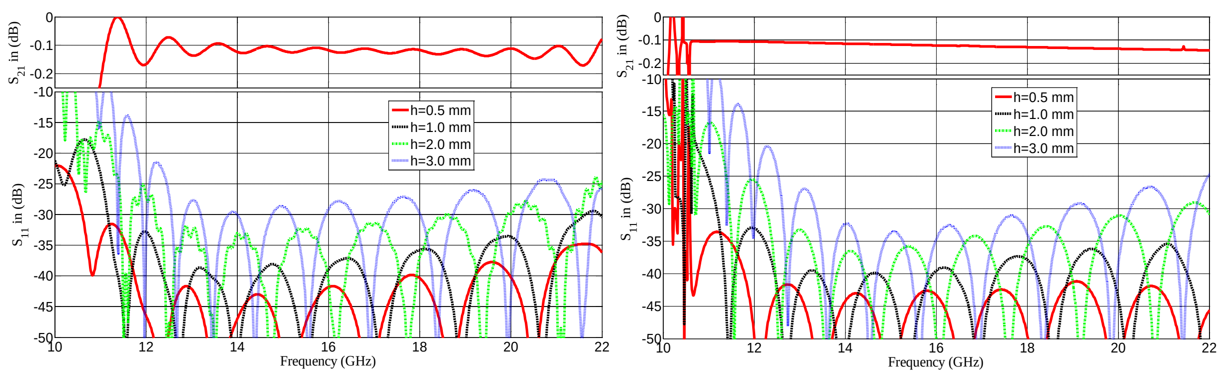


Figure 4.4: S-parameters of the waveguide for the gap, using CST (left) and HFSS (right)

The numerical waveguide ports are defined to attach with the equivalent hollow waveguides with the same dimensions as the gap waveguides, shown in Figure 4.3. We have also done simulations on both gap waveguides by making ridge waveguide interface on both sides of ridge gap waveguide as well as rectangular waveguide interface on both sides of groove waveguide. These arrangements also produce similar S-parameters as in the case shown in Figure 4.3, where the numerical ports are directly attached to the gap waveguides. Figure 4.4 and Figure 4.5 show the S-parameters for the two cases, i.e., ridge gap waveguide and groove gap waveguide, respectively. The dimensions of the ridge gap waveguides are the same as the structure used for the dispersions diagrams in Figure 4.1. We see that the reflection coefficients S_{11} of the ridge gap waveguide is below –35 dB if the gap above the ridge is smaller than or equal to 1 mm, whereas it increases when the gap increases. As predicted in all the dispersion diagrams, the stopband start from 11 GHz. Similarly, for the groove gap waveguide S_{11} is below -30 dB over most of the parallel-plate stopband, except in the beginning of it. Thus, the equivalence between the gap and hollow waveguide structures is good except near the beginning of the stopband. The periodic nature of the reflection coefficient comes from the two interfering reflections, from the port at each of the two ends of the waveguides. This means that each waveguide transition has a reflection coefficient that is 6 dB lower than the peaks of the combined reflection coefficient S_{11} that we measure [45].The simulations have been done with copper as material which explains that the transmission coefficient S_{21} is about -0.2 dB within the part of the stopband where the reflection coefficient is low.

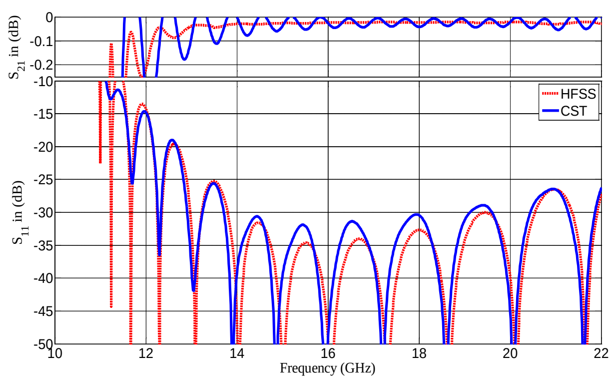


Figure 4.5: S-parameters of the groove gap waveguide using CST and HFSS.

4.3 Characteristic Impedance of Ridge Gap Waveguide

The characteristic impedance of the ridge gap waveguide depends mainly on the height of the ridge [46]. There are some methods available to calculate the approximate value of the characteristic impedance of ridge gap waveguide[47][48]. However, the results in all these methods are subject to approximations. We have already found that the reflection coefficients compared to a numerical hollow waveguide flange is -30 dB. This indicates that the ports are well matched to the gap waveguides, and we can use as a first approximation the characteristic impedance provided by the numerical field simulator for the equivalent hollow waveguide cases. We can even correct this by using the simulated reflection coefficients when we use standard numerical flanges corresponding to hollow waveguides like in the geometries in Figure 4.3. The reflection coefficient **r** of one such transition is given by

$$\mathrm{r} = \frac{Z_o - Z_{port}}{Z_o + Z_{port}}; \quad Z_o = Z_{port}\frac{1+\mathrm{r}}{1-\mathrm{r}}$$

where Z_{port} is the impedance of the numerical classical waveguide port, and Z_o is the characteristic impedance of the gap waveguide. The waveguide port impedances are provided both in CST and in HFSS over the whole frequency range. The reflection coefficient S_{11} of the whole simulation geometries in Figure 4.3 can be estimated by using theory of small reflections by using [45]

$$S_{11} \approx \mathrm{r}\left(1 - e^{-j2\beta l}\right)$$

where 'l' is the length of ridge waveguide. The absolute value of **r** will exactly be half of the magnitude of S_{11} at the peak values. This approximation is valid for small reflections, typical $|\mathbf{r}| < 0.2$. The above two equations enable us to calculate the characteristic impedance of the gap waveguides from the simulated S_{11} and the port impedances available from HFSS. With the values of the characteristic impedance at the peaks of S_{11}, the whole impedance curves can be interpolated and are shown in Figure 4.6 for ridge gap waveguide of different air gap heights. The width of the ridge is kept the same in all cases.

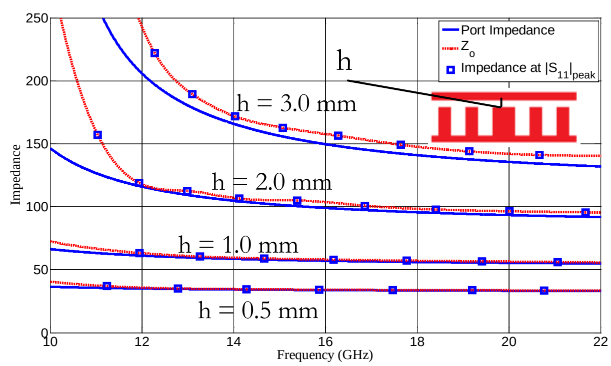


Figure 4.6: Characteristic impedance of ridge gap waveguide from HFSS port model and our corrected result (Z_0).

The characteristic impedance for the case when the air gap above the ridge is 0.5 mm is exactly the same as that of the port impedance over the whole frequency band gap. This is because the reflection coefficient is below -40 dB for this case. Slight variations appear as the air gap increase to 1 mm and more. Again referring back to Figure 4.1, it can be observed that within the stopband, the fundamental modes of both the ridge waveguide and ridge gap waveguide are exactly the same for the gap equal to 0.5 mm and 1 mm. However, for the gap equal to 2 mm and 3 mm, they are slightly different both at the beginning and at the end of the stopband. This might be the possible reason of the impedance variation shown in Figure 4.6 when the gap above the ridge is 2 mm and 3mm. The same procedure was adopted for the case of groove gap waveguide, where the wave impedance for the waveguide has been calculated through the port impedances. The results are shown in Figure 4.7.

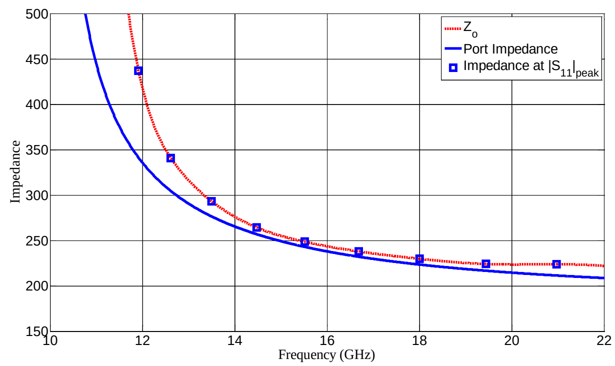


Figure 4.7: Wave impedance in groove gap waveguide from HFSS port model and our corrected result (Z_0).

4.4 A Low Loss Rat Race Balun in Gap Waveguide Technology

Hybrid 3dB coupler (0º /180º) is often needed in feeding networks for antennas [49]. However, for high frequency applications, such as in millimetre wave, sub-millimetre wave and up to terahertz, it is very challenging to realize a low-loss and low-cost hybrid. Therefore, a prototype at 16 GHz is designed and manufactured. Simulations and measurements for this prototype are presented for the verification of the design. The geometry of the new hybrid is a gap-waveguide ring surrounded by metal pins which provide the parallel-plate stop band; see Figure 4.8. In this work, the operating frequency band of 15-18 GHz has been chosen.

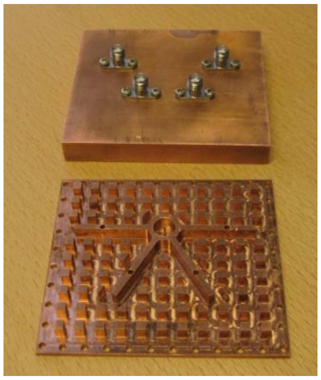


Figure 4.8: Photo of the 15 – 17.5 GHz 3dB hybrid in ridge gape waveguide. The texture plate showing the ring hybrid surrounded by metal pins.

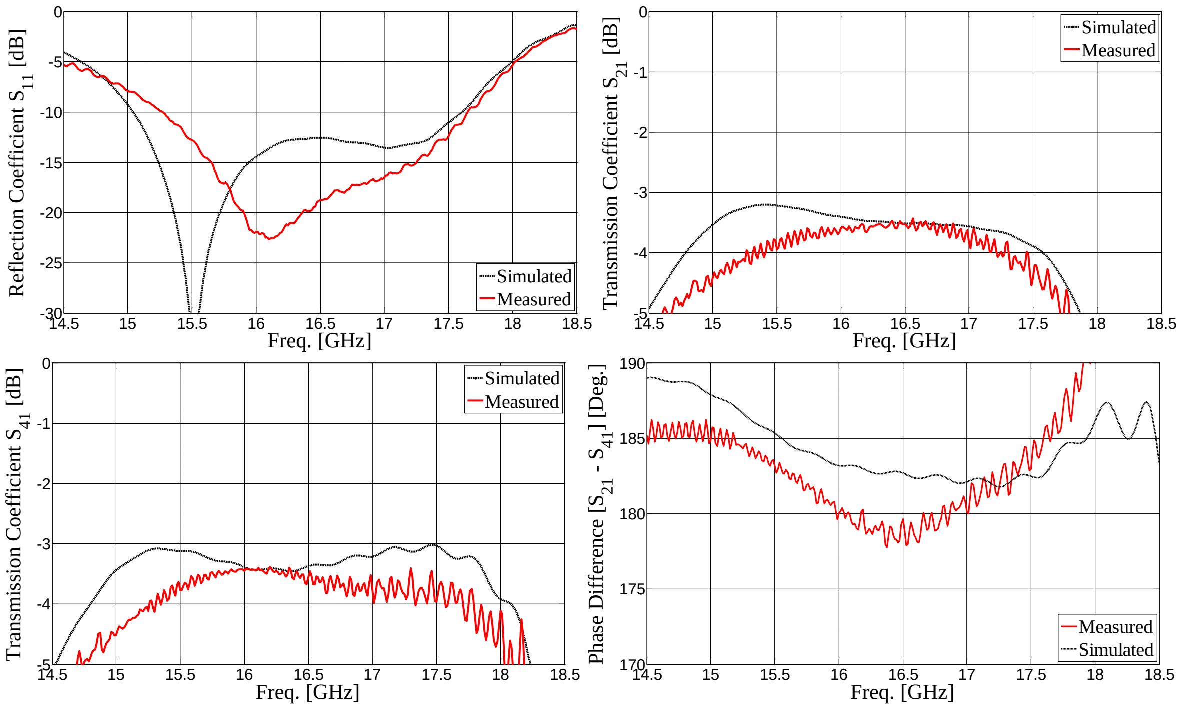


Figure 4.9: Simulated & Measured S-parameters and phase difference of port 2 and port 4, when port 1 is input port.

Figure 4.9 shows the simulated and measured reflection and transmission coefficients, respectively, when port 1 is the input port. It can be observed that the gap-waveguide ring hybrid has a very promising performance over the band of 15.25 – 17.75 GHz: the reflection coefficient is below -10 dB; the ohmic loss is very low, the transmission loss is mainly due to the mismatch loss; and the phase difference between the two output ports 2 and 4 is about 180º ± 5º as shown in Figure 4.9. Similarly, Figure 4.10 show the simulated and measured reflection and transmission coefficients, respectively, by considering port 3 as the input port, obtained by using CST MS. Again, it can be observed that the gap-waveguide ring hybrid has a very promising performance over the band of 15.25 – 17.75 GHz. The measurement results follow the simulation results quite well in all figures. The reflection coefficient is below -10 dB; the ohmic loss seems very low; and the phase difference between the two output ports 2 and 4 is about ± 5°. An empirical formula for design of the 3dB hybrid by gap waveguide was introduced in [50] after this work.

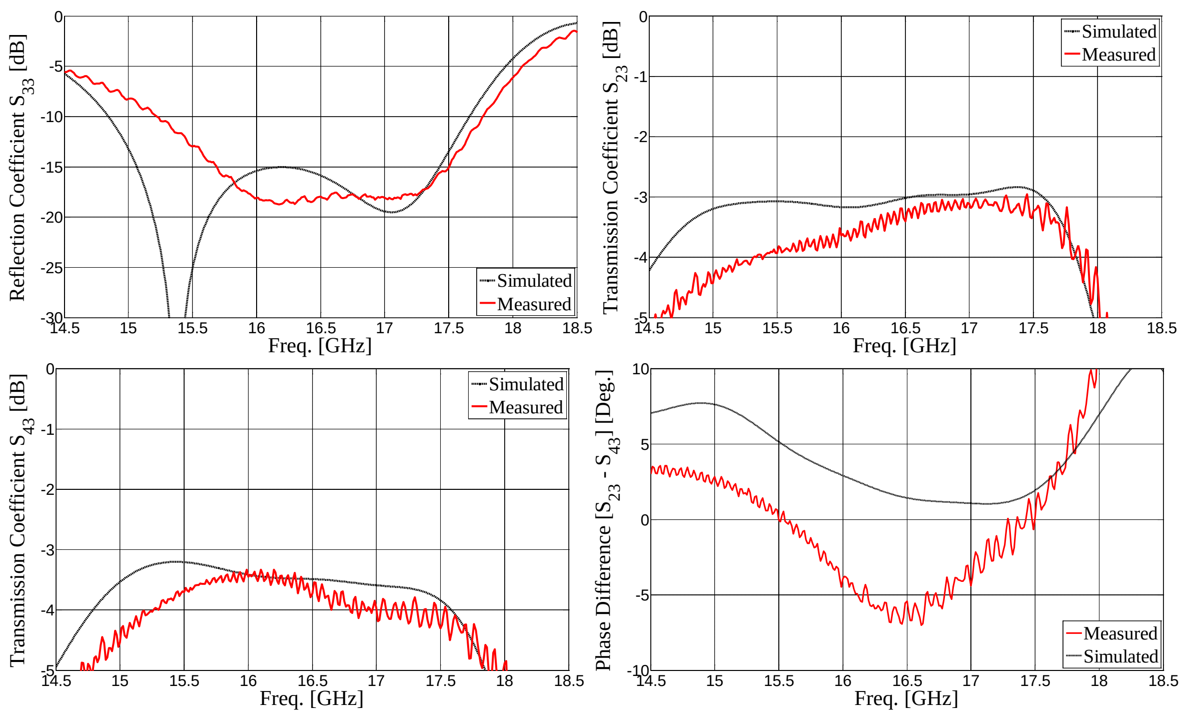


Figure 4.10: Simulated & Measured S-parameters and phase difference of port 2 and port 4, when port 3 is input port.

Chapter 5

UWB POWER DIVIDER, PACKAGED WITH GAP WAVEGUIDE TECHNOLOGY

The feeding network is always a critical part in the design of the Eleven antenna. Good UWB performance of a low reflection coefficient and low ohmic loss, a compact simple geometry and low cost make the design a real challenge. One alternative of feeding networks for the Eleven antenna is to employ the UWB passive baluns [51], plus two UWB 3-dB power divider, shown in Figure 5.1. By this feeding network, the 4 differential ports of the Eleven feed are transformed to 2 single ended ports, one for each polarization.

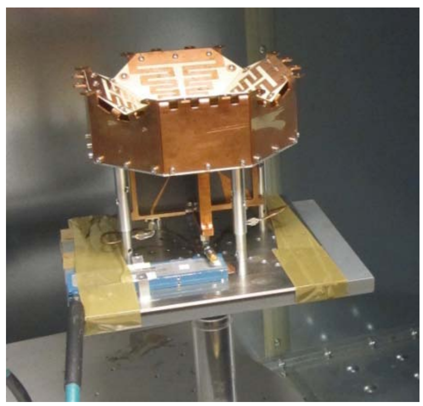


Figure 5.1: Eleven Antenna with passive balun solution and power combiner.

In addition to the high loss, the conventional packaging for UWB power dividers, simply enclosing the device in a metallic box, may also cause resonances. This will eventually degrade the performance of the whole system. The purpose is to develop a small and low loss power divider with a bandwidth in the order of 10:1. By reducing the size, when it is enclosed in a metal box, the power divider may have resonances only at higher frequencies of the bandwidth. These resonances can then be suppressed by using the gap waveguide technology, constructed of a bed of nails [34], see Figure 5.2. Both simulated and measured reflection and transmission coefficients are in good agreement between 1–12 GHz, see Figure 5.3-Figure 5.5. Although simulated results extend the behavior up to 13.5 GHz. This may be because of the mechanical inaccuracy in the milling process for the manufactured one.

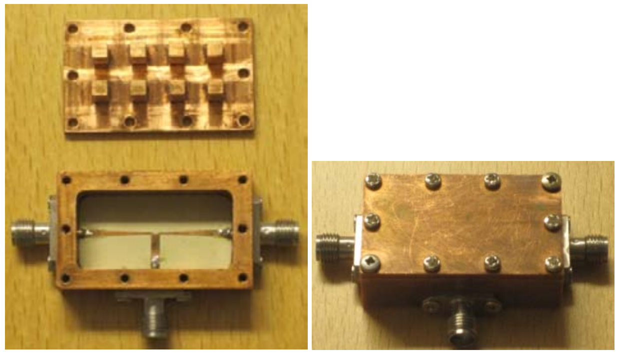


Figure 5.2: Photo of the 1–13.5 GHz power divider with bed-of-nails.

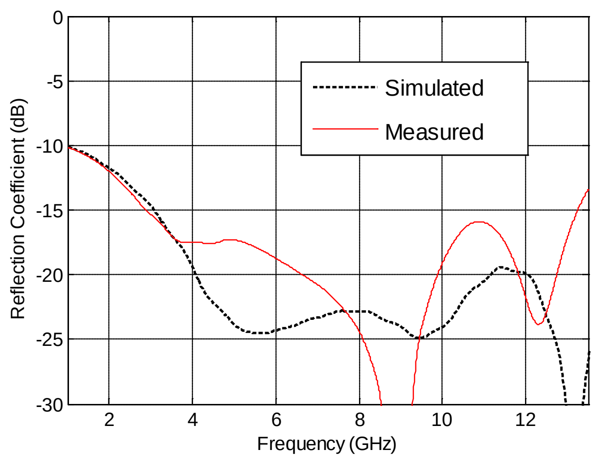


Figure 5.3: Simulated and measured reflection coefficient of the power divider package with bed-of-nails.

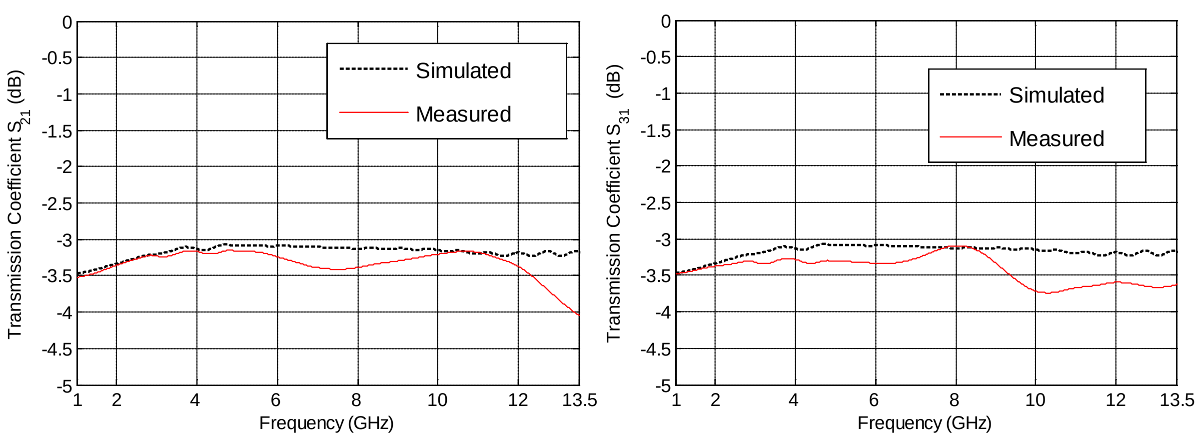


Figure 5.4: Simulated and measured transmission coefficient of the power divider package with the bed-of-nails.

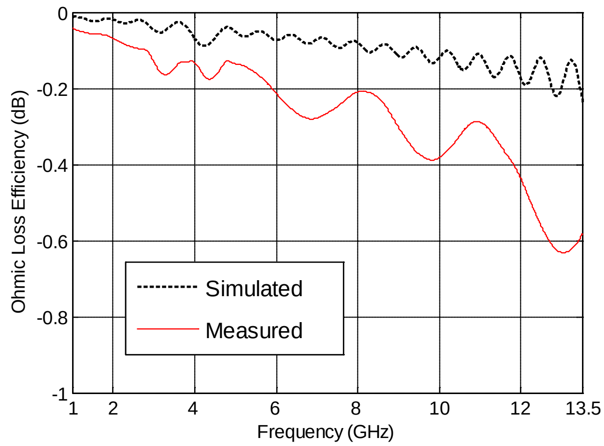


Figure 5.5: Simulated and measured ohmic losses of the power divider.

CONCLUSION

In this thesis, new compact UWB passive balun for the decade bandwidth Eleven antenna has been presented, based on using four printed circuit boards located vertically to the ground plane on the back side of it. The solution seems feasible for the requirements of the Eleven feed. The lower frequency of the operation is limited by the length of balun. There were two versions of the feeds that have been developed. The first version, linear tapering, has aperture efficiencies better than -2.5 dB between 2 – 12 GHz and around -3 dB between 12 – 14 GHz. Although the second version, K-tapering, has lower efficiency at few frequency points between 8–9 GHz, but it has better reflection coefficient than the linear tapering balun, i.e. less than -10 dB. This formation of balun will transform four differential ports to four single ended 50 ohm ports. It is also then possible to reduce these number of ports to two single ended ports, one for each polarization. A UWB power combiner is also presented to fulfill this requirement. This power combiner is based on simple T-junction packaged by using gap waveguide technology. A rigorous feeding network correction method for determining the radiation efficiency of a multiport antenna based on measurements has also been presented. This method removes the losses in the multiport feeding network from the total radiation efficiency, when there are multiple reflections between the antenna and the feeding network due to the mismatch between them. The measured radiation efficiency is about -0.5 dB.

Similarly, in the second part of this thesis, numerical studies have been done to determine the characteristic impedance and wave impedance of ridge gap waveguide and groove gap waveguide, respectively. This analysis is based on the port impedance and reflection coefficient provided by the commercial code for waveguide ports. The results show that within most of the stopband of the parallel-plate modes, both gap waveguides and equivalent hollow waveguides are performing relatively close to each other. The air gap between the top metal and the ridge is the key parameter and should be made as smaller as possible for the mode to propagate more closely to that of light line.

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