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Design and optimization of broadband planar baluns and dipole antennas

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Dedication

To my Mother Fidelina de Melais Rodriguez and my Father Sergio E. Melais L.
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# Table of Contents

List of Tables iii  
List of Figures iv  
Abstract x  

## Chapter 1 - Introduction
1.1 Overview 1  
1.2 Thesis Organization 3  
1.3 Contributions 4  

## Chapter 2 - Low-Profile Broadband Strip-Line Balun
2.1 Introduction 5  
2.2 Background Theory 6  
2.3 Comparison to Prior Balun Designs 9  
2.4 Multilayer Structure use for the Balun 12  
2.4.1 Broadside Coupled Strip-Line Transmission Line 12  
2.5 Procedures and Results 14  
2.5.1 Description of the Balun Design 14  
2.5.2 Comparison Between Simulated and Measured Data 22  
2.5.3 Statistical Monte Carlo Simulation 25  
2.6 Summary and Conclusions 27  

## Chapter 3 - Detector Circuit
3.1 Introduction 29  
3.2 Background Theory 30  
3.3 Detector Circuit Design 38  
3.3.1 Specifications for Material and Components 38  
3.3.2 Single Diode Detector 39  
3.3.3 Voltage Doubler or Dual Diode Detector 40  
3.4 Procedures and Results 41  
3.4.1 Comparison Between Simulated and Measured Data 42  
3.4.2 Monte Carlo study on Responsivity 50  
3.5 Summary and Conclusion 53  

## Chapter 4 – Broadband Printed Dipole Antenna
4.1 Introduction 54  

---

"i"
4.2 Background Theory 55
4.3 Description of the End-Loaded Planar Open Sleeve Dipole (ELPOSD) 57
4.4 Simulation Results 58
   4.5.1 Optimization and Parametric Analysis of the Antenna 59
4.5 Summary and Conclusions 71

Chapter 5 – Summary and Recommendations 72
   5.1 Conclusions 72
   5.2 Recommendations for Future Work 73

References 75

Appendices 77
   Appendix A: Even and Odd Mode Impedance Calculations 78
   Appendix B: Two Wire Transmission Line Equations 80
   Appendix C: Low-Profile Balun Simulation Results 83
   Appendix D: SOLT Calibration 93
   Appendix E: Capacitors and Inductors Plots 97
   Appendix F: TRL Calibration 102
List of Tables

Table 2.1 Rogers Material 15
Table 2.2 Ideal Values for the Schematic Figure 2.9 16
Table 2.3 Balun Dimensions on the Strip-Lines and Broadside Coupled Lines 19
Table 2.4 Dimensions for the Three Port Characterized Design 21
Table 3.1 Components Values for the Detector Circuits 39
Table 4.1 Geometry of the ELPOSD Before and After Optimization 60
Table 4.2 Comparison Between Optimized and Parameterized Results 69
Table B.1 Parameters for Calculation of Two Wire Transmission Line 80
Table B.2 Results of the Two Wire Transmission Line Equations 82
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Current Flow of a Section of Coaxial Line</td>
<td>6</td>
</tr>
<tr>
<td>2.2</td>
<td>Current Flow of a Two Wire Transmission Line Section</td>
<td>7</td>
</tr>
<tr>
<td>2.3</td>
<td>Marchand Compensated Balun, Coaxial Cross Section</td>
<td>8</td>
</tr>
<tr>
<td>2.4</td>
<td>(a) Robert’s Nomenclature for Coaxial Balun</td>
<td>10</td>
</tr>
<tr>
<td></td>
<td>(b) Microstrip Transmission Line</td>
<td></td>
</tr>
<tr>
<td>2.5</td>
<td>Lange Coupler</td>
<td>11</td>
</tr>
<tr>
<td>2.6</td>
<td>Microstrip Balun with Three Coupled Lines</td>
<td>12</td>
</tr>
<tr>
<td>2.7</td>
<td>Cross-Section of the Broadside Coupled Strip-Line</td>
<td>13</td>
</tr>
<tr>
<td>2.8</td>
<td>Model of the Broadside Coupled Strip-Line Balun</td>
<td>15</td>
</tr>
<tr>
<td>2.9</td>
<td>Ideal Transmission Line Schematic of the Low Profile Balun</td>
<td>16</td>
</tr>
<tr>
<td>2.10</td>
<td>Low-Profile Broadband Strip-Line Balun Schematic</td>
<td>18</td>
</tr>
<tr>
<td>2.11</td>
<td>Electromagnetic (Momentum) Model of the Low Profile Balun</td>
<td>20</td>
</tr>
<tr>
<td>2.12</td>
<td>Electromagnetic Model of the Balun for 3-port Characterization</td>
<td>21</td>
</tr>
<tr>
<td>2.13</td>
<td>Low Profile Broadband Strip-Line Balun</td>
<td>22</td>
</tr>
<tr>
<td>2.14</td>
<td>Comparison of the Return Loss for the Low Profile Balun</td>
<td>23</td>
</tr>
<tr>
<td>2.15</td>
<td>Comparisons of the Insertion Loss and Coupling for the Low-Profile Balun</td>
<td>24</td>
</tr>
<tr>
<td>2.16</td>
<td>Comparison of the Phase Difference of the Low-Profile Balun</td>
<td>24</td>
</tr>
<tr>
<td>2.17</td>
<td>Monte Carlo Analysis for S11 on the Broadside Coupled Strip-Lines</td>
<td>25</td>
</tr>
<tr>
<td>Figure</td>
<td>Description</td>
<td>Page</td>
</tr>
<tr>
<td>--------</td>
<td>-----------------------------------------------------------------------------</td>
<td>------</td>
</tr>
<tr>
<td>3.19</td>
<td>Responsivity of the Voltage Doubler</td>
<td>49</td>
</tr>
<tr>
<td>3.20</td>
<td>Comparison of Output Voltage Between Single Diode and Dual Diode</td>
<td>50</td>
</tr>
<tr>
<td>3.21</td>
<td>Statistical Monte Carlo Analysis Diagram</td>
<td>51</td>
</tr>
<tr>
<td>3.22</td>
<td>Monte Carlo Analysis on Resposnivity</td>
<td>52</td>
</tr>
<tr>
<td>3.23</td>
<td>Monte Carlo Analysis on Output Voltage</td>
<td>52</td>
</tr>
<tr>
<td>4.1</td>
<td>Open-Sleeve Dipole Reflector Assembly</td>
<td>56</td>
</tr>
<tr>
<td>4.2</td>
<td>ELPOSĐ Geometry</td>
<td>58</td>
</tr>
<tr>
<td>4.3</td>
<td>Optimized ELPOSĐ in HFSS</td>
<td>59</td>
</tr>
<tr>
<td>4.4</td>
<td>ELPOSĐ Fed by a Low-Profile Balun, Design in HFSS</td>
<td>61</td>
</tr>
<tr>
<td>4.5</td>
<td>Return Loss (S11dB) from Optimization on the ELPOSĐ</td>
<td>62</td>
</tr>
<tr>
<td>4.6</td>
<td>Peak Gain (dB) from the Optimization of the ELPOSĐ</td>
<td>62</td>
</tr>
<tr>
<td>4.7</td>
<td>VSWR from Optimization to the ELPOSĐ</td>
<td>63</td>
</tr>
<tr>
<td>4.8</td>
<td>Parametric Results on the Length (L) of the ELPOSĐ</td>
<td>64</td>
</tr>
<tr>
<td>4.9</td>
<td>Parametric Results on the Width (W) of the ELPOSĐ</td>
<td>65</td>
</tr>
<tr>
<td>4.10</td>
<td>Parametric Analysis on the Length of the Sleeves (Lp) of the ELPOSĐ</td>
<td>66</td>
</tr>
<tr>
<td>4.11</td>
<td>Parametric Analysis on the Width of the Sleeve (Wp) for the ELPOSĐ</td>
<td>66</td>
</tr>
<tr>
<td>4.12</td>
<td>Parametric Analysis on the Length (Ls) of the Loaded Section for the ELPOSĐ</td>
<td>67</td>
</tr>
<tr>
<td>4.13</td>
<td>Parametric Analysis on the Width (Ws) of the Loaded Section for the ELPOSĐ</td>
<td>68</td>
</tr>
<tr>
<td>4.14</td>
<td>Parametric Analysis on the Separation (S) Between the Sleeves and the Dipole Section</td>
<td>68</td>
</tr>
<tr>
<td>4.15</td>
<td>Comparisons Between Optimized and Parameterized Return Loss</td>
<td>70</td>
</tr>
</tbody>
</table>
Figure 4.16 Comparisons in Peak Gain Between the Optimized and Parameterized Simulation

Figure 4.17 Radiation Pattern for the Parameterized Antenna at 2 GHz

Figure B.1 Two Wire Line

Figure C.1 Return Loss (S11dB) of the Ideal Schematic

Figure C.2 Insertion Loss (S21dB) and Coupling (S31dB) of the Ideal Schematic

Figure C.3 S21 and S31deg. on the Ideal Schematic

Figure C.4 Phase Difference (S31-S21 deg.) of the Ideal Schematic

Figure C.5 Return Loss (S11dB) for the schematic of Balun 1

Figure C.6 Insertion Loss (S21 dB) and Coupling (S31dB) for Schematic of Balun 1

Figure C.7 S21 and S31 deg. for the Schematic of Balun 1

Figure C.8 Phase Difference (S31-S21 deg.) for the Schematic of Balun 1

Figure C.9 Return Loss (S11 dB) for the Schematic of Balun 2

Figure C.10 Insertion Loss (S21dB) and Coupling (S31dB) for schematic of Balun 2

Figure C.11 S21 and S31 deg. for the Schematic of Balun 2

Figure C.12 Phase Difference (S31-S21 deg.) for the Schematic of Balun 2

Figure C.13 Return Loss (S11dB) in Momentum from Balun 1

Figure C.14 Insertion Loss (S21dB) and Coupling (S31dB) in Momentum for Balun 1

Figure C.15 S21 and S31 deg. in Momentum for Balun 1

Figure C.16 Phase Difference (S31-S21 deg.) in Momentum for Balun 1

Figure C.17 Return Loss (S11 dB) in Momentum for Balun 2
Figure C.18  Insertion Loss (S21 dB) and Coupling (S31 dB) in Mom. Balun 2 91
Figure C.19  S21 and S31 deg. in Momentum for Balun 2 92
Figure C.20  Phase Difference (S31 –S21 deg.) in Momentum for Balun 2 92
Figure D.1  Amplitude Response for the Thru Between Port 1 and Port 2 94
Figure D.2  Phase Response for the Thru Between Port 1 and Port 2 94
Figure D.3  Amplitude of a Short at Port and Port 2 95
Figure D.4  Amplitude of an Open at Port 1 and Port 2 95
Figure D.5  Amplitude of a Load at Port 1 and Port 2 96
Figure E.1  3.9pF Shunt Bypass Capacitor Return Loss (Single Diode Detector) 97
Figure E.2  3.9pF Shunt Bypass Capacitor, Forward Transmission (Single Diode Detector) 97
Figure E.3  3.9pF Shunt Bypass Capacitor, Reverse Transmission (Single Diode Detector) 98
Figure E.4  27n H Shunt Input Choke Inductor, Return Loss (Single Diode Detector) 98
Figure E.5  27n H Shunt Input Choke Inductor, Forward Transmission (Single Diode Detector) 99
Figure E.6  27n H Shunt Input Choke Inductor, Reverse Transmission (Single Diode Detector) 99
Figure E.7  5.6p F Series Input Bias Capacitor, Return Loss (Voltage Doubler) 100
Figure E.8  5.6p F Series Input Bias Capacitor, Reverse Transmission (Voltage Doubler) 100
Figure E.9  5.6p F Series Input Bias Capacitor Forward Transmission (Voltage Doubler) 101
Figure F.1  Amplitude of the Delay Line S21 104
Figure F.2  Amplitude of the Delay Line S11 104
| Figure F.3 | Amplitude of the Thru Line S21 | 105 |
| Figure F.4 | Amplitude of the Thru Line S11 | 105 |
| Figure F.5 | Phase of the Delay S21          | 106 |
| Figure F.6 | Phase of the Thru S21           | 106 |
| Figure F.7 | Reflect (Short) S11             | 107 |
| Figure F.8 | Reflect (Short) S22             | 107 |
ABSTRACT

Due to the rapid growth in the filed of RF and Microwaves circuits, the demand for smaller, faster, and more precise devices has increased. In this thesis, a compact shielded balun, a diode detector, and a broadband sleeve dipole are modeled and implemented resulting in optimal devices presenting accurate operation.

The balun circuit is designed to meet the need for a miniaturized device with broadband application. A balun is a transformer used to connect unbalanced to balanced circuitry, for example, the unbalanced connection from a power splitter (coax) to the balanced input (two-wire) of a dipole. The main objective for the balun discussed in this work is to serve as the feeding element of a printed dipole antenna operating from 1.35 GHz to 2.7 GHz. The main characteristics from this device that distinguished it from previous structures are its PCB compatibility, shielding (strip-line), and compact footprint. The effective output balancing capability of this balun is proven by the close agreement between the measured and simulated results.

Diode detectors are nonlinear devices used for extraction of DC from free space transmission, achieving RF to DC conversion. Detectors are of popular use today in RF-
ID tags, virtual batteries, power scavenging, and many other portable applications. The dominating factor on the performance for this device is responsivity (mV/mW), which depends on the input matching, and in the relation of power level to input impedance. Here, an accurate comparison between model measured detectors has been accomplished, and a sensitivity study has been applied to the input impedance to represent variations in responsivity due to input mismatch.

At last an End-Loaded Planar Open-Sleeve Dipole (ELPOSD) is realized resulting on a printed dipole structure offering broadband response. An ELPOSD consist of a radiating element exposed to the effects of sleeves (parasitic), loading plates, and a thick substrate which contributes to a broad bandwidth. The main characteristic of its structure is the large number of parameters that can be optimized for maximum performance.
Chapter 1

Introduction

1.1 Overview

This work presents the design of a low profile broadband balun, a diode detector, and an end-loaded planar open sleeve dipole. These structures are of great interest and offer numerous applications in the area of RF and microwave systems. Baluns are important components employed in balanced to unbalanced circuitry, realizing components such as push pull amplifiers, mixers, and antenna feeds. Diode detectors are nonlinear devices used to convert RF signals into video or D.C. voltage, suggesting attractiveness for RF-ID tags and power scavenging systems. An end-loaded planar open sleeve dipole is a type of planar microstrip dipole antenna allowing a very wide operating bandwidth. The simulations and measurements of these structures have shown the realization of effective designs by reaching all the goals appointed.

The implementation of these devices is not straightforward. They suffer from design limitations caused by tolerances within the system that diminish their performance. Examples of such drawback effects are the following: in the balun, variations to the dimensions of the coupled lines and output load impedance affect the amount of return loss attained; in the detector circuit, a mismatch at the input causes power losses in the system; and in the end-loaded planar open sleeve dipole, an incorrect dimensional parameter can significantly reduce the operating bandwidth. These
challenges have motivated interesting studies towards utilizing optimization and statistical analysis which will quantify the impact of these design limitations. For the optimization study a gradient revision is used to acquire the best results, while for the statistical study a Monte Carlo analysis is included for understanding of element restrictions.

Next, the specification requirements for the designs elaborated in this thesis are mentioned. The requirements are the same for the balun and the antenna because these elements were originally designed to function in tandem. The conditions followed for the balun design required a structure having tight coupling (3dB) and maximum operation over a frequency range from 1.35 GHz to 2.7 GHz. For this a broadside coupled strip-line transmission line is selected because it permits variations in coupling and impedance just by manageable alterations in line dimensions (width and length) and coupled line separation. The main application for this balun is to serve as the feeding element for an end-loaded planar open sleeve dipole.

For the detector circuit, the specification requires a design operating at low power levels and over the 802.11b frequency band. To achieve this goal, a Schottky diode is used because of its ability to react to small signals close to the noise level and monitor large signals well above the noise floor. For this application there are two types of circuits employed, a single diode detector and a voltage doubler, with the voltage doubler being preferred because it doubles the voltage of the single diode and gives a better match to the input impedance. An important condition from this system is that the diode for the detector circuit has the characteristic of being very sensitive to the source
impedance, and in addition its input impedance is dependent on the input power level. To understand this sensitivity effect, a statistical analysis (Monte Carlo) is carried on the detector structure to evaluate the relation between input impedance and changes in power levels.

Finally, for the end-loaded planar open sleeve dipole the specifications require an antenna radiating over the same frequency range as the balun, 1.35 GHz to 2.7 GHz. This antenna consists of a microstrip dipole with parasitic elements incorporated for broadband expansion. In addition, loading effects have been incorporated through capacitive sections added to the structure. The main characteristic of this type of antenna is the dependence of its broadband operation on the large number of geometric parameters that can be changed, making it favorable for optimization.

1.2 Thesis Organization

This thesis is organized into five chapters, with chapters one and five corresponding to the introduction and conclusion respectively, and chapters two through four describing the main contents of the devices designed.

Chapter 2 introduces an analysis and design of a low profile broadband strip-line balun. The material starts by theoretically defining the balun, explaining its common applications and comparing different topologies. Following this is a description of the multilayer structure selected, listing its advantages and disadvantages over other previous structures. Next, the simulated and measured results are presented. Finally, fabrication tolerance errors and changes on the output load impedance are analyzed using a statistical (Monte Carlo) analysis.
Chapter 3 talks about the design and implementation of a diode detector circuit, beginning with a background theory where a complete definition and explanation of the device and its components is given. Next, a discussion of the two types of detector circuits generated is incorporated, indicating the advantage of using one design (Voltage Doubler) over the other (Single diode detector). Also, since a diode detector is a very sensitive device where the input impedance is affected by the power level, a statistical (Monte Carlo) simulation is done to find the relation of impedance to input power levels.

Chapter 4 elaborates on describing an end loaded planar open sleeve dipole, initiating with its application and then mentioning the different parameters that control its operation. Due to the numerous parameters of this geometry, optimization and parametric studies are run on its dimensions, resulting in a favorable approach for obtaining best performance and understanding of the antenna design.

Chapter 5 is the closing chapter, finalizing by adding concluding remarks and recommendations for future work on these types of structures.

1.3 Contributions

The design of a low profile balun, a diode detector circuit, and end loaded planar open sleeve dipole has been presented, meeting design specifications and adding valuable studies on random variations to main circuit parameters. The simulations of these devices and their characteristics have allowed for an understanding of the limitations for each design. The main contributions from this work are the introduction of microwave devices achieving optimum operation while providing characteristics such as size reduction, simplicity, and repeatability of results.
Chapter 2

A Low-Profile Broadband Strip-Line Balun

2.1 Introduction

A balun is a type of transformer used to connect balanced to unbalanced transmission line circuits, providing a balanced output from an unbalanced input. Some types of unbalanced transmission lines are coaxial cables, microstrip lines, and coplanar waveguides (CPW). Balanced transmission lines are made of two wire conductors with the same geometry and equal potential but have 180 degrees phase difference. Examples of balanced transmission structures are two wire lines and dipole antennas. Baluns are used today for balanced mixers, push-pull amplifiers, and antenna feeds.

Over the years compensated Marchand baluns have been investigated to optimize the output power balance over wide frequency ranges. These baluns are typically made of coupled microstrip lines, Lange couplers, and spiral coils. Unfortunately, these structures possess certain disadvantages. Microstrip and Lange couplers have a relatively large geometry. Spiral baluns can be very complicated to design and fabricate, and may suffer from large output signal imbalance and high parasitics [1]. In this thesis a simpler type of structure is introduced that provides a low-profile, broadband design. It consists of a broadside multilayer coupled strip-line [2] appropriate for printed circuit board techniques (PCB). In addition, this balun has been fully packaged and conveniently integrated with coaxial connectors.

2.2 Background Theory
The first studies on balun design began in the early 1940’s with Nathan Marchand and his work on “Transmission Line Conversion,” introducing a method for joining a shielded balanced two wire line to an unbalanced coaxial line [3]. This technique considers important equilibrium conditions to maintain the current on the two types of transmission lines (coaxial and two wire) with proper relation.

Coaxial transmission lines are conducting wires concentrically arranged in a hollow conducting tube filled with dielectric (e.g., air). Figure 2.1 show a front and side view of a coaxial transmission line. This displays a perfect shield (no loss) outside the inner conductor preventing coupling between current I2 inside the shielding and current Io outside the shielding. This means that current I2 on the inside of the shield must be equal in amplitude but opposite in phase to the current I1 on the inner conductor. There are two aspects in a coaxial line that need to be taken into account: prevention of coupling between currents I2, I1, and the unbalancing current Io outside the shielding; and for power considerations, maintaining the surge impedance of the line constant [3].

Figure 2.1 - Current Flow of a Section of Coaxial Line
A shielded two wire transmission line consists of two parallel conductors in a dielectric surrounded symmetrically by a shield. Figure 2.2 shows four currents (Io, I1, I2, and I3) for the side view of the line. I3 and Io are the total resultant currents in the inside and outside of the shielding respectively. For the system to be perfectly balanced (180° out of phase) three considerations need to be met: First, a current I1 with equal amplitude but opposite phase to I2, resulting in I3 equal zero. This is achieved by keeping the impedance from each conductor to the shield equal; Second, a continuous shielding for prevention of the outside current (Io) from coupling into the lines; Third, a close impedance matching between the lines for maximum power transfer.

![Diagram of a Two Wire Transmission Line Section](image)

*Figure 2.2 - Current Flow of a Two Wire Transmission Line Section*

For the connection of these two types of transmission lines a transformer (Compensated Marchand balun) is generated consisting of a shielding box, a coaxial line, open and short stubs, and a two wire line. The evolution of this design started from single frequency transformation to wide-band conversion, and then to phase shift correction [3].
Figure 2.3 illustrates the structure for the Compensated Marchand Balun [4]. Here, \( Z_1 \) and \( Z_2 \) have equal impedance; \( Z_{S1} \) and \( Z_{S2} \) operate in series and are shunted to the balanced load; \( Z_B \) has a characteristic impedance matching the \( Z_L \) value and is usually used as a matching section. This study [3] on transmission line conversion pioneered research into these types of transformers which will later incorporate microstrip lines, Lange couplers, and spiral coils.

Figure 2.3 - Marchand Compensated Balun, Coaxial Cross Section
2.3 Comparison to Prior Baluns Designs

Baluns serve as important components of wireless systems such as double-balanced mixers and push-pull amplifiers where an unbalanced to balanced transformation is required. They are either active or passive with preference on the passive baluns which have lower power consumption. Examples of these types are the 180° hybrid, the lumped-element filter, and the previously described Marchand baluns. Among passive baluns, Marchand baluns are favored, providing excellent balanced output over frequency while the 180° hybrid and the filter type are impractical due to large sizing and complicated layouts with poor balancing. The Marchand balun can be made of spiral coils, Lange couplers, and coupled microstrip lines. The balun proposed in this work offers improvements to the previous Marchand designs by introducing compact footprint (size), integrated shielding (strip-line), coupling, and printed circuit board (PCB) compatibility.

Next, a brief description of Marchand type baluns is stated to show the importance that previous works have added to the development of this technology. First, a multilayer spiral transmission line balun is explained. This is a transformer made of two coupled spiral conductors on a multilayer structure and vertically offset by a dielectric layer. The multilayer structure benefits the design by canceling the high parasitic (capacitance) brought in by the spiral coils. Figure 2.4 (a) shows the coaxial representation of the balun, where the input impedance (Zin) at point d consists of an open circuit stub impedance (Zb) in series with the parallel combination of a short circuit stub impedance (Zab) and a load resistance (R) from f to g. Broadband response is
obtained when at certain frequencies the electrical length is $\lambda/2$, resulting in a small $Z_b$ and large $Z_{ab}$ characteristic. Figure 2.4 (b) is the monolithic planar representation of the coaxial balun, Figure 2.4 (a), with one strip-line related to the conductor of the coaxial line and the other to the shield. Strip-lines connecting to ground (c) are part of an input transmission line, an open circuit stub, and a short circuit stub. For input matching purposes the impedance at point d is matched back to the source impedance ($Z_s$) by a 90° transmission line.

Figure 2.4 - (a) Robert’s Nomenclature for Coaxial Balun. (b) Microstrip Transmission Line

On the spiral configuration, the lines of Figure 2.4 (b) are coiled on themselves, increasing the mutual capacitance and segments between the lines. Therefore, for the spiral device the resonant frequency is directly related to the length of the coils. It is important to point out that the dependence of this device on the length of the lines also adds some disadvantages to its application, such as high parasitic effects due to capacitive coupling between the spirals, and large output signal imbalance produced by capacitance formed through the substrate. Full diagrams and results for a coupled spiral balun can be found in [1].

The second class of the Marchand balun is the Lange Coupler [5] or interdigitated microstrip coupler [6], consisting of three or more parallel strip-lines with alternate lines
tied together by wire bridges and built on a single ground plane, a single dielectric, and a single metal layer. The fringing fields found at both edges of this structure increase the coupling with the use of several parallel lines. This is shown in Figure 2.5. The main disadvantages of the Lange Couplers are that the lines are very narrow, close, and the bonding is difficult to fabricate.

A third class of the Marchand type balun is the planar microstrip three coupled line balun [7]. This balun is simple to fabricate (one layer) and increases coupling factor in relation to the previous mentioned cases. Figure 2.6 shows the realization of this structure. The center conductor impedances Zb and Za represent, respectively, an open terminated line and an impedance transformation for a reduction of in band reflection; the outer conductors with Zab/2 impedances are the ground resonators across the balanced line. The lengths of the lines are about one quarter wavelength (90°) at the center frequency. The configuration of this design is also simplified from Figure 2.4 (a).
Unfortunately there is a drawback to this implementation, at low gigahertz frequencies it is relatively large making its operation very difficult.

![Microstrip Balun with Three Coupled Lines](image)

**Figure 2.6 – Microstrip Balun with Three Coupled Lines**

2.4 Multilayer Structure use for the Balun

The introduction of a multilayer structure in the balun design helped increase the coupling factor, leading to the selection of a Broadside Coupled Strip-Line. This multilayer topology allows adjustments in coupling and characteristic impedance without having to change the design of the conductors. This is done by controlled variations on the distance between the grounds and between the conductors.

2.4.1 Broadside Coupled Strip-Line Transmission Line

A broadside coupled strip-line transmission line consists of three parallel layers of dielectric substrate stacked between two ground planes, and within these layers there are two strip conductors separated by a middle layer of dielectric, Figure 2.7. In printed circuit board designs such as that used here, the middle dielectric layer is usually a bonding material. Using this type of structure offers the advantage of tighter coupling and
effective separation ($S \geq 4\text{mil}$) between the strip conductor lines in comparison to the edge-coupled configuration.

Two important excitations or propagation modes (TEM) result from parallel coupled transmission lines, the even and odd mode. In the even mode the voltage and current between the two conductors are equal and with same sign; in the odd mode the voltage and current are equal but with different signs. All other conditions (voltage, current, and field) on the conductors can be expressed as a linear combination of these two modes, therefore the complete performance may be calculated in terms of the characteristic impedance and propagation constants for these modes. The formulas for calculating the even and odd mode characteristic impedance are described in Appendix A.

![Figure 2.7 – Cross-Section of the Broadside Coupled Strip-Line](image-url)
2.5 Procedures and Results

The proposed plan required the design of a broadband balun for operation over a frequency range from 1.35 GHz to 2.7 GHz. For this, fixed goals needed to be met: return loss ($S_{11}\text{dB}$) less than -10 dB, insertion loss ($S_{21}\text{dB}$) and coupling ($S_{31}\text{dB}$) between -3 to -4 dB over frequency with a maximum 1 dB of separation, and phase difference between $S_{31}$ and $S_{21}$ equal 180 degrees (+-10 degrees of error). This will ensure minimum reflection and a balanced output for the design. These standards are also set as the optimizer goals for the simulations. On this procedure a gradient and a statistical study is performed on the transmission lines of the device through an optimization and a Monte Carlo analysis, testing for maximum device performance and sensitivity.

2.5.1 Description of the Balun Design

In this work the balun is transforming an unbalanced (coaxial) input of 50 Ohms to a balanced output (two wire transmission lines) of 56 Ohm load impedance. The broadside coupled strip-line structure consists of the RO4350B ($\varepsilon_r=3.48$) for the top and bottom layer and the prepreg bonding material RO4450 ($\varepsilon_r=3.54$) for the middle layer, Table 2.1. The two substrates (top and bottom) with thicknesses (H) of 30mil surround a prepreg (S) 4 mil layer, defining the distance between ground planes (B) to be 64 mils. For conductivity ($\sigma$), a copper gold plated structure is used with a value of $5.8\times10^7$ S/m and cladding of 0.5 ounces (which is 0.7 mils or 18 microns). The input signal (coaxial) is applied to the top strip conductor by a thru-via that goes through the layers and ground. (The extension of the thru-via to the opposite ground facilitates soldering of the center pin of the coaxial connector.) The output exits from the bottom strip conductor through a
two wire parallel transmission line, Figure 2.8. The separation between the pads of the two wires depends on the diameter of the wire and the type of substrate (Er of 3.5 or 10) to which the wire is connected. Appendix B shows the pertinent two wire transmission line equations for this calculation.

<table>
<thead>
<tr>
<th>Top and Bottom Layer RO4350B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Substrate Dielectric, (Er)</td>
</tr>
<tr>
<td>Substrate Thickness, (H)</td>
</tr>
<tr>
<td>tan δ</td>
</tr>
<tr>
<td>Conductor (Copper) σ</td>
</tr>
<tr>
<td>Standard Copper Cladding (T)</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Middle Layer RO4450 Prepreg Material</th>
</tr>
</thead>
<tbody>
<tr>
<td>Substrate Dielectric, (Er)</td>
</tr>
<tr>
<td>Substrate Thickness, (S)</td>
</tr>
<tr>
<td>tan δ</td>
</tr>
</tbody>
</table>

Figure 2.8 – Model of the Broadside Coupled Strip-Line Balun
The balun is first designed schematically using ideal transmission line elements to find the even and odd mode characteristic impedance of the broadside-coupled strips. A gradient optimization is set on the lines for a coupling of -3dB. When the simulation is completed, the lines are replaced by a broadside coupled strip-line and simulated schematically and electromagnetically (Momentum).

<table>
<thead>
<tr>
<th>Table 2.2 – Ideal Values for the Schematic of Figure 2.9</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Z</strong>-<strong>even</strong></td>
</tr>
<tr>
<td>TLIN 1</td>
</tr>
<tr>
<td>TLIN 2</td>
</tr>
<tr>
<td>TLIN 3</td>
</tr>
<tr>
<td>TLIN 4</td>
</tr>
<tr>
<td>TLIN 5</td>
</tr>
<tr>
<td>TLIN 6</td>
</tr>
<tr>
<td>TLIN 7</td>
</tr>
<tr>
<td>CLIN 1</td>
</tr>
<tr>
<td>CLIN 2</td>
</tr>
</tbody>
</table>

Figure 2.9 - Ideal Transmission Line Schematic of the Low Profile Balun
Figure 2.9 represents the ideal transmission line schematic used in the optimization and its dimensions are listed in Table 2.2. The advantage of running an optimization using this type of transmission lines is that it neglects losses (conductor, dielectric loss, fringing fields etc.) in the medium resulting in a faster and approximate simulation. In addition the even and odd mode impedances are obtained (as listed on Table 2.2) matching closely to those calculated by conformal transformation in Appendix A.

Following this optimization, the ideal and coupled transmission lines are transformed into strip-lines and broadside coupled strip-lines. Forty five degrees bends are added to the transmission lines allowing the lengths to be curved and reducing the size of the structure. A second optimization is done on the new modified geometry resulting on an effective designed (Balun1) meeting desired specification. Finally, an additional optimization is run on the lines of Balun1 to include the effects that an input coaxial connector brings into the system. As a result a second balun, Balun 2, is originated. The differences between the two baluns are minor, mostly widths and lengths for the broadside coupled strip-lines and the output strip-lines sections as listed in Table 2.3. However Balun2 does present improvements in the simulation results (return loss, insertion loss, and coupling) by accounting for the inductive effect from the length of the connector. Figure 2.10 shows the resultant schematic for the broadside coupled strip-line balun and its dimensions are shown in Table 2.3.
Figure 2.10 – Low-Profile Broadband Strip-Line Balun Schematic
Table 2.3 – Balun Dimensions on the Strip-Lines and Broadside Coupled Lines

<table>
<thead>
<tr>
<th></th>
<th>Balun 1</th>
<th></th>
<th>Balun 2</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>W</td>
<td>L</td>
<td>S</td>
<td>W</td>
</tr>
<tr>
<td>SLIN 1</td>
<td>47</td>
<td>10</td>
<td>45.34</td>
<td>10</td>
</tr>
<tr>
<td>SLIN 2</td>
<td>47</td>
<td>353.7</td>
<td>45.34</td>
<td>356.84</td>
</tr>
<tr>
<td>SLIN 3</td>
<td>20</td>
<td>20</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>SLIN 4</td>
<td>10</td>
<td>87.6</td>
<td>10</td>
<td>87.31</td>
</tr>
<tr>
<td>SLIN 5</td>
<td>20</td>
<td>20</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>SLIN 6</td>
<td>50</td>
<td>137</td>
<td>50.43</td>
<td>138.4</td>
</tr>
<tr>
<td>SLIN 7</td>
<td>50</td>
<td>137</td>
<td>50.43</td>
<td>138.4</td>
</tr>
<tr>
<td>SLIN 8</td>
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<td>50.43</td>
<td>159.3</td>
</tr>
<tr>
<td>SLIN 9</td>
<td>50</td>
<td>147</td>
<td>50.43</td>
<td>159.3</td>
</tr>
<tr>
<td>SLIN 10</td>
<td>50</td>
<td>147</td>
<td>50.43</td>
<td>159.3</td>
</tr>
<tr>
<td>SLIN 11</td>
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<td>147</td>
<td>50.43</td>
<td>159.3</td>
</tr>
<tr>
<td>SLIN 12</td>
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<td>147</td>
<td>50.43</td>
<td>159.3</td>
</tr>
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<td>SLIN 13</td>
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<td>147</td>
<td>50.43</td>
<td>159.3</td>
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<td>50.43</td>
<td>10</td>
</tr>
<tr>
<td>SLIN 15</td>
<td>50</td>
<td>10</td>
<td>50.43</td>
<td>10</td>
</tr>
<tr>
<td>SBCLIN 1</td>
<td>10</td>
<td>350</td>
<td>4</td>
<td>10</td>
</tr>
<tr>
<td>SBCLIN 2</td>
<td>10</td>
<td>333.4</td>
<td>4</td>
<td>10</td>
</tr>
<tr>
<td>SBCLIN 3</td>
<td>10</td>
<td>333.4</td>
<td>4</td>
<td>10</td>
</tr>
<tr>
<td>SBCLIN 4</td>
<td>10</td>
<td>350</td>
<td>4</td>
<td>10</td>
</tr>
</tbody>
</table>

NOTE: COAX, DIGIKEYJ501-ND

*W=Width (mil)
*L=Length (mil)
*S=Separation (mil)

After simulating and optimizing the baluns in schematic they are tested using an electromagnetic simulator, Momentum. This simulator is based on the method of moments, computing general S-parameters for planar circuits (strip-line, micro-strip line) and offering the advantage of identifying parasitic coupling between components. The dimensions for this new electromagnetic design (Figure 2.11) are the same as the schematics as listed by Table 2.3. A feature presented in this simulation is the addition of via holes, providing connection between the top and bottom ground, preventing any type of radiation or interaction between the curvatures of the lines, and maximizing return loss by their
optimized position. The simulation results for the two cases (Balun 1 and Balun 2) of the low profile baluns are included in Appendix C.

![Figure 2.11 – Electromagnetic (Momentum) Model of the Low Profile Balun](image)

Next for measurements purposes changes are made to the output strip-lines of the two baluns. These changes consisted of bending the lines for fitting coaxial connectors (Amphenol RFX, 901-144-8RFX). An optimization is also run on these bent lines and maximum performance is obtained for this geometry (Figure 2.12); the dimensions are listed on Table 2.4.
Figure 2.12 – Electromagnetic Model of the Balun for 3-port Characterization

Table 2.4 – Dimensions for the Three Port Characterized Design

<table>
<thead>
<tr>
<th></th>
<th>Balun 1</th>
<th>Balun 2</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>W</td>
<td>L</td>
</tr>
<tr>
<td>SLIN 1</td>
<td>47</td>
<td>10</td>
</tr>
<tr>
<td>SLIN 2</td>
<td>47</td>
<td>353.7</td>
</tr>
<tr>
<td>SLIN 3</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>SLIN 4</td>
<td>10</td>
<td>87.6</td>
</tr>
<tr>
<td>SLIN 5</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>SLIN 6</td>
<td>50</td>
<td>369</td>
</tr>
<tr>
<td>SLIN 7</td>
<td>50</td>
<td>369</td>
</tr>
<tr>
<td>SBCLIN 1</td>
<td>10</td>
<td>350</td>
</tr>
<tr>
<td>SBCLIN 2</td>
<td>10</td>
<td>333.4</td>
</tr>
<tr>
<td>SBCLIN 3</td>
<td>10</td>
<td>333.4</td>
</tr>
<tr>
<td>SBCLIN 4</td>
<td>10</td>
<td>350</td>
</tr>
<tr>
<td>COAX (2) , 901-144-8 RFX</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

*W=Width (mil)  
*L=Length (mil)  
*S=Separation (mil)
The schematic and electromagnetic results for this twist on the output line dimensions show excellent operation for the baluns under the design specifications. Although both designs show good implementation, balun 2 is preferred due to the overall expansion on return loss for the measured and simulated results and for maintaining the separation between insertion loss and coupling within the limits (1dB).

2.5.2 Comparison Between Simulated and Measured Data

The design balun was fabricated (Figure 2.13) and measured through a full two calibration on a vector network analyzer (HP 8753D). The calibration results for the device measurements are found in Appendix D. The topology used here is the same as from Figure 2.12.

![Figure 2.13 – Low Profile Broadband Strip-Line Balun](image-url)

The comparisons of simulated (schematic and momentum) and measured data for the balun are shown next (Figure 2.14, Figure 2.15, and Figure 2.16). Figure 2.14 illustrates the return loss (S11dB) comparison pointing out the operational increase in the measured design. The improvement comes as a result of the addition of connectors at the output ports, effect which was disregarded in the simulations.
Figure 2.14 – Comparison of the Return Loss for the Low Profile Balun

Figure 2.15 shows the relation between insertion loss (S21) and coupling (S31) response for the balun experiment. The behavior on magnitude of the results proves the efficiency of the design on meeting the specifications by staying within a desired range (e.g. +- 1dB apart). The results on phase difference (S31-S21 degrees) from Figure 2.16 show an excellent operation for balancing at the output of the device (180 ° with +/- 10°). Furthermore, nonlinearities are seen in the S21 and S31 response for the measured data (Figure 2.15) which are related to consequences of calibration issues, connectors, and the nonlinear nature of the balun. These effects are later discussed in Appendix D.
Figure 2.15 – Comparisons of the Insertion Loss and Coupling for the Low Profile Balun

Figure 2.16 – Comparison on the Phase Difference of the Low Profile Balun
2.5.3 Statistical Monte Carlo Simulation

A statistical (Monte Carlo) analysis is performed in order to study the tolerance of the broadside coupled strip-lines to manufacturing errors and how variations on the load impedance will affect the performance. A Monte Carlo test consists of multiple simulation trials in which the statistical variables have values which may vary randomly with respect to the design values. The advantage of performing a Monte-Carlo simulation is that its accuracy is independent of the statistical variables, requiring no assumption about the probability distribution of either component parameter value or performance response.

The first statistical (Monte Carlo) simulation consisted of applying variations of 1 mil to the width of the broadside coupled strip-lines. Figure 2.17 shows the results of the return loss simulation, the response reveals a reasonable acceptance to manufacture errors, the worst case response appears at the ends of the frequency band, which tend to be the areas more sensitive to variations due to the trade-off of acceptable bandwidth between the upper and lower end of the frequency range.

![Figure 2.17 – Monte Carlo Analysis for S11 on the Broadside Coupled Strip-Lines](image-url)
Figure 2.18 displays the results for insertion loss and coupling. As expected, small changes on the width of the coupled lines do not add any considerable alterations on the device achieving the tight coupling performance.

A second statistical (Monte Carlo) analysis was performed in order to study the impact of load impedance variations on the performance of the balun. This sensitivity tests consists of random variations of 20% on the nominal value of each output load, Port 2 and Port 3. This verifies that variations on the load impedance over frequency, e.g. due to the frequency-dependent nature of an antenna impedance, can be tolerated. As seen from the simulation results in Figure 2.19 minimum variations are seen in the return loss response. In the contrary, a non-desirable aspect is noticeable in Figure 2.20, the separation between insertion loss and coupling has been expanded at certain values for the load. The increase in deviation comes as a result of the balun being designed to operate at a particular load impedance, but by getting farther away from this load value the balun operation is affected.
2.6 Summary and Conclusions

An efficient design for a low profile broadband strip-line balun has been realized, progressing from an ideal element to a perfectly packaged structure. The main characteristics presented by this balun are its compact footprint, integrated shielding, coupling, PCB compatibility, and broadband response, making it suitable for tasks concerning unbalanced to balanced transformation. In the designed of this structure, to
obtain maximum response, the dimensions were optimized resulting in a design offering minimum size and maximum operation over the desire frequency band, 1.35 GHz to 2.7 GHz. In addition, to account for manufacturing errors and output load variations, a statistical (Monte Carlo) analysis is performed on the coupled lines and output load parameters. The results indicated that the balun designed is very dependant to variations in the output load impedance, defining that an accurate selection for this parameter is of great interest to achieved optimum performance. Another important concluding aspect for the balun is the optimum operation provided by the simulated and measured data, resulting on an accurate design performing properly over the whole system.
Chapter 3
Detector Circuit

3.1 Introduction

This chapter describes the design of a detector circuit for operation at the 802.11b frequency band; also known as the instrument, scientific, and medical (ISM) band ranging from 2.4 to 2.4835 GHz. A detector is a nonlinear device used to achieve frequency conversion from an input RF signal to a DC voltage. This device is also known as the virtual battery and is commonly used for extraction of DC power from free-space power transmissions, allowing other devices to receive power and exchange information. For this detector a Schottky diode (HSMS-2850) is selected due to its great advantage in comparison to the PN junctions on offering low-forward voltage drop and fast switching speed. Today, detector circuits are of great popularity in power scavenging and RF-ID tags applications by offering long lifetime, low cost, and small size. In this work a detector circuit is realized from simulation to fabrication into a prototype model for measurements. The results from this experiment demonstrate good detection or DC rectification by the diode, proving, how important it is to have an optimal matching network for the system so that maximum power is transferred and less is reflected. A sensitivity study (Monte Carlo) is done on the input impedance of the detector circuit to investigate how minimal variations of this will affect the matching of the system and as a result the amount of DC output voltage that can be obtained.
3.2 Background Theory

Since their earlier designs detector circuits have been traditionally used in many applications where power transmission or extraction is required, especially for the purpose of “Free Space” transmission from a transmitting to a receiving location where any interference is neglected. Power transmission is defined as a three step process: First, DC electrical power is converted into RF power, second, the RF power is transmitted through space to some distant point, and third, the power is converted back into DC power at the receiving point. For this power transferring process to occur it is necessary to have a receiving or transmitting antenna (e.g. printed patch or dipole) and a rectifier circuit (single or full wave) in charge of converting the transferred input signal into DC power. A good review in microwave transmission is given in [10].

A detector circuit is a non linear circuit consisting of a matching network, a diode circuit, a bypass capacitor, and a load resistance, Figure 3.1. The signal is received either from an antenna or provided from a source (Network Analyzer or Signal Generator). For maximum power transferring the impedance from the antenna or the source needs to match the input impedance of the diode circuit, which is the reason why an RF matching network is incorporated. The next element in this system is the diode circuit which is in charge of the detecting process by converting the RF received signal into DC power. The capacitor at the end of the diode circuit separates the RF from the video side of the circuit, assuring DC flow. Finally, the load impedance is set to increase the output voltage coming out of the detector circuit.
A diode circuit can be made of two types of nonlinear elements: single diode (half wave rectifier) and voltage doubler (full wave rectifier). The single diode has the advantage of simplicity and minimal cost while the voltage doubler offers higher output for a given input power and lower input impedance to the source, resulting in a reduction of the RF impedance matching network [11].

The type of diode commonly used in detector circuits is the Schottky barrier diode. A cross section for this device is displayed on Figure 3.2. This shows that a diode when stripped of its package consists of a metal semiconductor barrier, which is the deposition of metal layer on semiconductor. The diode is first fabricated on a (N or P type) substrate material made of single crystal of silicon with a thickness of a few hundred micrometers. A thin layer of semiconductor material (N or P doped) is deposited on top of the single crystal using standard epitaxial or ion implantation techniques. This section is then protected or insulated by an oxide layer. Next several metallic layers are sequentially evaporated, resulting on a layer made of silicon and another for use in external contact. For this Schottky structure the top metallic deposit constitutes the anode,
while a different deposition of metals in the base forms the cathode. When a deposited metal makes contact with the n-type layer of a semiconductor, electrons within this semiconductor diffuse into the metal and form a depletion layer of thickness $d$ which has minimal free carriers. This region contains only fixed positive charges creating a perfect insulator. Furthermore, as the semiconductor becomes more positive an electric field is produced, causing a drift on the electrons which will eventually causes an equilibrium known as the barrier height. This equilibrium can be disturbed by the application of an external voltage making the metal either positive or negative in respect with the semiconductor. This condition introduces the rapid switching speed for the device when alternating from conducting to non-conducting mode.

![Figure 3.2 – Structure of a Schottky Barrier Diode](image)

The equivalent circuit of a Schottky diode is displayed in Figure 3.3. Here there are two types of parasitic coming into place; the chip parasitic and the package parasitic
(SOT 23). The chip parasitic is made of a junction resistance and capacitance (Rj and Cj). The junction resistance also known as the video resistance is the section of the diode where the RF power is converted into video output voltage. This parameter is dependent of the total operating current flowing through the diode. In general the smaller this value the better the design. The junction capacitance is a parasitic element in charge of shorting out the junction resistance, shunting more RF energy to the series resistance (Rs). This parasitic series resistance (Rs) represents the resistances of the bond-wire, lead-frame resistance, and bulk layer of silicon. Finally, the package parasitics (Lp and Cp) consist of an inductance and capacitive impedance which can be easily tuned out with an external matching network.

![Figure 3.3 – Equivalent Circuit of a Schottky Diode](image)

For the design of a detector circuits there are two types of Schottky diodes to choose from, externally biased and zero biased diodes. Externally biased diodes are generally formed on n-type silicon with high barrier metal and low values of series
resistance while the zero bias diodes are formed on p-type silicon with low barrier metal and height series resistance.

For this work the family of Agilent’s HSMS-285X zero bias Schottky detector is selected as a result of its great operation at small signal applications (Pin<-20dBm). The Spice parameters for this diode are found in [12]. In forward conditions this Schottky diode obeys equation (3.1) which results in the I-V curve from Figure 3.4

\[
I = I_S(e^{\frac{q}{nkT}(V-I_R)} - 1) \quad (3.1)
\]

Figure 3.4 – Forward Characteristic of Detectors Diodes

An important parameter of Schottky detector diodes is voltage sensitivity ($\gamma$). This is expressed as the product of the current sensitivity ($\beta$) and the video resistance ($\frac{\partial I}{\partial V}$)$^{-1}$, which is the inverse of the derivative of current with respect to voltage. A detector circuit is referred a “perfect detector” when all its parasitic and reflections losses are neglected, (3.2). Practically, in an actual circuit this is not the case. The sensitivity response of a
detector Schottky diode is affected by components within it such as junction capacitance, lead resistance, and mismatch causing reflection losses, Figure 3.5. The following formulas describe the effects of these losses to voltage sensitivity [13].

\[
\gamma = \frac{0.52 V}{I_s W} \quad (3.2)
\]

\[
\gamma_1 = \frac{0.52}{I_s(1 + \omega^2 C_j^2 R_s R_j)} \frac{mV}{\mu W} \quad (3.3)
\]

\[
\gamma_2 = \frac{0.52}{I_s(1 + \omega^2 C_j^2 R_s R_j)(1 + R_j R_L)} \frac{mV}{\mu W} \quad (3.4)
\]

\[
\gamma_3 = \gamma_2(1 - \rho^2) \frac{mV}{\mu W} \quad (3.5)
\]

Figure 3.5 shows the sensitivity response of a perfect detector (3.2) and how its performance is diminished by the addition of different components, [13]. First, due to the addition of a junction capacitance only half of the RF current is going through, split between the junction capacitance and the junction resistance, (3.3). Second, the voltage across the output load resistance (RL) will be affected by its ratio with the junction resistance, (3.4). Finally, most of the incident power is absorbed by reflection losses due to mismatch, causing drastic deterioration on the result for the system.
Schottky diode detectors are characterized for operation over a span of detection called the square law region [14]. Here, the output DC voltage is proportional to the square of the input RF voltage or the input power, ranging from the ending of the noise level (-50dBm) to the beginning of the linear or saturation region (-20dBm); over this range the slope of the response stays constant. The following (3.6) represents the formula for square law detection:

\[
V = k(\sqrt{P})^\alpha \quad (3.6)
\]

Where P represent the input power levels, V is the output voltage, \( \alpha = 2 \) at small signal level (less than -20dBm), and k is just a constant representing the continuous slope. A plot representing this law will be displayed in the following section, Figure 3.18.
Depending on the utilities for a specific design the diode detector circuit has two main topologies, single diode and voltage doubler. The typical design of a single diode detector (Figure 3.6) consists of an input inductance (L), a bypass capacitor (C), and an output load resistance. The input inductance (L) acts as an RF choke, providing a current return path along with the bypass capacitor and presenting high impedance to the incident RF signal. The bypass capacitor (C) separates the RF signal from interfering with the video side of the circuit, allowing only DC to go through. The video load resistance (RL) is in charge of amplifying the output voltage.

![Figure 3.6 – Single Diode Detector](image)

The circuit design for the voltage doubler follows a similar structure as to the single diode, Figure 3.7. The only difference is that in this case the RF choke is replaced by a series bypass capacitor, allowing RF flow but blocking DC current from returning back into the source. Also, there is the addition of an extra diode which contributes on the doubling of the output voltage. If compared to the single topology the voltage doubler presents a higher output for a given input power (double of the single diode voltage) and has a lower input impedance to the source, simplifying the input matching network.
3.3 Detector Circuit Design

When designing a detector circuit operating in the square law region there are some requirements that need to be achieved; maximum output voltage, voltage sensitivity, and power efficiency. For these conditions to be attainable a perfect matching of the input impedance for the circuit is necessary. In this work these requirements are studied under supplied design specifications for two schematic structures of the detector circuit, a single diode detector and a voltage doubler.

3.3.1 Specifications for Material and Components

The detector circuits (single diode and voltage doubler) are fabricated on a 32 mil thick Roger’s (RO4003C) substrate material with copper cladding of 1oz. (34µm). The RO4003C is selected due to its great stability and electrical advantages over high frequencies, offering a low dielectric constant ($\varepsilon_r = 3.48$) with loss tangent (tan$\delta$) of 0.0021, which will reduce the amount of losses in the circuit.

The different components [15] (inductors and capacitors) of the detectors were characterized in a simulator (Agilent Technologies, ADS) and then plotted as shown in
Appendix E with the values from Table 3.1. These plots display the characteristics of each component for return loss, forward transmission, and reverse transmission, resulting from a manual tuning done on the simulator until optimum performance is accomplished at our specific design frequency (2.45 GHz). The components have specific functions at RF and DC; at RF the inductors behave as an open circuit while allowing free flow of DC, on the other hand, the capacitors will behave as a short circuit blocking the RF from feeding through the output and only letting DC to be outputted.

<table>
<thead>
<tr>
<th>Table 3.1 – Components Values for the Detector Circuits</th>
</tr>
</thead>
<tbody>
<tr>
<td>ATC 0603 600S - bypass capacitor</td>
</tr>
<tr>
<td>Toko 0603 - input RF choke inductor</td>
</tr>
<tr>
<td>KOA 0603 - output load resistor</td>
</tr>
<tr>
<td>Coilcraft 0805 - output RF choke inductor</td>
</tr>
<tr>
<td>ATC 0603 600S - bypass capacitor (Voltage Doubler)</td>
</tr>
</tbody>
</table>

3.3.2 Single Diode Detector

Single diode detectors are the most common configuration use for in video detection circuits as a result of its great reputation on accuracy for detection of RF/microwave power and inclusion of design simplicity. The model for a single diode design is shown in Figure 3.8. The process of detection starts when the RF input power enters the system and goes through the diode detector (HSMS2850), here this power is converted into DC. Next, the resulting DC bypasses a capacitor arriving an at output load resistance where its voltage is incremented. Notice that the inductors and capacitors function different at RF in contrast to DC. For example, at RF the input inductor is neglected because it serves only at DC as a returning path, in addition a capacitor at the output short out any RF from flowing into the video side of the system.
However the expansions that this configuration has brought into the dynamic range have resulted in problems to DC sensitivity and inaccuracy of measurements due to harmonics when operating above the diode square law region, [16]. An alternative solution to this problem is introduced next.

![Single Diode Detector (Layout)](image)

**Figure 3.8 – Single Diode Detector (Layout)**

3.3.3 Voltage Doubler or Dual Diode Detector

The voltage doubler consists of the series combination of two single diode detectors. An improvement on detection responsivity is introduced here by an achievement of higher output voltage and better input match. First, the increased output voltage responsivity is an effect of the two diode voltages added in series. Secondly, the impedances of the two diodes are added in parallel providing a simpler matching condition. The application of the dual diode detector establishes advantages over the single diode detector by reducing sensitivity to DC on the center conductor at the input, sensitivity on second harmonics, and an improved signal to noise ratio (3dB), [16]. The
reduction on sensitivity (second and all even harmonics) is a consequence of one diode detecting positively while the other negatively, leading to a peak to peak detection, contradicting to the single detector which is only peak detection. The improvements on the signal to noise ratio result from the inherent doubling of the output voltage, [16].

![Image of Voltage Doubler (Layout)](image)

**Figure 3.9 – Voltage Doubler (Layout)**

The voltage doubler design (Figure 3.9) is very similar to the single diode detector (Figure 3.8) differing only on the selected diode (HSMS 2582) and that there is a bypass capacitor at the input.

3.4 Procedures and Results

The following section elaborates on the procedures followed for simulation and measurement of the detector circuits. The results from these experiments demonstrate how sensitivity is affected at different power levels due to inaccuracies brought in by impedance mismatch. For best detecting results, the measurements are conducted in a shielded environment (“screen room”), counteracting any RF interferences from external sources. The comparison of simulated and measured data showed some discrepancies in
output voltage as a result of power losses due to mismatch when sweeping through different power levels. These discrepancies lead to a study on the input impedance of the detector to test how changes in power level cause impedance mismatch in the system. For the measurement the following equipment was used: HP 8719 Vector Network Analyzer, a double stub tuner (Maury Microwave Model No.2640C), HP 8648C Signal Generator, Anritsu ML2438A Power Meter, and a HP 3478A multimeter.

3.3.1 Comparison Between Simulated and Measured Data

The first step on this experiment consisted of measuring the input impedance of the diode detectors, both the single diode and the voltage doubler. For this a thru-reflect-line (TRL) calibration is prepared to measure the S-parameters for a device on a specific substrate dielectric material. A description of the TRL calibration kit along with the calibration data is found in Appendix E. The S-parameter measurements are using the HP 8719 network analyzer. The single diode data shows minimal differences on the real (Figure 3.10) and imaginary (Figure 3.11) input impedances for the measured and simulated data. The small variation results from signal power level effects being ignored in the simulation. The measurements were performed using a signal level of -30 dBm.
Figure 3.10 – Real Part of the Input Impedance for the Single Diode Detector

Figure 3.11 – Imaginary Part of the Input Impedance for the Single Diode Detector
In the voltage doubler case, the results illustrate an expected behavior (Figure 3.12, Figure 3.13) for the input impedance, presenting a higher capacitive reactance which facilitates the functioning of the matching structure (stub tuner).

Figure 3.12 – Real Part of the Input Impedance for the Voltage Doubler

Figure 3.13 – Imaginary Part of the Input Impedance for the Voltage Doubler
Next, following the considerations on input impedance for both devices (single diode detector and voltage doubler) a double stub tuner is added to the measurement system to increase the power that can be delivered to the detector. After achieving matching through tuning, power and impedance calculations [17] are done for every element added to the network; tuner and coaxial cable connector. Tracking these values will allow for power correction. Referring to Figure 3.14, the correction consisted of measuring the power at the tuner output into $50\, \Omega$ (PL), the reflection coefficients for the source ($\Gamma_s$), power meter ($\Gamma_{ps} = 1$) and the load ($\Gamma_L$). Notice that ideally $\Gamma_s$ should be equal to the complex conjugate impedance of $\Gamma_L$. This situation leads to finding the actual power available ($P_{av}$) from the output of the tuner by using (3.7).

$$P_{av} = PL \frac{\left|1 - \Gamma_s \Gamma_{ps}\right|^2}{(1 - \left|\Gamma_{ps}\right|^2) \cdot (1 - \left|\Gamma_s\right|^2)} \quad (3.7)$$

The actual power ($P_{in}$) input to the detector system is calculated from (3.8):

$$P_{in} = P_{av} \frac{\left|1 - \Gamma_s\right|^2 \cdot (1 - \left|\Gamma_L\right|^2)}{(1 - \left|\Gamma_s \cdot \Gamma_L\right|^2)} \quad (3.8)$$

It is important to state that there two other corrections that could be taken into account but were neglected for simplicity. The losses added by the female to female barrel used to connect the coaxial cable to the coax connector of the tuner. In addition there are also losses at the center of the SMA connector used for the diode. Inclusion of these factors will give the exact power received at the detector.
Figure 3.14 describes the set up used to measure the output voltage. A 2.45 GHz signal varying in power due to attenuation within the source is sent to the detector and the DC voltage is measures with a multimeter. These measurements are done inside a screening environment to avoid detection and rectification of other external signals.

Next, the design schematic (ADS) used for the simulation of the detector circuit is presented, Figure 3.15. Notice that as previously mentioned the input structure varies depending on the diode topology (single diode or dual diode) selected for the detector. It is important to mentioned that for the simulations, the used of substrate scalable models from the global model library provided by Modelithics Inc. facilitated the modeling of the component desired for the input (inductor or capacitor), [15].
Figure 3.15 – Diode Detector Schematic Simulated in ADS

The comparison of simulated and measured data for the detectors circuits (single diode and voltage doubler) are displayed from Figure 3.16 to Figure 3.19.

Figure 3.16 – Output Voltage for the Single Diode Detector
Figure 3.17 – Responsivity for Single Diode Detector

Figure 3.18 – Output Voltage for the Voltage Doubler
The previous figures suggest good comparison for both output voltage and responsivity for both detectors. Figure 3.16 and Figure 3.17 for the single diode case, present small disagreement on the results at the higher power levels, caused by the sensitivity of the device for input matching at high powers. For the voltage doubler plots, Figure 3.18 and Figure 3.19 show minimal difference for output voltage and responsivity display, corresponding to the aid added for matching by the higher capacitive reactance value of its input impedance. In addition, a comparison is proposed between the voltage doubler and the single diode, Figure 3.20, the results confirmed the advantage on using the dual diode by producing double the overall output voltage in respect to the single diode. A supported explanation for the effects of mismatch on the system will be explained next by a Monte Carlo study.
Figure 3. 20 – Comparison of Output Voltage Between Single Diode and Dual Diode

3.3.2 Monte Carlo Study on Responsivity

The goal of this study is to understand the cause of differences between the simulated and measured detector responsivity. It is suspected that when a stub tuning match is constructed, the measured impedance match is very sensitive to variations on input power (Vector Network Analyzer power level settings) because of the dynamic diode impedance, resulting in a degraded match. The result is a decrease in the responsivity (mV/mW).

The detector impedance is obtained experimentally from S-Parameter measurements done with a vector network analyzer (HP 8719D). This impedance is matched through the combination of a series line and a shunt (short-circuit) stub. For a shunt stub matching section, there are two parameters that need to be adjusted; the distance from the load to the beginning of the stub, which provides a complex value (Yo
+ jB) for the admittance looking into the line; the shunt stub length, which provides a susceptance value (–jB) that will cancel the imaginary part of the line resulting in a match for the system (Yo + jB - jB = Yo). Choosing between an Open and Short Circuit stub will depend on the distance of the load to the Open or Short Circuit location on the Smith Chart. The stub with the shortest length to load is usually preferred.

The Monte Carlo statistical study consisted of introducing an impedance equation (Z-Block) to the detector schematic in order to add random variations within a certain percentage (e.g. 15%) to the impedance of the load resistance (detector input). The addition of this variable parameter permitted an analysis on the sensitivity of the device by indicating at what point or range the match was loss. A block diagram representing this test is shown in Figure 3.21.

![Figure 3.21 – Statistical Monte Carlo Analysis Diagram](image)

Figure 3.22 and Figure 3.23 reveal the sensitivity of the device to changes in power levels, concluding that a minimal variation in power causes a significant mismatch.
for the system. The statistical test applied reproduces the variations that the input impedance of the detector is exposed to during the switching of input power level.

Figure 3.22 – Monte Carlo Analysis on Responsivity

Figure 3.23 – Monte Carlo Analysis on Output Voltage
3.5 Summary and Conclusion

In this work two diode designs are developed for voltage detection, a single diode detector and a voltage doubler, both resulting in an accurate modeling procedure providing good result comparison. From the study done in this work the voltage doubler design is preferred because it provides greater output voltage (twice that of the single diode) and the comparison of simulated and measured results agree better due to lower mismatch and less power losses for the system. There are many other factors that could be taken into account to improve performances for both devices. An addition of a filter stage at the input will improve performance by rejecting the additional harmonics that come along with the main RF signal (from the source). Also another improvement is to consider the equipment’s instability for doing lower power level measurements. For example, the multimeter used was very instable when measuring voltages lower than 0.1 mV.
Chapter 4

Broadband Printed Dipole Antenna

4.1 Introduction

An antenna radiating element providing a broadband return loss, VSWR, efficient gain, and unidirectional radiation pattern over nearly an octave bandwidth (1.35 GHz to 2.7 GHz) is explained. This element consists of a planar dipole with two parasitic elements (sleeves) fed by a broadband balun. The main characteristic introduced by this planar dipole structure in comparison to the conventional cylindrical dipole structure is the addition of two parasitic elements providing a wider bandwidth. This structure also features an end-loaded section increasing the width of the dipole, which will contribute by adding capacitive loading effects.

This chapter elaborates on a study done on an End-Loaded Planar Open Sleeve Dipole (ELPOSD) to study the effect of varying geometrical and material parameters on antenna performance. Since there are a large number of parameters for this structure the analysis is based on a robust but efficient optimization performed using ANSOFT HFSS. The results from the simulations are revised through a parametric analysis to estimate antenna limitations. Following this optimization and parameterization, a low-profile broadband balun is attached to serve as the feeding point. The return loss response from the supplemented feed is evaluated for consistency, meaning, the behavior of the antenna should not be altered dramatically.
4.2 Background Theory

A dipole (standing wave antenna) is frequently defined as a thin wire antenna or a lossless two wire flared transmission line in which the radiating fields do not cancelled each due to the separation of the wires, causing a net radiation in the transmission line system [18]. Dipoles are being used today for applications covering a wider range of bandwidth, driving the development of optimal designs for broadband operation, for example: biconical antennas, Bow-Tie antennas, cylindrical dipoles, folded dipoles, and sleeve dipoles. Frequently, cylindrical and sleeve dipoles are compared since both are based on the same structure, but the sleeve approach generally introduces a broader band characteristic[19].

A cylindrical dipole is a type of broadband dipole. Its bandwidth increases with the thickness of the cylinder diameter. Restrictions on bandwidth determine that a very thin linear dipole suffers from narrow band input impedance characteristics, and any perturbation to the operational frequency will result in large changes in its behavior. A procedure to increase this operational bandwidth consists of decreasing the length to diameter ratio. This is accomplished by keeping the length the same while increasing the diameter of the wire [19].

A cylindrical dipole radiator is referred to as a sleeve antenna when the exterior of the coaxial transmission feed line used to feed the antenna is also integrated as part of the radiating element. This antenna has an operational bandwidth that is wider than the conventional, thin cylindrical dipole. An open sleeve antenna which is a variation in the physical arrangement of the sleeve antenna consists of a dipole with two immediate
separated parasitic elements, with the length of the parasitics (sleeves) being roughly one half that of the center dipole. The main advantage of using an open sleeve dipole over the conventional dipole is that it is less sensitive to frequency and provides an even wider bandwidth.

Work on this type of antenna is documented in [20], which describes a balun fed open sleeve dipole mounted on top of a metallic reflector; the dipole reflector assembly is shown in Figure 4.1. This design presents two important aspects; the feeding of the dipole through a coaxial line; and the incorporation of a balun. In this case the dipole and the sleeve were constructed with cylindrical elements. A brief explanation of the study done on this structure by [20] is stated next.

![Figure 4.1 – Open-Sleeve Dipole Reflector Assembly](image)

A parametric study over a frequency range from 225 MHz to 400 MHz is carried out on the VSWR response of the open sleeve dipole (Figure 4.1) as a function of the dimensions for dipole and sleeve diameter (D), sleeve length (L), and sleeve to dipole spacing (S). Before running the parametric analysis the dipole to reflector separation (Sd)
was adjusted to a value with appropriate VSWR response, eliminating the addition of extra parameters into the simulation. A display of the wide band properties for this characterized open sleeve dipole antenna compared to a conventional dipole under even conditions (same balun as feed point) is shown in Figure 2 (from [20]). As expected the VSWR response becomes flatter as the dipole length (L) is decreased and the diameter (D) is increased, decreasing the ratio of length to diameter.

The impact of the substrate thickness is very significant and can be use towards improvements on broadband response. In the microwave region the thickness of the substrate used for printed antennas is a small fraction of a wavelength only. Thin substrates suggest resonator antennas with high Q, increasing the radiation resistance and leading to narrow bandwidth. The Q can be lowered by increasing the width of the dipole and using a thicker substrate, introducing advantages in bandwidth and in fabrication tolerance. The only problem with this structure is the generation of surface waves, adding power losses to the antenna system, [21].

4.3 Description of an End-Loaded Planar Open Sleeve Dipole (ELPOSD)

An illustration of the ELPOSD geometry analyzed in this chapter is shown in Figure 4.2. This structure is preferred because it offers several design parameters that can be varied to obtain a wider range of operating bandwidth. It is possible to have different versions of this antenna with broadband or multi-band response.

The ELPOSD antenna is a center fed planar microstrip dipole printed on a substrate ($\varepsilon_r = 10$) and backed by a Perfect Electric Conductor (PEC) metal ground plane. On occasion plates are placed at the end of the dipole adding capacitive loading effects.
The presence of this loading characteristic contributes to a uniform current flow along the entire length of a dipole, [18]. This antenna is designed to be fed by a low profile balun. On the different optimization and parameterization simulations the balun is replaced by a tapered feed defined with an excitation of 50 Ohms (Port1). Figure 4.2 lists the parameters varied in the simulations; $L_p$ and $W_p$, representing the sleeve length and width; $L$ and $W$ correspond to the dipole length and width, which are the most important parameters in determining the bandwidth, in addition, $W$ (equal to the width of the dipole) represents the distance between the 45 degree tapers; $W_s$ and $L_s$ are the sleeve width and length; and finally $S$, indicating the separation between dipoles and the sleeves.

![Figure 4.2 – ELPOSD Geometry](image)

4.4 Simulation Results

The design of this type of microstrip antenna requires the specification of several parameters for the geometry and material characteristics. Here, a robust optimization process is implemented on an ELPOSD to obtain maximum return loss ($S_{11} \text{dB}$) over a
defined frequency range (1.35 GHz to 2.7 GHz) and a peak gain greater than 0 dB, additionally, a parameterization analysis is applied to the optimized values to examine the behavior of each parameter.

4.5.1 Optimization and Parametric Analysis of the Antenna

Optimization and parameterization analysis on the ELPOSD are realized using the OPTIMETRICS feature from Ansoft HFSS. The dimensions used to start the optimization are listed in Table 4.1. For any antenna analysis it is required to define a radiation box enclosing the radiating element. The antenna metallization and ground plane are set as perfect electric conductors. The excitation (port) is set as a tapered line (50 Ohms) with a 45 degree angle placed in the center of the dipole.

![Figure 4.3 – Optimized ELPOSD in HFSS](image)

Figure 4.3 – Optimized ELPOSD in HFSS
Table 4.1 – Geometry of the ELPOSD Before and After Optimization

<table>
<thead>
<tr>
<th>ELPOSD Geometry Parameters</th>
<th>Start</th>
<th>Optimized</th>
</tr>
</thead>
<tbody>
<tr>
<td>Substrate Dielectric Constant ([E_r])</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>Substrate Thickness ([H]) (mm)</td>
<td>10.16</td>
<td>17.7</td>
</tr>
<tr>
<td>Substrate Loss Tangent, (\tan \delta)</td>
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<td>0.0027</td>
</tr>
<tr>
<td>Dipole Length ([L]) (mm)</td>
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<td>21.94</td>
</tr>
<tr>
<td>Dipole Width ([W]) (mm)</td>
<td>4.28</td>
<td>17.99</td>
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<tr>
<td>End-Load Length ([L_s]) (mm)</td>
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<td>8.86</td>
</tr>
<tr>
<td>End-Load Width ([W_s]) (mm)</td>
<td>2.35</td>
<td>2.17</td>
</tr>
<tr>
<td>Parasitic Element Length(sleeve) ([L_p]) (mm)</td>
<td>13.43</td>
<td>9.41</td>
</tr>
<tr>
<td>Parasitic Element Width(sleeve) ([W_p]) (mm)</td>
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<td>8.68</td>
</tr>
<tr>
<td>Spacing ([S]) (mm)</td>
<td>2.85</td>
<td>1.58</td>
</tr>
<tr>
<td>Dipole Feed Width ([F]) (mm)</td>
<td>1</td>
<td>0.72</td>
</tr>
</tbody>
</table>

The results from this optimization (Table 4.1) confirmed theoretical strategies for expanding bandwidths on dipole antennas. Here are some considerations: First, the thickness of the substrate material increased by approximately 70%, proving that a relatively thick substrate improves bandwidth extension. Secondly, the width of the dipole increased while the length of the dipole did not change, causing the ratio of length to width to decrease from 4.7:1 to 1.21:1. Third, the length of the sleeve is about one half the length \((L)\) of the dipole, which is common for this type of antenna [20].

After the optimization a low profile balun is incorporated to replace the taper feeding point, Figure 4.4. A new simulation is conducted having two wire lines coming from the output of a low profile balun as feeding point, furthermore, the input port is set at the input coaxial connector of the balun.
The simulation results from the ELPOSD tapered feed are compared to the ELPOSD balun feed to verify that there is not a major change due to the feed implementation. The first simulation on return loss resulted in an operational bandwidth (1.65 GHz to 2.3 GHz) of 32% for the ELPOSD tapered feed and 20% for the balun feed, Figure 4.5. These differences are caused by the nonlinear nature of the balun. A comparison is also done on the total gain of the antenna over frequency, as shown in Figure 4.6; the peak for both simulations is above 1dB meeting the requirements over frequency. Finally, the VSWR response (Figure 4.7), as expected from the return loss response, is below 2:1 from 1.65 GHz to 2.3 GHz.
Figure 4.5 – Return Loss (S11dB) from Optimization on the ELPOSD

Figure 4.6 – Peak Gain (dB) from the Optimization of the ELPOSD
In addition to a robust optimization on the dimensions of the ELPOSD, it is very useful to add a parametric analysis as part of the simulation process. A parametric analysis specifies all the design variations solved during the optimization, offering a better understanding of the effect that each parameter has on the antenna response. The simulated parametric analysis was defined to select every parameter and sweep its value from a minimum to a maximum range. For example, the study presented here sweeps the variables with a limit of +- 70% of the nominal value (optimized value). This type of analysis is also simulated in HFSS, selecting and setting the centered band (2.025 GHz) as the point to be parameterized. The simulation consisted of sweeping each individual parameter while the others were held at their nominal value (optimized value), once this process was completed all the new values from the parametric analysis were simulated.
together, resulting in a 22% improvement in the overall bandwidth in correspondence to the already-optimized antenna. Figure 4.8 through Figure 4.14 display the results for the different behavior for each parameter as swept by the parametric analysis.

Figure 4.8 displays the variations on length for the dipole. A length of 27.94 mm provides the return loss with the broadest bandwidth, 52%, improving the optimized response from Figure 4.5 by 20%. Even though it is expected for the response to improve as the length of the dipole is decreased, this is not the case here because as listed in Table 4.1 the width of the optimized response to the original value has increased considerably (75%), the new improve result just takes this value and chooses a better ratio for this ELPOSD structure in relation with its center frequency (2.025GHz).

Next, changes on the width of the dipole are analyzed, Figure 4.9, resulting in better return loss for a width of 17.78 mm with an operating bandwidth of 44%, also
improving the previous optimized results by 12%. From these plots it is established that finding an accurate ratio between length and width (e.g. 1.5:1 for this case) at a corresponding frequency will contribute to maximum return loss.

![Graph showing S11 (dB) vs Frequency (GHz) for different widths of the ELPOSD.]

**Figure 4.9 – Parametric Results on the Width (W) of the ELPOSD**

Figure 4.10 and Figure 4.11 show the response for the parameterization analysis run on the lengths and widths for the sleeves, respectively. In evaluation of the length, a maximum bandwidth of 25% is accomplished for a sleeve length of 8 mm. For the width sweep the operating bandwidth is 35% for a sleeve width of 11.55 mm. In comparison to the dipole dimensions, the length of the sleeve is about one half the size of one of the arms while the width of the sleeve is 65% of the dipoles width.
Figure 4.10 – Parametric Analysis on the Length of the Sleeves (Lp) of the ELPOSD

Figure 4.11 – Parametric Analysis on the Width of the Sleeve (Wp) for the ELPOSD
Figure 4.12 and Figure 4.13 show results for a sweep in length and width for the loading element added to the dipole. A length value of 7.62 mm provides an operating bandwidth of 38% and a width value of 0.5 mm provides an operating bandwidth of 25%. From the dimensions obtain here and from the optimized result it is concluded that a long but thin load element is more favorable on this particular topology.

![Graph showing S11 dB vs Frequency for different Ls values](image)

**Figure 4.12 – Parametric Analysis on the Length (Ls) of the Loaded Section for the ELPOSD**

The final parameter from the geometry sweep is the separation between the dipole and the sleeves. Figure 4.14 shows no major difference due to adjustments to this parameter. The separation selected is 0.5mm, presenting the same bandwidth as the optimized design, 35%. This parameter adds a beneficial factor for simulation, by recognizing that its variation is independent of performance, and thus less time is spent on varying it.
Figure 4.13 - Parametric Analysis on the Width (Ws) of the Loaded Section for the ELPOSD

Figure 4.14 – Parametric Analysis on the Separation (S) Between the Sleeves and the Dipole Section
A parameterization analysis has been done to the dimensions of an optimized ELPOSD. The new design is a combination of the best parameterized results, leading to improvements which have increased the operational bandwidth of the ELPOSD relative to the design obtained via optimization. As shown in Table 4.2 most of the improvements were related to the dipole length, loading element, and width of the sleeves.

<table>
<thead>
<tr>
<th>ELPOSD Geometry Parameters</th>
<th>Optimized</th>
<th>Parameterized</th>
</tr>
</thead>
<tbody>
<tr>
<td>Substrate Dielectric Constant [Er]</td>
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<td>Substrate Thickness [H] (mm)</td>
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<td>Substrate Loss Tangent, tanδ</td>
<td>0.0027</td>
<td>0.0027</td>
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<td>Dipole Length <a href="mm">L</a></td>
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<td>Dipole Width [W] (mm)</td>
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<td>End-Load Width [Ws] (mm)</td>
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</tr>
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<td>Parasitic Element Width(sleeve) [Wp] (mm)</td>
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<tr>
<td>Spacing [S] (mm)</td>
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<td>0.5</td>
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<tr>
<td>Dipole Feed Width [F] (mm)</td>
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<tr>
<td>Maximum Operating Bandwidth %</td>
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<td>60</td>
</tr>
<tr>
<td>Antenna Size (mm)</td>
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<td>88x59x17</td>
</tr>
</tbody>
</table>

The following figures illustrate a comparison between the optimized and the parameterized designs, showing improvement in both return loss (Figure 4.15) and peak gain (Figure 4.16) over the frequency range. Also the radiation patterns for the electric (phi = 90) and magnetic (phi = 0) field are shown for this element, Figure 4.17.
Figure 4.15 - Comparison Between Optimized and Parameterized Return Loss

Figure 4.16 – Comparisons in Peak Gain Between the Optimized and Parameterized Simulation
4.5 Summary and Conclusions

An optimization analysis is done on the dimensions of an end-loaded planar open sleeve dipole antenna for maximum return loss and gain over a specific frequency (1.35 GHz to 2.7GHz). The simulation results show the importance of the sleeve dimensions and how modifications on one end of the frequency band can affect the other end, e.g., if optimizing for optimum performance at the lower end, the mid band and the upper end will suffer.

The parametric analysis is used to predict the behavior of the simulator when optimizing the antenna geometry. This review will be of great help before performing an optimization because it what value ranges to choose during the optimization.
Chapter 5

Summary and Recommendations

5.1 Conclusions

In this work, the design of a low profile broadband strip-line balun, a diode detector, and an end-loaded planar open sleeve dipole have been successfully designed and tested confirming accurate design techniques.

During the simulations, dominating factors controlling the operation of the circuits were analyzed. The dominating factors were: the coupled line dimensions and the output load impedance in the balun; the relation of input impedance to input power level in the detector circuit; and the ratio of length to width of the dipole arms on the end-loaded planar open sleeve dipole. These parameters were all considered during the simulations via optimization and statistical analysis introducing an excellent method to study design tolerances.

A low-profile broadband balun consisting of three layers of dielectric material is presented. This simple structure (broadside coupled strip-line) provides a tight coupling factor and good return loss over a wide frequency range. In addition, the balun design has been fully packaged and integrated with connectors demonstrating excellent performance over the desire frequency band.

Next, a review of the designed and measured diode detector circuit is discussed. The comparisons of these results have shown the realization of a successful design
offering excellent RF to DC detection. However, there is a small discrepancy in the responsivity comparison caused by an impedance mismatch associated with changes in the input power level. These differences are studied through statistical analysis on the input impedance for the system. The evaluation concluded that a better response can be achieved if the power losses of extra components are subtracted from the system. In order to improve detection, the diode detector is measured in a shielded environment, and power losses are subtracted from the system using mathematical expressions.

The end-loaded planar open sleeve dipole optimization and parametric analysis have offered an antenna design with an accomplished bandwidth of 54 percent. The antenna structure consists of a microstrip dipole fed by either a tapered line or a balun. The bandwidth of the return loss was found to be from 1.7 GHz to 3 GHz, completing a broadband bandwidth response.

5.2 Recommendations for Future Work

The analysis and experiments presented on the previous structures have produced interesting conclusions and ideas for future research.

The balun simulation and experimental data suggested that a reduction on the overall size can be achieved by a tighter compression of the circuit layout. However, size reduction adds disadvantages by causing an increase in reflection, thus there exists a small trade-off between size and performance (return loss).

In the case of the detector circuits, there are two important recommendations for improvements on measured results. First, the addition of a filter at the input of the detector, which blocks all the harmonics that come along with the input signal from the
source instrument and also avoids any return from these harmonics into the source. The addition of this element will also permit the use of an antenna as a receiving source. Second, performing a calibration and input matching the detector at each individual power level results in maximum power transfer to the input of the diode.

The end-loaded planar open sleeve dipole has offered a wide bandwidth, but it has mainly emphasized the upper end of the frequency range. It will be remarkable to see the result on the bandwidth if an antenna with thick substrate but with a lower dielectric constant is analyzed. It is suspected that a dielectric constant reduction will improve the overall bandwidth, including the lower end of the frequency range.
References


Appendices
Appendix A: Even and Odd Mode Impedance Calculations

The following formulas give the even ($Z_{oe}$) and odd ($Z_{oo}$) mode characteristic impedance for the broadside coupled strip-line transmission lines. There are two methods used for these calculations; conformal mapping [8] and explicit expressions [5].

The conformal mapping method is based on Schwarz-Christoffel transformation. These formulas hold for any ratio of $w/b$ and $s/b$ as long as $w/s$ is greater than 0.35.

\[
j := \sqrt{1 - \frac{1}{2}}, \quad \varepsilon := 3.48, \quad k := 0.5
\]

\[
w := 10.25.4, \quad d := 4.25.4, \quad b := 64 \cdot 25.42
\]

\[
w = 254.2 \text{ um}, \quad d = 101.68 \text{ um}, \quad b = 1.627 \times 10^3 \text{ um}
\]

\[
k := \text{root} \left[ \frac{w}{b} - \frac{2}{\pi} \text{ atanh} \left( \frac{k \left( k - \frac{d}{b} \right)}{1 - k \frac{d}{b}} \right) - \frac{d}{b} \text{ atanh} \left( \frac{k \left( k - \frac{d}{b} \right)}{k \left( 1 - k \frac{d}{b} \right)} \right) \right], k \quad \text{(A.1)}
\]

\[k = 0.362 \quad \text{Parameter}
\]

\[kp := \sqrt{1 - k^2} \quad \text{(A.2)}
\]

\[
\text{kratio1}(x) := \frac{\pi}{\ln \left( \frac{2 \cdot 1 + \sqrt{x}}{1 - \sqrt{x}} \right)} \quad \text{(A.3)}
\]

\[
\text{kratio2}(x) := \frac{1}{\pi} \ln \left( 2 \cdot \frac{1 + \sqrt{x}}{1 - \sqrt{x}} \right) \quad \text{(A.4)}
\]

Kratio = [K (K)/K (Kp)] elliptic integral of the first kind.

\[
\text{kratio} := \text{if}(0 \leq k \leq 0.707, \text{kratio1}(kp), \text{kratio2}(k))
\]

\[
Z_{even} := 59.952 \frac{\pi}{\sqrt{\varepsilon r}} \frac{1}{\text{kratio}} \quad \text{(A.5)}
\]

\[
Z_{even} = 152.281 \text{ Ohms}
\]

\[
Z_{odd} := \frac{94.172 \pi \frac{d}{b}}{\text{atanh}(k) \sqrt{\varepsilon r}} \quad \text{(A.6)}
\]

\[
Z_{odd} = 26.179 \text{ Ohms}
\]
Appendix A (Continued)

\[ \text{Coupling} := \frac{(\text{Zeven} - \text{Zodd})}{(\text{Zeven} + \text{Zodd})} \quad (A.7) \]
\[ \text{Coupling} = 0.707 \]

Appendix A (Continued)

\[ C := 20 \log(\text{Coupling}) \]
\[ C = -3.016 \quad \text{dB} \]

The next method is based on using explicit expressions. When the values for the

\[ \text{Coupling} \text{ and the characteristic impedance are specified, } C = -3\text{dB and } \text{Zo} = 50\text{ohms.} \]

\[ C := 10^{-\frac{3}{20}} \]
\[ C = 0.708 \]

\[ \text{Zeven} := \text{Zo} \cdot \sqrt{\frac{1 + C}{1 - C}} \quad (A.8) \]
\[ \text{Zeven} = 120.914 \quad \text{Ohms} \]

\[ \text{Zodd} := \text{Zo} \cdot \sqrt{\frac{1 - C}{1 + C}} \quad (A.9) \]
\[ \text{Zodd} = 20.676 \quad \text{Ohms} \]
Appendix B: Two Wire Transmission Line Equations

A two wire transmission line is a type of transverse electromagnetic transmission line (TEM) on which its wave propagation is characterized by an electric and magnetic field transverse to the direction of propagation; Table B.1 list important parameters for the calculation of a two wire transmission line [5].

<table>
<thead>
<tr>
<th>Two Wire Transmission Lines</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency (f)</td>
<td>2</td>
<td>GHz</td>
</tr>
<tr>
<td>Permeability of Free Space (µ0)</td>
<td>4πe-7</td>
<td>H/m</td>
</tr>
<tr>
<td>Relative Permeability of Material (µr)</td>
<td>1</td>
<td></td>
</tr>
<tr>
<td>Permittivity of Free Space (ε0)</td>
<td>8.85E-12</td>
<td>F/m</td>
</tr>
<tr>
<td>Conductivity of the Material (σ)</td>
<td>0</td>
<td>S/m</td>
</tr>
<tr>
<td>Conductivity of the Copper (σ1)</td>
<td>5.80E+07</td>
<td>S/m</td>
</tr>
</tbody>
</table>

For an arbitrary dielectric material, the application of some electric field causes the addition of some polarization (Pe) increasing the electric flux density. For this, B.1 is substituted into B.2 resulting in the electric flux density B.3 from which the complex permittivity B.4 is obtained. Note that $\chi$e is defined as the electric susceptibility.

\[ D = \varepsilon_0 E + Pe \quad \text{(B.1)} \]
\[ Pe = \varepsilon_0 \chi e E \quad \text{(B.2)} \]
\[ D = \varepsilon_0 E + Pe = \varepsilon_0 (1 + \chi e)E = \varepsilon E \quad \text{(B.3)} \]
\[ \varepsilon = \varepsilon 1 - j\varepsilon 2 = \varepsilon_0 (1 + \chi e) \quad \text{(B.4)} \]

By including B.4 (Electric Flux Density) and B.5 (Ohms Law) into B.6 (Maxwell Curl) we obtain the loss tangent formula (B.7) which is the ratio of the real ($\varepsilon 1 = \varepsilon \varepsilon_0$) to the imaginary part ($\varepsilon 2 = \frac{\sigma}{2\pi f}$) of the total displacement current.
Appendix B (Continued)

\[ J = \sigma \ast E \]  \hspace{1cm} (B.5)

\[ \nabla \times H = \frac{\partial D}{\partial t} + J \]  \hspace{1cm} (B.6)

\[ \tan \delta = \frac{\varepsilon_2}{\varepsilon_1} \]  \hspace{1cm} (B.7)

Based on the conductivity value from Table B.1 the loss tangent of the system is calculated to be zero. Next, a set of equations ([4], [9]) are use to calculate the lumped parameters of a two wire transmission line (Figure B. 1) for a 50 Ohm system. The wire diameter dimensions are specified for a substrate dielectric of (Er) 3.5 with (diameter) \( 2a = 40 \), and for a substrate dielectric (Er) 10 with \( 2a = 20 \).

Self Inductance per unit length (H/m): \[ L = \frac{\mu}{\pi} \cosh^{-1} \left( \frac{D}{2a} \right) \]  \hspace{1cm} (B.8)

Capacitance per unit length (F/m): \[ C = \frac{\pi \varepsilon_1}{\cosh^{-1} \left( \frac{D}{2a} \right)} \]  \hspace{1cm} (B.9)

Resistance per unit length (\( \Omega \)/m): \[ R = \frac{R_s}{\pi a} \]  \hspace{1cm} (B.10)

Skin Depth (m): \[ \xi_s = \sqrt{\frac{2}{2\pi f \sigma}} \]  \hspace{1cm} (B.11)

Surface Resistance of the conductors (\( \Omega \)): \[ R_s = \frac{1}{\sigma_1 \xi_s} \]  \hspace{1cm} (B.12)

Shunt Conductance per unit length S/m: \[ G = \frac{\pi^2 2 \varepsilon \varepsilon_2}{\cosh^{-1} \left( \frac{D}{2a} \right)} \]  \hspace{1cm} (B.13)
Appendix B (Continued)

<table>
<thead>
<tr>
<th>Relative Permittivity or Dielectric Constant ((\varepsilon_r))</th>
<th>3.5</th>
<th>10</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Diameter (2a) of each wire</td>
<td>40</td>
<td>20</td>
<td>mil</td>
</tr>
<tr>
<td>Separation between center of the wires (D)</td>
<td>53</td>
<td>40</td>
<td>mil</td>
</tr>
<tr>
<td>Self Inductance per unit length (L)</td>
<td>3.14E-07</td>
<td>5.27E-07</td>
<td>H/m</td>
</tr>
<tr>
<td>Capacitance per unit length (C)</td>
<td>1.24E-10</td>
<td>2.11E-10</td>
<td>F/m</td>
</tr>
<tr>
<td>Skin Depth ((\xi_s))</td>
<td>1.48E-06</td>
<td>1.48E-06</td>
<td>m</td>
</tr>
<tr>
<td>Surface Resistance of the Conductors (Rs)</td>
<td>0.012</td>
<td>0.012</td>
<td>Ω</td>
</tr>
<tr>
<td>Series Resistance per unit length (R)</td>
<td>7.311</td>
<td>14.622</td>
<td>Ω/m</td>
</tr>
<tr>
<td>Shunt Conductance per unit length (G)</td>
<td>0</td>
<td>0.00E+00</td>
<td>S/m</td>
</tr>
<tr>
<td>Characteristic Impedance ((Z_o))</td>
<td>50.372-j0.047</td>
<td>49.941-j0.055</td>
<td>Ω</td>
</tr>
</tbody>
</table>

![Figure B.1 Two Wire Line](image_url)

Figure B.1 Two Wire Line
Appendix C: Low-Profile Broadband Balun Simulation Results

Figure C. 1 - Return Loss (S11 dB) of the Ideal Schematic

Figure C. 2 - Insertion Loss (S21 dB) and Coupling (S31 dB) of the Ideal Schematic
Appendix C (Continued)

Figure C. 3 - S21 and S31deg. on the Ideal Schematic

Figure C. 4 - Phase Difference (S31-S21 deg.) of the Ideal Schematic
Appendix C (Continued)

Figure C. 5 - Return Loss (S11 dB) for the Schematic of Balun 1

Figure C. 6 - Insertion Loss (S21 dB) and Coupling (S31 dB) for the Schematic of Balun 1
Appendix C (Continued)

Figure C. 7 - S21 and S31 deg. for the Schematic of Balun 1

Figure C. 8 - Phase Difference (S31-S21 deg.) for the Schematic of Balun 1
Appendix C (Continued)

Figure C. 9 - Return Loss (S11 dB) for the Schematic of Balun 2

Figure C. 10 - Insertion Loss (S21 dB) and Coupling (S31 dB) for the Schematic of Balun 2
Appendix C (Continued)

Figure C. 11 - S21 and S31 deg. for the Schematic of Balun 2

Figure C. 12 - Phase Difference (S31-S21 deg.) for the Schematic of Balun 2
Appendix C (Continued)

Figure C. 13 – Return Loss (S11dB) in Momentum from Balun 1

Figure C. 14 – Insertion Loss (S21dB) and Coupling (S31dB) in Momentum for Balun 1
Figure C. 15 – S21 and S31 deg. in Momentum for Balun 1

Figure C. 16 – Phase Difference (S31-S21 deg.) in Momentum for Balun 1
Appendix C (Continued)

Figure C. 17 – Return Loss (S11 dB) in Momentum for Balun 2

Figure C. 18 – Insertion Loss (S21 dB) and Coupling (S31 dB) in Mom. Balun 2
Figure C. 19 – S21 and S31 deg. in Momentum for Balun 2

Figure C. 20 – Phase Difference (S31 – S21 deg.) in Momentum for Balun 2
Appendix D: SOLT Calibration

In the results for the measured balun, there is the critical appearance of ripples degrading the response for insertion loss (S21) and coupling (S31) of the device. To understand the origin of these errors, either measurement related or due to the device, several test calibrations were performed using different network analyzers (Anritsu MS4623B Scorpion three ports and HP 8753D two ports).

The studies revealed that ripples are incorporated by measurement related factors such as cable movements, connector orientation, and the transition from one medium to another (coaxial to strip-line). Cable movements affect the response by varying the phase and adding uncertainties to the calibration correction. The vertical orientation of the connectors for the balun adds considerable bends to the cables causing constant phase movement and changes in the dielectric of the cable. Finally, the transition from coaxial to strip-line for the input and output results in discontinuities in the medium which could contribute to the nonlinear ripple effect.

The plots for the best calibration for our balun measurement is shown next, it is important to mention that these results were obtained in the HP 8753D network analyzer.
Figure D.1 – Amplitude Response for the Thru Between Port 1 and Port 2

Figure D.2 – Phase Response for the Thru Between Port 1 and Port 2
Appendix D (Continued)

Figure D.3 – Amplitude of a Short at Port 1 and Port 2

Figure D.4 – Amplitude of an Open at Port 1 and Port 2
Figure D.5 – Amplitude of a Load at Port 1 and Port 2
Appendix E: Capacitors and Inductors Plots

Figure E.1 - 3.9pF Shunt Bypass Capacitor Return Loss (Single Diode Detector)

Figure E.2 - 3.9pF Shunt Bypass Capacitor Forward Transmission (Single Diode Detector)
Appendix E (Continued)

Figure E.3 - 3.9pF Shunt Bypass Capacitor Reverse Transmission (Single Diode Detector)

Figure E.4 - 27n H Shunt Input Choke Inductor Return Loss (Single Diode Detector)
Appendix E (Continued)

Figure E.5 - 27n H Shunts Input Choke Inductor Forward Transmission (Single Diode Detector)

Figure E.6 - 27n H Shunt Input Choke Inductor Reverse Transmission (Single Diode Detector)
Appendix E (Continued)

Figure E.7 – 5.6p F Series Input Bias Capacitor Return Loss (Voltage Doubler)

Figure E.8 – 5.6p F Series Input Bias Capacitor Reverse Transmission (Voltage Doubler)
Appendix E (Continued)

Figure E. 9 – 5.6p F Series Input Bias Capacitor Forward Transmission (Voltage Doubler)
Appendix F: TRL Calibration

To measure the input impedance of the detector circuit a transition from coax to micro-strip is required, suggesting the implementation of a TRL (thru-reflect-line) calibration kit. The advantage of this type of calibration is that it allows for the addition of correction effects for a particular micro-strip substrate and freedom on the placement of the reference plane for the system. For the designed of this TRL calibration kit the following definitions were established: the length of thru line is arbitrarily selected and used to set the reference plane, the reflect line is fixed as a short with a total length of half the thru, and the delay line is equal to the length of the thru plus the length of a quarter wavelength the center frequency (2.45 GHz). In this measurement, due to the positioning of the diode detector, for proper calibration the reference plane had to be moved to the edge of the board. To explain this procedure, the dimensions of the calibrations standards are listed and the necessary equations for shifting the reference plane are elaborated.

1. Dimensions for the Calibration Standards:

L_thru := 2.54 cm
L_90 := 1.876 cm
L_delay := L_thru + L_90 (E.1)
L_delay = 4.416 cm
L_reflect := \frac{L_thru}{2} \quad (E.2)
L_reflect = 1.27 cm

2. Equations followed to shift the Reference plane from the middle of the thru to the edge of the board:
Appendix F (Continued)

\[ \text{Eff} := 2.65 \]

\[ \frac{3 \times 10^{10}}{\sqrt{\text{Eff}}} \quad (E.3) \]

\[ \text{vp} = 1.84 \times 10^{10} \text{ cm/s} \]

\[ \text{T}_{\text{delay}} := \frac{L_{\text{delay}}}{\text{vp}} \quad (E.4) \]

\[ \text{T}_{\text{delay}} = 2.4 \times 10^{-10} \text{ Seconds} \]

\[ \text{T}_{\text{thru}} := \frac{L_{\text{thru}}}{\text{vp}} \quad (E.5) \]

\[ \text{T}_{\text{thru}} = 1.38 \times 10^{-10} \text{ Seconds} \]

\[ \text{T}_{\text{reflect}} := \frac{L_{\text{reflect}}}{\text{vp}} \quad (E.6) \]

\[ \text{T}_{\text{reflect}} = 6.902 \times 10^{-11} \text{ Seconds} \]

Finally, after the standard definitions are calculated a TRL calibration kit is mill and implemented for error correction using the HP819D. The results from the calibration show adequate error correction for the kit and the results are displayed on the next figures (Figure F.1 through Figure F.8).
Appendix F (Continued)

Figure F.1 - Amplitude of the Delay Line S21

Figure F.2 - Amplitude of the Delay Line S11
Appendix F (Continued)

Figure F.3 - Amplitude of the Thru Line S21

Figure F.4 - Amplitude of the Thru Line S11
Figure F.5 - Phase of the Delay S21

Figure F.6 - Phase of the Thru S21
Appendix F (Continued)

Figure F.7 - Reflect (Short) S11

Figure F.8 - Reflect (Short) S22