Design And Characterization Of Multi-Layer Coplanar Waveguide Baluns And Inductors

Khaled Obeidat
University of South Florida

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Design And Characterization Of Multi-Layer Coplanar Waveguide Baluns And Inductors

by

Khaled Obeidat

A thesis submitted in partial fulfillment of the requirements for the degree of
Master of Electrical Engineering
Department of Electrical Engineering
College of Engineering
University of South Florida

Major Professor: Tom M. Weller, Ph.D.
Lawrance P. Dunleavy, Ph.D
Horace C. Gordon, Jr., M.S.E

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Keywords: model, marchand, antenna, compensation, spiral

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DEDICATION

To my mother Haia, my brother Omar, my sisters Rash and Loda, my wife Noor, and in memory of my father Ahmad Obeidat, who had been through a lot.
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DESIGN AND CHARACTERIZATION OF MULTI-LAYER COPLANAR WAVEGUIDE BALUNS AND INDUCTORS

Khaled Obeidat

ABSTRACT

This work examined the design and characterization of multilayer coplanar waveguide baluns and inductors. This work derives a design procedure that helps RF engineers design cost effective multilayer coplanar waveguide (CPW) spiral balun that works in the frequency range 1-8 GHz. The accuracy of the developed procedure has been proven by designing two balun circuits of different dimensions and simulating them using available commercial software, Momentum (MoM) and Empire (FDTD). The simulation results have shown good balun performance over the desired frequency range. Furthermore some of the designed balun circuits have been fabricated and measured and the results agree with the simulations. The smaller balun (2.4 mm x 1.4 mm) with a minimum spacing of 25µm works very good in the frequency range 4-8 GHz with a 4 GHz operational bandwidth (OBW) and 5° phase difference and 0.5 dB amplitude imbalance. The larger balun (5.6mm x 3.0 mm) with minimum spacing of 100µm works well in the frequency range 2-4 GHz with a 2 GHz operational bandwidth (OBW) and 10° phase difference and 0.5 dB amplitude imbalance. Such a large-size balun is suitable for a new fabrication technique called Direct-Write.
This thesis focuses on techniques that can be used to enhance balun performance, it has been shown through this work that adding some capacitance at certain points in the balun circuit will decrease both the phase difference and the amplitude imbalance of the balun. Some of these techniques were discovered through the thesis work and the other techniques were used before, but for different balun structures.

An additional study to the effect of the ground plane on the spiral inductor model is included herein. Formulas for the inductance nominal value in the existing CPW ground plane for some spiral inductors are derived here, in addition to the derivation of an RF spiral inductor model that is independent of the ground plane. The importance of this model lies in its necessity in designing an antenna dipole loaded with lumped elements (in the absence of ground plane) to control the antenna electrical length without changing its physical length.
CHAPTER 1 – INTRODUCTION

1.1 Introduction

Symmetric dipole antennas are symmetric resonant structures requiring a balanced feed as shown in Figure 1.1. However, the connection to the signal source is typically an unbalanced line such as a coaxial cable. Coaxial cable is inherently unbalanced because the currents on the inner and outer parts of the ground conductor are not the same – i.e., they are unbalanced. The matching network between the unbalanced cable and balanced antenna terminals is called a balun, derived from Balanced to Unbalanced. If the currents on the antenna element are not balanced, spurious back lobes and asymmetry will appear in the radiation pattern (Figure 1.2 and Figure 1.3). The balun job is to deliver equal current amplitude through its two output ports with 180° phase difference, return loss should be minimum to ensure proper matching and the insertion loss should be high to ensure most of the power is delivered to the output ports.

In order to control the antenna dipole electrical length without changing its physical length lumped elements can be used to load the antenna dipole. However, the lumped elements formulas should be independent of the effect of the ground plane since the dipole exists in free space (No ground).
Figure 1.1 - Simple dipole with balanced feed.

Figure 1.2 - Radiation diagram of a dipole with balun in free space.

Figure 1.3 - Radiation diagram of a dipole without balun.
1.2 Thesis Organization

The thesis is divided into eight chapters. The first chapter describes the general problem tackled in this research and defines the meaning and the importance of the balun. Brief summary to the thesis and all of its chapters is presented at the end of this chapter. The second chapter describes the Marchand balun and a CPW (Coplanar Wave Guide) multilayer version of it. The third chapter addresses the design procedure of the spiral multilayer CPW Marchand balun to help the designer design similar type of circuits. It also describes different compensation techniques that can be used to enhance the performance of the multilayer CPW Marchand spiral balun. The fourth chapter describes two different model of the spiral multilayer CPW Marchand balun and it provides comparative results obtained by using full-wave simulation (Momentum) and using the two models. The fifth chapter includes comparison of the measured results of certain balun structures and the results obtained from the full-wave EM simulation. The last section of this chapter also shows comparison of the baluns designed in this work and in previous works. The sixth chapter is a benchmark comparison of the simulation results of the balun circuit obtained using two commercial full-wave EM simulators, namely Momentum, which uses MoM (Method of Moment) and Empire, which uses the FDTD (Finite-Difference Time-Domain) method. The seventh chapter is an additional study of the effect of the ground plane on the spiral inductor model. A spiral inductor model independent of ground plane was derived. This model is important for designing a dipole loaded with inductors, since there is no physical ground at the dipole. The last chapter is a summary of the whole thesis work and some recommendations for future work in this area.
1.3 Research Contributions

The problems solved here are important for antenna design engineers. In this research we developed a design procedure for baluns that works in the frequency range 1-8 GHz. Also we developed a procedure for finding an accurate inductor model that is independent of the ground plane. Another issue addressed here that might be helpful for future work in measurements is the process of four ports TRL calibration.
CHAPTER 2 - BALUN BACKGROUND

2.1 Introduction

A balun (balanced-to-unbalanced) is a transformer used to connect balanced transmission line circuits to unbalanced transmission line circuits. Coaxial cable, microstrip and CPW lines are examples of unbalanced transmission lines, while a two wire transmission line is an example of balanced transmission line as shown in Figure 2.1 and 2.2 (a,b). Two conductors of the same geometry having equal potential with 180-degree phase difference constitute a balanced line; when this condition is not satisfied the transmission line is termed as unbalanced as shown in Figure 2.2 (a, b) (i.e, \( I_1 \neq I_2 \) and \( I_g \neq 0 \)) in this case \( I_g \) is finite and flows through the outer side of the grounded shield, since there is also potential voltage at the outer conductor.

![Figure 2. 1 - Balanced two wire transmission line, \(|I_1| = |I_2|.|](image)

Figure 2. 1 - Balanced two wire transmission line, \(|I_1| = |I_2|.|}
Figure 2.2 - Unbalanced transmission lines (a) coaxial cable and (b) microstrip.

Baluns are required for such circuits as balanced mixers, push pull amplifiers, balanced frequency multipliers, phase shifters, balanced modulators, and dipole antenna feeds. Several different kinds of balun structures have been developed, such as the coaxial balun, lumped-element balun and the Marchand Balun. In this work, the focus was on building a planar Marchand balun utilizing a multi-layer spiral coupler structure to feed a 6-GHZ dipole antenna.

In this chapter an overview of Marchand balun (one of the most commonly used broadband balun) will be presented. The overview will include the operation of Marchand balun and a study of its impedance analysis. An analysis of a monolithic planar version of the Marchand balun and some techniques used to enhance its performance over wide frequency range will also be discussed. A CPW multilayer spiral transmission-line balun will be presented at the end of this chapter.
2.2 Marchand Balun

The Marchand balun is one of the most commonly used components in broadband balanced circuit design. As compared with other baluns, the Marchand balun structure when implemented using couplers will have less strict requirements for $Z_{oe}$ (even mode characteristic impedance) \cite{1}. To obtain a balun with good performance it is sufficient to have $Z_{oe} \approx 3$ to 5 times larger than $Z_{oo}$ (odd mode characteristic impedance) \cite{1}. A wide bandwidth balun can be obtained by proper selection of the balun parameters. Figure 2.3 shows the original Marchand balun \cite{2}, which consists of an unbalanced, an open-circuited, and two short-circuited and balanced transmission line sections.

![Coaxial Cross section of compensated Marchand balun.](image)

Figure 2.3 - Coaxial Cross section of compensated Marchand balun.
In the Marchand balun topology, each section of transmission line is about a quarter-wavelength long at the center frequency of operation. The left-hand line (unbalanced) has the characteristic impedance of $Z_1$. The second line, which has a characteristic impedance of $Z_2$, is open circuited. The outer conductor of these transmission line sections combined with the housing make another two short-circuited $\lambda/4$ lines that are in series with each other and shunt the balanced line. The balanced line has characteristic impedance of $Z_B$ at locations a and b. The stubs $Z_{s1}$ and $Z_{s2}$ are in series and shunt the balanced lines.

A coaxial version [3] of a compensated Marchand balun is shown in Figure 2.4. The device is composed of two lengths of coaxial transmission line, $a$ and $b$. $Z_a$ and $Z_b$ represent the characteristic impedance of lines $a$ and $b$, respectively. $Z_{ab}$ is the characteristic impedance of the balanced transmission line $ab$ composed of the outer conductors of transmission lines $a$ and $b$.

Figure 2.4 - Wide band coaxial Marchand balun.

The external unbalanced source (or load) to the balun is located at $P$, while the terminals $O$ and $O'$ are the points of attachment of the balanced load (or source). $Z_L$ is the
external impedance which may be connected to the balun at $OO'$. Center conductors of lines $a$ and $b$ are connected at $C$ and $C'$, while outer conductors of $a$ and $b$ are connected at $D$.

Figure 2.5 shows an equivalent circuit of the balun shown in Figure 2.4 for the purpose of calculating the impedance. The terminals $P$, $O$, and $O'$ are the same as in Figure 2.4. $M$ is the impedance looking into coaxial line $b$ toward the open circuit. $N$ is the impedance looking from $OO'$ along the open transmission line $ab$ toward the short circuit at $d$.

If the transmission line losses are neglected,

$$M = -jZ_b \cot \theta_b \quad \cdots \quad 1$$

$$N = jZ_{ab} \tan \theta_{ab} \quad \cdots \quad 2$$

Where $\theta_b$ and $\theta_{ab}$ are, respectively, the electrical lengths of transmission lines $b$ and $ab$, taking into account their respective physical lengths and velocities of propagation. From the equivalent circuit:
\[ Z_{in} = \frac{Z_L N}{Z_L + N} + M \quad \ldots 3 \]

\[ Z_{in} = \frac{jZ_L Z_{ab} \tan \theta_{ab}}{Z_L + jZ_{ab} \tan \theta_{ab}} - jZ_b \cot \theta_b \quad \ldots 4 \]

\[
Z_{in} = \frac{Z_L}{Z_{L}^2 + \frac{Z_{ab}^2 \tan^2 \theta_{ab}}{Z_{ab}^2}} + 1 + \frac{jZ_L^2 Z_{ab} \tan \theta_{ab}}{Z_L^2 + Z_{ab}^2 \tan^2 \theta_{ab}} - jZ_b \cot \theta_b \quad \ldots 5
\]

If the electrical lengths of line segments \( b \) and \( ab \) are equal (\( \theta_b = \theta_{ab} = \theta \)), and the characteristic impedance \( Z_{ab} = Z_L \) and \( Z_b = Z_a = S = Z_o \), then

\[ Z_{in} = Z_L (\sin \theta)^2 + j(Z_L (\sin \theta)^2 - S) \cot \theta \quad \ldots 6 \]

When \( \theta = 90^\circ \) (\( \cot \theta = 0 \) and \( \sin \theta = 1 \)) the reactive component of \( Z_{in} \) is zero and \( Z_{in} \) equals \( Z_L \). While when \( \sin^2 \theta = \frac{S}{Z_L} \), \( Z_{in} \) equal to \( S \) (\( S = Z_o \)). As the lengths of the transmission lines approach \( \frac{\lambda}{4} \), i.e., \( \theta \) goes to \( \frac{\pi}{2} \) the impedance of the short circuit stub becomes larger while the open stub input impedance gets smaller, and hence the input impedance \( Z_{in} \) converges to \( Z_L \). Note that the operating frequency range of the balun is centered about its resonant frequency. When frequencies are slightly off the center frequency, the short stub impedance (\( N \)) mainly determines \( Z_{in} \) since the value of the open circuit stub impedance (\( M \)) is small. Thus larger values of \( Z_{ab} \) in Figure 2.4 make the balun more insensitive to frequency.
2.3 Monolithic Planar Marchand Balun

The concept of a coupled coaxial line balun is applied to the monolithic planar structures shown in Figure 2.6. This is a direct mapping of the coaxial balun with one stripline corresponding to the inner conductor of the coaxial line and the other stripline corresponding to the shield. A ground plane in the transmission line circuits provides the ground reference at $d$ in the coaxial balun.

As can be seen in Figure 2.6 the Marchand balun consists of two sets of coupled lines with each being $\lambda/4$ long at the center frequency of operation. The coupled lines are either side coupled or broadside coupled lines if a tighter coupling is required.

![Diagram of Monolithic planar Marchand balun](image)

Figure 2.6 - Monolithic planar Marchand balun.

The first coupler has one of its ports connected to ground and the other port is connected to the input impedance. This coupler serves as part of the input transmission line with characteristic impedance equal to $Z_a$. The second coupler has one of its ports open-circuited and other one is connected to ground. This coupler serves as an open
transmission line with characteristic impedance equal to $Z_b$. The two strip lines connected to the load impedance serve as a short-circuit stub (Figure 2.7).

![Figure 2.7 - The bottom two strip lines serve as a short-circuit stub.](image)

In an inhomogeneous medium, where there is partly air and partly dielectric such as in the case of microstrip and CPW lines, the odd- and even-mode phase velocities are unequal [4]. The even-mode effective dielectric constant is higher than the odd-mode effective dielectric constant because the former has less fringing field in the air region. The result is a lower phase velocity ($v = \frac{c}{\sqrt{\varepsilon_r}}$) in the case of the even-mode than in the case of the odd-mode.

In the case of a directional coupler, the phase velocity inequality will result in poor directivity [5]. Directivity is a measure of the coupler’s ability to isolate forward and backward waves, and its directivity performance becomes worse as the coupling is decreased or as the dielectric permittivity is increased. When analyzing the Marchand balun as two coupled line sections in cascade, the isolated ports of the couplers will actually be the two unbalanced ports of the balun. Therefore, any of the undesired effects
due to even and odd mode phase velocity inequities contribute to the balun’s amplitude and phase performance.

2.4 Enhanced Marchand Balun

Two compensation techniques that are used to improve the performance of baluns in an inhomogeneous medium will be presented. The two techniques proved to provide compact wide–band baluns.

2.4.1 Compensated Coupled Lines

Compensated coupled lines [6] can be used to design the enhanced Marchand balun by employing capacitors at each end of the coupled lines, as shown in Figure 2.8.

![Figure 2.8 - Capacitive-compensated Marchand balun.](image)
The added capacitors will not affect the even-mode (since the polarity on both strip lines comprising the coupler is the same) but effectively decrease the odd-mode phase velocity. This equalization of even and odd mode phase velocity will increase the directivity and thus provide broadband characteristics with good isolation. The compensation capacitor is given by [3]

$$C = \frac{1}{4\pi f_0 Z_{oo} \tan \vartheta_0}$$  ... 7

Where, \( \vartheta_0 = \frac{\pi \sqrt{\varepsilon_{effo}}}{2 \sqrt{\varepsilon_{effe}}} \), and \( f_0 = \) frequency of operation, \( Z_{oo} = \) odd-mode characteristic impedance, \( \vartheta_0 = \) odd-mode electrical length of the coupled section.

2.4.2 Short Transmission Line Between the Couplers

Compensation can also be accomplished by interconnecting a short transmission line (Figure 2.9) to a pair of couplers. The variation in the amplitude and phase of the transmission line will compensate for the amplitude and phase difference of the balun that is caused by the difference in phase velocities.

![Figure 2.9 - Compensated Marchand balun using a short transmission line between the two couplers.](image)
The short transmission line can be approximated as a capacitor connected between the ground plane (CPW or microstrip) and the connection point of both couplers (Figure 2.10). The existence of this capacitor will result in a virtual decrease in the odd-mode phase velocity of the coupler. This technique compensates for the amplitude and phase differences of the Marchand balun by creating a circuit that generates differences in electrical field opposite to those of the balun [7] and thus resulting in cancellation of the amplitude and phase differences of the balun.

![Compensated Marchand balun using a capacitor to ground between the two couplers.](image)

Figure 2.10 - Compensated Marchand balun using a capacitor to ground between the two couplers.
2.5 Parallel Connected Marchand Balun

In order to design a broadband Marchand Balun, the balun should have small odd mode characteristic impedance, which means large coupling capacitance between individual lines. \( Z_{oo} \propto \frac{1}{C_0} \), where \( C_0 \) is the coupling capacitance [8]. However, it is difficult to realize a large coupling capacitance because it requires very narrowly-spaced coupled lines. Two Marchand Baluns can be connected in parallel [8] to solve this problem (Figure 2.11). The parallel connection of the two baluns will decrease the odd and even characteristic impedances by 50% and thus provide the required small odd mode characteristic impedance.

![Figure 2.11 - Parallel-connected Marchand balun.](image-url)
2.6 Multilayer Spiral Transmission-Line Balun

A Marchand balun can be made of coupled microstrip lines, Lange couplers, CPW line or spiral coils. Because the length of each line should be equal to one-quarter wavelength, Marchand baluns of the microstrip or CPW line types operating at low gigahertz frequencies are relatively large (typically several centimeters), making them difficult to be integrated.

Coiling the transmission lines into a spiral configuration leads to several advantages. Turning the lines back on themselves results in an increase in the mutual capacitance and inductance between the lines, as well as between individual segments within the same line. Therefore, for the same length of conductor, the resonant frequencies of the spiral devices are significantly lower than those for simple straight lines. Thus the spiral type has the advantage of being more compact with shorter metal length for a given operating frequency, which means less metal loss. Figure 2.12 shows a drawing of a two-layer balun designed in this work; one of the spiral pairs is rotated by 180 degrees with respect to the other. The layout ensures that the directions of current flow in the two coils are the same, further enhancing the mutual inductive coupling. A coplanar ring surrounding the structure provides the ground for the device. Notice that the layout of the spiral coils in the design is symmetric to the ground connections, in order to minimize imbalance between the outputs due to asymmetric connections to ground. Figure 2.13 and 2.14 show a cross section of the multilayer spiral balun.
Figure 2.12 - (6-GHZ) multi-layer Marchand spiral balun.

Figure 2.13 - Cross section of multilayer spiral balun (2D).
Figure 2.14 - Cross section of a multi-layer CPW spiral balun. Notice that the input port and the output port were omitted from the drawing for simplicity.
Summary

This chapter defined the general problem tackled in the research, that is the symmetric resonant dipole antenna requires a balanced feed line that can’t be achieved with a direct connection between the dipole antenna and the regular coaxial transmission line, and how this problem can be solved using a device called balun (balanced-to-unbalanced). The work in this chapter gave a description of one of the commonly used types of baluns, the Marchand Balun. Next it described three of the techniques used to enhance the performance of the Marchand Balun. At the end of this chapter we presented the multilayer CPW (coplanar wave guide) spiral Marchand balun.
CHAPTER 3 – DESIGN AND ANALYSES OF A MULTILAYER CPW SPIRAL BALUN

3.1 Introduction

In the previous chapter we have introduced the multilayer spiral CPW balun (Figure 3.1), which consists of two spiral couplers connected together using an air bridge. Each coupler consists of two spirals on top of each other in order to obtain tight coupling (Figure 3.2).

Figure 3. 1 - Multi-layer Marchand spiral balun.
In this chapter we will present the design procedure of the multi-layer CPW spiral balun. Furthermore, three compensation techniques to enhance the balun performance will be presented. Those techniques are: introducing a ground plane below the air bridge and between the two couplers; increasing the capacitance between the air bridge and the spiral; and varying the capacitance value at the open port.
3.2 Design Procedure of the Multilayer CPW Spiral Balun

The required coupling factor \( k \) [9] for optimum balun performance can be found from the following equation:

\[
k = \frac{1}{\sqrt{\frac{2Z_L L}{Z_0} + 1}}
\]

For a 50 \( \Omega \) single-ended load impedance and 50 \( \Omega \) input impedance, \( k \) is -4.75dB.

The vertical distance between the two stacked spirals, as well as the horizontal center-to-center distance between the spirals, are the main parameters to determine the required amount of coupling. From the simulation for one spiral coupler, the required parameters (vertical and horizontal offset) can be determined (Figure 3.3).

Figure 3.3 - Broadside spiral coupler, in the drawing the bottom spiral is horizontally offset to get the required coupling factor.
From equation (2.4) (previous chapter):

$$Z_{ab} = \frac{jZ_L Z_{ab} \tan \theta_{ab}}{Z_L + jZ_{ab} \tan \theta_{ab}} - jZ_b \cot \theta_b$$

The value of $Z_{ab}$ should be very large to ensure a broadband balun. Looking from the two output ports (Figure 3.4) $Z_{ab}$ can be considered, as the characteristic impedance exists between the lower two spirals. It should be mentioned here that the definition of the characteristic impedance doesn’t apply completely for this case, since the two spirals are not exactly a real transmission line.

![Figure 3.4 - Equivalent representation of the lower metal spirals.](image)

$$Z_{ab} \propto \sqrt{\frac{L_{12}}{C_{12}}}$$

where $C_{12}$ and $L_{12}$ are the mutual capacitance per unit length and mutual inductance per unit length, respectively, between the two spirals. In order to get a high $Z_{ab}$ value, $C_{12}$ should be small and $L_{12}$ should be high. To increase $Z_{ab}$ the distance between the two spirals should be large enough to decrease the value of $C_{12}$ and the direction of the
currents in the two spirals should be in the same direction to ensure large mutual inductance between the two bottom spirals.

3.3 Compensation Techniques

3.3.1 Ground plane below the air bridge and between the two couplers

When introducing a ground plane between the two couplers, the performance of the balun was improved compared to its performance without the ground plane. One valid explanation for the better performance is that the ground plane, along with the air-bridge that connects the two couplers and passes over the ground plane, created a capacitance to ground between the two couplers (Figure 3.5), and as was explained in the previous chapter, adding a capacitance to ground between the two couplers is one of the techniques used to enhance the performance of the balun.

![Figure 3.5 - Cross section of a multi-layer CPW spiral balun. Notice the area between the ground plane and the air bridge.](image)
The capacitance to ground between the two couplers decreases the amplitude and phase difference over the operational frequency range with only a small change in the magnitude of $S_{11}$. When increasing the value of this capacitance, the frequency band where the acceptable output amplitude difference occurs (typically ~0.5 dB difference between $S_{21}$ and $S_{31}$) will be shifted to a lower frequency range, and the BW will become smaller. Varying the value of this capacitance is thus a useful tuning mechanism, and does not induce unwanted degradation in the return loss characteristics (Figure 3.6). In the following three sub–sections, the values of the parameters in the capacitance equation (area, height and permittivity) were varied in order to study their effect on the balun performance.
3.3.1.1 Changing the Ground Width and Air Bridge Width

Varying the width of the ground plane between the two couplers or the air bridge width is one of the methods that could be used to vary the value of the capacitance to ground. Increasing the capacitance to ground between the two couplers (by increasing either the ground width or the air bridge width) will decreases the amplitude and phase difference over the operational frequency range with only a small change in the magnitude of $S_{11}$ (Figure 3.6), also the frequency band where the acceptable output amplitude difference occurs (typically $\sim$0.5 dB difference between $S_{21}$ and $S_{31}$) will be shifted to a lower frequency range (Figure 3.6).

Figure 3.6 - Varying the value of the capacitance by varying the ground plane width or the air bridge width to get the desired performance. $S_{11}$ (solid lines) and $S_{21}$ (dashed lines).
As can be seen in Figure 3.7, by varying the value of $C$ from 0.07 pF to 0.1 pF, the frequency range where $|\text{dB} (S_{21}) - \text{dB} (S_{31})|$ is below 0.5dB shifts to a lower frequency range with some enhancement on the balun phase difference and the amplitude imbalance. A larger value of the capacitance will shrink both the amplitude and the phase bandwidth and will tend to degrade the phase performance (Figures 3.9 to 3.10). From the shown figures the value of $C$ at which the balun has the best amplitude imbalance and phase difference occurred at $C$ equal 0.1 pF. Other values of $C$ could also be chosen depending on the balun bandwidth requirements.

Figure 3.7 – Changing the amplitude imbalance by varying the ground plane width or the air bridge width (increasing $C$ from 0.07pF-to-0.1 pF).
Figure 3.8 - Changing the phase difference by varying the ground plane width or the air bridge width (increasing the value of $C$ from 0.1 pF to 0.08 pF).
Figure 3.9 – Changing the amplitude imbalance by varying the ground plane width or the air bridge width (increasing the value of $C$ from 0.14 pF to 0.18 pF.)
Figure 3. 10 - Changing the phase difference by varying the ground plane width or the air bridge width (increasing the value of $C$ from 0.14 pF to 0.18 pF).
3.3.1.2 Changing air bridge height

To study the effect of the middle capacitor, three similar baluns with different air bridge heights ($H = 2\mu$m, $H = 10\mu$m and $H = 25\mu$m) were simulated. The performance when using the highest air bridge balun (smallest capacitance value) was the worst, as can be seen in Figures 3.11 and 3.12.

![Graph showing changes in air bridge height](image)

Figure 3.11 - Shifting the amplitude imbalance by decreasing the height of the air bridge. For $H = 2\mu$m, $10\mu$m and $25\mu$m the BW is 3.63, 4.77 and 1.7 GHz, respectively.
Figure 3. 12 - Shifting the phase difference by decreasing the height of the air bridge.
3.3.1.3 Insulator permittivity

The permittivity constant (\(\varepsilon_r\)) of the material (polyimide 2, see Figure 3.13) that exists between the air bridge and the ground plane was varied to study its effect on the balun performance. A higher value of \(\varepsilon_r\) will increase the capacitance between the two couplers and shift the phase difference and the amplitude imbalance to a lower frequency range (Figures 3.14 and 3.15). However increasing the value of \(\varepsilon_r\) reduced the return loss BW (Figure 3.16) while the increases in the other capacitance-related parameters (see section 3.3.1.2 and 3.3.1.1) did not have same effect. An explanation for this difference is that changing the insulator material will affect not only the capacitance between the couplers but also the characteristics of the CPW lines comprising the balun, as the insulator material is uniformly deposited over the balun geometry.

![Figure 3.13 - Cross section of the multilayer spiral balun. Notice that polyimide 2 in the graph is uniformly deposited over the balun geometry since this balun was simulated using Momentum (ADS) that work better with uniform material depositing.](image)
Figure 3.14 - Shifting the amplitude imbalance by increasing the value of $\varepsilon_r$. 
Figure 3.15 - Shifting the phase difference by increasing the value of $\varepsilon_r$. 
Figure 3.16 - The change in the return loss by increasing the value of $\varepsilon_r$. 

Legend Title
- EpsonR = 1
- EpsonR = 8
- EpsonR = 5
- EpsonR = 2.6
3.3.2 The Capacitance between the air bridge and the spiral

A large capacitance between the air bridge and the spiral strips ($C_b$) (Figure 3.17) was used to compensate for the difference in the odd and the even mode phase velocities. Figure 3.18 presents a comparison between small capacitance and large capacitance values. The result was better for the case of large capacitance.

![Figure 3.17 - Balun implementing side capacitance techniques for compensation.](image-url)
Figure 3. 18 - Phase difference and amplitude imbalance comparison of the balun when varying the value of the capacitance exists between the air bridge and the spiral strips. Phase difference (dotted lines) and amplitude imbalance (solid lines).
3.3.3 Open circuit capacitance

A capacitance to ground at the balun open port (Figure 3.19) can be used to adjust the amplitude BW of the balun. Increasing the value of the capacitance will shift the amplitude BW to a lower frequency band, while the capacitance has less effect on the phase difference (Figure 3.20 and 3.21).

![Figure 3.19 - Implementing capacitance (C_{open}) at the output port to tune the balun.](image-url)
Figure 3.20 - Shifting the amplitude imbalance by increasing the value of the capacitor $C_{\text{open}}$. Notice that the amplitude imbalance at $C_{\text{open}} = 85 \text{ fF}$ has the best BW performance.
Figure 3.21 - Phase difference of the balun. Notice that $C_{open}$ has less effect on the phase difference.
3.4 Design Specification

Table 3.1 - Design parameters of the small and large size baluns, see Figure 3.22 – Figure 3.24 for visualization.

<table>
<thead>
<tr>
<th>Balun Type</th>
<th>Small size balun</th>
<th>Large size balun</th>
</tr>
</thead>
<tbody>
<tr>
<td>Balun size (mm²)</td>
<td>1.4 x 2.4</td>
<td>3 x 5.6</td>
</tr>
<tr>
<td>Air Bridge width (µm)</td>
<td>100</td>
<td>300</td>
</tr>
<tr>
<td>Ground width (µm)</td>
<td>200</td>
<td>500</td>
</tr>
<tr>
<td>Spiral strip line width (µm)</td>
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<td>100</td>
</tr>
<tr>
<td>Spiral gap width (µm)</td>
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<td>100</td>
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<tr>
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<td>340</td>
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<td>635</td>
</tr>
<tr>
<td>Horizontal offset (µm)</td>
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<td>150</td>
</tr>
<tr>
<td>Vertical offset (polyimide 1 thickness) (µm)</td>
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<td>20</td>
</tr>
<tr>
<td>Polyimide 2 thickness (µm)</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>Polyimide 1 and 2 permittivity</td>
<td>2.6</td>
<td>2.6</td>
</tr>
<tr>
<td>Quartz permittivity</td>
<td>3.8</td>
<td>3.8</td>
</tr>
<tr>
<td>Quartz thickness (mil)</td>
<td>25</td>
<td>25</td>
</tr>
<tr>
<td>Spiral line length λ/4 (mm)</td>
<td>6.7</td>
<td>7.0</td>
</tr>
</tbody>
</table>

Figure 3.22 - (6 GHz) multilayer CPW spiral Marchand balun.
Figure 3.23 - Spiral line coupler.

Figure 3.24 - Spiral line part of the spiral coupler.
Summary

In this chapter the design procedure of a multilayer CPW spiral Marchand balun and three techniques used to enhance the balun performance were presented. The first technique makes use of the capacitance to ground introduced between the ground plane and the short interconnect transmission line that is used to connect between the two couplers. The second technique makes use of the capacitance between the same short interconnect transmission line and the spiral strip lines. The last technique involves introducing an extra capacitance to ground at the open port.

The capacitance to ground provided by the short interconnect transmission line between the two couplers and the ground plane was shown to decrease the phase difference and the amplitude imbalance between the two balun’s output ports, one valid explanation for that enhancement is that the capacitance to ground equalized the phase velocity difference between the odd and the even mode and thus increased the balun bandwidth. To study the effect of this capacitance on the balun performance, the values of the parameters in the capacitance equation (area, height and permittivity) were varied. The study showed that by increasing the capacitance value the operational bandwidth of the balun was shifted to a lower frequency range and the amplitude and the phase difference over the operational frequency were minimized with only a small change in the magnitude of $S_{11}$ (Return Loss). However, A larger value of the capacitance to ground will shrink both the amplitude and the phase bandwidth and will tend to degrade the phase performance.

The other compensation technique, changing the capacitance between the short interconnection (Air bridge) and spiral strips proved to have good enhancement on the
balun performance where the amplitude imbalance bandwidth was doubled by increasing
the capacitance value for example from 16fF to 50 fF. Also increasing the capacitance to
ground at the balun open port proved to be a good technique to tune the amplitude
bandwidth with just small impact on the balun phase difference.
CHAPTER 4 – EQUIVALENT MULTILAYER CPW SPIRAL BALUN MODEL

4.1 Introduction

Modeling a microwave device is a useful technique for the design engineer for its ease in studying the performance of the device under design without the need to run full wave simulation for each design iteration. In this chapter two models that can be use in studying the effect on the balun performance due to varying the width of the air bridge, varying the ground plane width between the two couplers and increasing the capacitance at the open port were derived. In the first model the two couplers that comprise the balun were studied by analyzing the two couplers individually using a full wave simulation software (Momentum) and then combining the results in a circuit simulator. In the second model, a lumped elements model for each coupler was derived through circuit optimization against the full wave simulation results obtained using Momentum (ADS) and then the two models were combined in one single model that represent the balun.
4.2 Equivalent Split model for the Multilayer spiral Balun

The two couplers that comprise the balun were analyzed individually and the derived [S] parameters were combined in a circuit simulator. The representation developed from this approach will be termed the ‘split model’. Figure 4.1 shows a diagram of the full balun, Figure 4.2 shows the two separated couplers, and Figure 4.3 shows the circuit representation of the split model.

Figure 4.1 - Full balun diagram.

Figure 4.2 - The two Separated Couplers (a) left coupler (b) right coupler. The arrows point to the CPW transmission line added to enable proper simulation of the circuits in Momentum; these lines were de-embedded when circuit simulations were performed.
Figure 4.3 - Split model representation of the Marchand balun.

One of the main purposes behind deriving the split model was to study the effect of the capacitance to ground between the two couplers without the need to run the full wave simulation for each changes in the capacitance parameters. Hence a variable capacitor to ground $C$ was added at the connection point between the two couplers (Figure 4.4), the results shown in Figure 4.5, 4.6 and 4.7 show that the performance of the split model matched the full-wave simulation results of the balun when $C$ is equal to zero (no ground plane bellow the air bridge in the balun circuit). However, increasing the line width of the air-bridge from 25µ to 100µ above the ground plane, which corresponds to adding a 50 fF capacitance to ground at the connection point between the two couplers in the split model didn’t achieve the same result even so it enhanced the balun performance.
The reason for the difference in the performance between the full-wave results and the split model is due to the fact that increasing the air-bridge width will also introduce additional capacitance between the air bridge and the spiral strips and as was shown in section 3.3.2 this type of capacitance served as a compensation technique. Simply changing the capacitance value in the split model didn’t not introduce the same effect.

Figure 4.4 - Split model representation of the Marchand balun for the purpose of studying the effect of the capacitance to ground between the two couplers. The de-embed boxes negated the CPW lines denoted in Figure 4.2.
Figure 4.5 Similar phase difference performance between the split model and the full-wave simulation result for $C=0$ pF.
Figure 4.6 Similar amplitude imbalance performance between the split model and the full-wave simulation result for $C=0$ pF.
Figure 4. 7 Similar return loss performance between the split model and the full-wave balun simulation result.
Figure 4.8 Comparison in the amplitude imbalance performance between the split model and the full-wave balun simulation result when adding a 0.05 pF capacitor to ground between the two couplers.
To get an optimum capacitance value, the capacitance value in the split model can be tuned using the circuit schematic in ADS. The performances of the balun return loss, phase difference and amplitude imbalance can be derived as a function of the capacitance value as shown in Figure 4.9, Figure 4.10 and Figure 4.11.

Figure 4.9 The change of amplitude imbalance performance by varying the value of $C$ from 0.05 pF to 0.2 pF.
Figure 4.10 The change of phase difference performance by varying the value of $C$ from 0.05 pF to 0.2 pF.
Figure 4.11 The change in the magnitude of $\text{dB}|S_{11}|$ by varying the value of $C$ from 0.05 pF to 0.2 pF.

As can be seen in figure 4.9 and 4.10 there is a point at which the amplitude imbalances and phase difference are optimum. Notice that $S_{11}$ (Figure 4.11) is less sensitive than the phase difference and the amplitude imbalance to the change in the value of $C$. This fact can be used to adjust the center frequency by shifting the frequency range where the amplitude imbalance ($|\text{dB} (S_{21}) - \text{dB} (S_{31})|$) is less than 0.5 dB and the phase difference ($|\text{Phase} (S_{21}) - \text{Phase} (S_{31})|$) is $180 +/− 5$ degrees or less to the range where $\text{dB}|S_{11}|$ is minimum.
4.3 Equivalent Lumped Elements Model for the Multilayer Spiral Balun

A lumped element model that represents the performance of the multilayer CPW spiral balun over the desired frequency range (1-8 GHz) was developed and characterized. The lumped element model is an important tool in understanding different parameters that comprise the balun. The lumped element model can also help analyzing the effect of previously discussed compensation techniques such as varying the capacitance to ground at the point between the two couplers, varying the air bridge width over the spiral strips and adding extra capacitance to ground at the open port of the balun.

Since the balun consists of two couplers, a lumped element model was developed for each. The first model represents the left hand side coupler that has three ports (Figure 4.12 and 4.14). The second model represents the right hand side coupler (Figure 4.13 and 4.15) that has two ports. Figure 4.16 shows the balun equivalent lumped element model after connecting the two couplers models together; an ideal capacitor was added between the connection point between the two models and ground to represent the capacitance to ground resulted from the air bridge and the ground plane.

The lumped element model for a one side spiral coupler consists of seven broadside couplers each one representing a half turn in the real spiral couplers, The capacitors ($C_h$) represent the capacitance between half portion of the air bridge that connect between the two couplers and each strip of the upper spiral below that air bridge. The capacitors ($C_a$) represent the capacitance between each strip of the upper spiral and the air bridge that connects the bottom spiral to the output port. $C_{open}$ represents the capacitance to ground at the open port of the balun. Table 4.1 shows the values of the equivalent model optimized-parameters.
Figure 4.12 - Left side coupler that is part of the balun.

Figure 4.13 - Right side coupler that is part of the balun.
Figure 4. 14 - Equivalent circuit model of the left hand side coupler. The capacitors ($C_b$) represent the capacitance between the right air bridge and every strip of the upper spiral that passes below that air bridge. The capacitors ($C_a$) represent the capacitance between the strip lines of the upper spiral and the left air bridge that connects the bottom spiral to the output port.
Figure 4. 15 - Equivalent circuit modal of the right hand side coupler. Notice that the capacitors \((C_b)\) represent the capacitance resulted between the left air bridge and every single strip of the upper spiral that pass below that air bridge, also notice that the capacitors \((C_a)\) represent the capacitance between every single strip of the upper spiral and the right air bridge that connects the bottom spiral to the output port, the capacitance \(C_{\text{open}}\) represent the equivalent open circuit capacitance.
Figure 4.16 - The full balun equivalent circuit model consisting of the two optimized coupler models.

Table 4.1 - The optimized parameter values of the equivalent model.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_a$ ($C_{a_1} = C_{a_2} = C_{a_3} =$)</td>
<td>15.0328 fF</td>
</tr>
<tr>
<td>$C_b$ ($C_{b_1} = C_{b_2} = C_{b_3} =$)</td>
<td>15.0 fF</td>
</tr>
<tr>
<td>$C_{open}$</td>
<td>30.0 fF</td>
</tr>
<tr>
<td>$C$</td>
<td>50 fF</td>
</tr>
<tr>
<td>$Z_{oo}$</td>
<td>45.6124 Ω</td>
</tr>
<tr>
<td>$Z_{oe}$</td>
<td>477.901 Ω</td>
</tr>
<tr>
<td>$\varepsilon_{oo}$</td>
<td>3.79986</td>
</tr>
<tr>
<td>$\varepsilon_{oe}$</td>
<td>2.39395</td>
</tr>
<tr>
<td>$L$ (length of each coupler)</td>
<td>.855 mm</td>
</tr>
<tr>
<td>$Attenuation_{even}$</td>
<td>0.0251977</td>
</tr>
<tr>
<td>$Attenuation_{odd}$</td>
<td>68.6089</td>
</tr>
</tbody>
</table>
When applying the same capacitance to ground between the two couplers for both the equivalent model and the split model the results of both models matched very well with each other. When applying the corresponding capacitance value to the full balun circuit its full-wave simulation results also matched with the two models results. Figures 4.17 – 4.21 show comparisons of the value of dB ($S_{11}$), dB ($S_{12}$), dB ($S_{13}$), phase ($S_{12}$) and phase ($S_{13}$) when the value of the capacitance to ground equal .05 pF. The physical dimensions of the full small size balun are presented in Table 3.1. Notice from the result that at high frequency range both models result don’t match completely with the full-wave simulation result, since lumped element models usually work better at low frequency.
Figure 4. 17 – Comparisons of the value of dB ($S_{11}$), equivalent model versus full balun and split model.
Figure 4.18 – Comparisons of the value of dB ($S_{12}$), equivalent model versus full balun and split model.
Figure 4.19 – Comparisons of the value of $\text{dB} (S_{13})$, equivalent model versus full balun and split model.
Figure 4.20 – Comparisons of the value of Phase ($S_{12}$), equivalent model versus full balun and split model.
Figure 4.21 – Comparisons of the value of Phase ($S_{13}$), equivalent model versus full balun and split model.
Summary

In this chapter two multilayer CPW spiral Marchand balun models were presented. The two models helped in analyzing the balun physical parameters and in studying the effect of introducing different lumped elements on the balun circuit to derive the techniques required to compensate between the difference in the phase velocity of the odd mode and the even mode. In the first model (split model) the two couplers that comprise the balun were studied by analyzing the two couplers individually using a full wave simulation software (Momentum) and then combining the results in a circuit simulator, this model was used to study the effect of the capacitance to ground (exist between the two couplers) on the balun performance. The study showed good agreement between the model (split model) results and results obtained using the full wave simulation (Momentum). In the second model (equivalent model) the lumped element model for each coupler were combined in one lumped element model. Each coupler model was derived through circuit optimization against the full wave simulation results obtained using Momentum (ADS). The equivalent model result were compared with the results obtained from both the full-wave simulation (Momentum) and the split model, the comparisons showed good agreement.
CHAPTER 5 – MEASUREMENT RESULTS

5.1 Introduction

In order to verify the conclusion we draw out from the simulation results certain balun circuits have been manufactured by American Technical Ceramics (ATC). Four ports measurements were performed on the balun circuits. The results showed close agreements between the predicted balun performance and the measurement results. The following discussion includes comparisons between the result obtained from the full-wave EM simulation (Momentum) and the four ports measurement results of the 6 GHz small balun. Four different 6 GHz small-balun circuits designed during this work (labeled from 1 – 4) will be presented in this chapter. For each balun circuits the geometrical dimensions are presented in Appendix A. The four-port TRL calibration and the measurements procedure are presented in Appendix B.

It should be mentioned here that the performed measurements were referenced to 75Ω system, since by mistake we used 75Ω TRL calibration lines, the simulation result were also re-referenced to 75Ω to match the measured results. This affected the return loss of the balun, which should be better around 6 GHz when using 50Ω reference impedance.
5.2 Balun number 1

For a small balun circuit working at 6 GHz (Figure 5.1) that has 304µm ground width and 100µm air bridge width (capacitance to ground between the two couplers equal 70 fF), the comparison results are shown in Figure 5.2 - 5.9.

Figure 5.1 - (6-GHZ) multi-layer spiral Marchand balun.
Figure 5.2 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.3 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data on a different scale.
Figure 5.4 - Phase difference of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.5 - Return loss ($S_{11}$) of the balun circuit, measurements versus full-wave EM simulation data.
Figure 5.6 – dB(S12) of the balun circuit number 1, measurements versus full-wave EM simulation data.
Figure 5.7 – dB(S13) of the balun circuit number 1, measurements versus full-wave EM simulation data.
Figure 5.8 – Phase(S12) of the balun circuit number 1, measurements versus full-wave EM simulation data.
Figure 5.9 – Phase(S13) of the balun circuit number 1, measurements versus full-wave EM simulation data.
5.3 Balun number 2

For a small balun circuit working at 6 GHz (Figure 5.1) that has 200μm ground width and 50μm air bridge width (capacitance to ground between the two couplers equal 23 fF), the comparison results are shown in Figure 5.10 - 5.17.

Figure 5.10 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.11 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data on a different scale.
Figure 5.12 - Phase difference of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.13 - Return loss ($S_{11}$) of the balun circuit, measurements versus full-wave EM simulation data.
Figure 5.14 – $\text{dB}(S_{12})$ of the balun circuit number 2, measurements versus full-wave EM simulation data.
Figure 5. 15 – dB(S13) of the balun circuit number 2, measurements versus full-wave EM simulation data.
Figure 5.16 – Phase(S_{12}) of the balun circuit number 2, measurements versus full-wave EM simulation data.
Figure 5.17 – Phase($S_{13}$) of the balun circuit number 2, measurements versus full-wave EM simulation data.
5.4 Balun number 3

For a small balun circuit working at 6 GHz (Figure 5.18) that has 204µm ground width and 300µm air bridge width (capacitance to ground between the two couplers equal 140 fF), the comparison results are shown in Figure 5.19 - 5.26.

Figure 5.18 - (6-GHZ) multi-layer Marchand spiral balun. Notice the air bridge cross shape above the ground plane.
Figure 5.19 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.20 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data on different scale.
Figure 5.21 - Phase difference of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.22 - Return loss ($S_{11}$) of the balun circuit, measurements versus full-wave EM simulation data.
Figure 5.23 – $\text{dB}(S_{12})$ of the balun circuit number 3, measurements versus full-wave EM simulation data.
Figure 5. 24 – $\text{dB}(S_{13})$ of the balun circuit number 3, measurements versus full-wave EM simulation data.
Figure 5.25 – Phase($S_{12}$) of the balun circuit number 3, measurements versus full-wave EM simulation data.
Figure 5.26 – Phase(S\textsubscript{13}) of the balun circuit number 3, measurements versus full-wave EM simulation data.
5.5 Balun number 4

For a small balun circuit working at 6 GHz (Figure 5.27) that has 204µm ground width and 50µm air bridge width (Capacitance to ground between the two couplers equal 23 fF), but with capacitance between the air bridge and the spiral strips equal to 16fF at each side. The comparison results are shown in Figure 5.28 - 5.34.

Figure 5. 27 - Balun implementing side capacitance techniques for compensation.
Figure 5.28 - Amplitude imbalance of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5.29 - Phase difference of a balun circuit, measurements versus full-wave EM simulation data.
Figure 5. 30 - Return loss ($S_{11}$) of the balun circuit, measurements versus full-wave EM simulation data.
Figure 5. $\text{dB}(S_{12})$ of the balun circuit number 4, measurements versus full-wave EM simulation data.
Figure 5.32 – dB(S_{13}) of the balun circuit number 4, measurements versus full-wave EM simulation data.
Figure 5.33 – Phase($S_{12}$) of the balun circuit number 4, measurements versus full-wave EM simulation data.
Figure 5.34 — Phase(S_{13}) of the balun circuit number 4, measurements versus full-wave EM simulation data.
5.6 Balun Performance.

A previous paper on Marchand baluns by Yeong J, Yoon and Robert C. Frye reported a small size CPW multilayer spiral fabricated on silicon with good performance - RL (return loss) better than 17dB in the frequency range from 1.6 GHz – 4.1 GHz, and better than 39dB RL around 1.9 GHz. The amplitude imbalance was less than 0.3 dB through the whole band, and the phase difference was less than 2° over the frequency range 1.6-2.6 GHz. The overall size is 1.2 x 2.3 mm² [10]. The same authors reported a similar topology spiral balun, fabricated on glass and high resistivity silicon [11]. The frequency response was similar in both cases since the mutual coupling capacitance occurred in the insulator between the spirals of the couplers. The ratio of the conductor width to pitch (the conductor strip width plus the width of the gap between two adjacent spiral strips) fell between 0.4-0.6 to get good return loss and good amplitude and phase difference for a 50Ω input impedance and a 50Ω load impedance. The center frequency of the 3dB BW ranged from 1.2 to 3.5 GHz, while the relative BW normalized by the center frequency was typically ~ 1.48. Return loose (RL) at the center frequency in all cases was in the range of 13-18 dB. Amplitude imbalance within the OBW (operational bandwidth) was less than 0.9dB for the glass substrate balun and less than 0.5dB for the silicon substrate balun.

In [12] a balun built using a microstrip transmission line and a short transmission line between the two couplers to compensate for the phase and amplitude imbalance was described. They reported a 2-6 GHz bandwidth with amplitude imbalance of 0.2dB and phase difference of 10°, and an overall size equal to 0.7 x 1.5 mm².
We developed a new CPW multilayer spiral Marchand balun based on the techniques in [10, 12]. Both the simulation and measurement results showed a 75% bandwidth around the center frequency of 5.5 GHz with return loss better than 14dB. The amplitude and phase difference are less than 0.5dB and 3°, respectively. Overall balun size is 1.4 x 2.4 mm² the balun has 25µm minimum line width and 25µm minimum spacing between adjacent strips. The balun Operational BW (OBW) is 4 GHZ. The center frequency λ/4 (6 GHz) corresponds to a spiral line length equal to 6.7 mm. The vertical offset distance that gives the required coupling factor was found to be 0.01 mm without the need to vary the horizontal offset. Another larger balun size (3 x 5.6 mm²) was designed with minimum line and gap widths of 100µm, the balun OBW equal to 2.0 GHZ around 4 GHZ with return loss (RL) better than 11 dB. The amplitude imbalance and phase difference are 0.5 dB and 10°, respectively.

Table 5.1 - Comparison between the previously reported balun performance results and our balun design results (the small and the large size baluns).

<table>
<thead>
<tr>
<th>Balun type</th>
<th>OBW (GHz)</th>
<th>Phase Difference (Degree)</th>
<th>Amplitude Imbalance (+/- dB)</th>
<th>Frequency Range (GHz)</th>
<th>Size (mm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Robert CPW balun [10]</td>
<td>2.5</td>
<td>3° - 6°</td>
<td>0.2</td>
<td>1.6 – 4.1</td>
<td>1.2 x 2.3</td>
</tr>
<tr>
<td>Microstrip balun [12]</td>
<td>4</td>
<td>10°</td>
<td>0.2</td>
<td>2 - 6</td>
<td>0.7 x 1.5</td>
</tr>
<tr>
<td>Small balun Range 1</td>
<td>4</td>
<td>3°</td>
<td>0.5</td>
<td>4 - 8</td>
<td>1.4 x 2.4</td>
</tr>
<tr>
<td>Small balun Range 2</td>
<td>5.5</td>
<td>5°</td>
<td>0.5</td>
<td>5.5 -10</td>
<td>1.4 x 2.4</td>
</tr>
<tr>
<td>The large size balun</td>
<td>2</td>
<td>10°</td>
<td>0.5</td>
<td>3 -5</td>
<td>3 x 5.6</td>
</tr>
</tbody>
</table>
Summary

This chapter presented comparisons between the full-wave EM simulation results, obtained using Momentum, and the results of the four ports measurements of the small size balun circuit. Both results showed good agreement with each other, which supports the validity of other results obtained using Momentum. At the end of this chapter previously published balun designs were compared with the two balun designs (small balun and large balun) presented in this work. The comparison showed that the small size balun has very good performance over the desired frequency range.
CHAPTER 6 – COMPARISON OF MoM AND FDTD

6.1 Introduction

Before the advent of high-speed computers, it was advantageous to spend considerable effort to manipulate solutions analytically into a form, which minimized the subsequent computational effort. Nowadays with the high-speed computers it is more convenient to use simple analytical methods without worrying about the amounts of computation. Many numerical techniques have been derived to solve Maxwell’s differential or integral equations [13]- [15] such as, finite-difference time-domain (FDTD) and method of moments (MoM).

In this chapter two commercial softwares, Momentum (MoM) and Empire (FDTD), were applied to the 6 GHz balun and compared in terms of accuracy of scattering parameters, CPU speed and memory usage. This chapter is organized in the following way: Sections 6.1 and 6.2 briefly introduce the FDTD and MoM techniques and, section 6.3 discusses the numerical results.
6.2 Finite-Difference Time-Domain method (FDTD)

The FDTD as introduced by Yee [16], has been proven to be convenient and simple tool for time domain analysis and for various electromagnetic scattering problems. The basis of the FDTD algorithm is the Maxwell’s curl equations. For uniform, isotropic and homogenous media Maxwell’ s curl equation is given as:

\[ \nabla \times E = -\mu \frac{\partial H}{\partial t} \]

\[ \nabla \times H = \sigma E + \varepsilon \frac{\partial E}{\partial t} \]

where \( E, H, \varepsilon, \mu \), and \( \sigma \) are the electric field, magnetic field, permittivity, permeability, and conductivity, respectively. The central difference approximation is given by the following equation as a function of time and space [16]

\[ H_{x(i,j,k+1)}^{n+1/2} = H_{x(i,j,k+1)}^{n-1/2} + \frac{\Delta t}{\mu \Delta z} [E_{y(i,j+1,k+1)}^{n} - E_{y(i,j+1,k)}^{n}] - \frac{\Delta t}{\mu \Delta y} [E_{z(i+1,j,k+1)}^{n} - E_{z(i-1,j,k+1)}^{n}] \]

for \( H_x \) and similar equations for the \( E_x, E_y, E_z, H_y \) and \( H_z \) components. The scattering parameters as a function of the frequency can be obtained from the time-domain solutions \( E \) and \( H \) fields using Fourier transform (FT).

Stability restriction limits the time increment \( \Delta t \). The maximum time step that can be used to avoid numerical instability as a function of the grid sizes \( \Delta x, \Delta y, \) and \( \Delta z \) is given by the following formula:

\[ \Delta t < \frac{1}{v_{\text{max}}} \left[ \frac{1}{\Delta x^2} + \frac{1}{\Delta y^2} + \frac{1}{\Delta z^2} \right]^{-1/2} \]

where \( v_{\text{max}} \) is the maximum speed of the electromagnetic wave inside the modeled material.
Since it is not possible to use infinite grid size the solution domain for the FDTD must be terminated probably by absorbing boundaries around the grid [17][18], in order to reduce the computational cost.

6.3 Method of Moments

The method of Moments (MoM) [14] can be used for solving both differential and integral equations. After discretizing the required integral equation into a matrix equation, evaluates the matrix elements, and solves the matrix equation.

6.4 Discussion of the results

The FDTD method (Empire) used a non-uniform grid with total 1.5124 x 10^6 cells; see Table 6.1 for the cells description. Refer to balun #1 in chapter five to see the simulated structure physical dimensions.

<table>
<thead>
<tr>
<th>Number of lines</th>
<th>Minimum cell size (µm)</th>
<th>Maximum cell size (µm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>X- direction</td>
<td>199</td>
<td>6</td>
</tr>
<tr>
<td>Y- direction</td>
<td>95</td>
<td>.8</td>
</tr>
<tr>
<td>Z- direction</td>
<td>80</td>
<td>3</td>
</tr>
</tbody>
</table>

The computational domain was truncated using PML that was 4 cells deep. Both, the source line and the receiver line were truncated into the PML. The source has excitation frequency content up to 10 GHz. A total of 160,000 time steps were used in simulations. The MoM method (Momentum) is easy to set-up, the steps involved are:

1. Creating a physical design of the balun.
2. Defining the substrate characteristics, which included the number of layers in the substrate and the position of the layer of the balun.

3. Solving the substrate for a range of frequencies from 1 to 10 GHz. With 50 cells per wavelength at 10 GHz

4. Specifying input and output ports on input and output lines, respectively.

Empire (FDTD) full-wave simulation results showed good agreement with the balun measurement results and with the result obtained using the full-wave simulator Momentum (MoM), as shown in Figures 6.1 – 6.5. Below are some of the differences between Empire and Momentum:

1. Empire simulates finite substrates, while substrates extend to infinity in Momentum.

2. Momentum assumes zero metallization thickness (After simulation it account for the skin effect), while Empire is capable of simulating the exact metallization thickness.

3. Momentum is easy to use, while Empire is not, which could lead to some simulation mistakes using Empire.

4. For our specific structure (the Balun) and using 2.8 GHz CPU, the CPU time for running Empire was four hours, while it took Momentum two hours to finish the simulation. However, Momentum consumed large amount of memory (225,380 KB) compared to Empire (40,000 KB).
Figure 6.1 - Return loss ($S_{11}$) of the balun circuit #1, FDTD versus MoM versus measurement.
Figure 6.2- Insertion loss $\text{dB}(S_{12})$ of the balun circuit #1, FDTD versus MoM versus measurement.
Figure 6.3 - Insertion loss Phase($S_{12}$) of the balun circuit #1, FDTD versus MoM versus measurement.
Figure 6. 4 - Insertion loss ($S_{13}$) of the balun circuit #1, FDTD versus MoM versus measurement.
Figure 6.5 - Insertion loss Phase($S_{13}$) of the balun circuit #1, FDTD versus MoM versus measurement.
Summary

A full-wave FDTD and MoM have been used to analyze the multilayer spiral transmission-line balun; the results obtained were compared to the result obtained from the four-port measurements. In all cases the output matched reasonably well. The difference between Empire (FDTD) and Momentum (MoM) were compared, every software has it is own advantages and disadvantages. The geometrical dimension of the balun under comparison was presented in Appendix A (balun number 1).
CHAPTER 7 – THE EFFECT OF THE GROUND PLANE ON THE SPIRAL INDUCTOR MODEL

7.1 Introduction

The main goal of this work was to derive a lumped element model for spiral inductors that is independent of the ground plane effect. The importance of this model lies in its necessity in designing an antenna dipole loaded with lumped elements (in the absence of ground plane) to control the antenna electrical length without changing its physical length. However this technique required the design engineer to have good knowledge about the lumped element model at high frequency, since parasitic effect will not have negligible effect at high frequency and the inductor performance will not be the same as when a ground plane is present. In order to derive a ground plane independent inductor model a CPW line was used to connect between the two ports of the inductor such that the model elements as a function of ground plane spacing can be derived by increasing the distance to ground ($S_g$) between the CPW ground plane and the spiral inductor (Figure 7.1). At every distance $S_g$ the inductor was simulated using Momentum (ADS) and the simulation result was then used to obtain a lumped element model equivalent to the full wave simulation. Knowing the lumped elements values of one inductor at different distances from the ground plane, equations for the nominal inductance value, as well as the other model parameters, were derived as a function of the distance to ground.
Rectangular CPW spiral inductors were designed using existing free space inductance formulas [19]. A replica of all the inductance values were simulated using 50 Ω, 70 Ω, 80 Ω, and 90 Ω CPW feed transmission lines. In order to vary the characteristic impedance values the strip line width of the CPW line was fixed and the distance \( S_g \) to ground between the CPW ground plane and the spiral inductors was varied.

As was expected by decreasing the distance between the spiral inductor and the ground plane the value of the inductance was decreased from the free space nominal inductance value. By knowing the inductance values of one inductor at different distances from the ground plane, a formula for the inductance nominal value as a function of the distance to ground was derived. Two inductance values 3 nH and 4.5 nH were simulated four times, and at each time the distance to ground \( (S_g) \) was increased. For each full wave simulation results new values for the elements in the model were extracted and used to develop expressions dependent upon the ground distance, as shown in Figure 7.2 for the 4.5 nH inductor and Figure 7.3 for the 3nH inductor. Other parameters in the lumped
element model were also affected by varying the distance to ground, such as the mutual capacitance between the spiral turns, and the capacitance between the ground plane and the spiral turns (which should approach zero at infinite distance to ground), (Figures 7.4 to 7.7).

Figure 7. 2 - Inductance versus ground plane distance for the 4.5nH inductor $L = \frac{4.41}{1 + .49 e^{-2.6x}}$. 
Figure 7.3 - Inductance versus ground plane distance for the 3nH Inductor $L = \frac{2.9}{1 + 0.43e^{-40.2x}}$. 
Figure 7.4 – Capacitance to ground versus ground plane distance for the 4.5nH inductor

\[ C_g = \frac{0.033}{1 - 0.771e^{-2.208x}}. \]
Figure 7.5 – Capacitance to ground versus ground plane distance for the 3nH inductor

\[ C_g = \frac{0.00311}{1 - 0.981 e^{-0.077x}}. \]
Figure 7.6 - Mutual capacitance between the spiral strips versus ground plane distance for the 4.5nH inductor.

\[ C_s = \frac{0.144}{1 + 0.406e^{-4.004x}} \]
Figure 7.7 - Mutual capacitance between the spiral strips versus ground plane distance for the 3nH inductor: 

\[ C_s = \frac{0.148}{1 + 0.291e^{-2.201x}} \]
It should be mentioned that square and circular spiral inductors are more practical than the rectangular shape spiral inductors since they exhibit a higher Q-factor and higher resonant frequency. However, the rectangular geometry is more convenient in this research for studying the effect of ground spacing since for certain characteristic impedance the distance to ground is fixed and to get different inductance values all that has to be changed is the length of the spiral inductor.
7.2 Spiral Inductor

The CPW spiral inductor modeled here is shown in Figure 7.8. It should be noted that the spiral under consideration is a rectangular one of width $d_{o1}$ and length $d_{o2}$. An air-bridge is used to connect the inner port of the spiral to the output CPW (Figure 7.1). The number of turns ($N$) is two. Only the length of the spiral ($d_{o2}$) was varied to vary the inductance value.

Figure 7.8 - Two-turn rectangular spiral inductor.
When analyzing a CPW rectangular spiral inductor the following parameters need to be specified:

1. $d_{o1}$: the outer width of the spiral.
2. $d_{o2}$: the outer length of the spiral.
3. $W$: the spiral strip width.
4. $S$: the spacing between the spiral turns.
5. $S_g$: the distance between the outermost turn and the ground plane (Ground spacing).
6. $H_a$: the height of the air-bridge.
7. $W_a$: the width of the air-bridge.
8. $t$: the metallization thickness.
9. $\rho$: The metallization resistivity.

After a thorough literature search and numerical experiments, the lumped element equivalent circuit shown in Figure 7.9 was chosen to model the CPW spiral inductor [20]. This circuit should be applicable up to the first resonant frequency of the inductor, which is the frequency range of interest as it represents the region where the spiral acts as an inductor.
The optimized lumped elements nominal values should be converged to the values obtained from the free space analytical formulas listed below when increasing the value of the distance to ground \( S_g \). The values obtained using the analytical formula can also be used to set the range by which the lumped element values will be varied in the optimizer. The lumped element values of the spiral inductor model are obtained as follows:

1- To evaluate the inductance \( L \) in free space, the simpler expression in [19] was used.

\[
L = \frac{4}{10} N^2 d_{m1} \left[ \ln\left( \frac{2d_{m1}d_{m2}}{(d_{m1} + g)(b + t)} \right) + \frac{d_{m2}}{d_{m1}} \ln\left( \frac{2d_{m1}d_{m2}}{(d_{m2} + g)(b + t)} \right) - \frac{d_{m1} + d_{m2}}{2d_{m1}} + \frac{2g}{d_{m1} + 0.477 \frac{b + t}{d_{m1}}} \right]
\]

Where,
\[
a = w + s \\
b = (N - 1)a + w \\
g = \sqrt{d_{m1}^2 + d_{m2}^2} \\
d_{a1} = d_{m1} + b \\
d_{a2} = d_{m2} + b \\
d_{m1} = d_{i1} + b \\
d_{m2} = d_{i2} + b
\]

With \( L \) in nH and all dimension in mm.

2- The resistance \( R \) was computed using the expression from [20], which takes the skin effect into account: \( R(f) = a + b * f \), (where \( f \) is the frequency). The values of the variables \( a \) and \( b \) were extracted from the EM full wave simulation result (obtained from ADS- Momentum) through circuit optimization.

3- The series capacitor \( C_s \) models the parasitic capacitive coupling between the input and the output ports of the inductor, as well as the turn-to-turn capacitance in the spiral itself. Both the crosstalk between the adjacent turns and the overlap between the spiral and air-bridge contribute to this capacitance \( C_s \). The following approximate expression is used to evaluate this capacitance:

\[
C_s = C_{\text{coupling}} + C_{\text{air-bridge}}, \text{ where } C_{\text{air-bridge}} \text{ is approximated as a parallel-plate capacitance between the air-bridge and the spiral metallization. This approximation is valid since the height of the air bridge is very small compared to the bridge width and the spiral strip width.}
\]

\[
C_{\text{air-bridge}} = N\varepsilon_o \frac{WW_a}{H_a}
\]

\[
C_{\text{coupling}} = \frac{N-1}{N^2} * 2\varepsilon_o (\varepsilon_r + 1)(d_{m1} + d_{m2})(\frac{K(k)}{K(k)})
\]
\[ k = \left[ \tan \left( \frac{\pi}{4} \left( \frac{w}{a} \right) \right) \right]^2, \quad k' = \sqrt{1 - k^2} \]

Where \( K(k) \) and \( K(k') \) are the complete elliptic integral of the first kind of \( k \) and \( k' \), \( \varepsilon_r \) is the relative dielectric constant of the substrate. [19]

4- The parallel capacitors \( C_g \) is used to model the capacitance between the spiral (mainly the outermost turn) and the ground plane. Usually the distance to the ground \( S_g \) is large to reduce its effect on the spiral response. The expression for \( C_g \) can be determined by considering every turn of the spiral as an asymmetric CPW transmission line with line width equal to \( W \) [21], \( C_g \) is the sum of the individual self capacitances of all the turns since these capacitance are connected in parallel.

5- \( L_{\text{Air-Bridge}} \) can be evaluated if we consider the air bridge as a strip conductor in free space. Since the air bridge strip width is small and relatively far from the ground plane its inductance value can be assumed to be constant even with varying the distance to ground. Using a free space closed-form expression for the strip conductor in free space [24] the inductance of the air-bridge (Figure 7.10) can be found as following:

\[ L(nH) = 2 \times 10^{-4} \times l \times \left[ \ln \left( \frac{l}{w + t} \right) + 1.193 + 0.02235 \times \frac{w + t}{l} \right] \]

For a strip line of width 100\( \mu \)m, length 700\( \mu \)m and 3\( \mu \)m thickness the inductance is equal 0.44 nH.

![Figure 7.10 - Strip conductor in free space. \( l \) = length, \( w \) = width and \( t \) = thickness.](image-url)
7.3 Lumped Element Model

Two spiral inductors 3nH and 4.5nH fed with 50Ω, 70Ω, 80Ω and 90Ω CPW transmission lines were designed and simulated. The simulated results were then optimized to extract the lumped element values in the equivalent circuit model as a function of the characteristic impedance (distance to ground) as shown in Figure 7.11.

Figure 7.11 - Schematic diagram showing the optimization of the lumped element model and the data obtained from full wave simulation (Momentum - ADS).
Figure 7.12 shows the schematic diagram of the lumped element model and Figure 7.13 shows the schematic diagram of the simulated data blocks. Notice that the left and the right taper, used to connect between the spiral inductor and the CPW line from both side, were de-embedded to get correct optimized values. Figure 7.14 shows a drawing of a 4.5nH CPW spiral inductor and the location of the tapered-line sections. Figure 7.15 and 7.16 show comparisons between the insertion loss (IL) and the return loss (RL) of both the full-wave EM simulation data and the result of the optimized equivalent circuit model.

Figure 7.12 - Schematic diagram of the 4.5nH spiral lumped elements model.
Figure 7. 13 - Schematic diagram of the 4.5nH spiral data blocks with two de-embed blocks to cancel the effect of the right and the left taper.

Figure 7. 14 - CPW spiral inductor, notice the left and right taper.
Due to a miscommunication with the company that was contracted to manufacture the spiral inductors, the circuits were not properly fabricated. As such, only the full-wave simulation data and the associated equivalent circuit model results are shown herein. Actual performance should be very close to the simulated result due to the simplicity of the structure.

Figure 7.15 - Comparison of the insertion loss (S12) between the simulated data and the optimized circuit model for the 4.5 nH inductor operated in 50Ω CPW line.
Figure 7.16 - Comparison of the return loss ($S_{11}$) between the simulated data and the optimized circuit model for the 4.5 nH inductor operated in 50Ω CPW line.
Summary

The elements in the lumped elements model of a spiral inductor have been discussed and their nominal formulas have been presented. A lumped element model that is independent of the ground plane has been derived for different spiral inductor values. Formulas for the nominal inductance value, calculated in free space, as a function of distance to ground in CPW line were also derived. Since the existing inductor formulas for rectangular spiral inductors that can be found in literature assume a free space condition, we have derived a formula that can be used to obtain the value of the nominal inductance in the presence of the CPW ground plane. The formula is a function of the distance to ground plane from the spiral inductor edge. Furthermore formulas for the mutual capacitance between the spiral strips and for the capacitance to ground of the spiral inductor were derived as a function of distance to ground in CPW line.
8.1 Conclusion

In this work we derived a design procedure to help RF engineers design cost effective multilayer CPW spiral balun that works in the frequency range 1-8 GHz. Furthermore three techniques to enhance the balun performance were developed. The importance of the balun for feeding symmetric resonant antenna dipole was discussed, in addition to the balun background and its impedance analysis. The accuracy of the developed procedure has been proven by designing two balun circuits of different dimensions and simulating them using available commercial software, Momentum (MoM) and Empire (FDTD). The simulation results have shown good balun performance over the desired frequency range. It has been also shown through this work that introducing some capacitance at certain points in the balun circuit decreases both the phase difference and the amplitude imbalance of the balun. Furthermore some of the designed balun circuits have been fabricated and measured and the results agreed with the simulations.

A procedure to derive independent of ground lumped element model, which is independent of ground for spiral inductor, has been discussed at the end of this work. Formulas for lumped elements in the spiral inductors model were extracting as a function of distance to ground. The result showed that the nominal inductance value in the inductor model decreased by reducing the distance to ground (slot width in the
CPW line) and the capacitance to ground decreased by increasing the CPW line slot width.

**8.2 Recommendations**

Using the two derived lumped element models for the balun circuit, further work in future can be done to study the effect of varying physical parameter values in the balun circuit on the balun performance. More detailed analysis can be performed on the derived compensation techniques to understand its operation. The derived independent of ground lumped element model of the spiral inductor can be tested in the design of an antenna dipole loaded with spiral inductor.
REFERENCES


APPENDICES
### APPENDIX A - BALUN GEOMETRICAL DIMENSIONS

Table A. 1 - Common geometrical dimension of the small size balun. See (Figure 3.22 – Figure 3.24) in section 3.4 for parameters visualization.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Spiral strip line width</td>
<td>25µm</td>
</tr>
<tr>
<td>Spiral gap width</td>
<td>25µm</td>
</tr>
<tr>
<td>Spiral inner radius</td>
<td>212µm</td>
</tr>
<tr>
<td>Spiral outer radius</td>
<td>338µm</td>
</tr>
<tr>
<td>Horizontal offset</td>
<td>10µm</td>
</tr>
<tr>
<td>Vertical offset (polyimide 1 thickness)</td>
<td>10µm</td>
</tr>
<tr>
<td>Polyimide 2 thickness</td>
<td>10µm</td>
</tr>
<tr>
<td>Polyimide 1 and 2 permittivity</td>
<td>2.6</td>
</tr>
<tr>
<td>Quartz permittivity</td>
<td>3.8</td>
</tr>
<tr>
<td>Quartz thickness</td>
<td>25 mil</td>
</tr>
<tr>
<td>Spiral line width</td>
<td>6.7 mm</td>
</tr>
<tr>
<td>CPW line width</td>
<td>0.21 mm</td>
</tr>
<tr>
<td>CPW line slot width</td>
<td>.03 mm</td>
</tr>
</tbody>
</table>
APPENDIX A (Continued)

Figure A. 1 Balun number 1, dimensions are in mm.
Figure A. 2 Balun number 2, dimensions are in mm.
APPENDIX A (Continued)

Figure A. 3 Balun number 3, dimensions are in mm.
Figure A. 4 - Balun number 4, dimensions are in mm.
APPENDIX B – FOUR-PORT ON-WAFER TRL CALIBRATION AND BALUN MEASUREMENT PROCEDURE

B.1 Equipments Configuration

The four-port on-wafer TRL calibration and the balun measurements were performed using the following equipments:

- Anristu Lightening Network Analyzer
- Flexible Coaxial Cables (4)
- SUSS Semi-automatic Probing Station
- PC with a GPIB Port
- GPIB Cable (1)
- NIST-Cal (Multi-Port NIST) Installed on PC
- 2-Port to 4-Port Switch Box (1)
- Semi-flex Coaxial Cables with K connectors (2)
- CGB Wafer Pico-probes (4)
- 6 GHz Balun Wafer fabricated by ATC
- SMA to K connector (2)
- TRL Calibration Substrate (1)

Figure B.1 - General equipment configuration for the measurement.
Appendix B (Continued)

B.2 Calibration Kit

The Coplanar Waveguide width dimensions for the small size balun measurements are as follows:

1. Signal Line width 0.21 mm
2. Gap width 30um

![Figure B. 2 – CPW calibration kit dimension.](image)

The Cal kit consists of the following standards:

1. Thru line connection with a length of 0.4mm (thru =0.2mm)
2. 2 delay line lines:
   a. At 4GHz with a length of 12.5mm.
   b. At 6GHz with a length of 8.5mm.
3. Open standard with a pad length of 0.2mm.

B.3 Calibration Setup

Before starting the Multi-Port NIST program, it is important to set up a number of folders for the data to be stored. We will need one folder for calibration standards, and one folder for every Balun that will be measured. It is also important to go ahead and set up the VNA to default specifications. This needs to be done before proceeding to the following steps.

1. Open the NIST Multi-port Calibration Software.
2. Click on 8510 Tab from main menu and change the selection to Anritsu Lightening

3. Click on Error Box Configuration from main menu and make sure all ports are transparent.

4. Click on the Perform 4 Port Calibration tab from main menu

5. Click on Measure and then go to Browse Cal.

6. Open the folder that we have designated as our calibration data folder and click on the Select Current Directory.

7. Click on Browse Device and open the folder we’ve designated as the Balun folder. Click on Select Current Directory. This folder should be the Balun we are measuring first.

Now enter the names of our cal standards, i.e. thru, line, open/short.

The name that entered does not matter, but where we put that name does matter. All these files will be stored in the calibration directory we have selected.

Note: make sure the GPIB address field is set to 16.

8. For the line standard enter a name in the 1st row to the right of the Start tab, select a name such as delay1SB. This will make locating the file easier. If there are multiple delay lines, enter it into the next row under the 1st line. Make sure a different name is given.

9. For the reflect standard, enter a name in any of the remaining rows except the thru row. Once a name is assigned, click on the reflect dot to the left of the reflect name. It does not matter if we have entered and open or a short name.

10. Finally, enter a name for the thru standard in the standard row.
Setting up the file names is now completed. It is time to make calibration measurements. There is no need at this point to define calibration standard. This is taken care of in the later steps.

**B.4 Calibration Measurement**

1. Make sure that the tab at the top of the measure window is set at the measure West/East Calibration STD’s.

2. Connect the line standard to the west and east ports, and click the start tab corresponding to the line standard. Wait until measurement is through. The measurement was done with “probe-tip” calibration reference planes, since in the calibration the reference was not at zero but it was shifted to the outside of the circuit by 0.185mm.

![Figure B. 3 - Insertion loss- $S_{21}$- for delay line at 6GHz (West-East).](image-url)
Figure B. 4 - Return loss - $S_{11}$ - for delay line at 6GHz (West-East).

Figure B. 5 - Insertion loss - $S_{21}$ - for delay line at 8GHz (West-East).
3. Connect the open standard and click the corresponding tab. if the reflect measurement is made using only one port at a time then click the 2-P tab before starting the measurements.
Figure B. 7 - Insertion loss- $S_{21}$- for Open (West-East).

Figure B. 8 - Return loss- $S_{11}$- for open (West-East). Notice that the decay in phase is due to the amount of calibration reference shift.

Figure B. 9 - Insertion Loss- $S_{21}$ - for Thru (North-South).

Figure B. 10 - Return loss- $S_{11}$ - for Thru (North-South).
5. Click the Measure West/East Calibration STD’s tab till it changes to Measure North/East Cal STD’s. Repeat steps 2-4 but make sure we use the North and South port.

Figure B. 11 - Insertion Loss- $S_{21}$- for delay line at 6GHz (North-South).
Figure B. 12 - Return loss- $S_{11}$ - for delay line at 6GHz (North-South).

Figure B. 13 - Return loss- $S_{11}$ - for delay line at 6GHz (South-North).
6. Click continue. This concludes the measurements of our cal standards.
B.5 Calculating Correction Factor

1. Click on Calibrate West/East tab. We should see all the line standards that we have measured in the menus, but without line lengths. Enter the physical line lengths at this time.

2. The software has a reflect standard defaulted as a short. Change to an open and record the reflect length.

3. Enter the estimated dielectric constant (Er=2.25).

4. When done recording values, click on Calculate Calibration Coefficients and Error Boxes.

5. Click Display Corrected Data.

6. Look thru this thoroughly to check for good calibration. If the calibration is bad, repeat the calibration steps. There is no need to repeat the Balun measurement.

7. Click Continue to exit out of window.
9. Save the corrected data. Saving the corrected data might take a few minutes depending on the PC.

10. Repeat steps 1-9 for the North/South, and South/North configurations.

B.6 Making The Four-Port Measurement

1. In the measurement setup window verify the selection of measuring single ended S-parameters and not the differential S-parameters

2. From Four Port main menu, click Measure.

3. Change the device path for which you want the Balun to be located.

4. Attach the Balun to the ports making note which measurement port is attached to which device port. Make sure you are extremely careful when landing the probes on the wafer. Stop the descending of the probes after a maximum of 30micron of skating. Click Measure 4-Port. When lifting the probes observe the skating marks on the metal and make sure all probe tips contacted properly with the circuit.

5. Repeat steps 2-3 making sure we change the dive path.

B.6 De-embedding The Four-Port Measurement

1. Go to Measure from the 4-port main menu and change the device path to the one our Balun is currently located.

2. Repeat the steps in the section Calculating Correction Coefficients. This step is so that our calibration coefficients are placed in the device directory for de-embedding. When saving the coefficients in these steps, replace any files when prompted to.

3. From the main menu click on 4-Por Calibrate.
4. Click De-embed Four Port from the 4-port main menu.

5. Make sure that the device directory is set to the Balun that we have measured.
   Also make note of the port definitions. We will use this when looking at our data.

6. If using ADS for the s4p data, change the format to Touchtone

7. If we want to look at the data in the NIST program, we will have to de-embed the
   device in Column (dB-phase) format. The NIST program will only read Column
   (dB-phase).

8. Click De-embed and save Four Prot S-Parameters.

9. Click Exit when finished.

**B.7 Display corrected Four-Port Data**

1. From 4-port main menu, click Display corrected 4-port data.

2. Choose the text file in which we named previously for the device.