



School of Applied Science and Engineering

Department of Electrical and Computer Engineering

**Investigation and Enhancement of The Phase Shift of 90°
Hybrid**

By

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**A Thesis Submitted to the Department of Electrical and Computer Engineering
in Partial Fulfillment of the Requirements Master of Science**

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DEDICATION

Dedication to my Dr. Bashir Mohamed .Khamoudi who was very helpful with his support .

Dedication to my brother Mnsour Saeed Mansour who was very helpful to me.

Dedication to precious soul my Grand Mother and to precious soul my Father , both were very helpful with providing both motivational and patience, support as I worked to complete my thesis and degree.

Dedication to precious soul my martyr brother Salah Saeed Mansour and to all martyrs of Libya.

Dedication to all my family and my friends they all encouraged me and supported me to complete this thesis .

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ABSTRACT

Most of the studies reporting on the 90° hybrid coupler assume that the coupler is ideal and concentrate on the analysis and design.

A 3dB 90° or 180° hybrid splits an input signal into two equal amplitude outputs.

A realistic coupler will not have a perfect 3 dB power split or provide a 90 degree phase shift across the entire band; this will cause additional attenuation of the phase shifted signal as well as introduce more reaction at the input port due to imperfect cancellation of the reacted signals.

The goal of this thesis is optimization matching of a 90° hybrid coupler for a smaller phase error and greater bandwidth by using improvement methods and investigate of them.

Simulation performance is carried out using advanced system design (ADS2009) software. Simulation gives a very good promising results for fixing the phase of 90° hybrid and improvement of the bandwidth by using two methods which are the extended transmission line and single stubs at the output ports.

Conventional 90° hybrid, 90° hybrid cascade and 90° hybrid extended TL outputs are fabricated, and tested using Hewlett - Packard 8510C vector network analyzer (VNA) which are used as references of phase and bandwidth .

Measurements of these hybrids such as, (S-parameters , Return Loss, coupling vector, Isolation , and Phases are presented.

Simulation and measurement results of these hybrids with the used two methods for improvement of phase and the bandwidth of the 90° hybrid are giving a good agreement results which is means the phase and the bandwidth of the 90° hybrid are improved.

الملخص

أغلب الدراسات المقدمة عن المقرن الهجين ذو التسعين درجة تفترض أن ذلك المقرن له سلوك مثالي وتركز على التحليل والتصميم.

نظرياً إشارة الدخل للمقرن الهجين ذو التسعين درجة نوع 3 ديسيبيل يقسم قدرة الخرج إلى نصفين متساويين. عملياً المقرن ليس له قدرة 3 ديسيبيل مثالية لتقسيم قدرة الخرج بالتساوي أو تزود المقرن بزاوية طور تسعين درجة بين منفذي الخرج خلال النطاق الترددي الكلي وهذا يؤدي إلي توهين إضافي لإشارة الطور المزاحة بقدر ما يكون تأثير تلك القدرة أكثر تعقيداً عند مدخله مسببةً إلغاءً غير مثالي لتلك الإشارات المنعكسة.

الهدف من هذه الرسالة هو الموائمة المثالية لتصميم المقرن الهجين ذو التسعين درجة للوصول لأقل خطأ في زاوية الطور والحصول على أكبر نطاق تردد باستخدام طرق التحسين والتحقق منها.

تم تنفيذ المحاكاة باستخدام برنامج النظام المتقدم للتصميم (ADS2009).

المحاكاة أعطت نتائج جيدة جداً وواعدة لتثبيت زاوية الطور عند التسعين درجة بدقة كبيرة للمقرن الهجين ذو التسعين درجة وكذلك زيادة عرض النطاق الترددي باستخدام طريقتين وهما طريقة تمديد خط النقل لمنافذ الخرج و طريقة الأعقاب المنفردة عند منافذ الخرج.

المقرن الهجين ذو التسعين درجة التقليدي و المقرن الهجين المتعاقب ذو التسعين درجة و المقرن الهجين ذو خطوط النقل الممددة لمنافذ الخرج تم تصنيعها و تجربتها عملياً على جهاز محلل الشبكات المتجهي

(HP 8510C VECTOR NETWORK ANALYZER) الذي يستخدم في قياس الطور والنطاق الترددي.

المعاملات المهمة والتي تم قياسها لهذه المقرنات الهجينة والمتمثلة في معاملات التشتت والفقد الراجع ومعامل الاقتران و العزل والطور والنطاق الترددي).

نتائج المحاكاة والنتائج العملية لهذه المقرنات الهجينة باستخدام الطرق المحسنة للطور والنطاق الترددي كانت متوافقة جداً مما يثبت أن الطور والنطاق الترددي تم فعلاً تحسينهم مقارنة بالمقرن الهجين ذو التسعين درجة التقليدي.

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LIST OF NOMENCLATURE

λ_0	Wavelength in free space
λ	Wavelength, m
[S]	Scattering - Parameter matrix
BW	Bandwidth
f	Frequency, Hz
f_c	Center frequency
f_L	Lowest resonance or lower edge frequency
f_U	Upper resonance or upper edge frequency
GHz	Gigahertz, 10^9 hertz
c	Speed of light, m/s
ϵ_0	Permittivity of free space, 8.854×10^{-12} [F/m]
ϵ_r	Dielectric constant
ϵ_{eff}	Effective dielectric constant
$\tan\delta$	Loss Tangent
μ	Permeability of the medium
σ	Conductivity
μ_0	Permeability of free space, $4\pi \times 10^{-7}$ [H/m] V Volts
h	Thickness of substrate, mm
w	Width of microstrip lines, mm

L	Length of microstrip lines, mm
L	Length With Frequency.
Ω	Ohm
Z_o	Characteristic impedance, Ω
Z_{in}	Input impedance, Ω
dB	Decibels
(VSWR)	Voltage Standing Wave Ratio
S_{11}	Reflection Coefficient, dB
S_{21}	Insertion Loss, dB
S_{31}	Coupling Balance, dB
S_{41}	Isolation, dB
ABCD	Matrices To Cascade The Stubs And Quarter-Wavelength Section
Y	Admittance
Z	Impedance
L	Inductance
C	Capacitance
R	Resistance
$C(dB)$	Coupling Factor(C in dB)
RL	Return Loss
(DUT)	Device Under Test .
HP8510C (VNA)	Hewlett -Packard 8510C Vector Network Analyzer
(VNA)	Vector Network Analyzer
C	Coupling
D	Directivity

I	Isolation
IL	Insertion loss
V^-	Reflected voltage wave vector .
V^+	Incident voltage wave vector .
T	Transmission Coefficient.
Γ	Reflection Coefficient .
Γ_e	Reflection Coefficient for even mode analysis
Γ_o	Reflection Coefficient for odd mode analysis
T_e	Transmission Coefficient for even mode analysis
T_o	Transmission Coefficient for odd mode analysis
Π	Pi (3.142)
α	Attenuation Constant
γ	propagation constant
β	Beta Phase constant
K_0	(wave number) of a plane wave in free space
v_{ph}	Phase Velocity
Φ	Phase
$\Delta\phi$	Difference outputs Phase
ω	Angular Frequency
h	Height
l	Length
w	Width
$P1$	Power Reflected
$P2$	Power Transmitted

P_3	Power Coupled
P_4	Power Isolated ISM - Industrial, Scientific and Medical
η_0	Intrinsic impedance for free space
η	Intrinsic impedance of the medium
ADS	Advance Design System
FR4	Flame Retardant 4
PCB	Printed Circuit Board
RF	Radio Frequency
SFG	Signal Flow Graph

Chapter 1

Introduction

Chapter 1

Introduction

1.1 Introduction

Couplers and hybrids are devices in which two transmission lines pass close enough to each other for energy propagating on one line to couple to the other line. A 3dB 90° or 180° hybrid splits an input signal into two equal amplitude outputs. A directional coupler normally splits an input signal into two unequal amplitude outputs. 90° Hybrids or hybrid couplers are basically 3 dB directional couplers in which the phase of the coupled output signal and the output signal are 90° apart. Since -3 dB represents half power, a 3 dB coupler divides the power equally (within a certain tolerance) between the output and coupled output ports.

The 90° phase difference between the outputs makes hybrids useful in the design of electronically variable attenuators, microwave mixers, modulators and many other microwave components and systems.

The 90° hybrid coupler is a four-port microwave device which takes an input on one port and provides an even power split on two output ports with a 90 degree phase shift between them. The fourth port on the device is isolated from the first port.

The 90 degree hybrid coupler also known as quadrature hybrids or simply quad hybrids ,This device exhibits reciprocal behavior because any port can act as an input, and the device will operate in the same manner .

Common uses of quadrature couplers include:- Antenna feed systems, the combination of power division/combining and 90° phase shift can simplify the feed network of phased array antenna systems, compared to alternative networks using delay lines.

A coupler can be categorized by five main parameters:- bandwidth, insertion loss, coupling ratio, phase shift between ports, and isolation.

Bandwidth is:- defined as the frequency range, this is the range over which specifications are guaranteed for the particular device.

Insertion loss is:- the additional loss within the device above the loss due to splitting. This can be caused by reactions of signals, dielectric losses, and conductor losses. Coupling ratio is:- defined as the ratio of the lower of the two output powers to the input power.

Isolation:- refers to the ratio between the input power and the leakage power at the isolated port .

It will be necessary to have an idea of the typical specifications that commercially available hybrid couplers possess in order to gauge the performance of the constructed coupler.

Hybrid couplers tend to have bandwidths of about 13:1 at a maximum. Over this bandwidth, couplers will typically exhibit about 3 dB of insertion loss, a coupling ratio of 3 dB and isolation of 20 dB .

The hybrid coupler can be analyzed using an even-odd mode technique. This technique can be used because the coupler is a linear device. Linearity allows for the use of superposition to split an input source into a sum of even (in phase) and odd(out of phase) sources.

This split allows the symmetry of the coupler to be preserved which simplifies analysis.

1.2 Problem Statement A realistic coupler will not have a perfect 3 dB power split or provide a 90 degree phase shift across the entire band; this will cause additional attenuation of the phase shifted signal as well as introduce more reaction at the input port due to imperfect cancellation of the reacted signals.

1.3 Motivation The goal of this thesis is to improve phase shift of 90 degree hybrid coupler which will provide a 90 degree phase shift between two output to a signal at center frequency of 2.4 GHz and bandwidth enhancement of all cases .

1.4 Proposed Solution The improvement of the phase and bandwidth by using two methods:- the first method by design 90 degree hybrid coupler with stub lengths ,the stubs on ports 2 and 3 to provide the proper phase shift over the correct bandwidth. The

second method by design 90° coupler with extended TL on ports 2 and 3 to provide the phase shift which will more closely meet the desired results and present details of methods. In this Thesis, the design, simulation, implementation and testing of 90° hybrid with center frequency of 2.4GHz will be carried out.

Investigation of how to enhance the phase shift and bandwidth of the designed 90° hybrid coupler by last methods.

1.5 Organization Of Thesis

This thesis will be divided into five Chapters. In chapter 1 the background and introduction of the topic will be given. Whereas in Chapter 2, the theoretical part related to the work will be covered such as hybrid types, design and analysis of the 90° hybrid and important parameters (S-parameters). Also the stub techniques will be covered.

In chapter 3, the calculation of all dimensions of 90° hybrid will be demonstrated.

In chapter 4 will present the design of the 90 degree hybrid with the simulated, fabricated and tested S-parameters results. Also investigation and enhanced phase shift and bandwidth of the designed 90° hybrid coupler by improvement methods such as, extended TL method and single stub method .

In chapter 5 is concerned with, the analysis and discussion of the obtained results. Comparison of the simulated and measured resulted will be presented.

In chapter 6, some conclusions of this work and suggested future work will be given.

Chapter 2

Theory of Hybrid Couplers

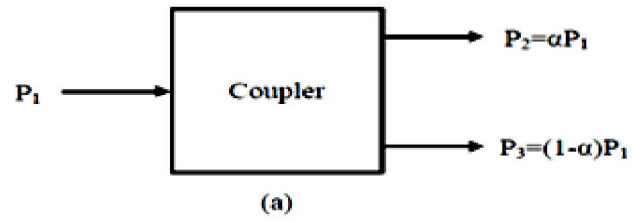
2.1 Introduction

This chapter will cover the theory related to the directional couplers, hybrids and important measure parameters such as coupling factor, isolation, directivity, phase balance and S-parameters.

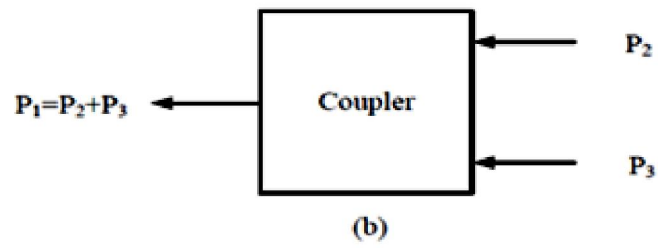
Applications of these microwave components will be stated. Stub techniques and mechanism also will be given.

2.2 Directional Couplers Overview

Power dividers and directional couplers are passive microwave components used for power division or power combining, as illustrated in Figure 2.1. In power division, an input signal is divided into two (or more) output signals of lesser power, while a power combiner accepts two or more input signals and combines them at an output port. The coupler or divider may have three ports, four ports, or more, and may be (ideally) lossless. Three-port networks take the form of T-junctions and other power dividers, while four-port networks take the form of directional couplers and hybrids. Power dividers usually provide in-phase output signals with an equal power division ratio (3 dB), but unequal power division ratios are also possible. Directional couplers can be designed for arbitrary power division, while hybrid junctions usually have equal power division. Hybrid junctions have either a 90° or a 180° phase shift between the output ports.[1]



Figure(2.1.a) An arbitrary power division.



Figure(2.1.b) power combining.

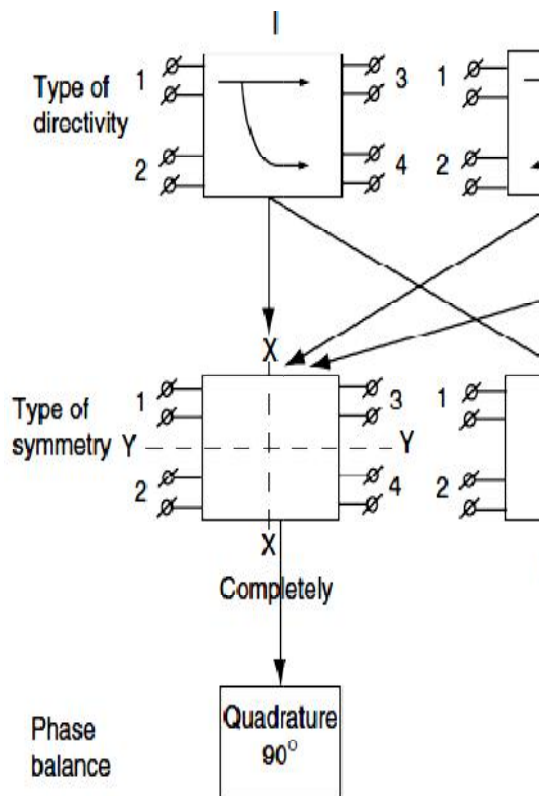
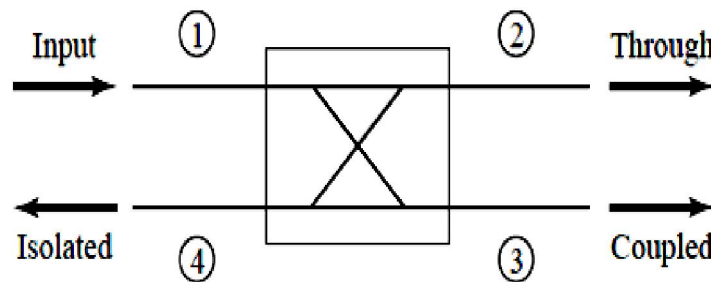


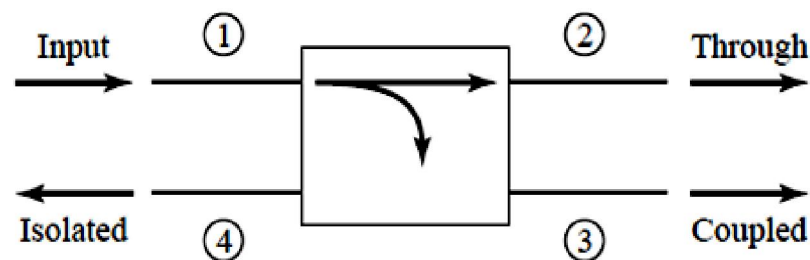
Figure (2.2) Directional couplers characteristics.

Figure (2.2) display various kinds of directivity, symmetry, and phase balance that a directional coupler can have.[2]

The most general definition for a directional coupler describes it as a multi-port (usually 4 port) matched, lossless, reciprocal circuit that has an isolated port which depends upon the incident signal port. Four port directional couplers are usually represented schematically by two single line representations of transmission lines with a crossed line between them to represent the coupling between the lines, as shown in Figure(2.3,a,b) We will see that this coupling between lines can occur by providing an electrical connection between the lines (branch line or Wilkinson's coupler) or by bringing two lines close together so that their magnetic and electric fields interact (parallel line or Lange couplers).[3]



Figure(2.3.a) Symbol for directional couplers with port 4 terminated with a matched load. .



Figure(2.3.b) symbol for a directional coupler with port 4 terminated with a matched load.

With reference to Figure (2.3.a), which is a commonly used symbol for directional couplers, the ideal directional coupler has the property that a wave incident in port 1 couples power into ports 2 and 3 but not into port 4. For the wave incident in port 4, the

power is coupled into ports 2 and 3 but not into port 1. Thus ports 1 and 4 are uncoupled. Similarly, ports 2 and 3 are also uncoupled. However, port 4 is usually terminated with a matched load and not accessible to the user. This results in a three-port component as shown in Figure(2.3.b). The performance of a directional coupler is characterized by two parameters, the coupling and the directivity. The coupling factor indicates the fraction of the input power that is coupled to the output port, and the directivity is a measure of how well the power is coupled in the desired direction . Let P_1 be the power supplied to port 1, P_3 be the coupled power in the forward direction in port 3 and P_4 be the power coupled in the backward direction in port 4. The coupling in decibels is given by

$$C = 10 \log \frac{P_1}{P_3} \quad (2.1)$$

The directivity is defined as

$$D = 10 \log \frac{P_3}{P_4} \quad (2.2)$$

Ideally, the power P_4 coupled in port 4 should be zero, and therefore the isolation of the coupler would be infinite.

$$I = 10 \log \frac{P_1}{P_4} \quad (2.3)$$

The ideal directional coupler, as illustrated in Figure(2.2.a), is a lossless reciprocal four-port network matched at all ports with $S_{14}=S_{23}=0$. Thus the scattering matrix has the form[1]:-

$$[s] = \begin{bmatrix} 0 & S_{12} & S_{13} & 0 \\ S_{12} & 0 & 0 & S_{24} \\ S_{13} & 0 & 0 & S_{34} \\ 0 & S_{24} & S_{34} & 0 \end{bmatrix} \quad (2.4)$$

The above definition of a directional coupler requires that a number of the S-parameters be equal to zero. These are shown together with the part of the definition that requires that they vanish.

$$\text{Matched} \leftrightarrow S_{11} = S_{22} = S_{33} = S_{44} = 0 \quad (2.5)$$

$$\text{Isolated output} \leftrightarrow S_{41} = S_{23} = 0 \quad (2.6)$$

$$\text{Reciprocal} \leftrightarrow S_{ij} = S_{ji} \quad (2.7)$$

A passive lossless circuit implies that the matrix of s-parameters must be a unitary matrix. This means that the matrix when multiplied by its transposed complex conjugate results in the identity matrix (i.e. ones on the diagonal and zeros otherwise). This is a mathematical condition which results from requiring that the total power exiting the circuit exactly equals the total power entering the circuit regardless of the load or source configurations.[3] Such a product is:-

$$[S] = \begin{bmatrix} 0 & S_{12} & S_{13} & 0 \\ S_{12} & 0 & 0 & S_{24} \\ S_{13} & 0 & 0 & S_{34} \\ 0 & S_{24} & S_{34} & 0 \end{bmatrix} \begin{bmatrix} 0 & S_{12}^* & S_{13}^* & 0 \\ S_{12}^* & 0 & 0 & S_{24}^* \\ S_{13}^* & 0 & 0 & S_{34}^* \\ 0 & S_{24}^* & S_{34}^* & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (2.8)$$

Carrying out the matrix multiplication results in a set of equations for the various s-parameters. Two such equations are shown:-

$$S_{12} S_{24}^* = -S_{13} S_{34}^* \quad \text{and} \quad S_{12} S_{13}^* = -S_{24} S_{34}^* \quad (2.9)$$

with the first resulting from the top row multiplied with the last column. Division of the two equations results in(2.10):-

$$\frac{S_{24}^*}{S_{13}^*} = \frac{S_{13}}{S_{24}} \leftrightarrow S_{24} S_{24}^* = S_{13} S_{13}^* \leftrightarrow |S_{24}| = |S_{13}| \quad (2.10)$$

so that the coupling between ports 1 and 3 equals that between ports 2 and 4. Similarly, another pair of equation(2.2) can be obtained to show that:-

$$|S_{12}| = |S_{34}| \quad (2.11)$$

so that the coupling between ports 1 and 2 is the same as that between ports 3 and 4. Then the product of the first row with its conjugate in (2.8) yields

$$|S_{12}|^2 + |S_{13}|^2 = 1 \quad (2.12)$$

similarly,

$$|S_{12}|^2 + |S_{24}|^2 = 1 \quad (2.13)$$

By properly choosing the terminal planes on ports 1 and 3, we can adjust the phase angle of scattering parameters, so that:-

$$S_{12} = S_{34} = C_1 \quad (2.14)$$

$$S_{13} = C_2 e^{j\alpha} \quad (2.15)$$

$$S_{24} = C_2 e^{j\beta} \quad (2.16)$$

where C_1 and C_2 are real, and α and β are phase constants to be determined. Use of (2.14) and (2.15) in (2.12) gives

$$C_1^2 + C_2^2 = 1 \quad (2.17)$$

Substituting (2.14) to (2.16) into (2.9) yields a relation between the two phase constants as:-

$$\alpha + \beta = \pi \pm 2n\pi \quad \text{or} \quad \theta + \phi = \pi \pm 2n\pi \quad (2.18)$$

If we ignore integer multiples of 2π , there are two particular choices that commonly occur in practice[1], there are two particular choices:-

2.2.1 The 90° coupler

A Symmetric Coupler, the phase constants are chosen equal, $\alpha = \beta = \pi / 2$. Then the scattering matrix has the form

$$[S] = \begin{bmatrix} 0 & C_1 & jC_2 & 0 \\ C_1 & 0 & 0 & jC_2 \\ jC_2 & 0 & 0 & C_1 \\ 0 & jC_2 & C_1 & 0 \end{bmatrix} \quad (2.19)$$

2.2.2 The 180° coupler

An Ant-symmetric Coupler , the phase constants are chosen to be 180° apart, $\alpha = 0$, $\beta = \pi$. Then the scattering matrix has the form

$$[S] = \begin{bmatrix} 0 & C_1 & -C_2 & 0 \\ C_1 & 0 & 0 & -C_2 \\ -C_2 & 0 & 0 & C_1 \\ 0 & -C_2 & C_1 & 0 \end{bmatrix} \quad (2.20)$$

2.3 Performance Measures

Coupler performance is characterized by five parameters defined below

$$\text{Return Loss} = -20 \log |S_{11}| \quad (2.21)$$

$$\text{Coupling} = C = 10 \log \frac{P_3}{P_1} = -20 \log \beta \text{ dB} \quad (2.22)$$

$$\text{Directivity} = D = 10 \log \frac{P_4}{P_3} = 20 \log \frac{\beta}{|S_{41}|} \text{ dB} \quad (2.23)$$

$$\text{Isolation} = I = 10 \log \frac{P_4}{P_1} = -20 \log |S_{41}| \text{ dB} \quad (2.24)$$

$$\text{Insertion loss} = L = 10 \log \frac{P_2}{P_1} - 20 \log |S_{21}| \text{ dB} \quad (2.25)$$

2.3.1 Return Loss

Return loss is a measure of the input impedance match. The term loss is somewhat anomalous. If the input match is good then the input reflection coefficient is small and only a small amount of incident power is reflected back. Abstractly, the power which passes into the circuit can be thought of as power lost to the feeding circuit.

2.3.2 Insertion Loss

Insertion is, a measure of the amount of power which enters the coupler but fails to make it out of the direct port. A large fraction of the power coupled away would represent a large loss of power to the direct port. When couplers are inserted into a circuit for the purpose of measuring the propagating energy, a low insertion loss is desired to insure that the original circuit is not modified by the measuring system.[3]

2.3.3 Coupling

Coefficient is a measure of the amount of power coupled away from the direct port .This coefficient is often used alone to identify a coupler. A 3 dB coupler is one in which half the power is coupled away from the direct port, i.e., the power is split between the coupled and direct port. A 20 dB coupler is one in which one hundredth of the power is coupled away.[3]

2.3.4 Isolation

Isolation, is a measure of how well the coupler prevents power from exiting the isolated port .For an ideal coupler the isolation would equal $(-\infty)$. [3]

2.3.5 Directivity

Directivity, is a measure of the fraction of coupled power which exits from the coupled port when compared to that which exits from the isolated port. [3]

2.4 Hybrid Couplers Overview

The hybrid coupler, or 3 dB directional coupler, in which the two outputs are of equal amplitude takes many forms. Not too long ago the quadrature (90 degree) 3 dB coupler with outputs 90 degrees out of phase was what came to mind when a hybrid coupler was mentioned. Now any matched 4-port with isolated arms and equal power division is called a hybrid or hybrid coupler. Today the characterizing feature is the phase difference

of the outputs. If 90 degrees, it is a 90 degree hybrid. If 180 degrees, it is a 180 degree hybrid. Even the Wilkinson power divider which has 0 degrees phase difference is actually a hybrid although the fourth arm is normally imbedded. [4]

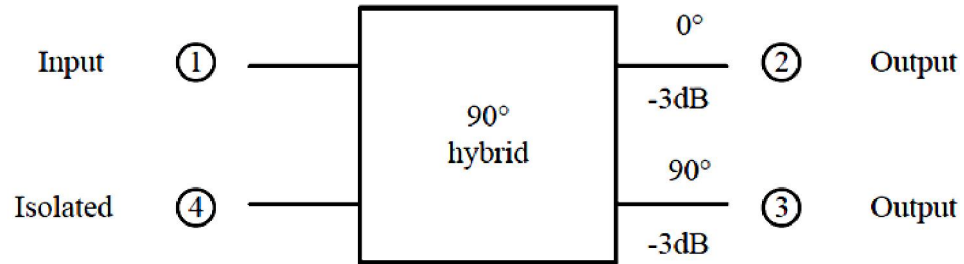
Directional couplers can be designed for arbitrary power division. However, hybrid couplers are special cases of directional couplers with equal power division (3dB), which implies

$$C_1 = C_2 = 1/\sqrt{2} \quad (2.26)$$

Hybrid couplers have either a 90° or a 180° phase shift between the output ports.

2.4.1 90° Hybrid Couplers Functional Description

The 90° hybrid coupler, also known as the quadrature hybrid coupler, is a 3dB directional coupler with 90° phase difference in the two output ports. With reference to a 90° hybrid symbol shown in Figure (2.4) , a signal applied to port 1 will be evenly split into two components with a 90° phase difference at ports 2 and 3, and port 4 will be isolated.



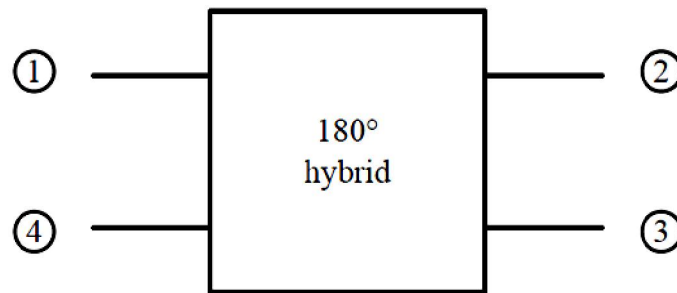
Figure(2.4) Symbol of a 90° hybrid coupler.

The quadrature hybrid has a 90° phase shift between ports 2 and 3 ($\theta = \phi = \pi/2$) when fed at port 1, and is an example of a symmetric coupler. Its scattering matrix has the following form: [1]

$$[s] = \frac{-1}{\sqrt{2}} \begin{bmatrix} 0 & j & 1 & 0 \\ 1 & 0 & 0 & j \\ j & 0 & 0 & 1 \\ 0 & 1 & j & 0 \end{bmatrix} \quad (2.27)$$

2.4.2 180° Hybrid Couplers Functional Description

The 180° hybrid coupler is a 3dB directional coupler with a 180° phase shift between the two output ports. It can also be operated so that the two outputs are in phase. With reference to Figure (2.5) , a signal applied to port 4 will be split equally into two components with 180° phase difference at ports 2 and 3, and port 1 will be isolated. If the input is applied to port 1, it will be evenly divided into two in-phase components at port 2 and 3, and port 4 will be isolated. Conversely, signals input into ports 2 and 3 will add at port 1 and the difference of the two signals will appear at port4.



Figure(2.5) Symbol of a 180° hybrid coupler.

The magic-T hybrid and the rat-race hybrid have a 180° phase difference between ports 2 and 3 when fed at port 4, and are examples of an anti-symmetric coupler. Its scattering matrix has the following form: [1]

$$[S] = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & -1 \\ 1 & 0 & 0 & 1 \\ 0 & -1 & 1 & 0 \end{bmatrix} \quad (2.28)$$

The branch-line coupler Figure(2.4) has dimensions of a quarter-wave length by quarter-wavelength at the center frequency, while a rat-race coupler Figure(2.5) has dimensions a half-wavelength by quarter-wavelength. Due to these large electrical lengths of the transmission-line elements, branch-line and rat-race couplers occupy a significant amount

of circuit area. This problem worsens with decreasing frequency. These couplers are often realized in a circular rather than a rectangular.

2.4.3 Cascaded Hybrid Functional Description

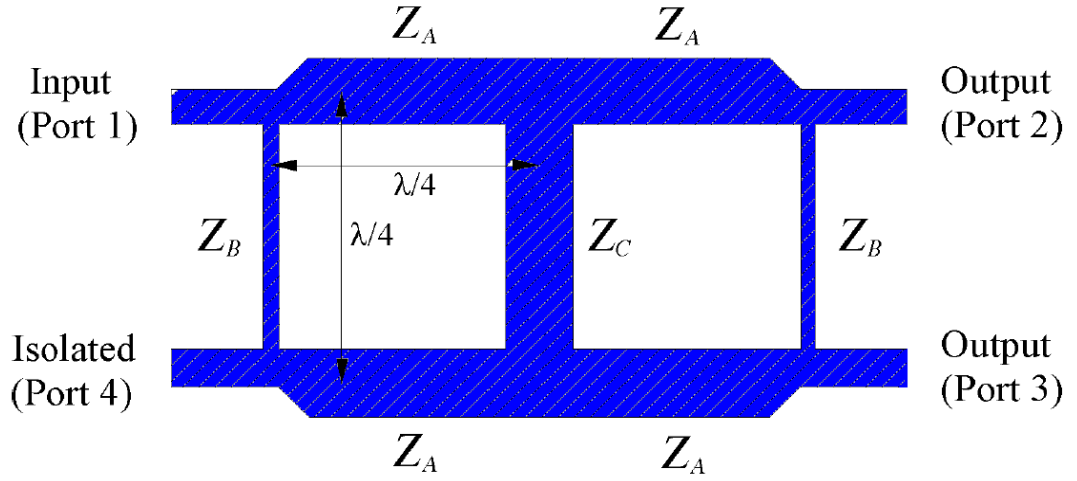
Previously in this thesis study on bandwidth enhancement of branch-line couplers, include using a multi-section cascaded.

If one pair of terminating resistors has different values compared to the other pair, the resulting branch-line hybrid can operate as a directional coupler and an impedance transformer simultaneously.[5]

The bandwidth of the branch coupler can be enlarged by increasing the number of branches. Young has given a formulation to design branch-line couplers of two to eight branches. Levi and Lind have given calculations for branch-line couplers of three to nine branches and synthesis for maximally flat responses and Chebyshev characteristics. Structure as conventional method for bandwidth enhancement .[2]

The three-branch coupler has larger bandwidth than the two-branch coupler. Additional branches can expand the bandwidth even further. However, couplers with more than four branches are difficult in microstrip because the end branches require impedances that reach the upper limits of practical realization.[2]

For conventional microstrip designs the limit is around 4 branches, more sections than this and the outer branch impedances are too high to be fabricated.[5]



Figure(2.6) Geometry of cascaded 90° hybrid .

Figure (2.6) shows two-section branch-line coupler with uniform impedances of Z_A , Z_B and Z_C and electrical lengths of $\frac{\lambda}{4}$. Using the even–odd-mode analysis technique, the current work determines characteristic impedances by:- [7]

$$Z_B = Z_0 \frac{C}{1 - \sqrt{1 - C^2}} \quad (2.29)$$

$$Z_C = \frac{Z_A^2}{Z_0 C} \quad (2.30)$$

For this it is possible to develop three or more branches. This technique allows to increase the width of the bandwidth and to keep the insulation on a frequency band higher.

where C is the coupling coefficient of two output ports. The value Z_A of is chosen arbitrarily and generally, $Z_A = Z_C$ gives maximum bandwidth [7]. When equal power division is required $C = \frac{1}{\sqrt{2}}$, from (2.29), we obtain

$$Z_B = 120.7 \, \Omega \text{ and } Z_A = Z_C = \frac{50}{\sqrt{2}} \, \Omega \text{ [7]}$$

The thickness of the substrate was chosen based on the resulting impedance values after optimization, this is because of the relation between substrate thickness and the width of the microstrip transmission lines. The fabrication limits govern the minimum line width

(maximum impedance) which is set to 0.2 mm (0.15 mm can be accepted in some cases) and the maximum width (minimum impedance) is decided by the quarter wavelength of the top frequency in the bandwidth. The width should ideally not exceed $\frac{\lambda}{8}$ at the top frequency.[6]

The characteristic impedance range of the microstrip line is 20 Ω to 120 Ω . The upper limit is set by production tolerances and the lower limit by the appearance of higher-order modes [2].

2.4.4 Hybrid and Directional couplers Common Applications

Directional couplers are used frequently in dividers, combiners, attenuators ,phase shifters, discriminators, balanced and double-balanced mixers ,balanced amplifiers, and feed networks in antenna arrays.[2]

The branch coupler has the advantage of adjacent output ports which permits combining them in the planar design. For example, a balanced mixer that has the two-branch hybrid and two diodes in output ports 3 and 4 can have one planar output intermediate-frequency (IF) port. Also, the two-branch coupler has a net electrical path length of one wavelength as compared with the ring coupler with 1.5 wavelengths.[2]

Couplers are used in circuits to generate separate signal channels with desirable characteristics . Several examples serve to illustrate such applications. A sum and difference coupler can be used in a monopoles receiver/direction finding system. The input and isolated ports are connected to two identical antennas. A received signal then produces a sum (Σ) and difference (Δ) signals which depend upon the angle of incidence of the receive signal. If the signal is incident from a normal direction, for example, the Δ signal would be zero.[3]

A 3dB hybrid is important in the application of negative resistance amplifiers. Negative resistance devices have a reflection coefficient which is greater than one, i.e., the reflected power exceeds the incident power. The 3dB hybrid can be used to separate the input and the output channels for such devices.[3]

2.5 90° Hybrid Analysis

The hybrid coupler will be characterized by its S matrix which can be found using S-parameter analysis. This matrix will give the coupler's performance at the center frequency, but can also be found across a frequency band to understand the coupler's performance over a bandwidth.

2.5.1 S-Parameters

RF and microwave range test equipment is not capable of measuring the voltages and currents of networks. For the given range, power in inputs (or outputs) of networks is a more convenient characteristic to measure.

In order to account for the phase relationships, it is necessary to think of power as the product of complex values of current and voltage.

Wave processes at the inputs and outputs of networks are usually described by the normalized wave matrices. It is helpful to define normalized voltages, currents, and incident and reflected waves as S matrix.

S-parameters relate incident and reflected voltage waves at the different ports of a microwave network. This relationship is defined to be

$$[V^-] = [S][V^+] \quad (2.31)$$

Where V^- is the reflected voltage wave vector, V^+ is the incident voltage wave vector, and S is the S matrix. The elements of these matrices are all complex numbers to account for both magnitude and phase.

The elements of the S matrix are defined by:-

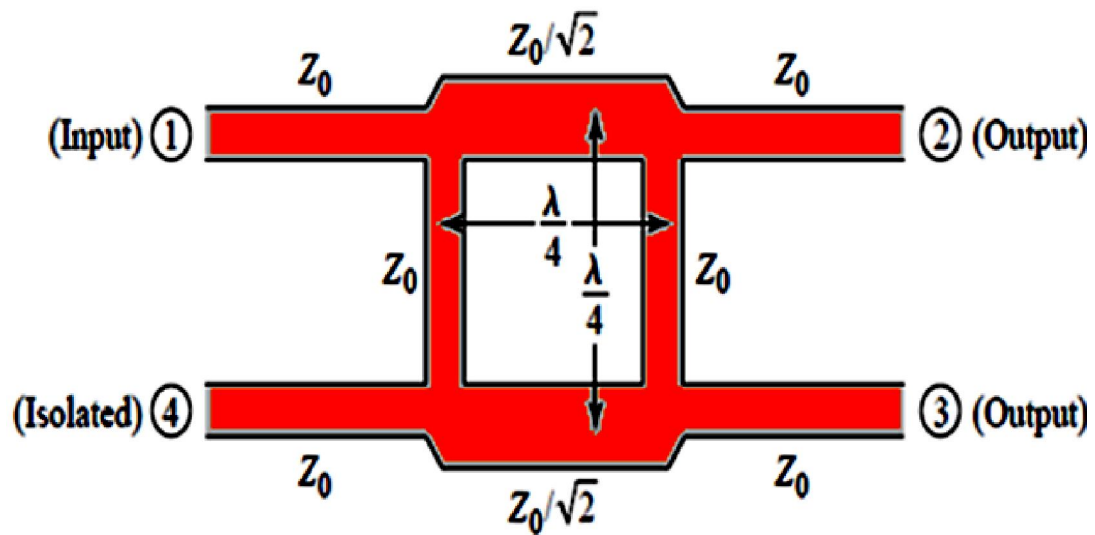
$$[S_{ij}] = \frac{V_i^-}{V_j^+} \text{ } | v_k^+ = 0, k \neq j \quad (2.32)$$

Where i is the port of the reflected wave and j is the port of the incident wave.

2.5.2 Even-Odd Mode Analysis

Quadrature hybrids are 3 dB directional couplers with a 90° phase difference in the outputs of the through and coupled arms. This type of hybrid is often made in microstrip line or stripline form as shown in Figure(2.6) and is also known as a branch-line hybrid. Other 3 dB couplers, such as coupled line couplers or Lange couplers, can also be used as quadrature couplers; these components will be discussed in later sections.

Here we will analyze the operation of the quadrature hybrid using an even-odd mode .[1]



Figure(2.7):Geometry of a branch-line coupler.

We first draw the schematic circuit of the branch-line coupler in normalized form, as in Figure (2.7), where it is understood that each line represents a transmission line with indicated characteristic impedance normalized to Z_0 . [1]

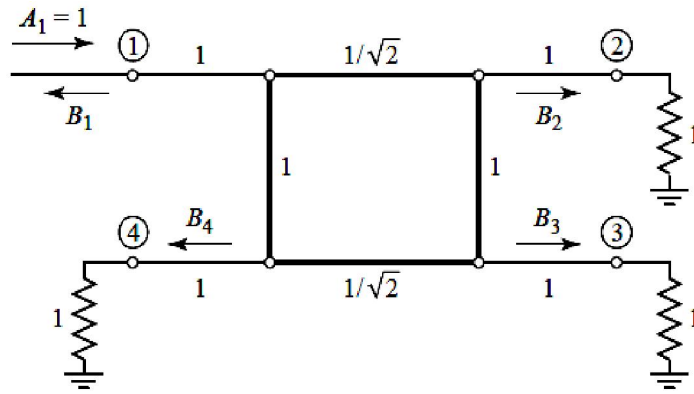


Figure (2.8) Circuit of the branch-line hybrid coupler in normalized form.

The common ground return for each transmission line is not shown.

We assume that a wave of unit amplitude $A_1 = 1$ is incident at port 1.

The circuit of Figure (2.7) can be decomposed into the superposition of an even-mode excitation and an odd-mode excitation [1].

As shown in Figure (2.8). Note that super imposing the two sets of excitations produces the original excitation of Figure (2.8), and since the circuit is linear, the actual response (the scattered waves) can be obtained from the sum of the responses to the even and odd excitations.

Because of the symmetry or anti-symmetry of the excitation, the four-port network can be decomposed into a set of two decoupled two-port networks, as shown in Figure (2.8). Because the amplitudes of the incident waves for these two-ports are $\pm 1/2$.

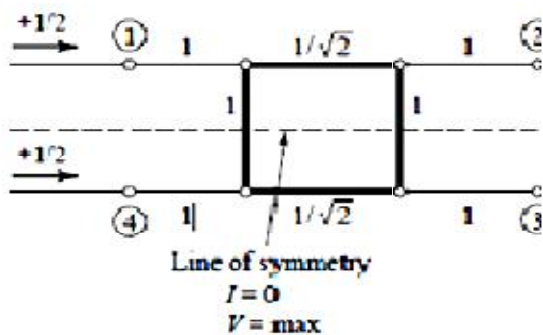


Figure (2.9.a) Even mode (e). Decomposition of the branch-line coupler into even- and odd-mode excitations.

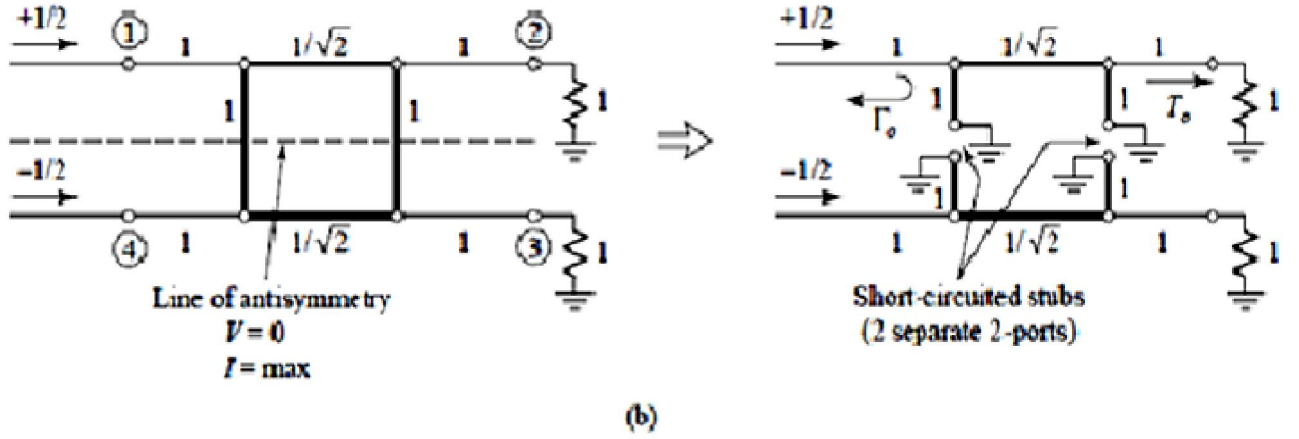


Figure (2.9.b) Odd mode (o). Decomposition of the branch-line coupler into even- and odd-mode excitations.

2.5.2.1 For Even Mode Analysis

In figure(2.9.a), the voltage sources are equivalent. The equivalent sources imply that there is no current at the line of symmetry.

Having no current at this point allows the vertical traces to be represented as open circuited stubs with length $=\lambda/8$. Now, there are two two-port networks (a network with ports 1, 2 and another with ports 3, 4); these two networks are identical which means only one of them needs to be analyzed. The easiest way to find the S-parameters of one of the two-port networks is to use transmission (or ABCD)matrices to cascade the stubs and quarter-wavelength section. Additional information on transmission matrices and the conversion to S-parameters can be found in Appendix A. The transmission matrix between ports 1 and 2 is:-

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = [T_1][T_2][T_1] = \frac{1}{\sqrt{2}} \begin{bmatrix} -1 & j \\ j & -1 \end{bmatrix} \quad (2.33)$$

Where subscript e represents a parameter associated with the even mode analysis, T1 is the transmission matrix for the open circuit $\lambda/8$ length stub and T2 is the transmission matrix of the quarter wavelength section.

T1 is defined to be

$$\begin{bmatrix} 1 & 0 \\ j & 1 \end{bmatrix} \quad (2.34)$$

T₂ is defined to be

$$\begin{bmatrix} 0 & j/\sqrt{2} \\ j/\sqrt{2} & 0 \end{bmatrix} \quad (2.35)$$

converting (2.33) to S- parameters gives the reflection (Γ_e) and transmission (T_e) coefficients for the even mode as:-

$$\Gamma_e = \frac{A + B - C - D}{A + B + C + D} = 0 \quad (2.36)$$

$$T_e = \frac{2}{A + B + C + D} = \frac{-1}{\sqrt{2}} - \frac{j}{\sqrt{2}} \quad (2.37)$$

A,B,C, and D are the elements of the transmission matrix from (2.33).

2.5.2.2 For Odd Mode Analysis

in figure(2.9.b), the two voltage sources are equal in magnitude but out of phase. This odd symmetry of the sources leads to zero voltage at the line of anti symmetry.

The vertical transmission lines in this case are short circuit stubs with length $\lambda/8$. Using the same approach as the even mode, the S-parameters can be found by using the transmission matrices as

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = [T_1] [T_2] [T_1] = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & j \\ j & 1 \end{bmatrix} \quad (2.38)$$

where subscript o represents a parameter associated with the odd mode analysis and T₃ is the transmission matrix for the short circuit $\lambda/8$ length stub. T₃ is defined to be

$$\begin{bmatrix} 1 & 0 \\ -j & 1 \end{bmatrix} \quad (2.39)$$

Converting this result to S parameters gives the reflection (Γ_o) and transmission(T_o) coefficients for the odd mode as

$$\Gamma_o = \frac{A + B - C - D}{A + B + C + D} = 0 \quad (2.40)$$

$$T_o = \frac{2}{A + B + C + D} = \frac{1}{\sqrt{2}} - \frac{j}{\sqrt{2}} \quad (2.41)$$

Where A,B,C, and D are the elements of the transmission matrix from (2.38).

Based on the polarity of the odd and even mode excitations, the total response of the system can be found to be

$$S_{11} = \frac{1}{2}\Gamma_e + \frac{1}{2}\Gamma_o = 0 \quad (2.42)$$

$$S_{21} = \frac{1}{2}T_e + \frac{1}{2}T_o = -\frac{j}{\sqrt{2}} \quad (2.43)$$

$$S_{31} = \frac{1}{2}T_e - \frac{1}{2}T_o = -\frac{1}{\sqrt{2}} \quad (2.44)$$

$$S_{41} = \frac{1}{2}\Gamma_e - \frac{1}{2}\Gamma_o = 0 \quad (2.45)$$

The last four equations show that, at the center frequency, this design will have no reflection at the input port (from (2.42)), no output at the isolation port (from (2.45)), and outputs that are shifted 90 and 180 degrees from the input at the output ports (from (2.43) and (2.44), respectively). The output ports are also shifted 90 degrees in

phase from each other. These are the desired results from a hybrid coupler.

Based on the symmetry of the circuit, the rest of the S- parameters can be deduced. The final S matrix for the hybrid coupler is:-

$$[S] = \frac{-1}{\sqrt{2}} \begin{bmatrix} 0 & 1 & j & 0 \\ 1 & 0 & 0 & j \\ j & 0 & 0 & 1 \\ 0 & j & 1 & 0 \end{bmatrix} \quad (2.46)$$

This S matrix is very idealized. Practically, there will always be some reflection at the input port and some power delivered to the isolated port. Also, there is loss in the

transmission line sections. Together, these losses will cause the power delivered to the output ports to be less than the -3dB that is expected in the ideal case.

2.6 Signal Flow Graph (SFG)

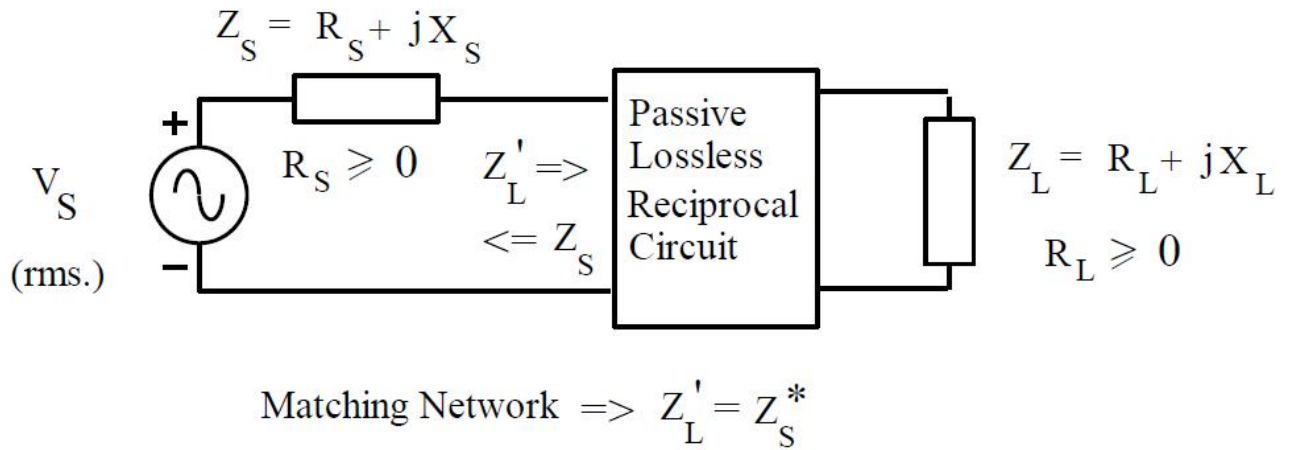
We have seen how transmitted and reflected waves can be represented by scattering parameters, and how the interconnection of sources, networks, and loads can be treated with various matrix representations. In this section we discuss the signal flow graph, which is an additional technique that is very useful for the analysis of microwave networks in terms of transmitted and reflected waves. We first discuss the features and the construction of the flow graph itself, and then present a technique for the reduction, or solution, of the flow graph.[1]

The primary components of a signal flow graph are nodes and branches:

- Nodes: Each port i of a microwave network has two nodes, a_i and b_i . Node a_i is identified with a wave entering port i , while node b_i is identified with a wave reflected from port i . The voltage at a node is equal to the sum of all signals entering that node.[1]
- Branches: A branch is a directed path between two nodes representing signal flow from one node to another. Every branch has an associated scattering parameter or reflection coefficient.[1]

2.6.1 Two-port passive lossless (SGF)

circuit inserted between a source and load can only be conjugate matched at its output if it is conjugate matched at its input, and vice versa. See Figure(2.10)



Figure(2.10) Illustrating a passive, lossless, reciprocal circuit insert between a load and a source to provide a conjugate match

The SFG describing this situation is shown below in figure (2.11) (a.) .

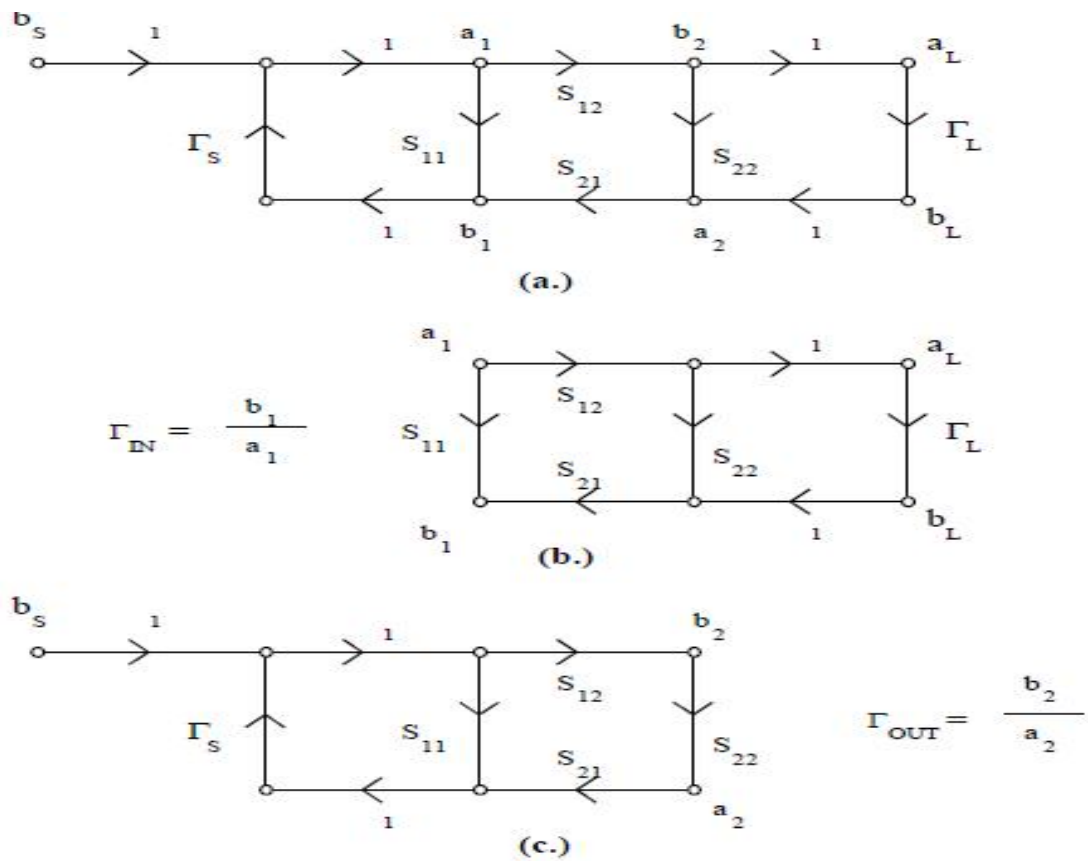


Figure (2.11) : (a.)Signal flow graph showing a circuit inserted between a source and a load, and the (b.) input and (c.) output reflection coefficients.

The input reflection coefficient is defined as the reflection coefficient that exists at the input of the circuit when the load is connected to the output. This is illustrated in figure (2.11) (b.). In this case

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \quad (2.47)$$

The output reflection coefficient is defined as the reflection coefficient that exists at the output of the circuit when the source is connected to the input. This is illustrated in figure (2.11) (c.) and

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \quad (2.48)$$

2.6.2 Four-port (two two-port circuits passive lossless(SGF))

- 1- Because of the symmetry or anti-symmetry of the excitation, the four-port circuit can be decomposed into a set of two decoupled two-port circuits, as shown in figure (2.12). Because the amplitudes of the incident waves for these two-ports are $\pm 1/2$.
- 2- Now, there are two two-port circuits (circuits with ports 1, 2 and another with ports 3, 4); these two circuits are identical which means only one of them needs to be analyzed. The easiest way to find the S-parameters of one of the two-port networks is to use transmission (or ABCD) matrices to cascade the stubs and quarter-wavelength section.
- 3- When two port circuits are cascaded it is often convenient to organized the signal parameters so that the input (port 1) signal parameters, a_1, b_1 are grouped together and output (port 2) parameters, a_2, b_2 are grouped together. This can be accomplished by starting with signal parameters arranged to emphasize their scattering characteristic .
- 4- When considering two circuits which are to be cascaded together to form a new combined circuit. Let X and Y represent the transmission matrices for the respective circuits shown in figure (2.12) (a.) Connecting the circuits yields the composite circuit illustrated in figure (2.12) (b.).

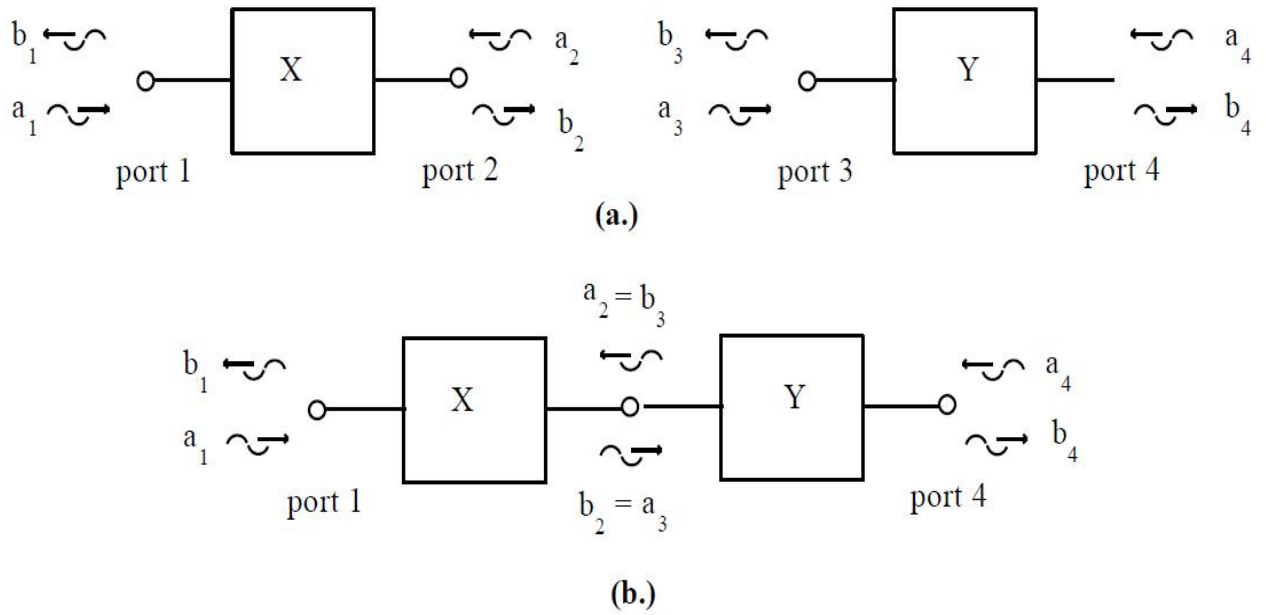


Figure (2.12): (a.) Circuits with a transmission matrices X and Y . (b.) Composite circuit formed by cascading the circuit X and Y .

- 5- Suppose a circuit with S-parameters $\{U_{11}, U_{21}, U_{12}, U_{22}\}$ is cascaded with a circuit with S-parameter $\{V_{11}, V_{21}, V_{12}, V_{22}\}$. Find the s-parameters for the combined circuit using signal flow graph reduction. Then will solve the same problem using transmission matrix approach..
- 6- The signal flow graph for the combined circuit is shown in the figure (2.13) below. The S-parameters for the combined circuit will be:

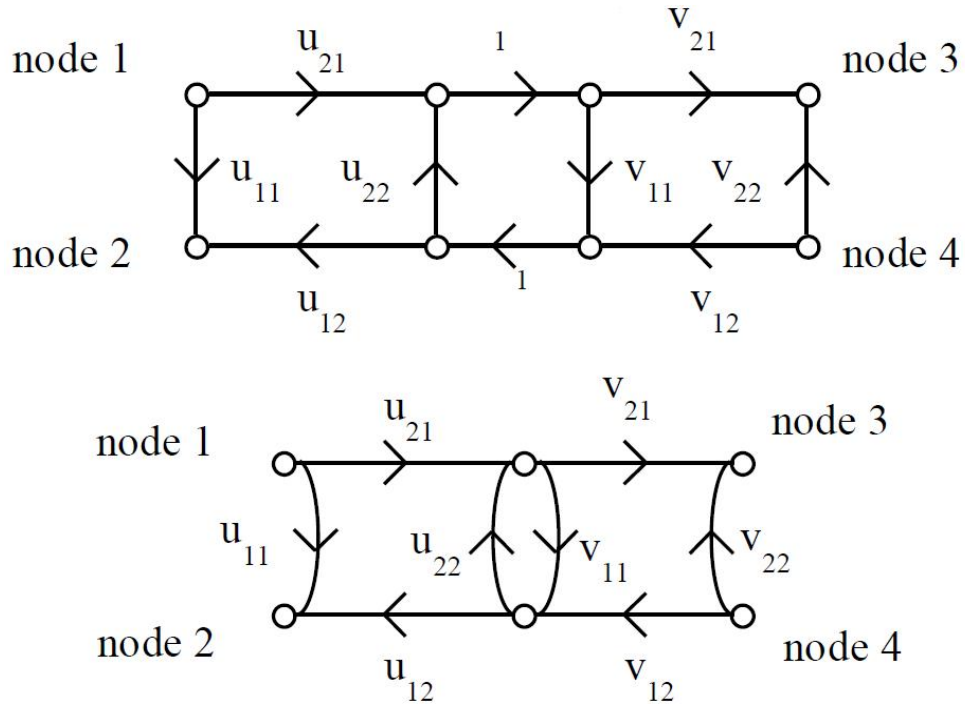
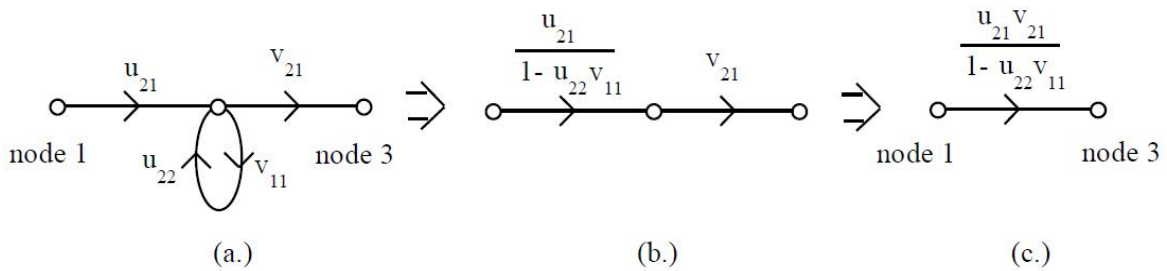


Figure (2.13): The SFG for two cascaded circuits

7- designated by, $\{w_{11}, w_{21}, w_{12}, w_{22}\}$. Determination of the s-parameters for the combined circuit is best undertaken in steps. First w_{21} is determined by recognizing it as a transfer ratio between nodes “1” and “3” if figure (2.13). With node “1” taken as an independent node then the SFG can be reduced as follows in figure (2.14).



Figure

(2.14): The SFG reduction to obtain the transfer ratio representing w_{21} .

$$w_{21} = \frac{U_{21}V_{21}}{1 - U_{22}V_{11}} \quad (2.49)$$

- 8- In a similar manner the SFG of figure (2.12) ,can be reduced to determine w_{21} which is the transfer ratio from node “4” to node “2.” The reduction is illustrated in figure (2.15).

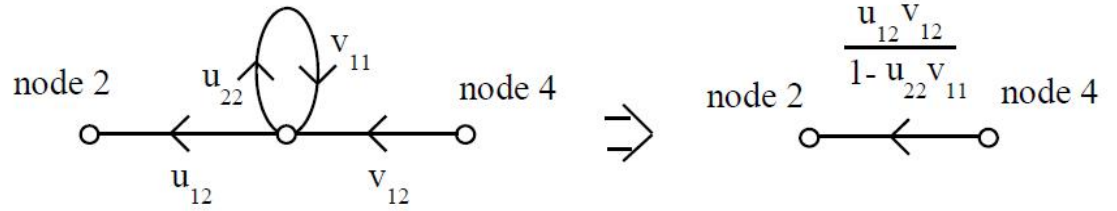


Figure (2.15): The SFG reduction to obtain the transfer ratio representing w_{12} .

$$w_{12} = \frac{U_{12}V_{12}}{1 - U_{22}V_{11}} \quad (2.50)$$

- 9- The s-parameter, w_{11} , is found from the transfer ration from node “1” to node “2.” The SFG is shown below in figure (2.16).

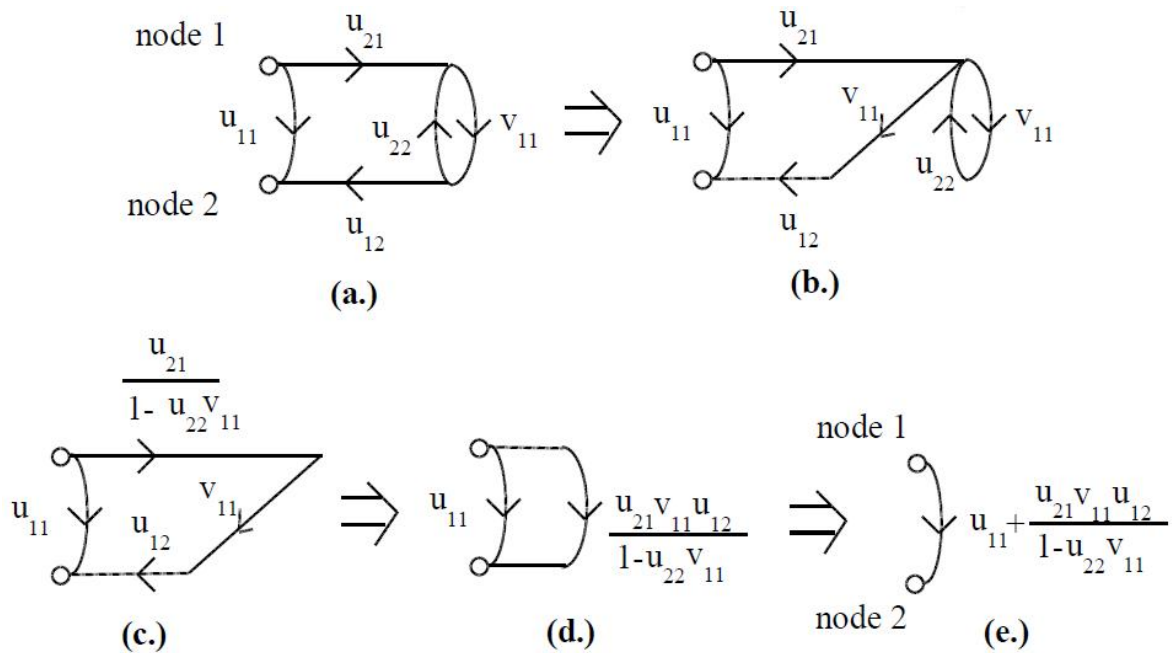


Figure (2.16) : Sequence of SFG reductions to obtain w_{11} .

$$w_{11} = U_{11} + \frac{U_{21}V_{11}U_{12}}{1 - U_{22}V_{11}} \quad (2.51)$$

- 10- In a similar manner w_{22} is found.

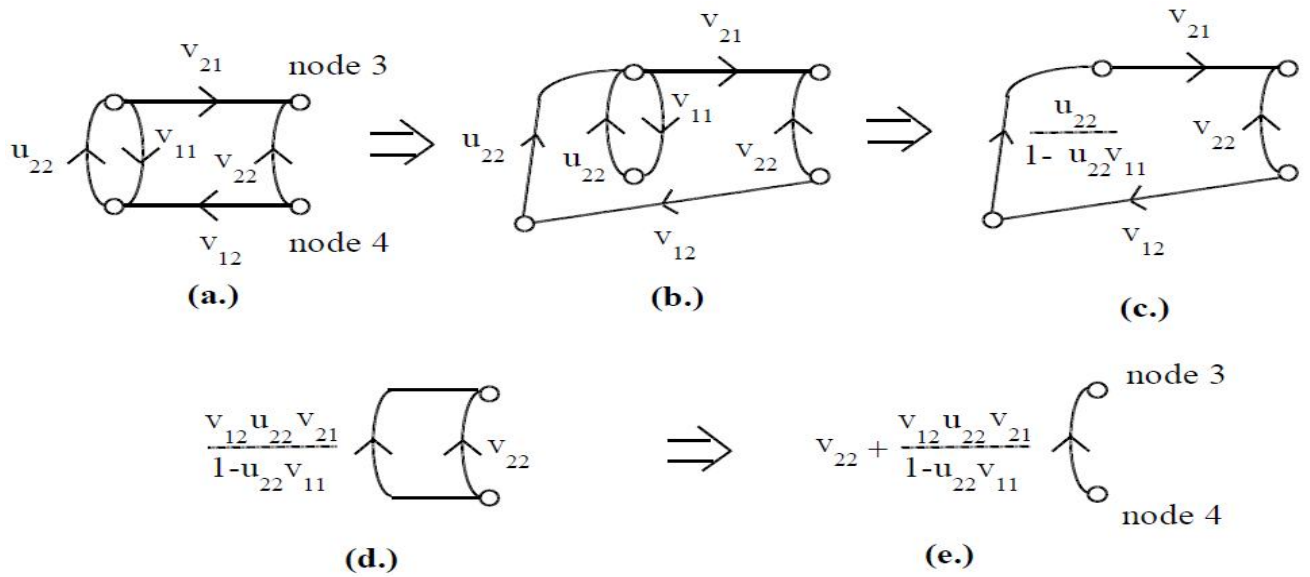


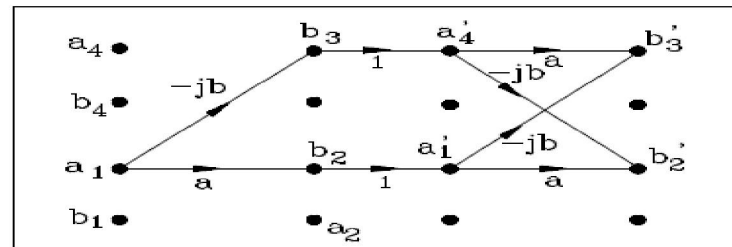
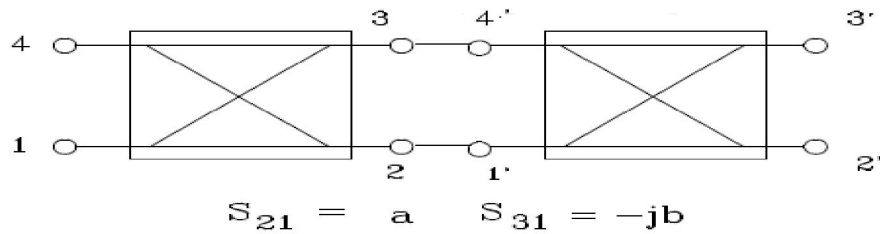
Figure (2.17): Sequence of SFG reductions to obtain W_{22} .

$$w_{22} = v_{22} + \frac{V_{12}U_{22}V_{21}}{1 - U_{22}V_{11}} \quad (2.52)$$

2.6.3 Cascaded Directional Coupler (SFG)

This arrangement in figure(2.18) is useful if tight coupling is difficult to achieve with a single coupler, e.g. parallel line coupler.[3]

Two cascaded 8.34 dB couplers = 3dB coupler.



$$\begin{aligned} S_{21} &= a^2 - b^2 = 1/\sqrt{2} \Rightarrow a = -.68 \text{ dB} \\ S_{31} &= -j2ab = -j1/\sqrt{2} \Rightarrow b = -8.34 \text{ dB} \end{aligned}$$

Figure (2.17.a): The SFG Cascaded Directional Coupler .

2.7 Parameters Effects on The Phase Shift and Bandwidth

To study phase shift we focus on significant points :-

2.7.1 Amplitude Balance

This terminology defines the power difference in dB between the two output ports of a 3 dB hybrid. In an ideal hybrid circuit, the difference should be 0 dB. However, in a practical device the amplitude balance is frequency dependent and departs from the ideal 0 dB difference.[4]

2.7.2 The Coupling Balance

For S_{21} and S_{31} must be equal or less than 1dB which means ± 0.5 dB.

Coupling flatness is the amount (in decibels) of the maximum variation in coupling that may be expected over a specified frequency range . [2]

2.7.3 Phase Balance

The phase difference between the two output ports of a hybrid coupler should be 0, 90, or 180 degrees depending on the type used. However, like amplitude balance, the phase difference is sensitive to the input frequency and typically will vary a few degrees.[4]

2.7.4 Power Split

For directional couplers is calculated as the difference (in decibels) between the coupling and insertion loss.

The power split in the two-branch hybrid, [$C_{31} - C_{21} = 0$] , unlike the hybrid-ring power split , does not depend on losses. This is generally characteristic of full symmetry couplers.[2]

2.7.5 Matching

Matching is determined by voltage standing-wave ratio (VSWR) or reflection coefficient of the input (or output) port, while the other ports are connected with reflection less terminations.

2.7.6 Bandwidth

Bandwidth is the range of frequencies for which a parameter falls within a specified limit with respect to certain characteristics and a hundred percent of bandwidth from :-

$$BW = F_U - F_L \quad (2.53)$$

Where F_U the upper limit for frequency, and F_L the lower limit frequency .

2.8 Parameters of Lossless Transmission Line

There are Parameters of lossless transmission line which effect on the medium performance over the variants frequency band and bandwidth.

2.8.1 Wave Reflection in General Medium

This is a general solution for reflection and transmission of a normally incident wave at the interface of an arbitrary material,

The propagation constant is[1] :

$$\gamma = \alpha + j\beta = j\omega\sqrt{\mu\epsilon}\sqrt{1 - j\sigma/\omega\epsilon} \quad (2.54)$$

The transmission coefficient and the reflection coefficient relationship is:-

$$1 + \Gamma = T \quad (2.55)$$

Where T is the transmission coefficient and where Γ is the reflection coefficient .

Then the intrinsic impedance relationship of the loss medium:-

$$\frac{1 - \Gamma}{\eta_0} = \frac{T}{\eta} \quad (2.56)$$

$$\eta = \frac{j\omega\mu}{\gamma} \quad (2.57)$$

Then solving these equations for the reflection and transmission coefficients gives:-

$$\Gamma = \frac{\eta - \eta_0}{\eta + \eta_0} \quad (2.58)$$

$$T = 1 + \Gamma = \frac{2\eta}{\eta + \eta_0} \quad (2.59)$$

2.8.2 Lossless Medium

The propagation constant in this case is purely imaginary and can be written

as :

$$\gamma = j\beta = j\omega\sqrt{\mu\epsilon} = jK_0\sqrt{\mu_r\epsilon_r} \quad (2.60)$$

Where $K_0 = \omega\sqrt{\mu_0\epsilon_0}$ is the propagation constant (wave number) of a plane wave in free space .The wavelength in the dielectric is

$$\lambda = \frac{2\pi}{\beta} = \frac{2\pi}{\omega\sqrt{\mu\epsilon}} = \frac{\lambda_0}{\sqrt{\mu_r\epsilon_r}} \quad (2.61)$$

the phase velocity is

$$v_p = \frac{\omega}{\beta} = \frac{1}{\sqrt{\mu\epsilon}} = \frac{c}{\sqrt{\mu_r\epsilon_r}} \quad (2.62)$$

(slower than the speed of light in free-space) and the intrinsic impedance of the dielectric is

$$\eta = \frac{j\omega\mu}{\gamma} = \sqrt{\frac{\mu}{\epsilon}} = \eta_0 \sqrt{\frac{\mu_r}{\epsilon_r}} \quad (2.63)$$

For this lossless case, η is real, so both Γ from (2.58) and T from (2.59) are real, and E and H are in phase with each other in both regions.

2.8.3 Wave Propagation on Transmission Line

In general transmission line propagation, including loss effects, and it was seen that the propagation constant and characteristic impedance were complex by :-

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (2.64)$$

a characteristic impedance, Z_0 , can be defined as

$$Z_0 = \frac{(R + j\omega L)}{\gamma} = \sqrt{\frac{(R + j\omega L)}{(G + j\omega C)}} \quad (2.65)$$

we find that the wavelength on the line is

$$\lambda = \frac{2\pi}{\beta} \quad (2.66)$$

and the phase velocity is

$$v_p = \frac{\omega}{\beta} = \lambda f \quad (2.67)$$

2.8.4 Wave Propagation in Lossless Line

In many practical cases, however, the loss of the line is very small and so can be neglected, resulting in a simplification of the results [1].

Setting $R = G = 0$ in (2.64) gives the propagation constant as

$$\gamma = \alpha + j\beta = j\omega = \sqrt{LC} \quad (2.68)$$

or

$$\beta = \omega\sqrt{L} \quad (2.69)$$

Where, $\alpha=0$

As expected for a lossless line, the attenuation constant α is zero. The characteristic impedance :-

$$z_0 = \sqrt{\frac{L}{C}} \quad (2.70)$$

The wavelength is

$$\lambda = \frac{2\pi}{\beta} = \frac{2\pi}{\omega\sqrt{\mu\epsilon}} \quad (2.71)$$

and the phase velocity is

$$v_p = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}} \quad (2.72)$$

2.9 Stub Matching

Stubs can be used to match a load impedance to the transmission line characteristic impedance. The stub is positioned a distance from the load. This distance is chosen so that at that point the resistive part of the load impedance is made equal to the resistive part of the characteristic impedance by impedance transformer action of the length of the main line. The length of the stub is chosen so that it exactly cancels the reactive part of the presented impedance. That is, the stub is made capacitive or inductive according to whether the main line is presenting an inductive or capacitive impedance respectively. This is not the same as the actual impedance of the load since the reactive part of the load impedance will be subject to impedance transformer action as well as the resistive part. The matching circuits, which incorporate mixed lumped and transmission line elements, are widely used both in hybrid and monolithic design technique.[5]

Matching stubs can be made adjustable so that matching can be corrected on test , there is two types of this stub:-

2.9.1 Single Stub Tuning

Another popular matching technique uses a single open-circuited or short-circuited length of transmission line (a stub) connected either in parallel or in series with the transmission feed line at a certain distance from the load, as shown in Figure(2.18.a.b). Such a single-stub tuning circuit is often very convenient because the stub can be fabricated as part of the transmission line media of the circuit, and lumped elements are avoided. Shunt stubs

are preferred for microstrip line or stripline, while series stubs are preferred for slot line or coplanar waveguide.[1] In single-stub tuning the two adjustable parameters are the distance, d , from the load to the stub position, and the value of susceptance or reactance provided by the stub. For the shunt-stub case is shown in Figure(2.18.a) , the basic idea is to select d so that the admittance, Y , seen looking into the line at distance d from the load is of the form $Y_0 + j B$. Then the stub susceptance is chosen as $-j B$, resulting in a matched condition. For the series-stub case is shown in Figure(2.18.b) , the distance d is selected so that the impedance, Z , seen looking into the line at a distance d from the load is of the form $Z_0 + j X$. Then the stub reactance is chosen as $-j X$, resulting in a matched condition.[1]

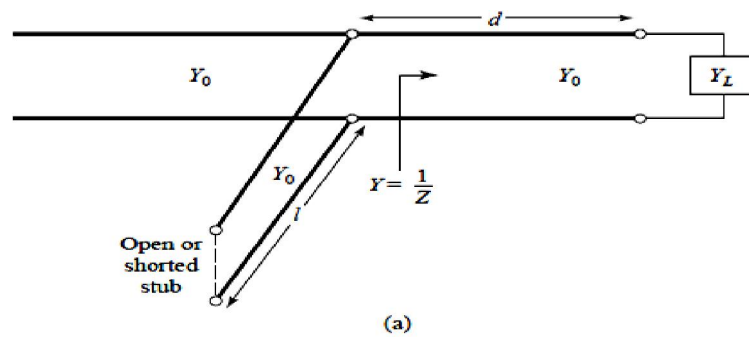


Figure (2.18.a) Shunt stub. Single-stub tuning circuits.

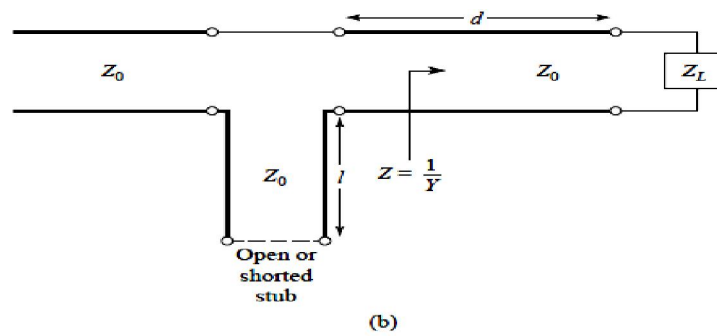


Figure (2.18.b) Series stub. Single-stub tuning circuits.

For transmission line media such as microstrip or stripline, open-circuited stubs are easier to fabricate since a via hole through the substrate to the ground plane is not needed. For

lines like coax or waveguide, however, short-circuited stubs are usually preferred because the cross-sectional area of such an open-circuited line may be large enough (electrically) to radiate, in which case the stub is no longer purely reactive.[1]

2.9.2 Double Stubs Tuning

The single-stub tuner of the previous section is able to match any load impedance (having a positive real part) to a transmission line, but suffers from the disadvantage of requiring a variable length of line between the load and the stub. This may not be a problem for a fixed matching circuit, but would probably pose some difficulty if an adjustable tuner was desired.[1]

In this case, the double-stub tuner, which uses two tuning stubs in fixed positions, can be used. Such tuners are often fabricated in coaxial line with adjustable stubs connected in shunt to the main coaxial line.[1]

We will see, however, that a double-stub tuner cannot match all load impedances.[1]

The double-stub tuner circuit is shown in Figure(2.19.a) , where the load may be an arbitrary distance from the first stub. Although this is more representative of a practical situation, the circuit of Figure (2.19.b), where the load $Y'L$ has been transformed back to the position of the first stub, is easier to deal with and does not lose any generality.

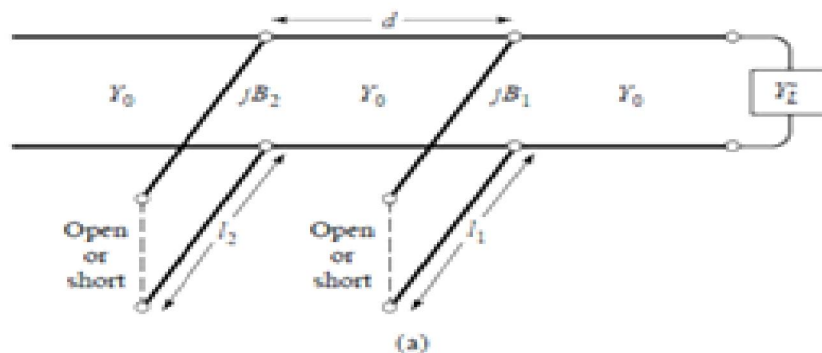


Figure (2.19.a) Double-stub tuning. Original circuit with the load an arbitrary distance from the first stub.

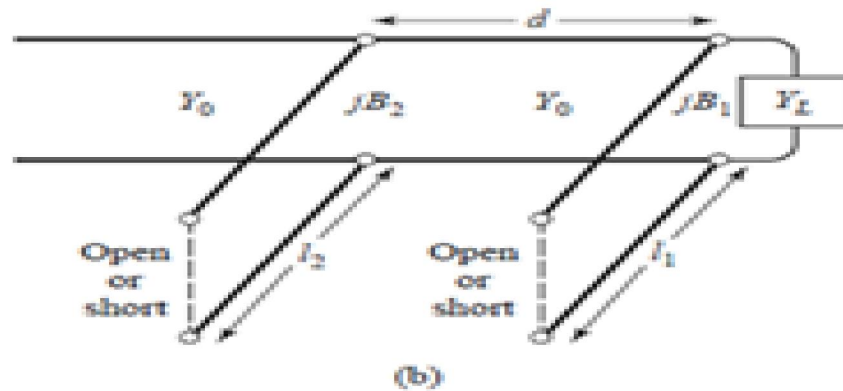


Figure (2.19.b) Double-stub tuning. Equivalent circuit with the load transformed to the first stub.

2.10 The Smith Chart

The Smith chart, shown in figure (2.20), is a graphical aid that can be very useful for solving transmission line problems. Although there are a number of other impedance and reflection coefficient charts that can be used for such problems.[1]

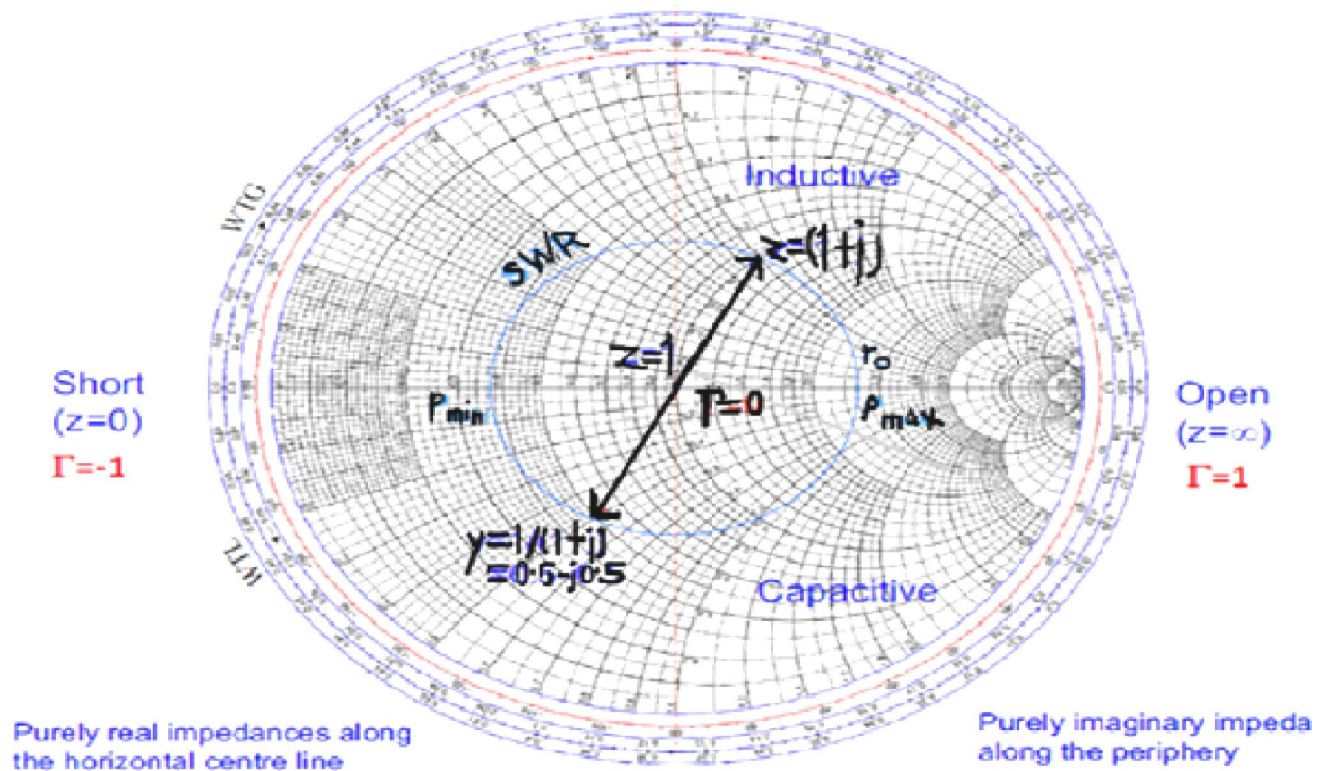


Figure (2.20) Smith Chart

2.10.1 Using the Smith Chart

1. First normalize the load impedance by dividing by the characteristic impedance, and find this point on the chart.
2. When working in terms of reactance, an inductive load will be located on the top half of the chart, a capacitive load on the bottom half. It's the other way around when working in terms of susceptance.
3. Draw a straight line from the center of the chart through the normalized load impedance point to the edge of the chart.
4. Anchor a compass at the center of the chart and draw a circle passing through the normalized load impedance point. Points along this circle represent the normalized impedance at various points along the transmission line.
5. Clockwise movement along the circle represents movement from the load toward the source, with one full revolution representing $1/2$ wavelength as marked on the outer edge of the Smith chart. The two points where the circle intersects the horizontal axis are the voltage maxima (right) and the voltage minima (left).
6. The point opposite the impedance (180° around the arc) is the admittance.
7. The reason admittance (or susceptance) is useful is because admittances in parallel are simply added.
(Admittance is the reciprocal of impedance; susceptance is the reciprocal of reactance.). See figure (2.21).

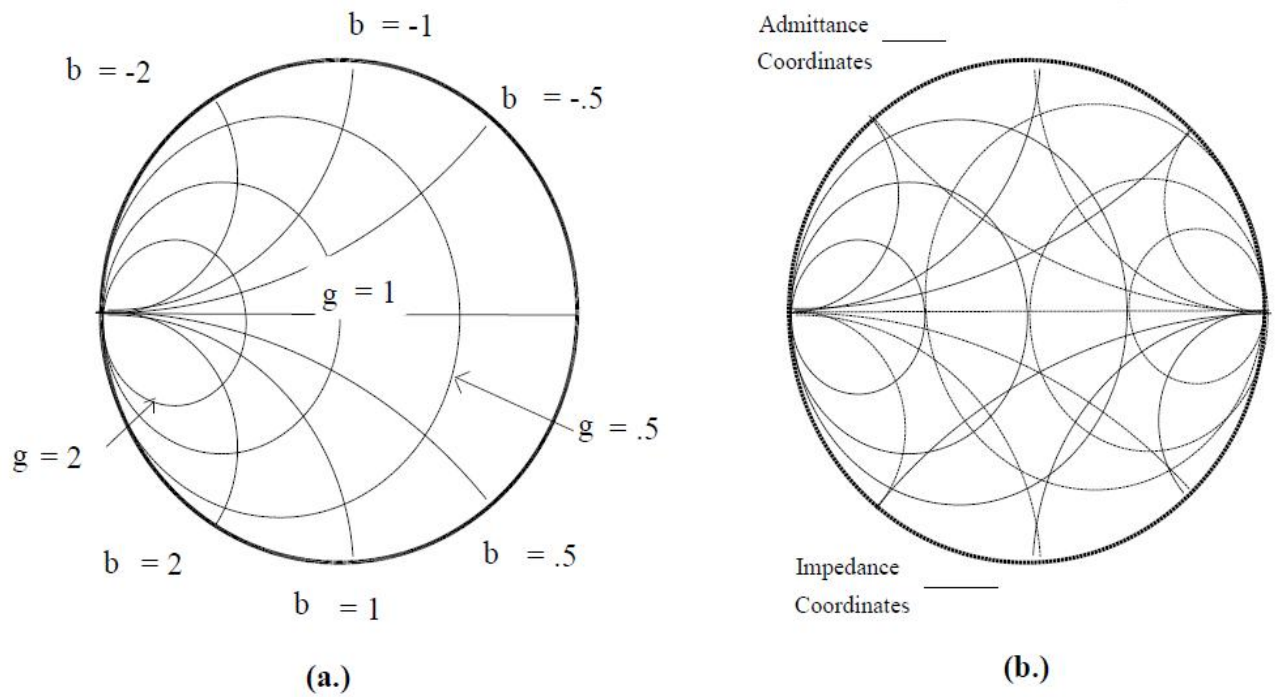


figure (2.20): (a.) The circles of constant conductance and constant suseptance.
 (b.) A Y-Z Smith Chart shows both admittance and impedance circles

Chapter 3

Design of 90° Hybrid

3.1 Introduction

Design method of 90 degree hybrid using microstrip technology will be demonstrated in details from theoretical specification microstrip until calculation lengths and widths of transmission line by calculator and compared with that results by calculation of software (linecalc ADS).

Also this chapter handled in details the enhancement methods using designs by smith chart ADS.

The single stub is first method , designs by smith chart ADS by the way hybrid 90 is all ports matched , but in this design single stub of same ZL the hybrid 90 , and constructed on port 2 and port 3 outputs .The extended TL is second method , designs by smith chart ADS.

3.2 DESIGN of 90° HYBRID

The branch-line coupler consists of two transmission lines linked by two or more transverse $\lambda/4$ branches located at $\lambda/4$ distance one from each other. It can be realized in both waveguide and printed circuit (i.e.microstrip, stripline, CPW, etc.) technologies.[8]

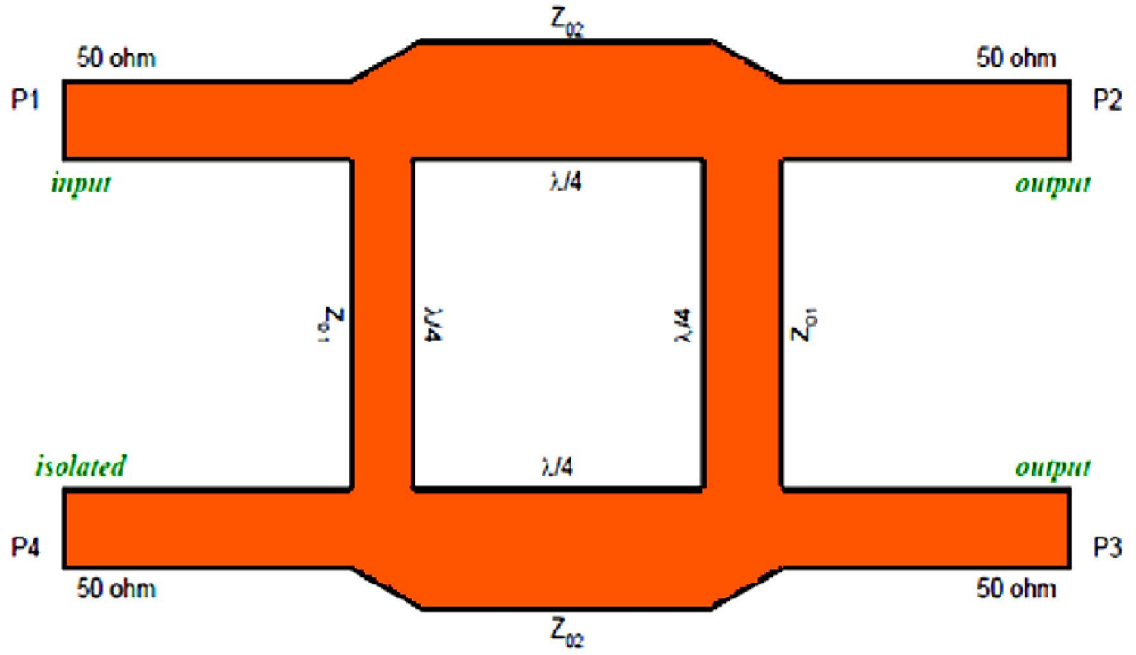


Figure (3.1) The branch-line coupler consists.

As figure (3.1) indicated there are two pairs of quarter-wave transmission line (TL) sections in 90 ° branch- line hybrid where each pair has a different characteristic impedance than the input or output transmission lines.[9]

The design equations for general 90 ° branch -line hybrid coupler with a coupling factor (C in dB) are given by [9] :-

$$C(\text{dB}) = -10 \log_{10} \left[1 - \left(\frac{Z_{02}}{Z_0} \right)^2 \right] \quad (3.1)$$

$$Z_{01} = \frac{Z_{02}}{\sqrt{1 - (Z_{02}/Z_0)^2}} \quad (3.2)$$

To calculate a 3 dB coupler's impedance value for 50 Ω input and output transmission lines then by equation (3.1)

$$C(\text{dB}) = 3 = -10 \log_{10} [1 - (Z_{02}/Z_0)^2]$$

$$\text{Where } \rightarrow Z_{02} / Z_0 = 1/\sqrt{2}$$

Thus $Z_{o2} = 50/\sqrt{2} = 35.36\Omega$

And by using Equation (3.2) we get $Z_{o1} = \frac{35.36}{\sqrt{1-(1/\sqrt{2})^2}} = 50\Omega$.

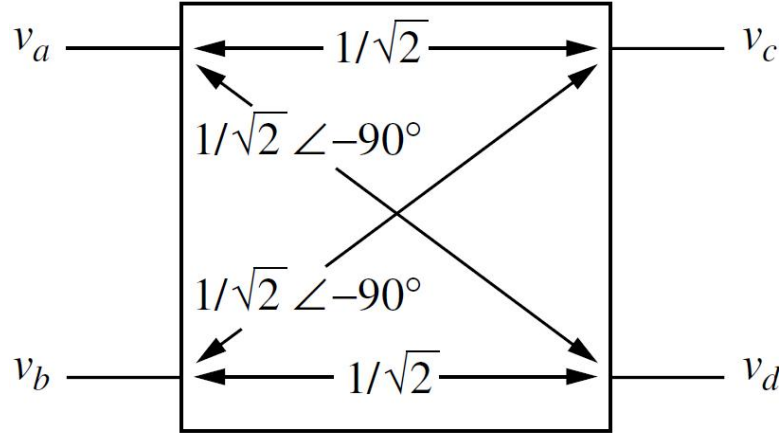


Figure (3.2) Ideal transfer characteristic for 90° hybrid coupler.

Figure (3.2) shows the ideal transfer characteristic of a 90° hybrid. There will be, in addition, a time delay that produces the same phase shift in each of the four paths without altering the ideality of the hybrid.

In practice, due to the quarter-wave length requirement, the bandwidth of a branch-line hybrid is limited to 10%–20%. However, as with multisession matching transformers and multirole directional couplers, the bandwidth of a branch-line hybrid can be increased to a decade or more by using multiple sections in cascade. In addition, the basic design can be modified for unequal power division and/or different characteristic impedances at the output ports. Another practical point to be aware of is the fact that discontinuity effects at the junctions of the branch-line coupler may require that the shunt arms be lengthened by 10°–20°.[1]

3.3 The Microstrip Line

The micro strip line, or simply micro strip, is the most common transmission line employed in microwave and RF engineering. This is because of a number of advantages such as its simplicity of manufacturing, low cost, low weight and excellent integrability with active devices. As shown in Figure (3.3), it is a planar transmission line like the stripline, but, in contrast, it is an open structure; it is not fully embedded in a dielectric but printed on top of a dielectric substrate metalized on the opposite side. The EM field guided by the metal strip propagates in an inhomogeneous medium consisting of the dielectric substrate and the air on top of it.[9]

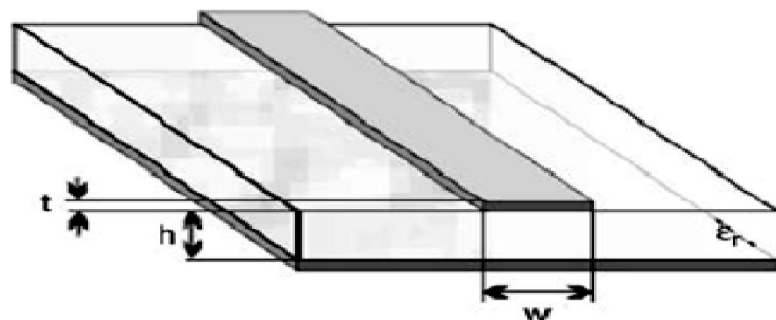


Figure (3.3) The microstrip line

In actuality, the exact fields of a microstrip line constitute a hybrid TM-TE wave and require more advanced analysis techniques than we are prepared to deal with here. In most practical applications, however, the dielectric substrate is electrically very thin ($d \ll \lambda$), and so the fields are quasi-TEM. In other words, the fields are essentially the same as those of the static (DC) case. Thus, good approximations for the phase velocity, propagation constant, and characteristic impedance can be obtained from static, or quasi-static in figure(3.4), solutions.[1]

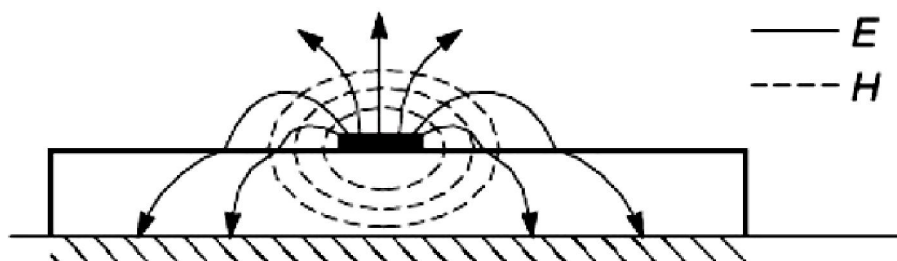


Figure (3.4) quasi-static.

3.3.1 Effective Dielectric Constant

Because some of the fields exist in the dielectric and some exist in the air, the structure cannot support a true TEM wave. However, in practical applications, when the substrate is much smaller than a wavelength, a quasi-TEM mode exists (i.e. the fields in the cross-section behave approximately like the static case). Using static results, approximations for phase velocity and propagation constant can be written as:-

$$v_{ph} = \frac{c}{\sqrt{\epsilon_e}} \quad (3.3)$$

$$\beta = k_0 \sqrt{\epsilon_e} \quad (3.4)$$

where k_0 is the free space propagation constant, v_{ph} is the phase velocity, and ϵ_e is the effective dielectric constant. These equations both assume that the material is non-magnetic (i.e. μ is equal to μ_0).

The effective dielectric constant accounts for some of the fields being in the air and the rest being in the dielectric substrate. Its value is between 1 (the relative dielectric constant for air) and ϵ_r (the dielectric constant of the substrate). The effective dielectric constant is a function of the height of the substrate and the width of the conductor [1]. The approximation for ϵ_e at low frequency is given as

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} + \frac{1}{\sqrt{1 + 12d/w}} \quad (3.5)$$

where d is the height of the substrate and W is the width of the conducting strip.

The ratio of W/d can be found for a given characteristic impedance, Z_0 , and a relative dielectric constant to be

$$\frac{w}{d} = \left\{ \frac{\left[\frac{8e^A}{e^{2A}-2} \right]}{2/\pi [B - 1 - \ln(2B - 1) + \epsilon_r - \frac{1}{2\epsilon_r}] [\ln(B - 1) + 0.39 \frac{0.61}{\epsilon_r}]} \right\} \quad (3.6)$$

Where

$$A = \frac{50}{60} \sqrt{\frac{\epsilon_r + 1}{2}} + \frac{\epsilon_r - 1}{\epsilon_r + 1} \left(0.23 + \frac{0.11}{\epsilon_r} \right) \quad (3.7)$$

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}} \quad (3.8)$$

Then among all planar transmission lines, microstrip line has gained much popularity and importance in microwave planar circuit technology, and so we will consider and analyze microstrip lines in this section. A microstrip is a transmission line consisting of a strip of conductor of thickness(t),width(w),and a ground plane separated by a dielectric medium of thickness(h).

3.3.2 Technical Specification of 90° Hybrid

The technical specifications of the hybrid are listed in Table (3.1). Some of these parameters are with a certain tolerance.

Table (3.1) Technical Specification of Design 90° Hybrid

Parameter	Value
Center Frequency	f = 2.4 GHz
Input/output Matching	50Ω
Amplitude Balance	Closed to 0dB
Return Loss	Expected lower than -12dB
Trans. Coefficient Magnitude	C<1

Coupling balance in dB	Must be equal or less than 1dB which means ± 0.5 dB.
FR-4 substrate	$\epsilon_r = 4.4$
Substrate thickness	$h = 1.58$ mm
Loss tangent	$\tan \delta = 0.001$
Thickness	0.035 mm
Reciprocal	$(S_{ij} = S_{ji})$

3.3.3 90° Hybrid Coupler Calculation

From figure (3.1) for a 3dB coupler

$$Z_{01} = Z_0 = 50 \text{ ohm}$$

$Z_{02} = Z_0 / 2 = 35.36$ ohm Calculate the length L and W of each element by the basis of equations of microstrip(3.4), (3.5), (3.6) ,(3.7)and (3.8).

To find the L1 and W1 for $Z_0 = 50 \Omega$ then, from same equations will find the L2 and W2 for $Z_0 / 2 = 35.36$ ohm.

Table (3.2) Comparison of dimensions of the hybrid by normal calculation and Line Cal. in ADS software.

Impedance Value	(L mm) Calculated	(L mm) line Cal. (ADS)	(W mm) Calculated	(W mm) Line Cal.(ADS)
$Z_0 = 50 \Omega$	L1 16.682	L1 16.718	W1 3.02096	W1 2.9819
$Z_0 / \sqrt{2} = 35.36 \Omega$	L2 13.6535	L2 13.621	W2 5.15252	W2 5.119

3.4 Enhancement Methods

In order to improve the phase and bandwidth of design 90° hybrid at the two output ports a single stub will be first method .There are methods for implementation of stubs are applied at the branch line coupler to increase the frequency bandwidth and to obtain different phase shifts between outputs as ideally as 90 degree.

Those methods have important points which are considered in this thesis:-

- 1- Methods have been studied to modify the conventional coupler structure to become a wide-band coupler , e.g., adding open stubs at the symmetry-planes of the coupler, splitting a low impedance line into two paralleled high-impedance lines, and connecting an open stub at the branching junctions .
Although these methods can avoid low- impedance lines, they result in reduction of the coupler bandwidth .[10]
- 2- Putting a matched network, i.e., a series transmission line and a parallel open stub, in the junction of the coupler can improve the bandwidth of a branch-line coupler.[10]
- 3- Stubs create reflection zeros at equal distance on both sides of the original perfectly matched points which can be seen in the resulting plots or with a Smith chart representation.[6]
- 4- Type of stub determines type matching circuits , if The stub contains open-circuited terminated and have series resistance and capacitance(SRC) load. It is sufficient to use a fast linear optimization process, which will take only a few minutes to complete the matching circuit design.[5]

The second method which will be designed is extended transmission line at the two output ports (open circuit terminated).

For two reasons choose open circuit terminated:

- 1- Open-ends are encountered any time a Microstrip is open terminated. Typical devices where open ends are encountered are Microstrip Filters and Matching Stubs.[11]
- 2- At the open-end of a Microstrip line with a width of w , the fields do not stop abruptly, but extend slightly further because of the effect of fringing field. This effect can be modeled either with an equivalent shunt capacitance C , or with an equivalent length of transmission line Δl . [11]

3.4.1 Single Stub Design

Stubs are shorted or open circuit lengths of transmission line which produce a pure reactance at the attachment point. Any value of reactance can be made, as the stub

length is varied from zero to half a wavelength. A stub consists of a side section of line attached in series with (or in shunt with) the main transmission line, at a point to be determined. The length of the line, and the point of attachment, both need to be calculated from the load impedance, smith chart is the best method to calculate and design any type of stub. The stub needs to be nearly lossless; it may be open circuit, or short circuit, or indeed terminated in a pure reactance.

In practice, usually shunt stubs, short stub for waveguides, open stub for microstrip.

Open or shorted stub (TL).

Z_{in} = pure reactance.

Y_{in} = pure susceptance, if $l < \lambda / 4$, shorted stub is inductive open stub is capacitive.

3.4.1.1 Steps of Single Stub Design

The type stub is parallel (shunt single stub) with the line. Where the stub contains open-circuited terminated and have series resistance and capacitance (SRC)load.

Steps of design by smith chart utility by ADS2009 software and find two solutions using open circuited stubs and choose appropriate one.

1- Launch ADS, and create a new project.

2- Place a term (S-Par palette) and series RC load  (Lumped Components palette) in the schematic. ($C = 0.9$ pf, $R = 51.75 \Omega$).

3- From the palette menu, select Smith Chart Matching , then click Smith Chart symbol and place it on the schematic. A message will appear saying that you must first use a certain utility for it to work. Next, wire the term to the smith chart, and smith chart to load, see figure (3.4. a).

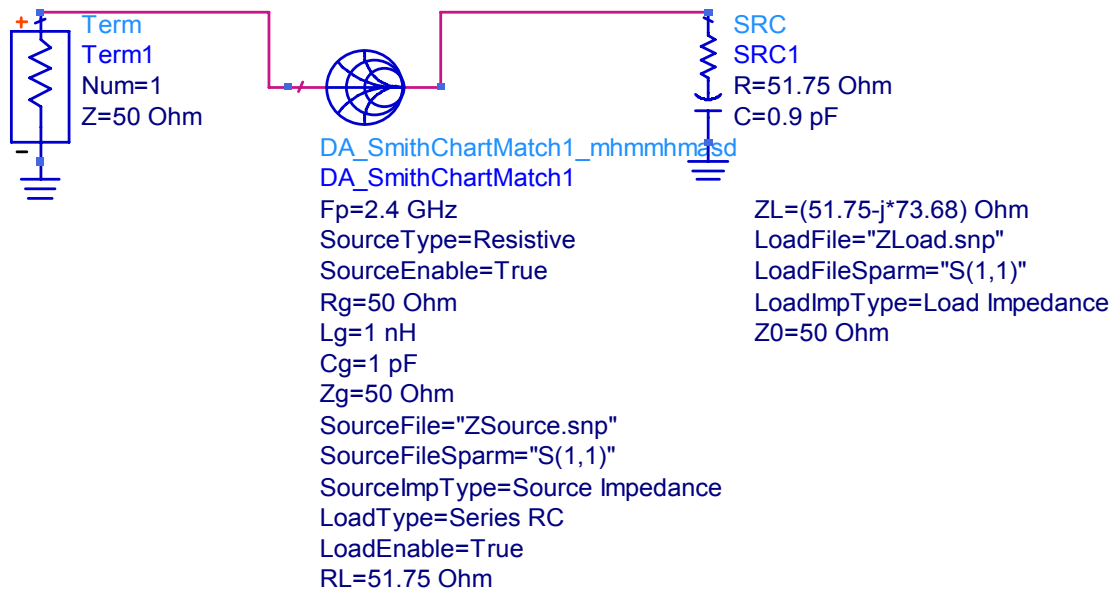
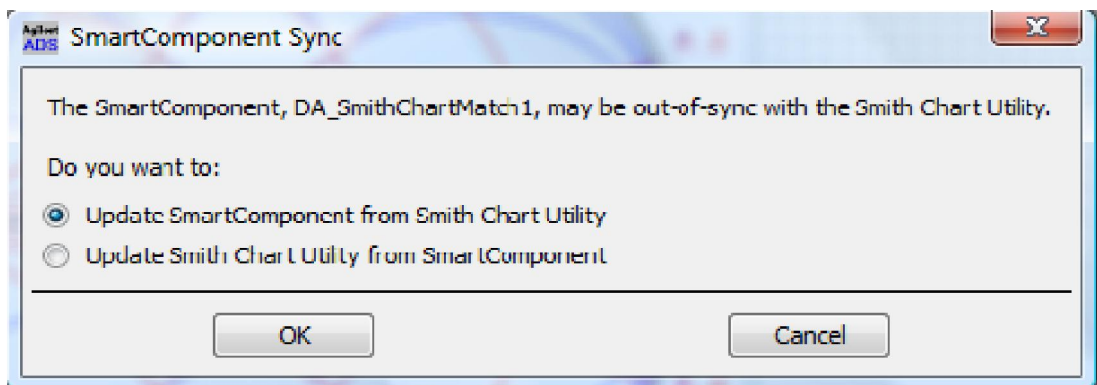


Figure (3.4.a) smith chart of single stub.

- 4- Select the Smith Chart, and go to Tools/Smith Chart. The following window will pop up(see figure(3.4.b)). Select the first option, and hit ok.



Figure(3.4.b) Option window.

The following window(see figure(3.4.c)).Using the drawing area (DA) to design single stub. Steps are relatively simple, first define normalized source, then normalized load impedance. Finally you want to add stubs or T-lines to make the load 'travel' to the source, accomplish this with the following steps(see figure (3.4.c)).

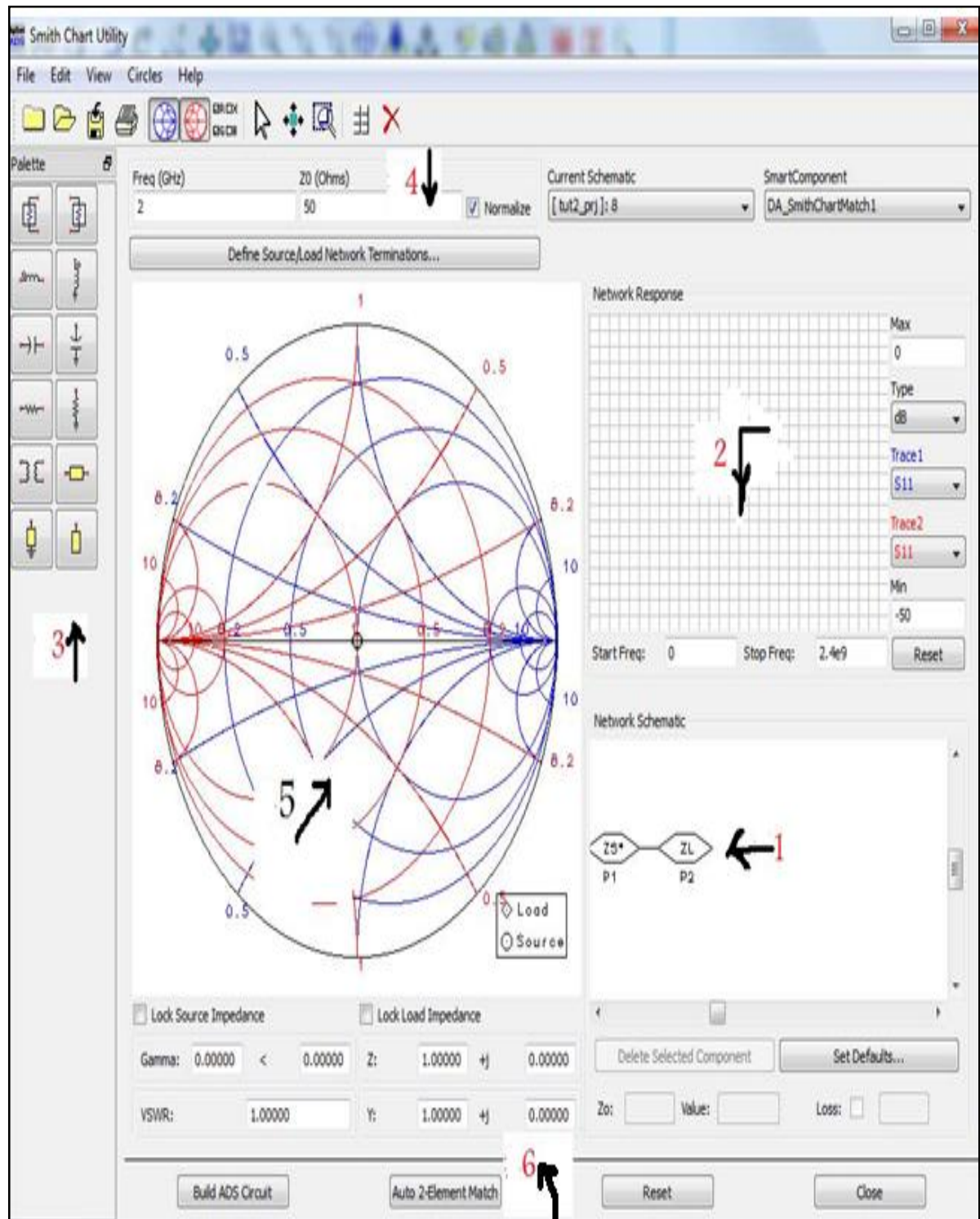


Figure (3.4.c) Drawing area(DA) of smith chart.

- 1) This is your circuit. Right now it has only a source and load, represented by hexagons. As we add more components, more items will appear.
- 2) This is the plot window. Here is where you will complete one of your objectives, finding

S11, as a function of frequency.

- 3) These are the components you may add. Among these are series inductor, capacitor, stubs, etc. We will only use open circuited stubs and t-lines.
- 4) Main toolbar. Here you specify frequency, set impedances. You may also choose to turn off admittance chart, or switch colors around.
- 5) The Smith Chart. A graphic illustration of your impedances. Right now, there is a load and a source, plotted on top of one another at the origin.
- 6) Here, set our source and load impedances. Also lock them, to prevent from accidentally moving them.

5- So, now we are ready to design your stub tuner. First, select ZL and set the impedance to ($Z = 1.035 - j0.018$). Lock the impedance. Similarly, do the same for source, setting it to $(1+j0)$. add line length. This is a standard transmission line. (see figure (3.4.d)).

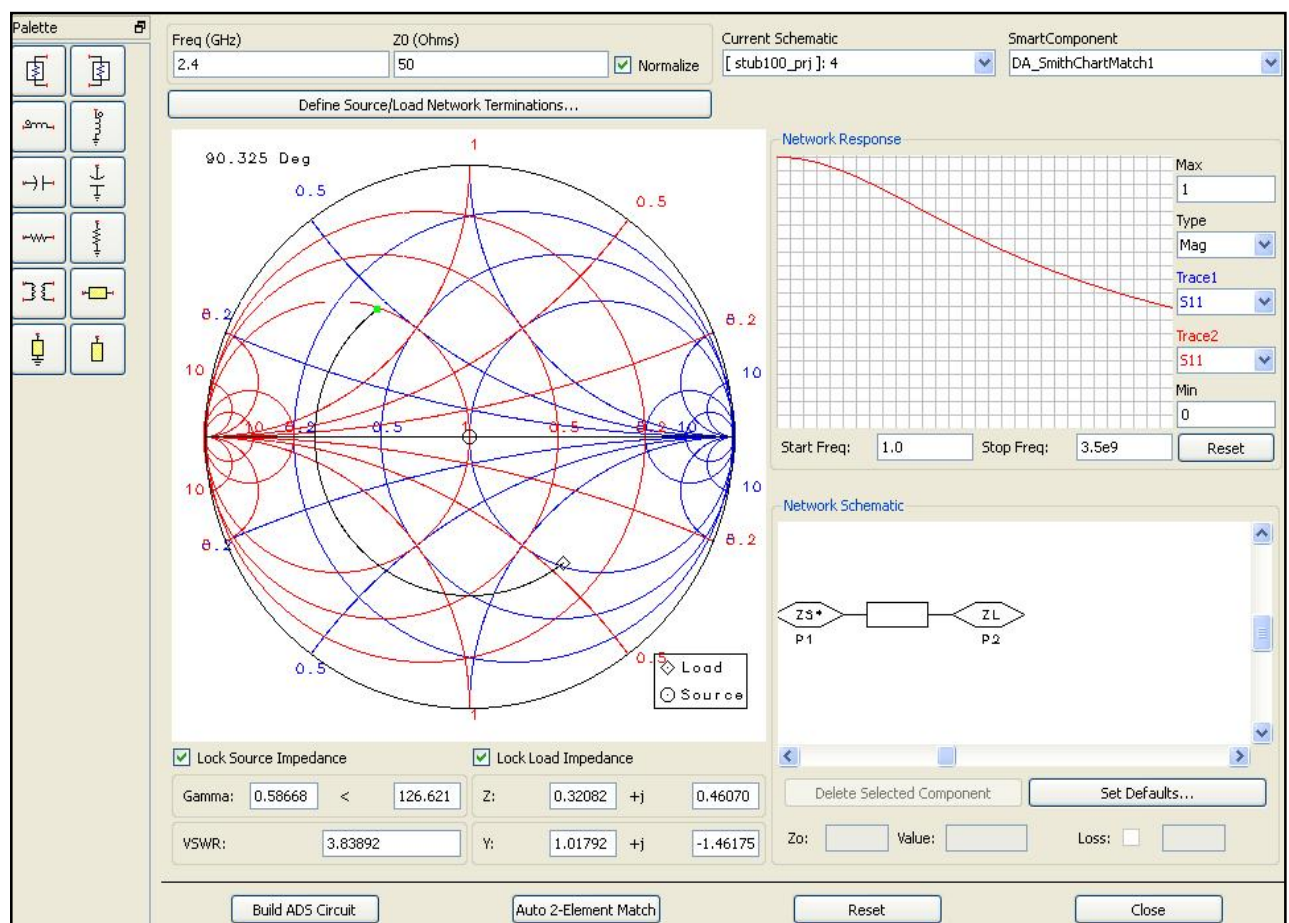


Figure (3.4.d) Drawing area(DA) of smith chart.

- 6- Notice on the smith chart that your cursor draws the clockwise (toward the generator). If cursor draws counterclockwise, you will see that this utility prevents you from making that mistake.

- 7- After selecting line length, make it almost half wavelength long. The cursor should draw almost a complete circle, and it should end up 'almost' back at the load. Now, select an open circuit stub, and again, make it long enough to draw out another, almost complete circle. Notice the path that cursor takes. Also, observe that our design (Figure (3.4.e1)) consists of load, line, stub, and source. Right now, our impedance is the pink square, but we want it to be at the center, to match the source .

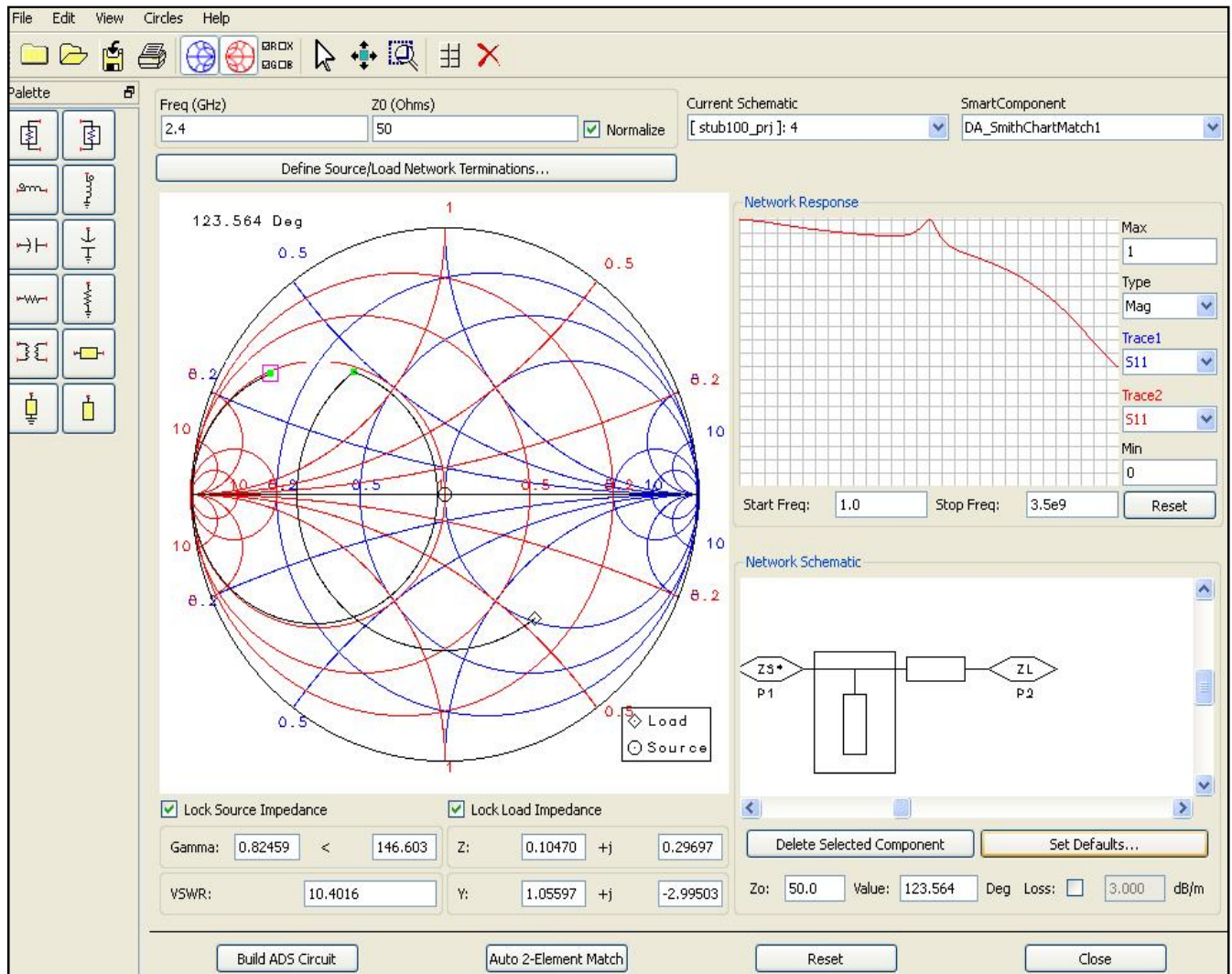


Figure (3.4.e1) Drawing area(DA) of smith chart.

- 8- Select the T-line. Now you can manipulate its length, either numerically, or graphically on the chart. As you change length, the circle drawn out by the stub moves along with the termination of the T-line. Reduce the T-line length until the 'circle' passes thru the origin. Like(Figure (3.4.e2) and (Figure (3.4.e3) respectively) .

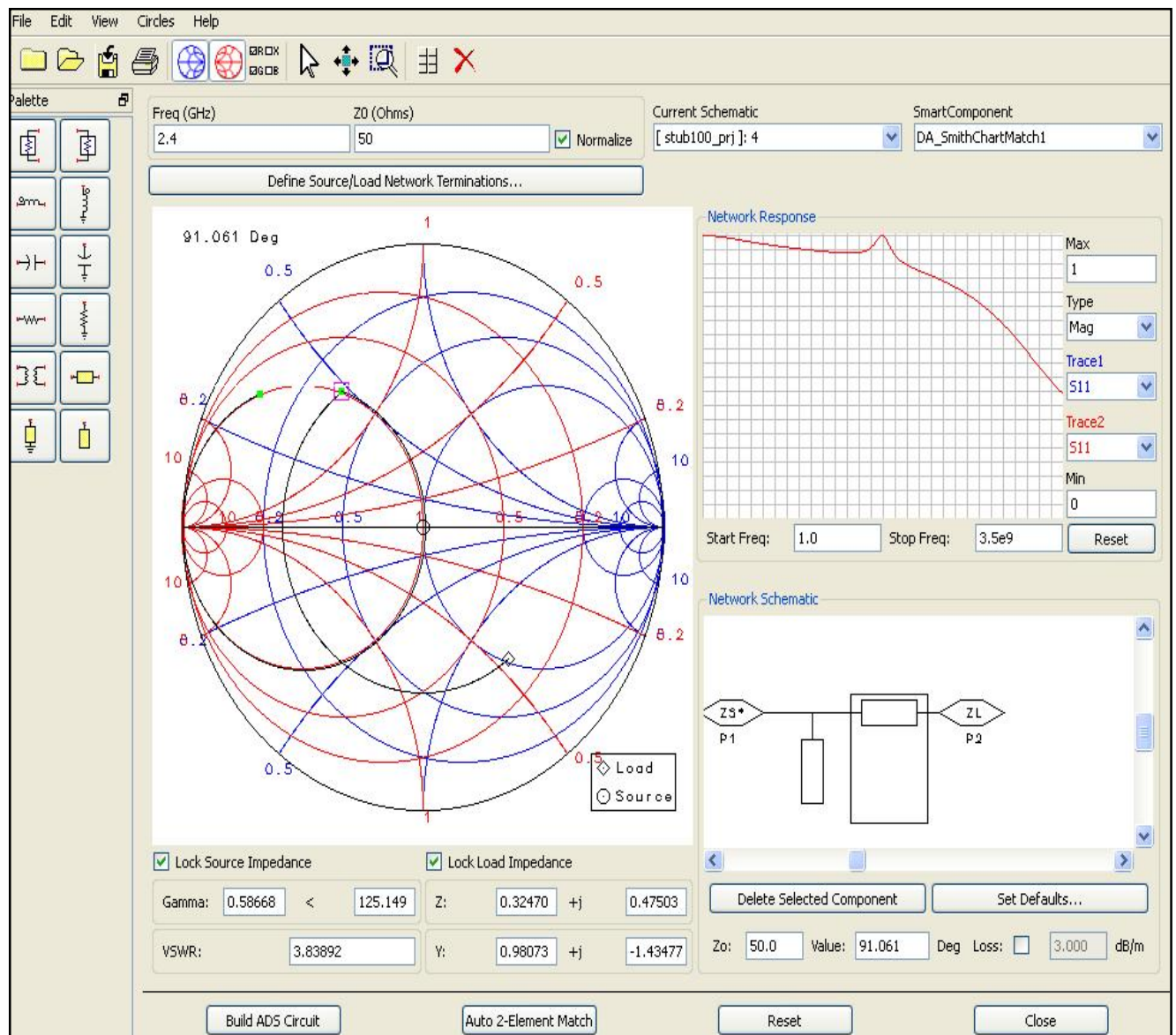


Figure (3.4.e2) Drawing area(DA) of smith chart.

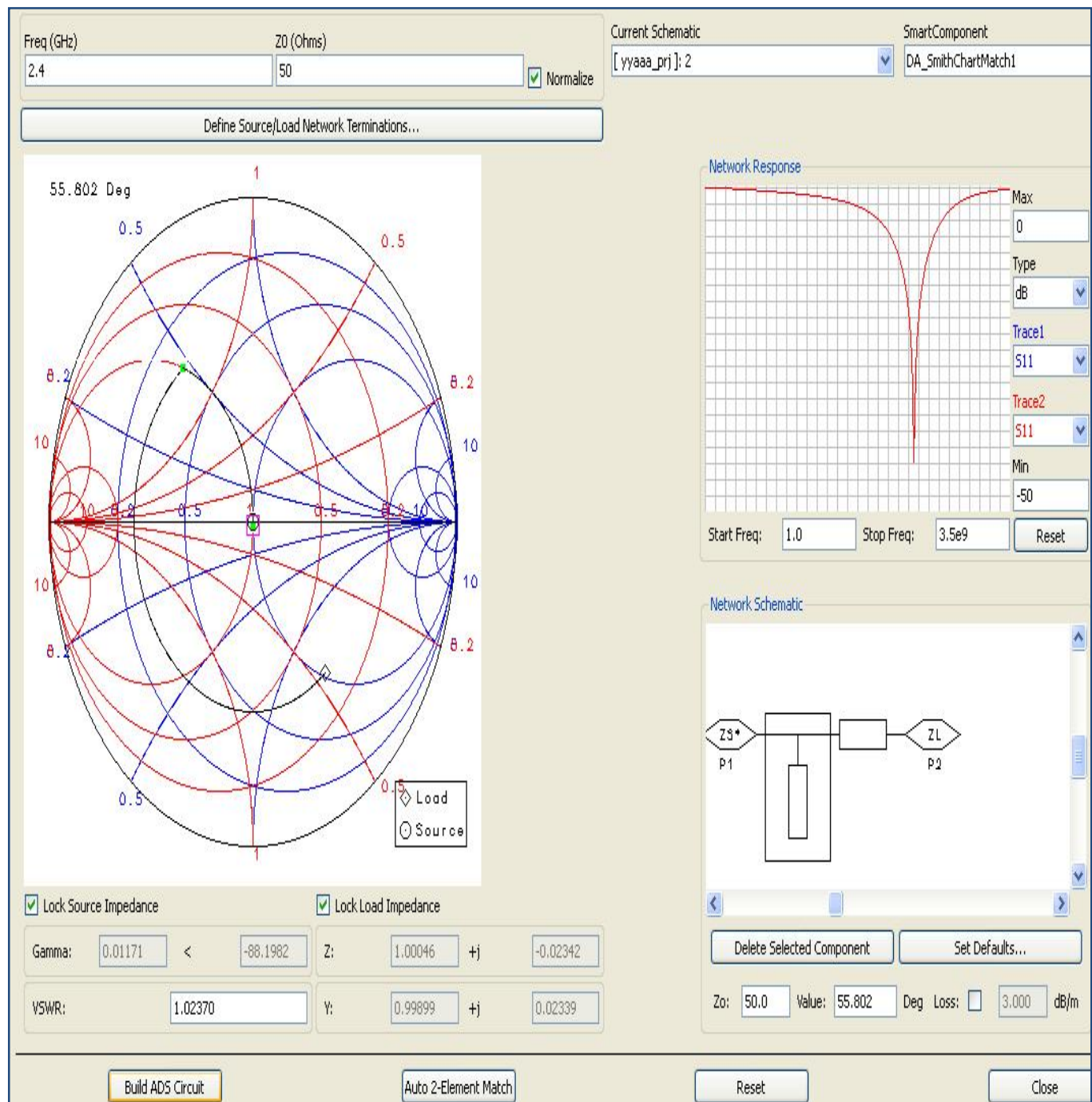


Figure (3.4.e3) Drawing area(DA) of smith chart.

So we need two solutions of stubs, then repeated the Procedure by make the stub long again on DA smith chart and we taking the degrees value for line and stub .

- 9- The question is 'Do we need the T-line to be that long'? No!. Then after we compared between two solutions ,we chosen the shorten stub(My values were 55.80 , and 90.69 degrees for line and stub respectively and look at the plot and see the reflection

coefficient. Make sure that frequency is set to 2.4 GHz. (see Figure (3.4.e3)).

- 10- Now we want to design the actual circuit. To do this, click (Build ADS Circuit). We will notice that the smith chart in our schematic is now updated. Also another schematic is created (see Figure (3.4.f1)).

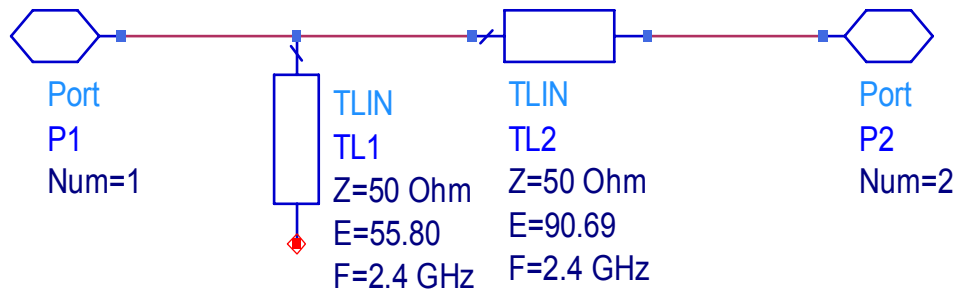


Figure (3.4.f1) Diagram Actual circuit of Single stub .

- 11- Notice that it consists of t-lines instead of microstrips that we desire. We can create an equivalent circuit by compensation data Table (3.1) Technical Specification into (Line-Calc) technique of ADS software and given (E-eff) , (Zo) and frequency in figure (3.4.f1) to determine the dimensions of the microstrips. Make sure to change port1 to term, and port2 to our load values. Place Msub and S-Parameters and run simulations to get the final Schematic of single stub design (see figure (3.4.f2)).

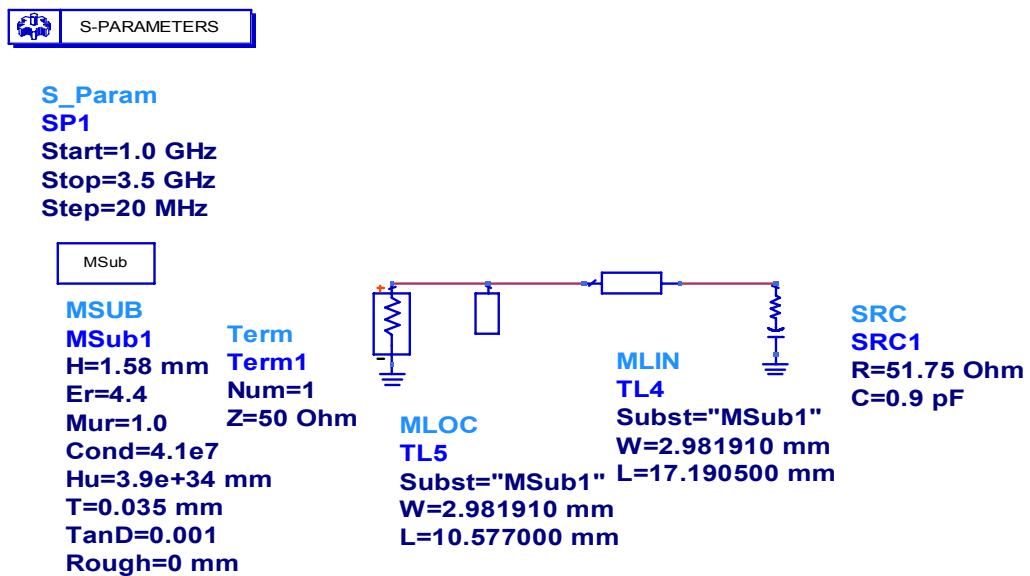


Figure (3.4.f2) Schematic of single stub design.

- 12- Figure (3.5) shows layout of a single stub .

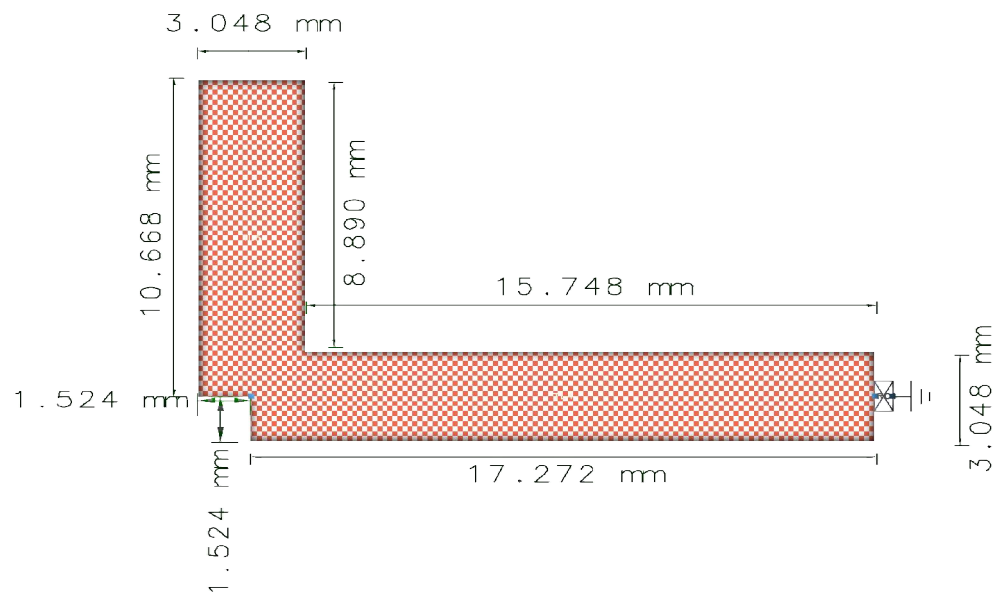


Figure (3.5) Layout of single stub.

- 13- The return losses is measured at designed frequency 2.4 GHz as indicated by the marker m1(S11= -35.37) in figure(3.6) .

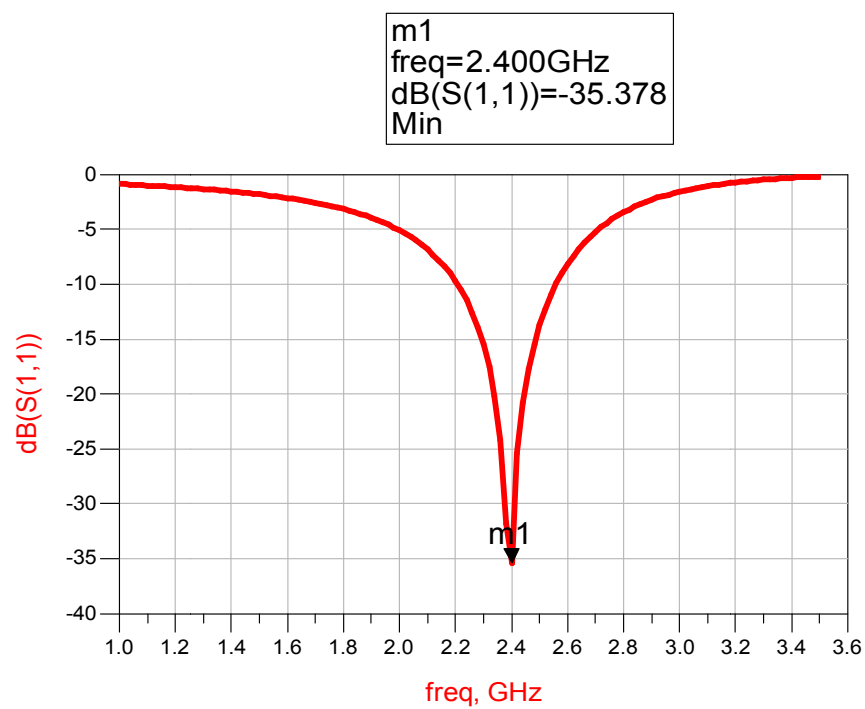


Figure (3.6) Simulated return losses of single stub.

14- figure(3.7) gives the input impedance for single stub on smith chart as pointed by the marker m2 which is equal 49.75Ω and its very closed to 50Ω .

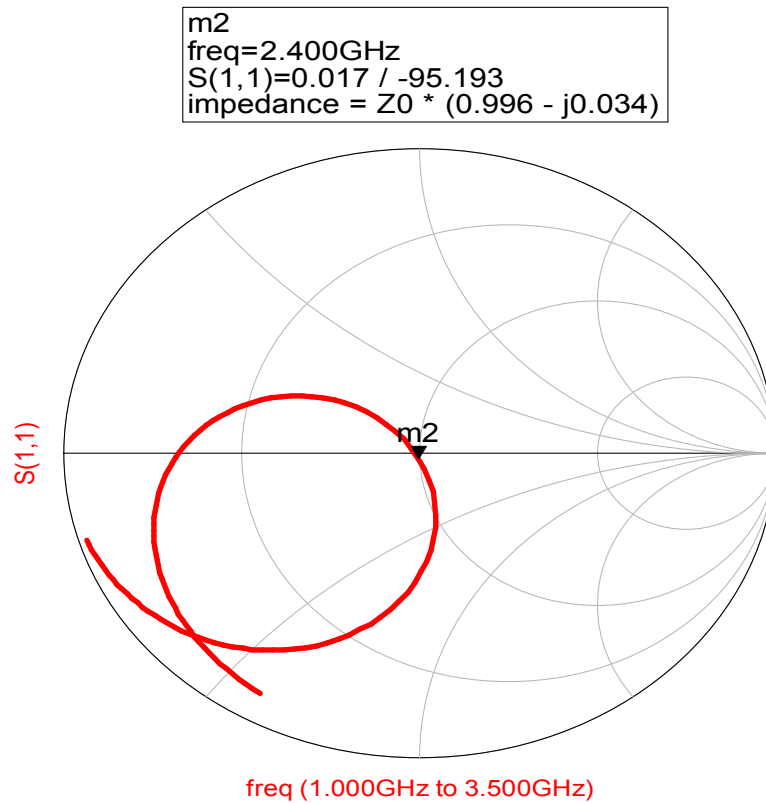


Figure (3.7) Input impedance of a single stub .

3.4.2 Design Extended Transmission Line (TL) Method

In this method design by smith chart is used to design open circuit terminated matching at resonant frequency 2.4GHz and it is constructed at the output ports of 90° hybrid, then tuning on them until obtained the probe length of design.

Open-ends are encountered any time a Microstrip is open terminated.

3.4.2.1 Steps of design

1- Launch ADS, and create a new project.

2- From the palette menu, select Smith Chart Matching , then click Smith Chart symbol and place it on the schematic. Where using open circuit only in this design. See figure (3.8.a) and figure (3.8.b).

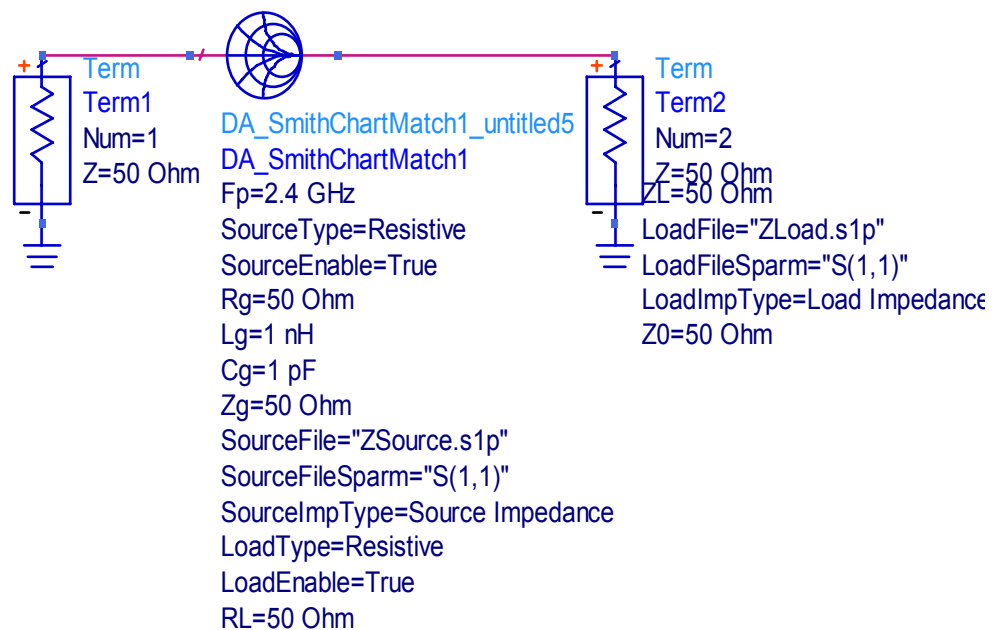


Figure (3.8.a) Smith chart utility of extended TL.

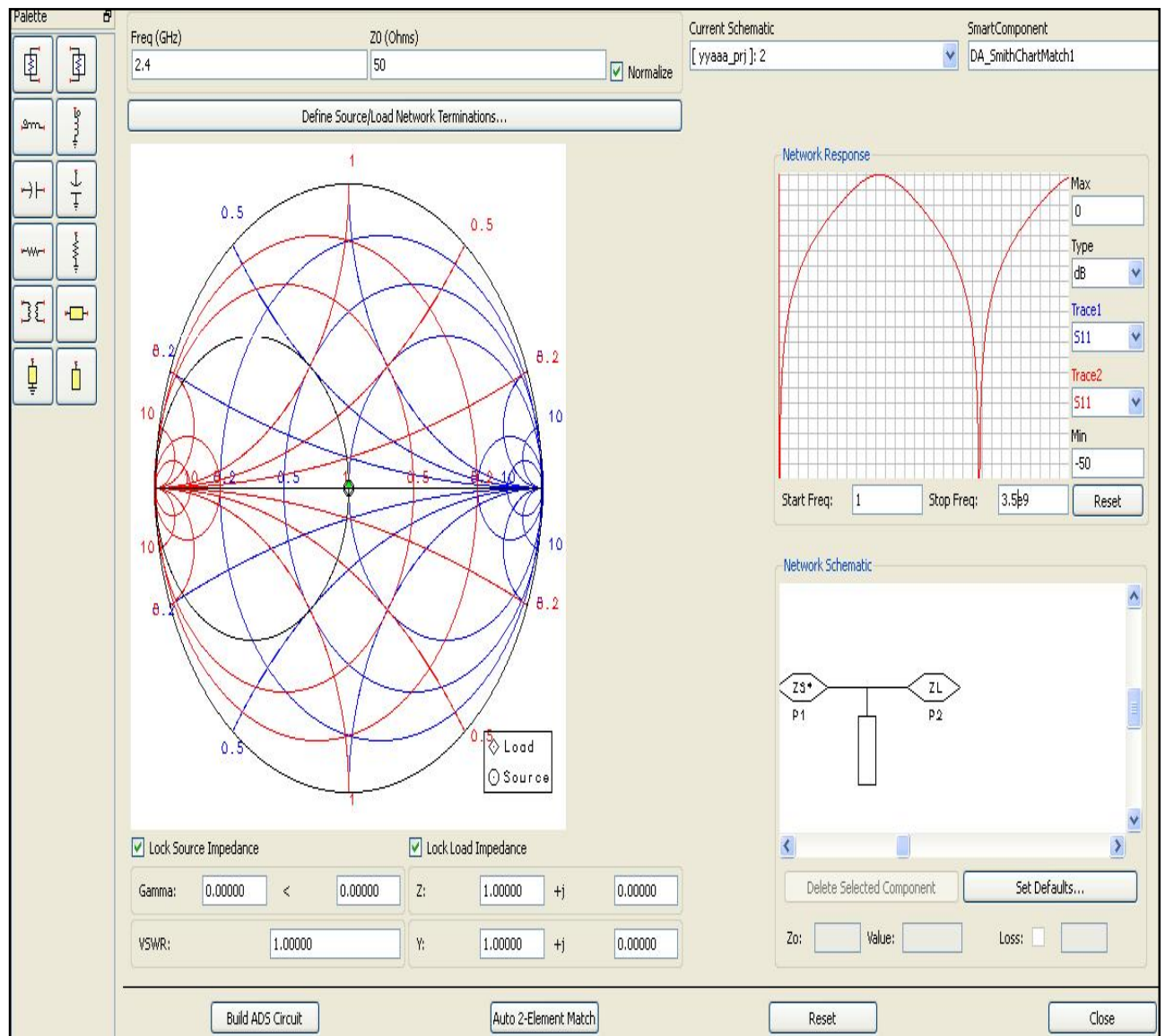


Figure (3.8.b) Drawing area(DA) of smith chart.

- 3- Now we want to design the actual circuit. To do this, click (Build ADS Circuit). We will notice that the smith chart in our schematic is now updated. Also another schematic is created (see Figure (3.9.a)).

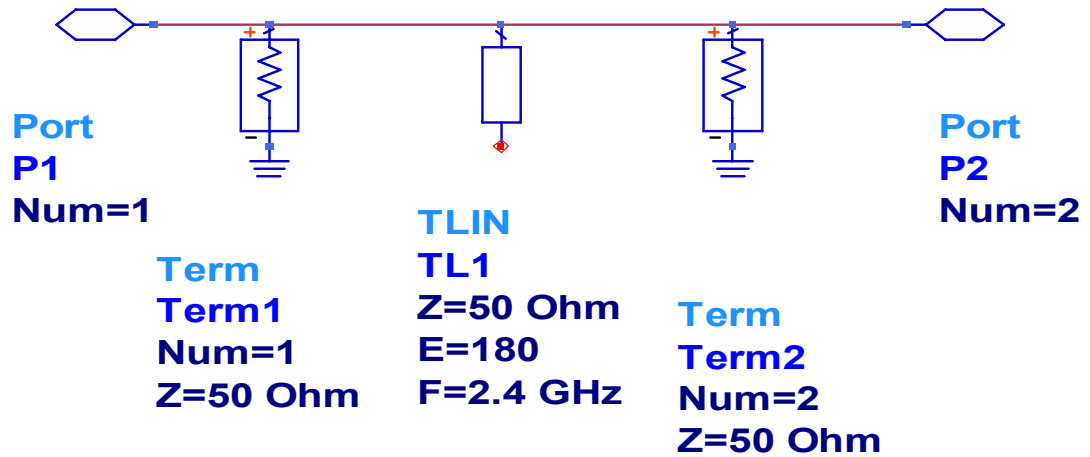


Figure (3.9.a) Diagram Actual circuit of extended TL .

- 4- Figure (3.9.b) Schematic diagram extended TL circuit by compensation data Table (3.1) Technical Specification into (Line-Calc) technique of ADS software and given (E_{eff}), (Z_o) and frequency in figure (3.9.a) to determine the dimensions of the microstrips. Make sure to change port1 to term, and port2 to our load values. Place Msub and S-Parameters and run simulations to get the final Schematic of single stub design (see figure (3.9.b)).

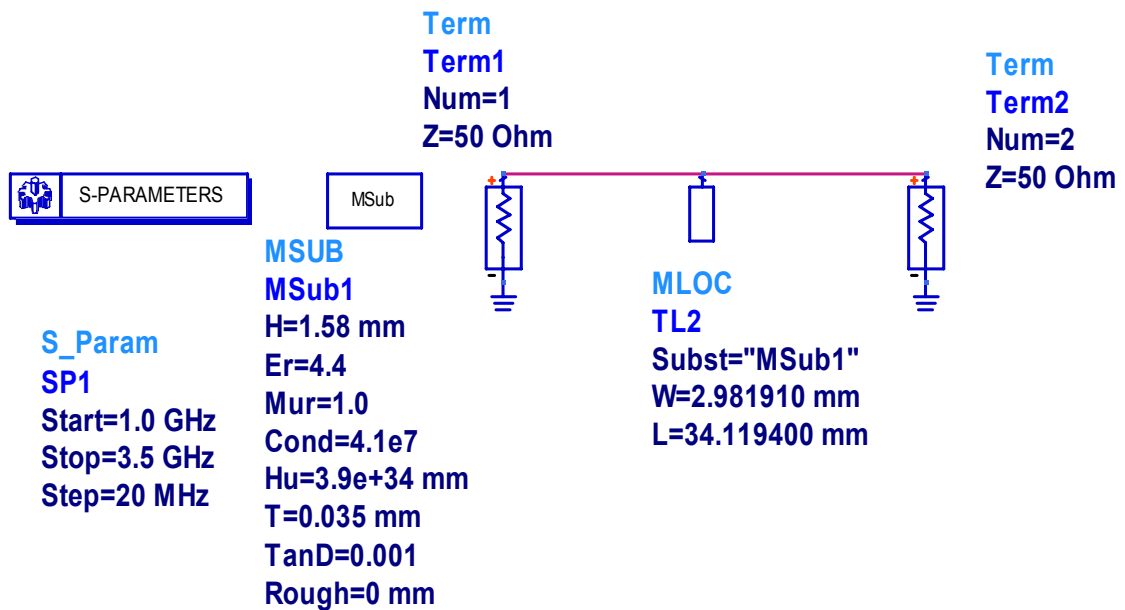


Figure (3.9.b) Schematic diagram extended TL.

- 5- figure (3.10) is the generated layout from the schematic circuit with all dimensions.

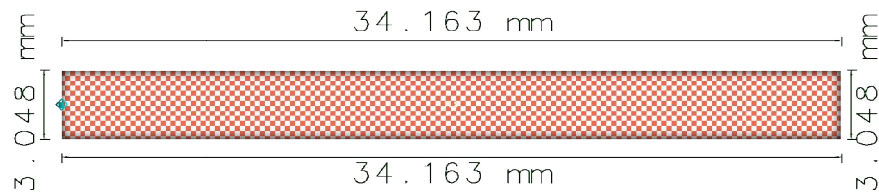


Figure (3.10) Layout dimension of the extended TL .

- 6- Plots of return losses (S_{11}) at resonant frequency 2.4GHz is shown in figure (3.11) . In this plot the marker m2 gives the return losses ($S_{11} = -53.637$ dB) at frequency 2.4GHz .

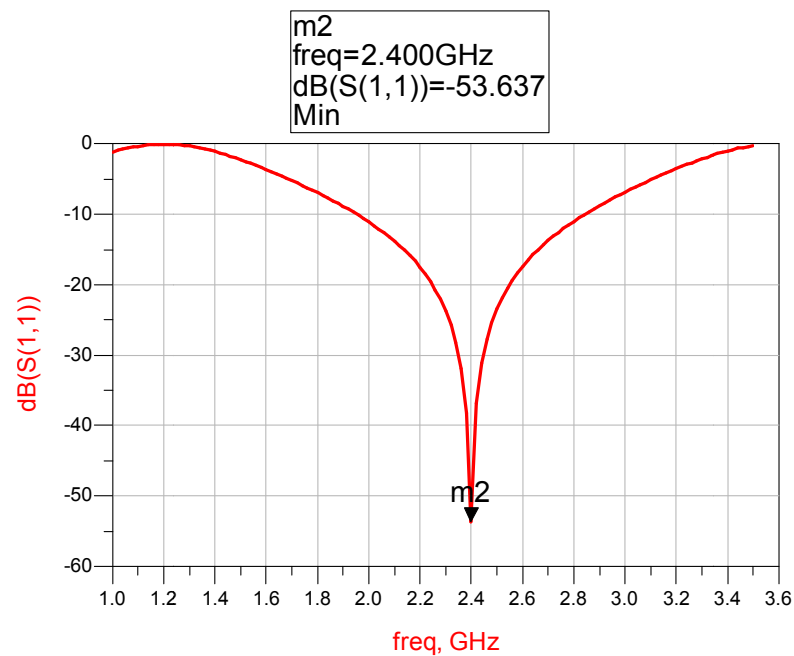


Figure (3.11) Return losses of extended TL.

- 7- Figure (3.12) gives the location of the input impedance on the smith chart ($Z_{in} = 49.8 - j0.1 \Omega$) which is very good match to 50Ω .

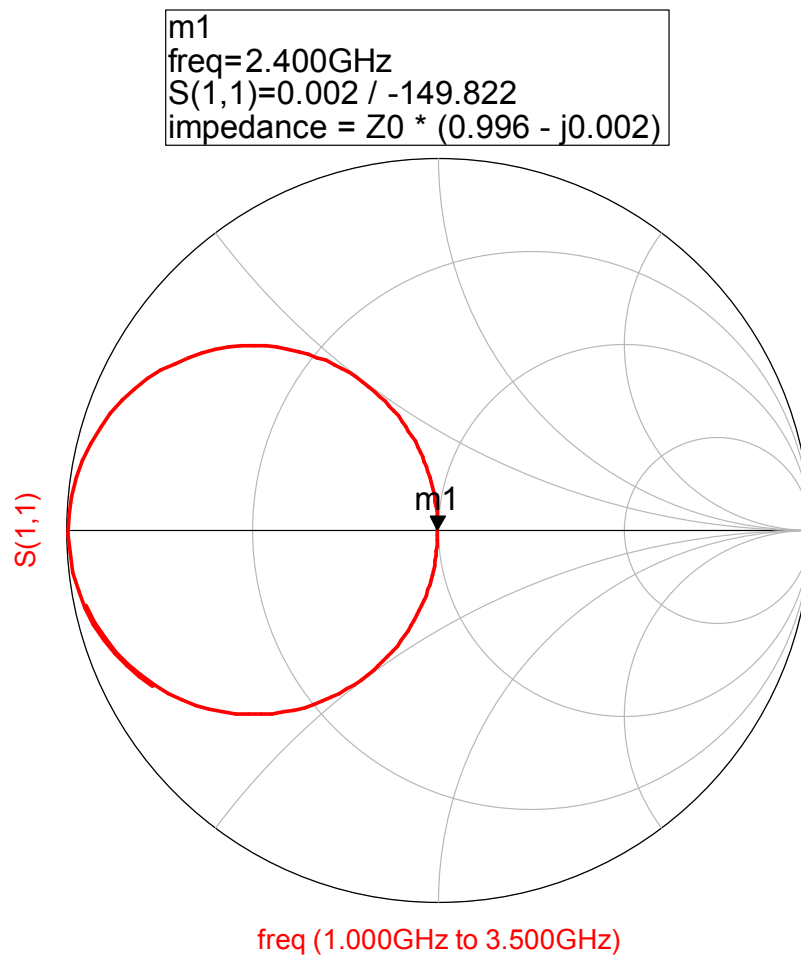


Figure (3.12) Input impedance of extended TL.

Chapter 4

Simulation of 90° Hybrid

4.1 Introduction

In this chapter Simulation of 90° hybrid parameters by software ADS (Advance Design System) such as S-parameters will be demonstrated, also the simulation of enhancement methods of phase and bandwidth will be covered. Fabrication and measurement of 90 hybrid with its methods of improve the phase and bandwidth will be highlights.

4.2 General Guide Lines of Hybrid Coupler

One of the most important challenges in the coupler design is equal power dividing and accurate phase difference. To achieve the optimum design for hybrid coupler ,certain tolerance value can be use as guidance :-

1-Matching, is determined by voltage standing-wave ratio (VSWR) or reflection coefficient of the input (or output) port, while the other ports are connected with reflection less terminations.

Matched ($S_{11}=S_{22}=S_{33}=S_{44}$).

Isolated output ($S_{41}=S_{23}=0$). .

2 - Return loss, expected lower than -10 dB which means lower RL due to higher power can be transmitted. Actually, the loss for hybrid coupler must be considering the combination of coupling loss, dielectric loss, conductor loss .

3 - Reciprocal ($S_{ij} = S_{ji}$), the device exhibits reciprocal behavior because any port can act as an input, and the device will operate in the same manner.

4- Coupling balance for S_{21} and S_{31} must be less than 1dB and the coupling flatness is the amount (in decibels) of the maximum variation in coupling that may be expected over a specified frequency range. Ideally flat coupling is the best design for hybrid coupler but is impossible to achieve it.

5-Voltage coupling factor $C = 10^{\frac{-3dB}{20}} = |V_3^-/V_1^+| = 0.707$ (Transmission coefficient magnitude), where $C < 1$.

6- Insertion loss, is the ratio (in decibels) of input power and output power of the main line with reflection less terminations connected to ports of the hybrid . Theoretical mainline insertion loss due to coupling factor (3dB).

$$C_{21} = C_{31} = [10 \log P_3/P_1 = -20 \log |S_{21}| \text{ dB} = 10 \log P_3/P_1 = -20 \log |S_{31}| \text{ dB}].$$

7- Bandwidth is the range of frequencies for which a parameter falls within a specified limit with respect to certain characteristics. So bandwidth, $BW = F_U - F_L$.

8- Amplitude balance (Power split), is defines the power difference in dB between output ports which is ports 2 and 3. Ideally the amplitude balance difference is should be 0 dB. For hybrid couplers is calculated as the difference (in decibels) between the coupling and insertion loss ($C_{13} - C_{12} = 0$ in ideal case).

9- Phase balance, is the difference between the two output ports where it is should be 90° . Basically a broad side coupler (hybrid coupler) having 90° wave length will give 90° phase shift and like amplitude balance, the phase difference is sensitive to the input frequency and typically will vary a few degrees and it is called phase error.

10- Directivity, this is the signal level at an isolated port relative to the signal level at a coupled port when the signal is injected at an input or to state it as S-parameters.

As a rule-of-thumb the directivity of the coupler should be 20 dB greater than the return loss of the unknown load. This will keep directivity errors below 1 dB.

4.3 Case1. Simulation of 90° Hybrid

It consist of two lines which wave length $\lambda/4$ mm and the main line is coupled to secondary line through two shunt arms with a length of $\lambda/4$ mm each. This creating a square approximately one wave length in circumference. Impedance for two lines is $Z_0/\sqrt{2}\Omega$ and, $Z_0 \Omega$ for each arm. The length of each line of hybrid coupler is $\lambda/4$ mm.

In this work, hybrid coupler has been modified to get optimum result for 2.4GHz hybrid coupler.

The simulation of the hybrid is carried out using ADS. The basic schematic is shown in figure (4.1), the S-parameters simulation engine is used after terminating the four ports with 50 Ω termination and swept the frequency from 1GHz to 3.5GHz in steps of 20 MHz. Where each input line used 2.9819 mm for the width and 16.718 mm for the length. While the main line used 5.119 mm for the width and 13.621 mm for the length, and for the shunt arm dimension is 2.981 mm for the width and 16.718 mm for the length.

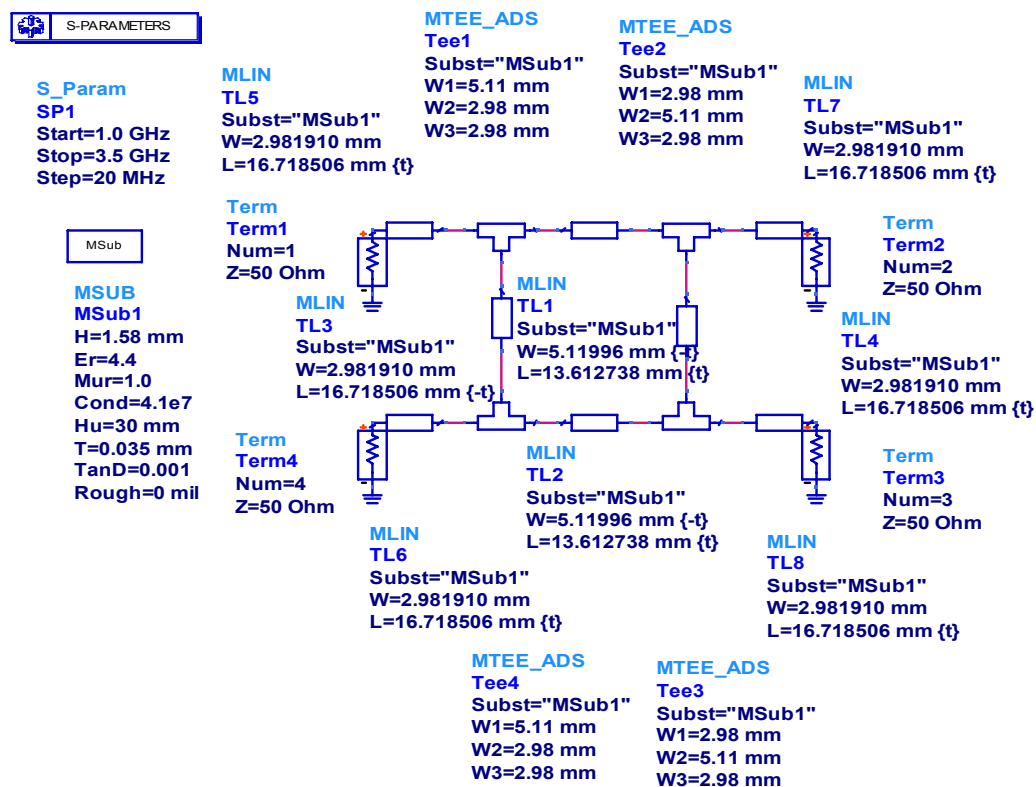


Figure (4.1) Schematic diagram of 90° hybrid coupler.

Figure (4.2), shows the simulated return losses at the resonant frequency of 2.4 GHz at $m1 = -34.386$ dB of a hybrid and similarly that return losses are the same at all ports ($S_{11} = S_{22} = S_{33} = S_{44}$) because they are symmetrical.

The plot shows a deep null at 2.4GHz which indicates a very good match at that frequency to 50 Ω . This is very desirable, because there will be maximum power delivered to the load (device) at this frequency.

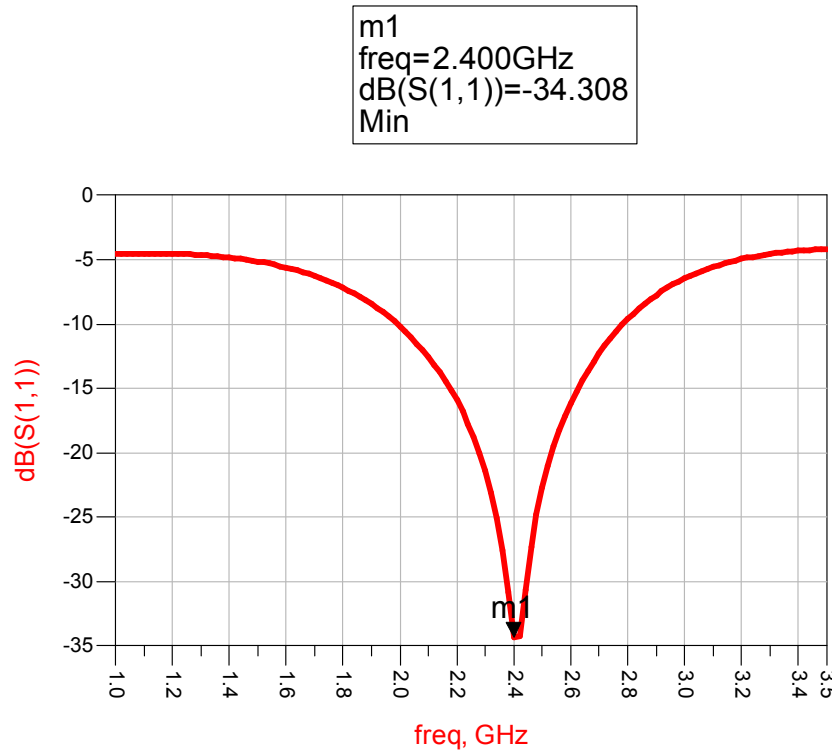


Figure (4.2) Simulation return losses of 90°Hybrid

Figure (4.3) shows, simulation of the isolation S_{41} at the marker $m2 = -36.458$ dB of hybrid coupler. The isolation level in dB from port 1 to port 4 over the frequency range between 1 to 3.5 GHz. There is a deep null at 2.4GHz which means that there will be negligible power at the isolated port near the center frequency.

Figure (4.4) shows, simulation the reciprocal of 90°hybrid coupler at all port of hybrid coupler at frequency 2.4 GHz, which means ($S_{ij} = S_{ji}$), that it does not make any difference which port is the input port as long as the relationship between ports remains. This is because the 90° hybrids are electrically and mechanically symmetrical about both the X and Y axes.

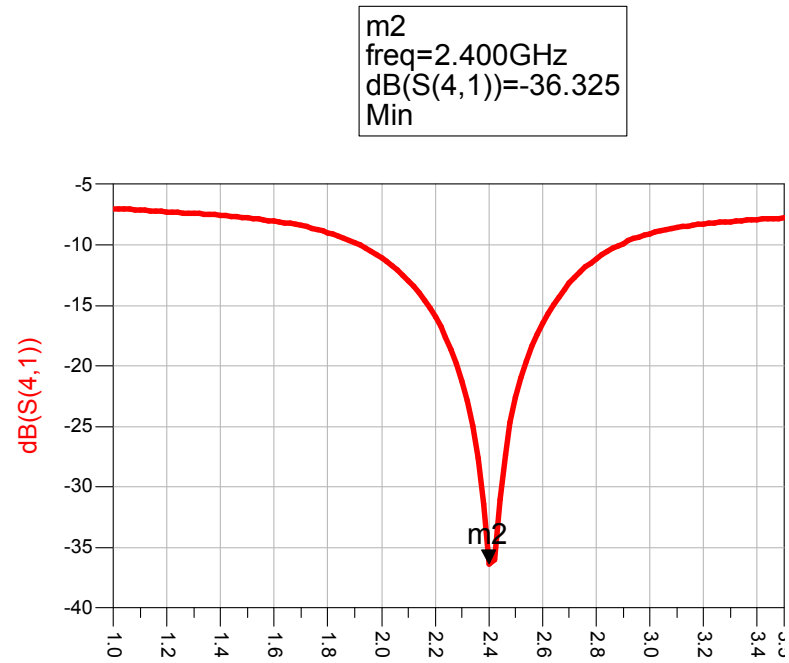


Figure (4.3) Simulation isolation of 90°hybrid coupler

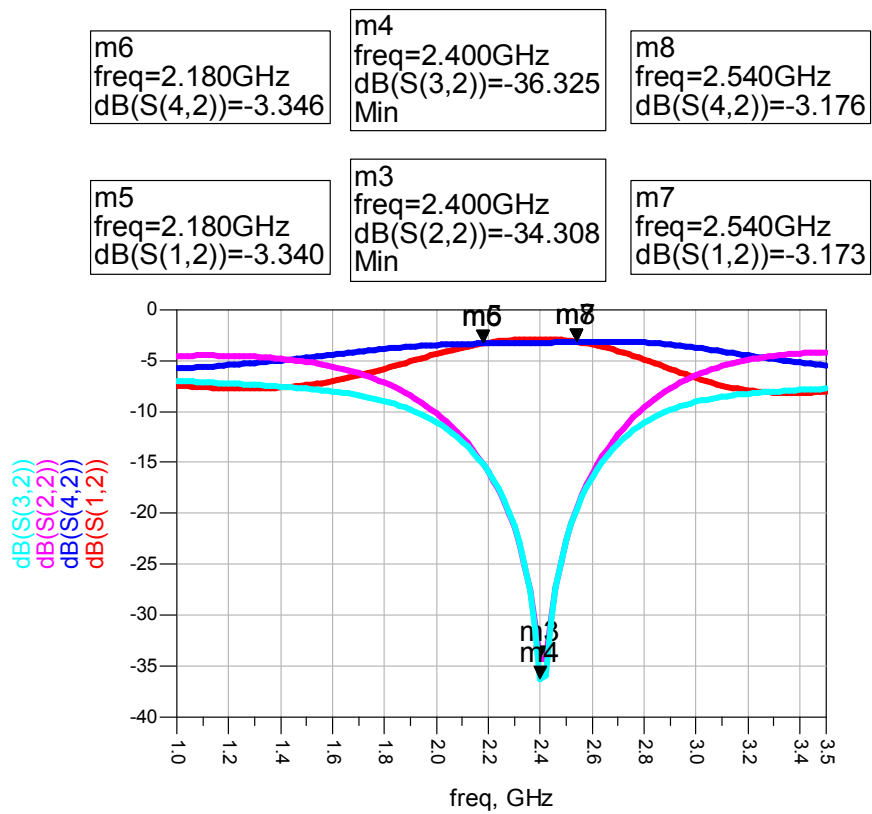


Figure (4.4) Simulation of reciprocal of 90°hybrid coupler

Figure (4.5), shows the transmission coefficient from port 1 to ports 2 and 3 (in dB) across the same frequency band as magnitude of coupling voltage at markers (m13) and (m14)= ± 0.707 V, the coupling voltage < 1 V as close as in ideally case. As can be illustrated in Figure (4.6), which gives the value of coupling balance as maximally flat $S_{31}=-3.273$ dB as indicated by the marker m19, its less than 1dB as close to the ideally case, then amplitude differences $(m17-m18)=(S_{31}-S_{21})= 0.008$ dB, it's close to 0 dB as in ideally case.

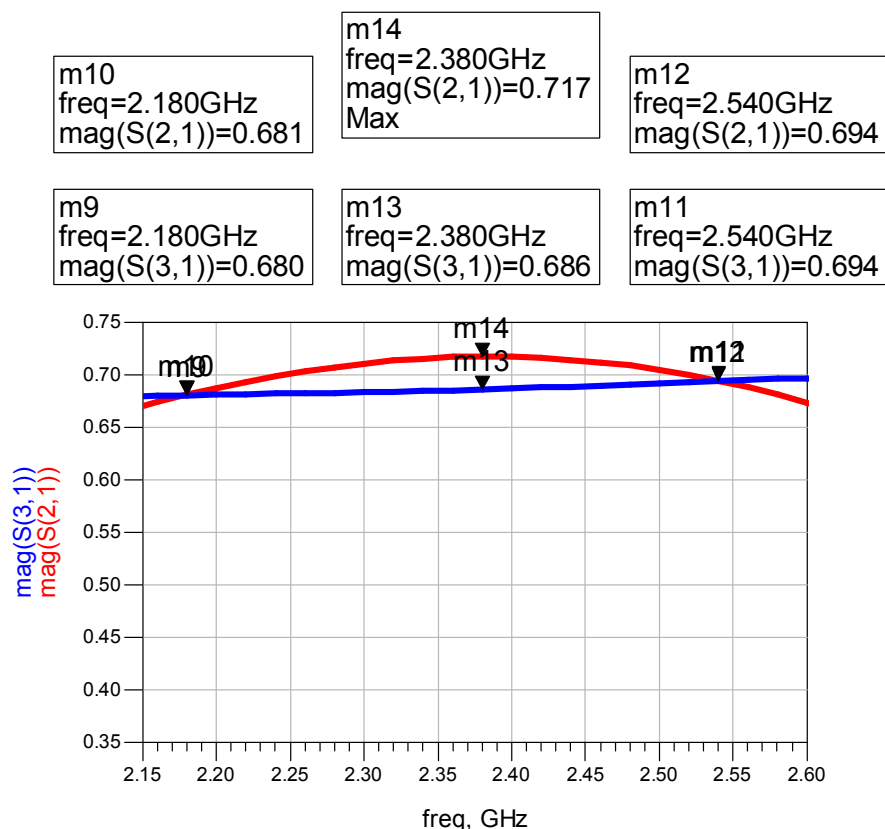


Figure (4.5) Simulation transmission coefficients of 90° hybrid

Both figures (4.5) and (4.6) show, same frequency bandwidth starting from 2.180 GHz until 2.540 GHz .Then the calculated bandwidth (BW)=360 MHz.

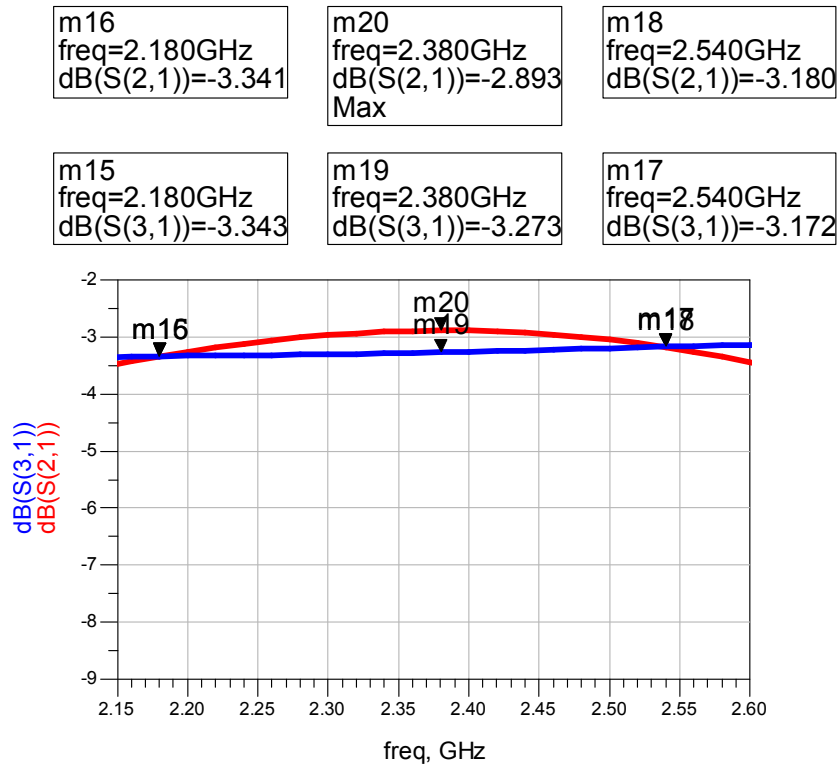


Figure (4.6) Simulation coupling balance of 90°hybrid coupler

Figure (4.7), shows the simulated phase at port 2 and port 3. The phase difference between port 2 and port 1 which marketed by m21 at the frequency 2.4 GHz is 82.213° where as the marker m22 gives the phase -7.82° , then the phase difference between ports 2 and 3 is

$$\text{Phase difference (m21-m22)}=90.033^\circ.$$

The ideal phase difference between these two ports should be 90 degrees.

Figure (4.8) is also a plot of phase difference between ports 2 and 3 just by other way, we have divided the phase of S(2,1) by phase of S(3,1). The marker m23 gives directly the phase difference between ports 2 and 3 at designed frequency ($m23 = 90.033^\circ$).

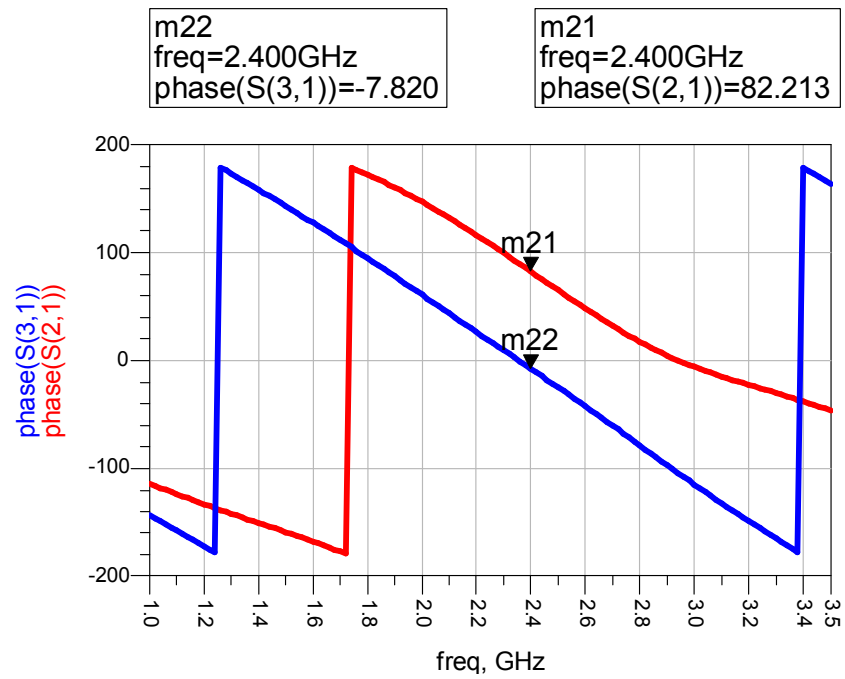


Figure (4.7) Simulation the phase difference at outputs of 90° hybrid

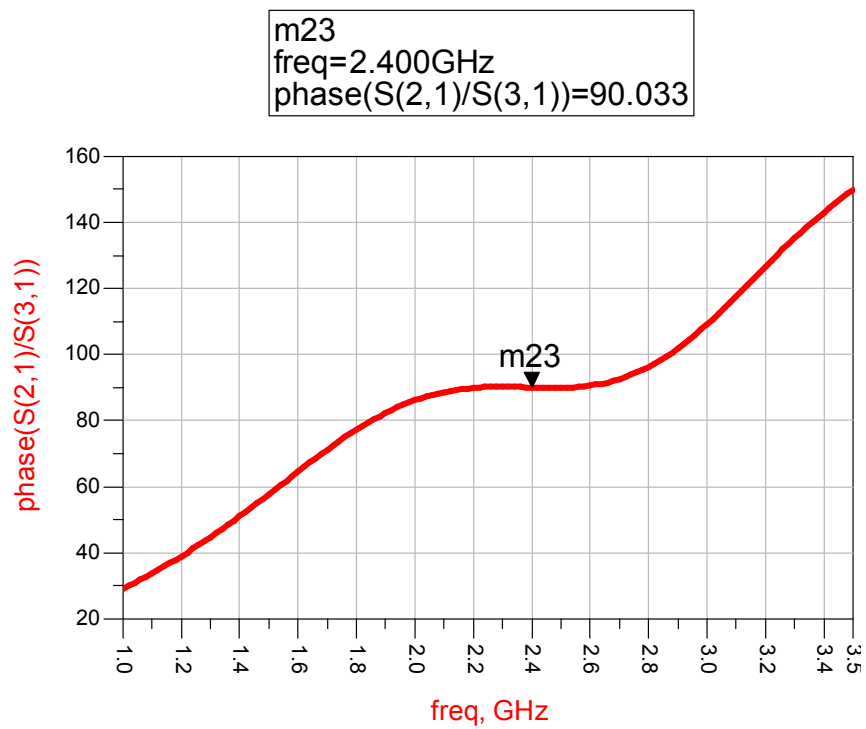


Figure (4.8) Simulation phase of 90° hybrid

4.3.1 Layout of 90°Hybrid Coupler

By using a command generate layout tin ADS software the schematic design of figure (4.1), is transformed to the layout design with all lengths and widths and substrate specifications as shown in figure (4.9).

Tuning process is used at the layout design in order to obtain the target specification of the design.

Figure (4.9) shows, the layout design with all dimension after using the tuning process which satisfies our specification after momentum (method of moment) simulation is used.

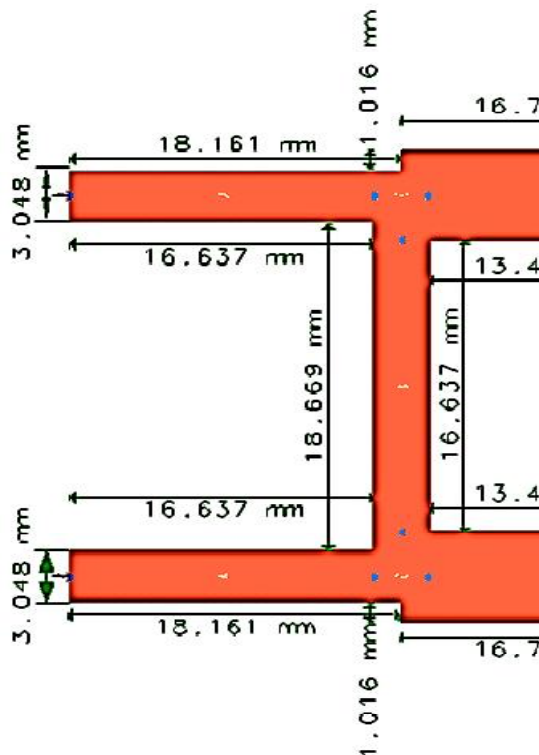


Figure (4.9) Layout of 90°hybrid coupler with dimensions .

Figure (4.10), shows the return losses of 90°hybrid coupler, at the resonant frequency of 2.4GHz as pointed by $m_{24} = -29.26$ dB and also at all ports $S_{11} = S_{22} = S_{33} = S_{44}$ are the same, the plot shows a deep null at 2.4GHz which indicates a very good match at that frequency to 50Ω , there will be maximum power delivered to the load (device) at this frequency. Figure (4.11), shows the isolation of 90°hybrid coupler, isolated output S_{41} on $m_{25} = -32.420$ dB.

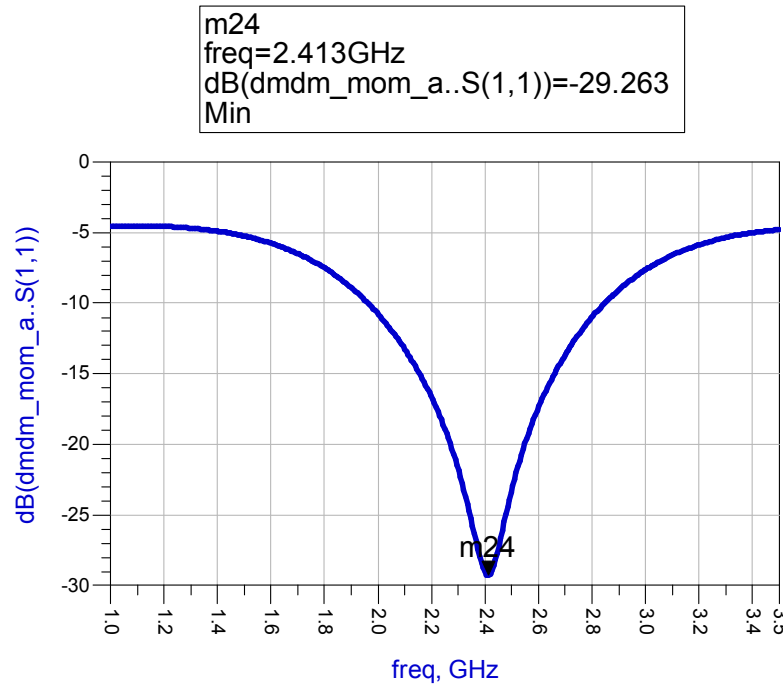
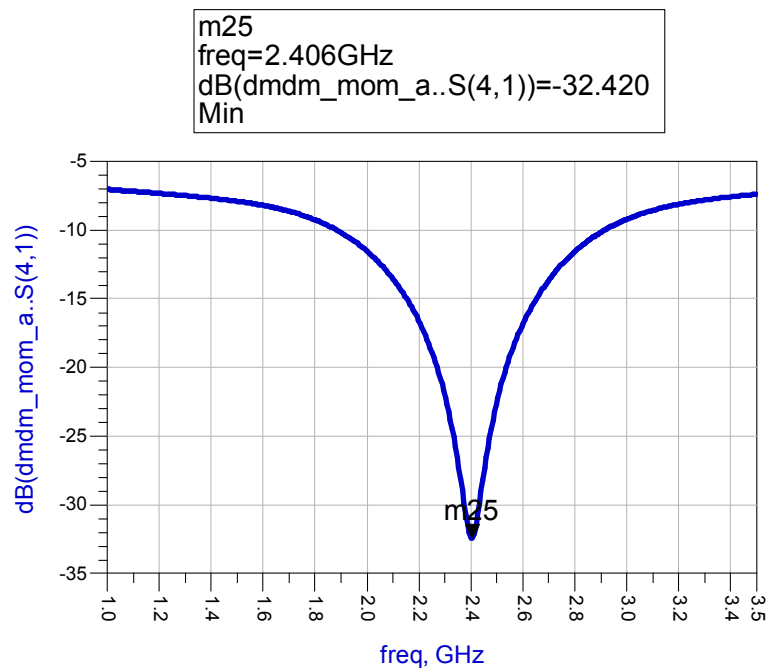


Figure (4.10)Return losses of 90° hybrid coupler.



Figure(4.11) Isolation of 90° hybrid coupler.

Figure (4.12), illustrate transmission coefficient magnitudes(voltage coupling factor) at the designed frequency 2.4 GHz as indicated by the markers m30 and m31 which are

0.661V and 0.735V respectively, these values are less than 1V, and it is closed to the typical values.

As been illustrated in Figure (4.13), the value of coupling balance as maximally flat $m_{36}=S_{31}=-3.593$ dB, its less than 1 dB as close as ideally case.

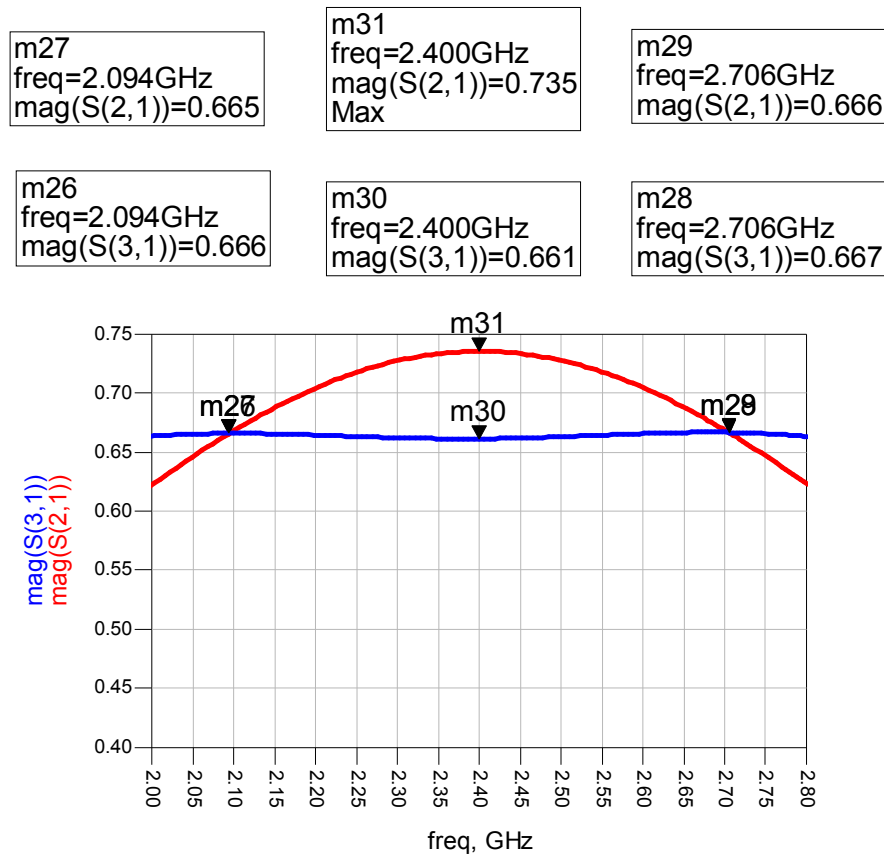


Figure (4.12) Transmission coefficients of 90°hybrid coupler.

Amplitude differences in decibels $(m_{32}-m_{33})=(S_{31}-S_{21})=0.003$ dB. it's close to 0 dB as in ideally case. This will provide an even power split at the designed frequency.

The bandwidth of the hybrid can be calculated from figure (4.13) as the intersect of the two curves at the two frequencies points indicated by m32 and m35 which gives BW=612 MHZ.

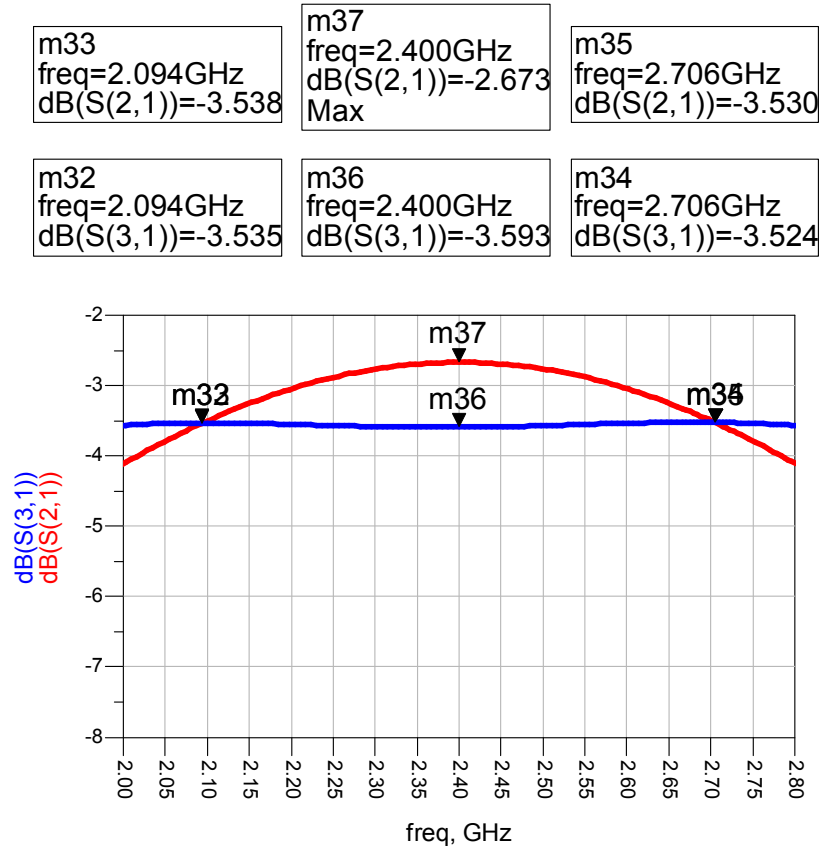


Figure (4.13) Coupling balance of 90°hybrid coupler.

From layout transformation, Figure (4.14) shows, phase difference as indicated by markers (m38 and m39), then phase difference:-

$$\text{Phase difference (S21-S31)} = (m38 - m39) = 89.65^\circ.$$

Figure (4.15) shows (m40) the phase of output ports where plot of phase difference between ports 2 and 3 just as divided the phase of S(2,1) by phase of S(3,1). The marker m40 gives directly the phase difference between ports 2 and 3 at designed frequency ($m40 = 89.65^\circ$).

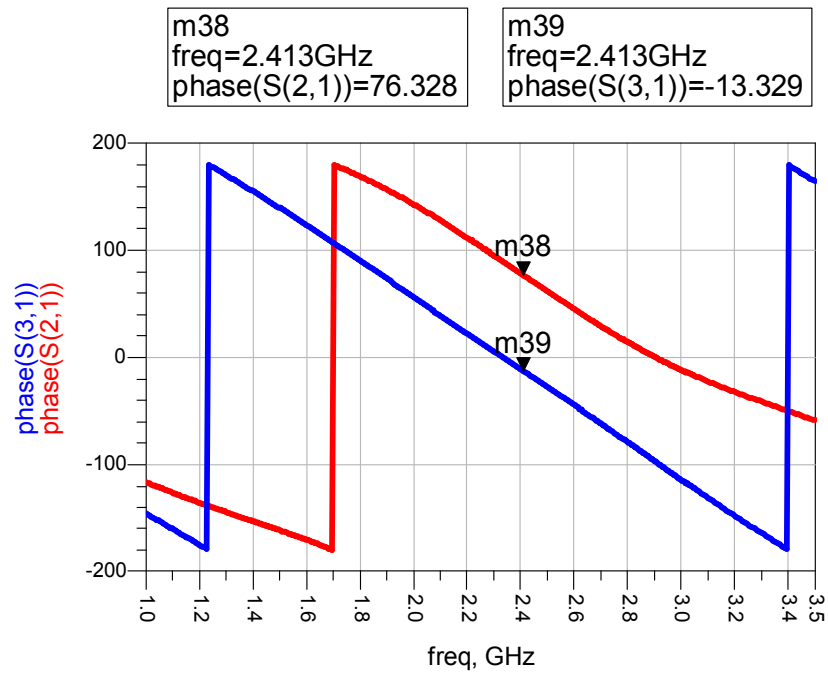


Figure (4.14) Phase difference of 90° hybrid coupler.

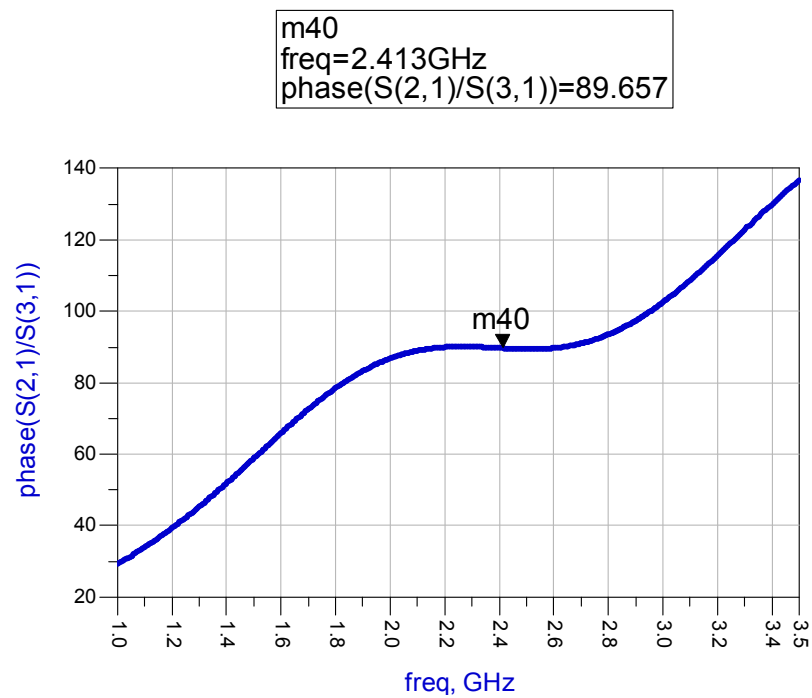


Figure (4.15) Phase of 90° hybrid coupler

4.3.2 3D Layout of 90° Hybrid

Figure (4.16), shows the 90° hybrid layout in 3-dimensions. According to our simulation, the height of the cover was tested and it was at least ($h_u=30\text{mm}$), in order not to disturb the simulated S-parameters.

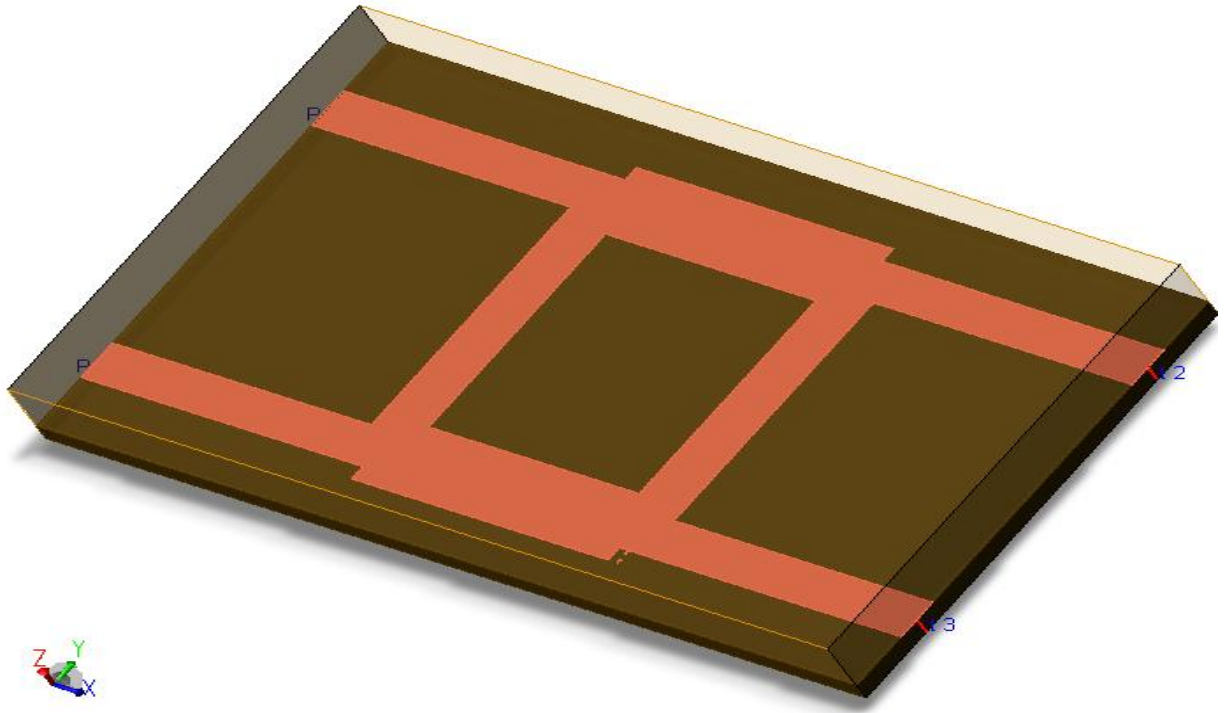


Figure (4.16) 3-Dimensional layout of 90° hybrid .

4.4 Case2. Simulation of Cascaded 90° Hybrid

Multiple quarter-wavelength coupling sections are used to obtain wide frequency bandwidth hybrid coupler. Typically this type of hybrid coupler is designed to a frequency bandwidth ratio and a maximum coupling ripple within the frequency band. Figure (4.17), shows schematic design 90° hybrid cascade coupler; it's easy way for a two-section coupler design, the dimension will be considered in order to meet the appropriate requirement (explained in chapter 2). To simplify the formulation, the electrical lengths of the series lines are kept to be equal and the shunt arms should ideally have impedance of 120Ω which is upper limit for practical construction, because it will give a width of the strips about 0.2 mm .

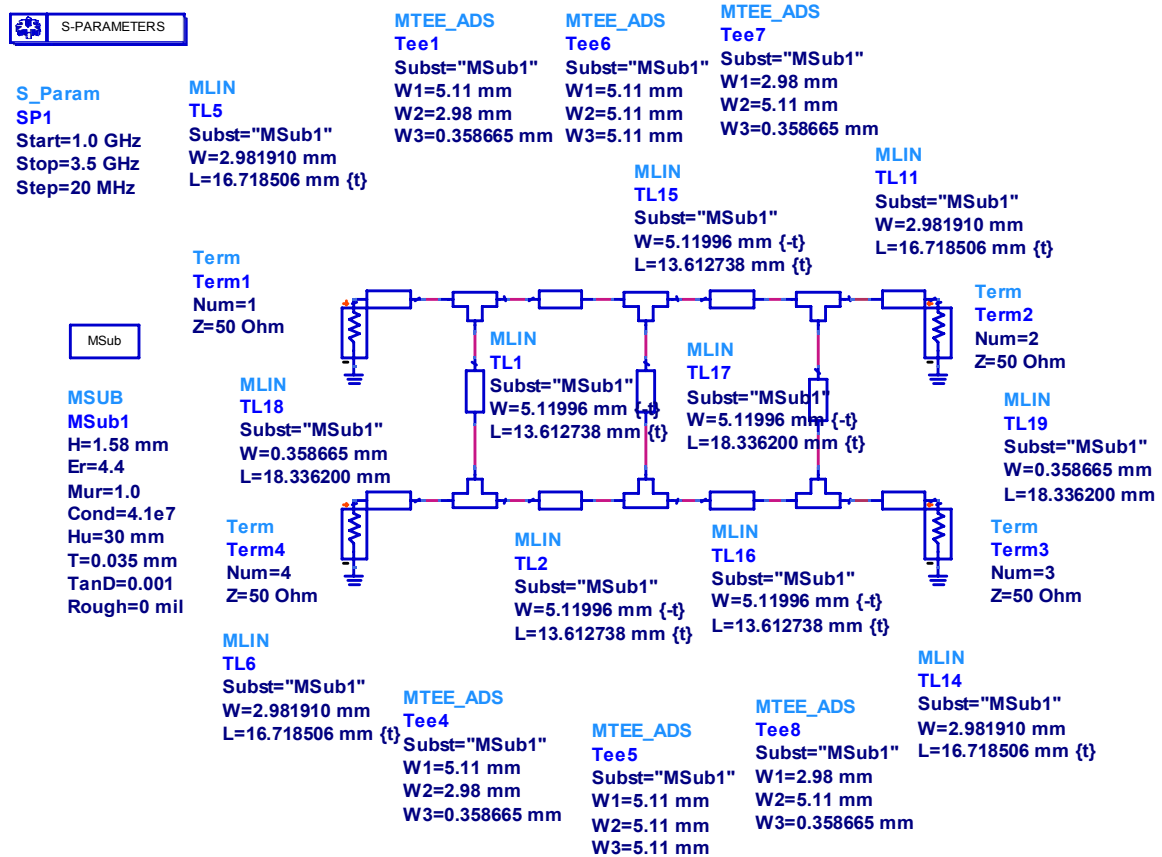


Figure (4.17) Schematic diagram of cascaded 90° hybrid coupler.

Figure (4.18) shows, simulation the return losses where the resonant frequency at 2.4GHz is (m1) of $S_{11}=-24.008\text{dB}$ of 90° cascade and also at all ports $S_{11}=S_{22}=S_{33}=S_{44}$. This is very desirable, because there will be maximum power delivered to the load at this frequency.

To obtain wide band performance, multi branch hybrid was tested, but this technique resulted in a wide return loss and not wide insertion loss within the working band .

Figure (4.19), shows simulation the isolation of 90° Hybrid coupler isolated output S_{41} at $m_2=-24.223\text{ dB}$. In this case, effective power entering into port 1 is distributed into ports 2 and 3. Port 4 is decoupled and without power if ports 2 and 3 are ideally matched.

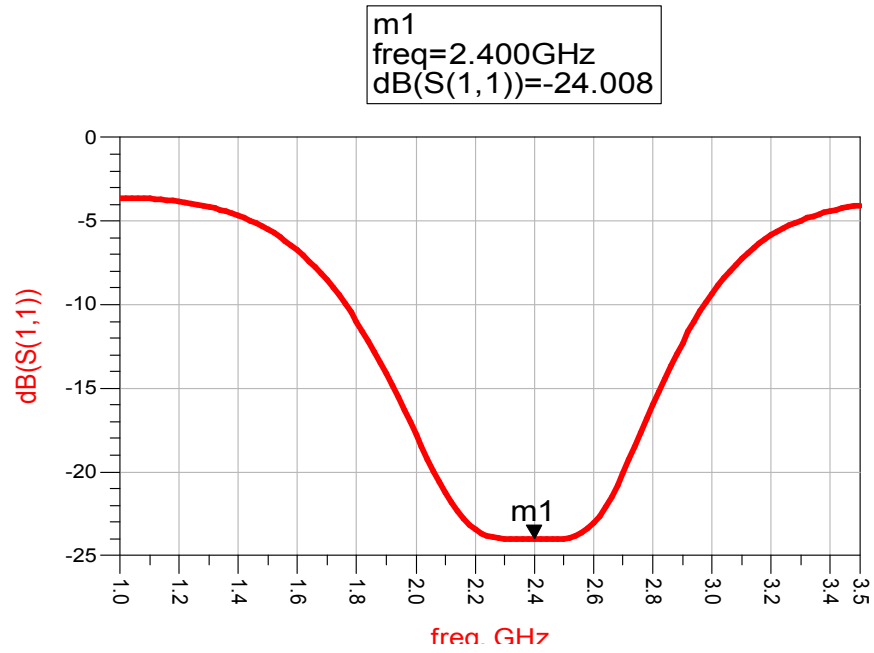


Figure (4.18) Simulation return losses of cascaded 90° hybrid coupler.

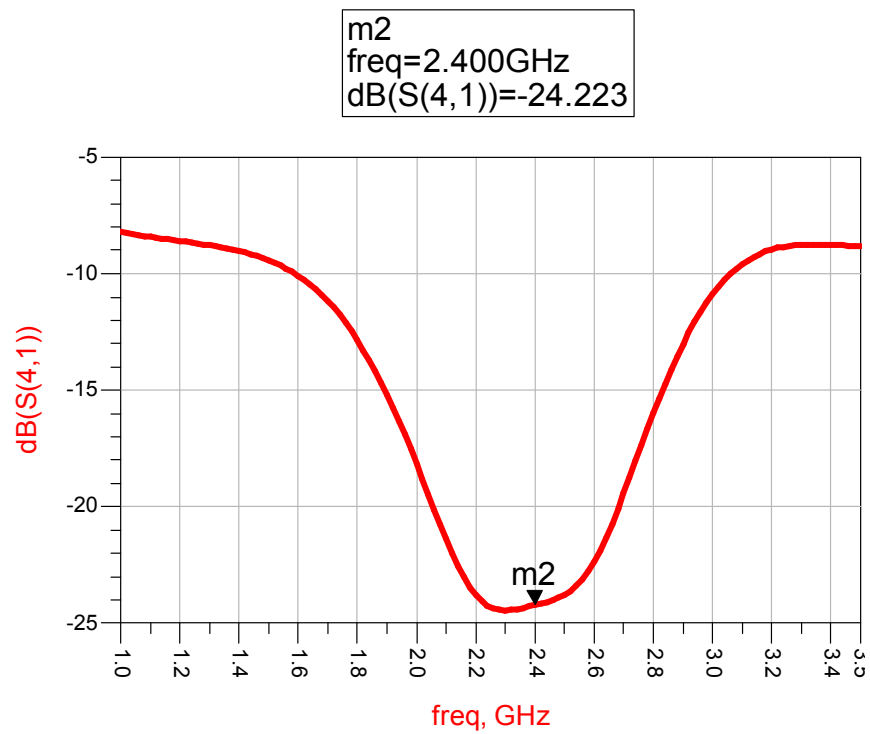


Figure (4.19) Simulation isolation of cascaded 90° hybrid coupler .

Figure (4.20), shows reciprocal ($S_{ij} = S_{ji}$) of cascade, which makes no difference which port is the input because the relationship at the outputs remains the same as these devices are electrically and mechanically symmetrical.

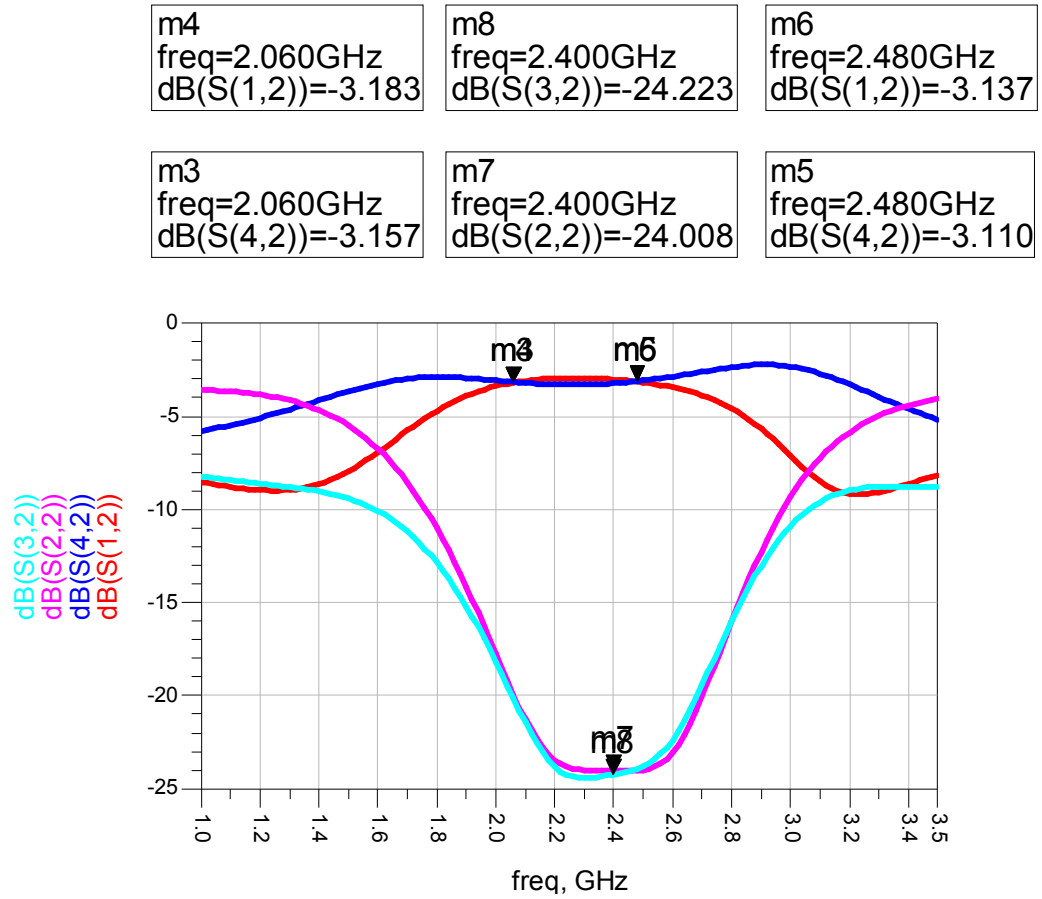


Figure (4.20) Simulation reciprocal of cascaded 90°hybrid coupler.

Figure (4.21) and figure (4.22) shows, the transmission coefficient from port 1 to ports 2 and 3 across the same frequency band as magnitude of coupling voltage at range $\pm 0.707 < 1$ as close as in ideal case, both figures provides bandwidth frequency starting 2.060GHz until 2.480 GHz, as pointed by markers m4 and m6 respectively. Then the calculated bandwidth (BW)=420 MHz, as pointed by markers m4 and m6 respectively.

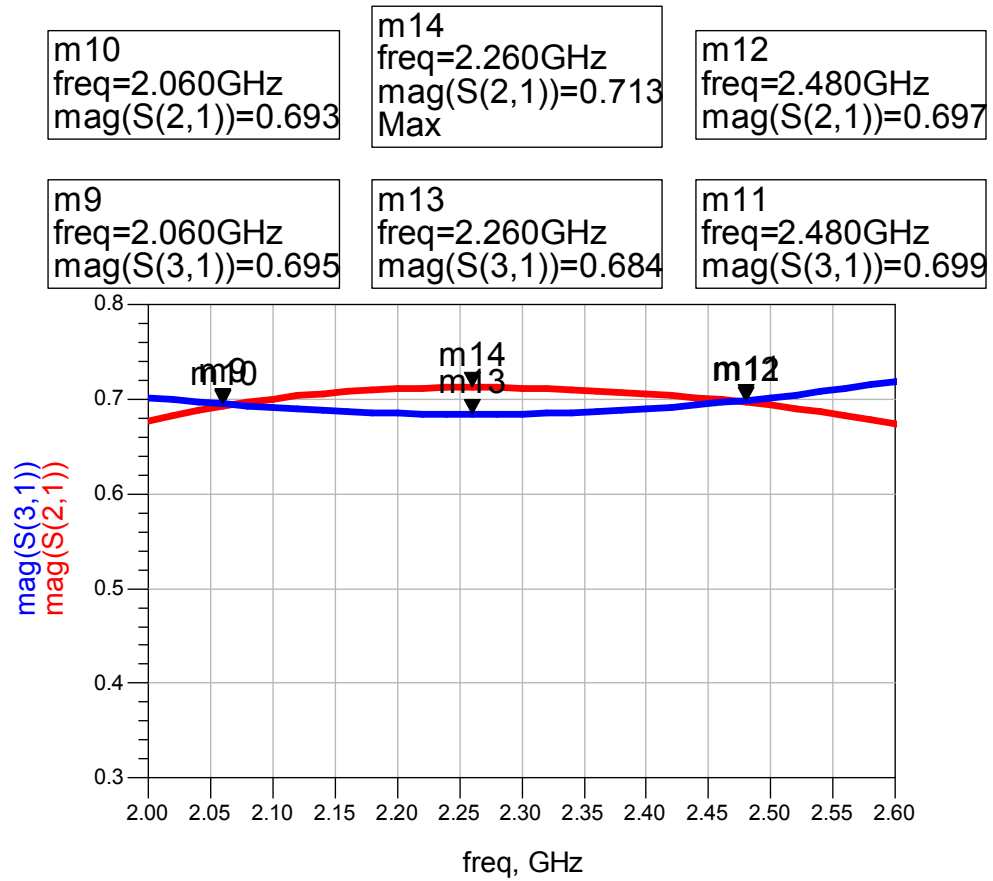


Figure (4.21) Simulation transmission coefficients of cascaded 90° hybrid coupler

As has been illustrated in Figure (4.22), the value of coupling balance as flat maximally $m19=S31=-3.297\text{dB}$, its less than 1dB as close as ideally possible and the amplitude differences $(S31-S21)=(m17-m18)=0.027\text{dB}$. its close to 0dB as in ideal case, means the two outputs are of equal amplitude. This will provide an even power split at the design frequency and close to even across the band.

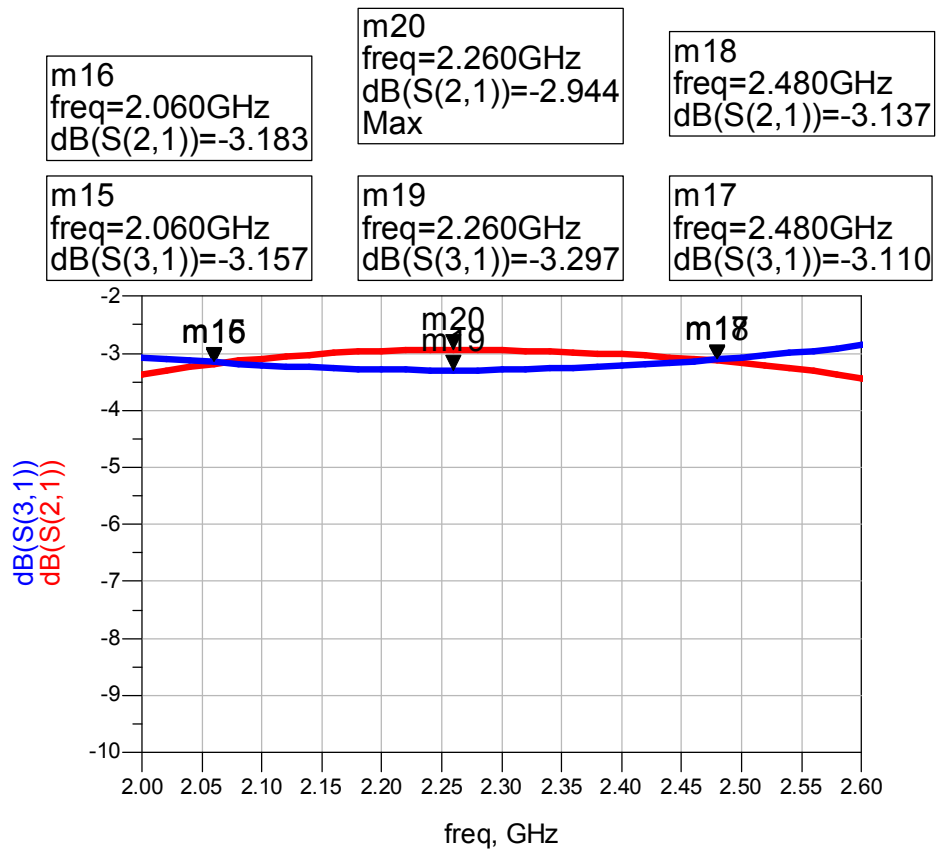


Figure (4.22) Simulation coupling balance of cascaded 90° hybrid coupler.

Figure (4.23) shows phase difference pointed by marker (m21 and m22) and figure (4.24) shows, marker (m23) where the phase output ports.

$$\text{Phase difference (S21-S31)} = (m21 - m22) = 89.923^\circ$$

Figure (4.24), shows, (m23) the phase of output ports where plot of phase difference between ports 2 and 3 just as we have divided the phase of S(2,1) by phase of S(3,1).

The marker m23 gives directly the phase difference between ports 2 and 3 at designed frequency (m23 = 89.923°).

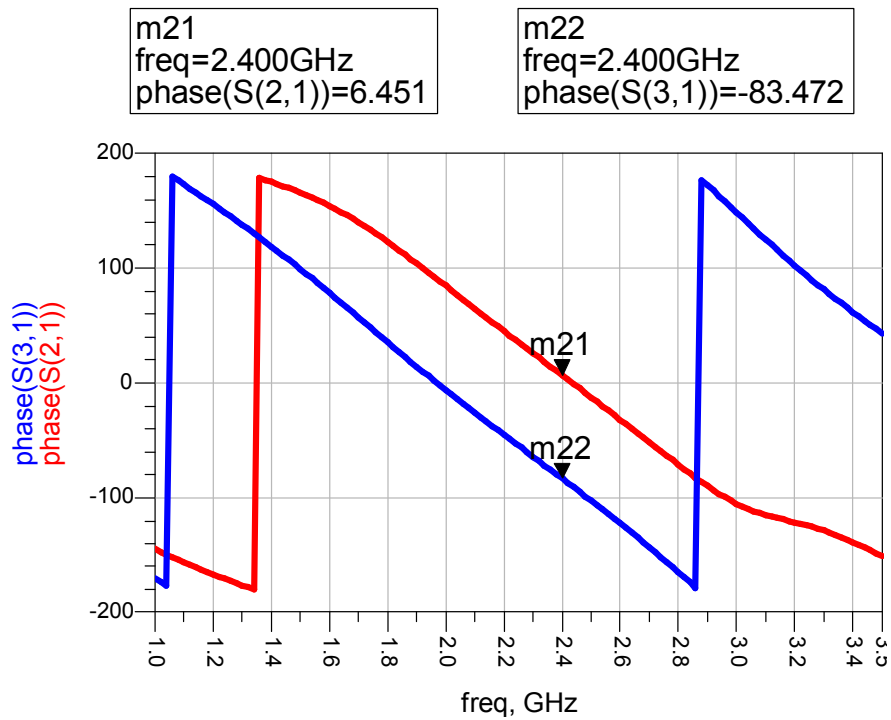


Figure (4.23) Simulation phase difference output of cascaded 90° hybrid coupler.

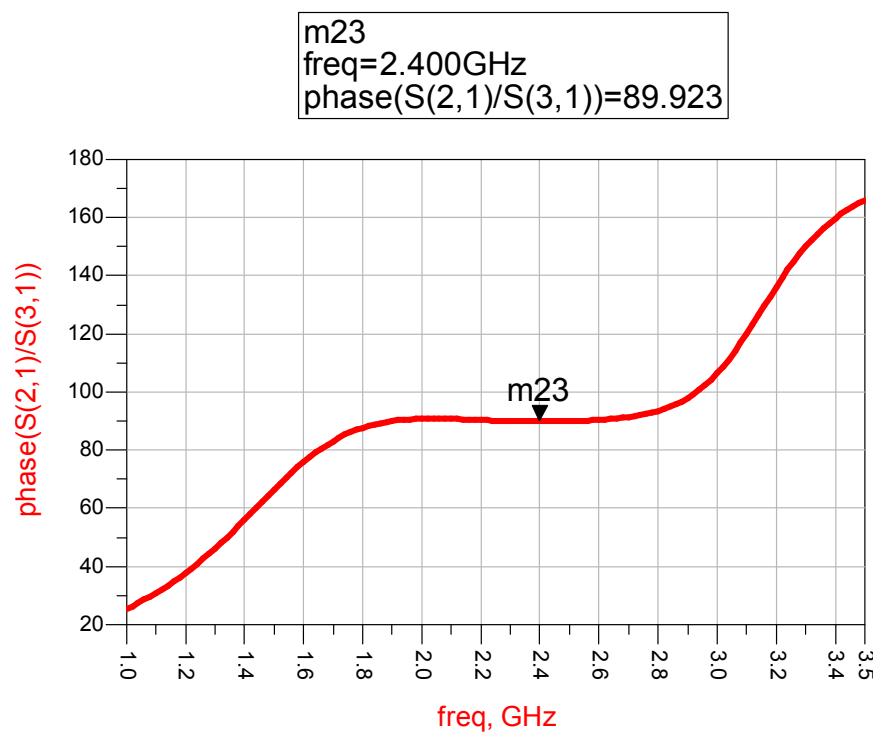


Figure (4.24) Simulation phase of cascaded 90° hybrid coupler .

4.4.1 Layout of Cascaded 90°Hybrid

In this type we use the conventional cascade as enhancement bandwidth. The multi-branch line coupler prototype and layout dimensions are illustrated in Figure (4.25).

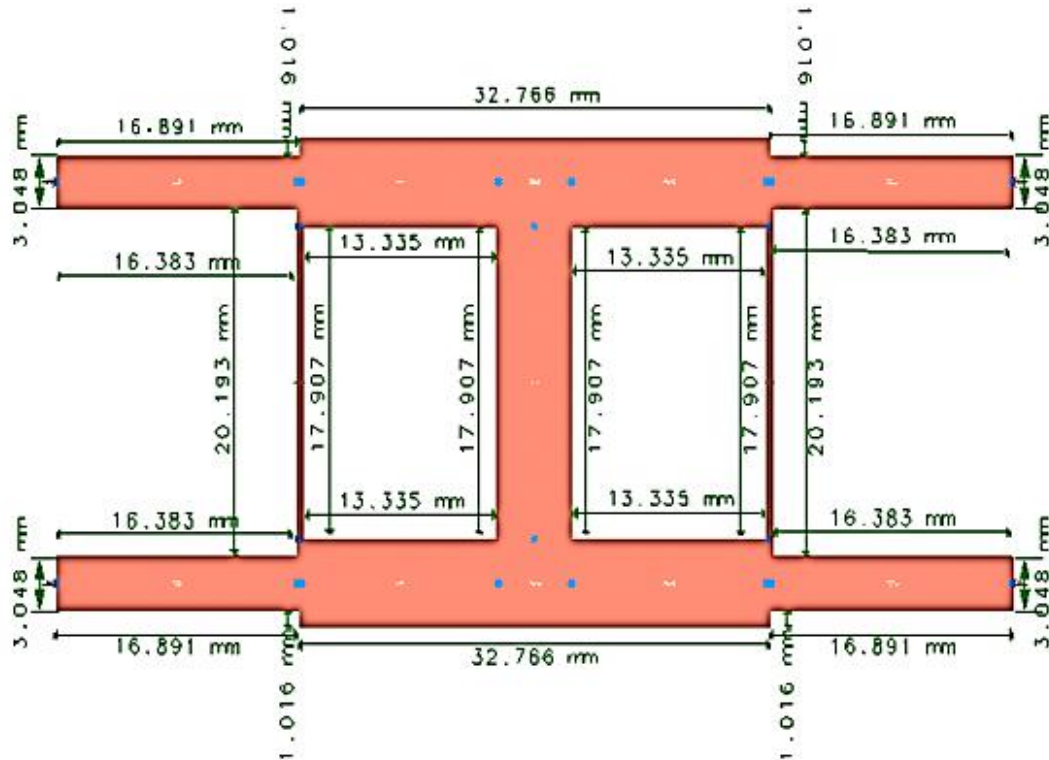
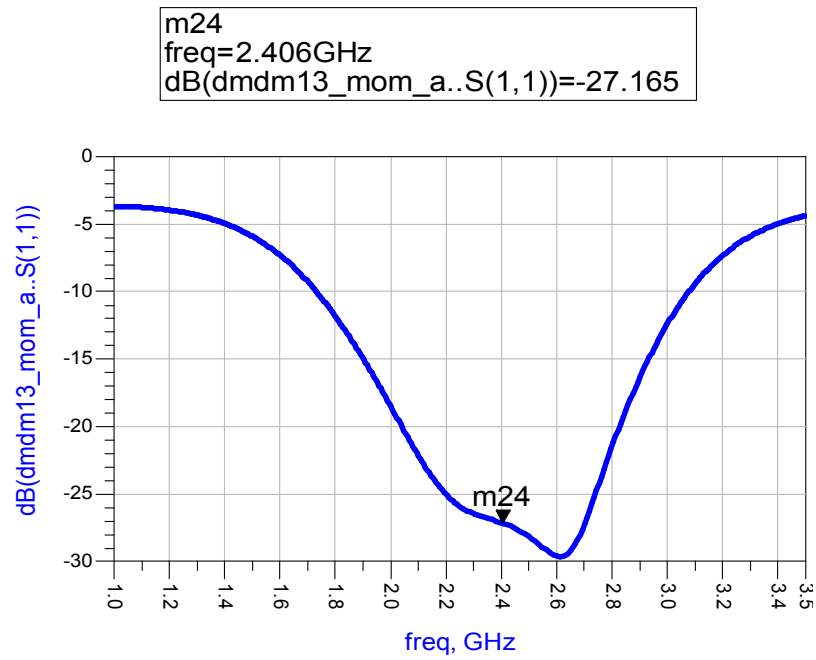


Figure (4.25) Layout of cascaded 90° hybrid with dimensions .

Figure(4.26), illustrate the simulated return losses of the cascade hybrid at resonance frequency of 2.4GHz which is S11 pointed by m24=-27.165dB. All simulation of the layout diagram were done by appealing a method of moment.



Figure(4.26) Return losses of cascaded 90°hybrid coupler.

Figure (4.27), shows the simulated isolation (S41) of 90° hybrid cascade at the designed frequency which is pointed by the marker m25=-28.736dB and it's a good isolation.

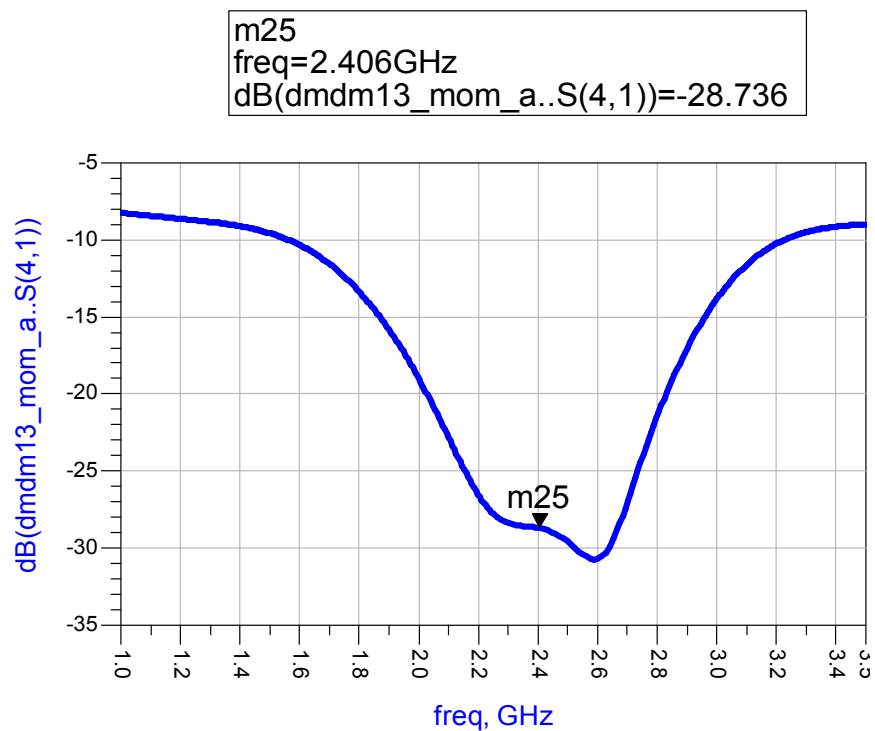


Figure (4.27) Isolation of cascaded 90°hybrid coupler.

Figure (4.28), shows the transmission coefficient from port 1 to ports 2 and 3 across the same frequency band as magnitude of coupling voltage at (m30 and m31) at range $\pm 0.707 < 1$. The simulated results of cascade hybrid depicting the amplitude very close to the ideal case of 0 dB. This will provide an even power split at the design frequency and close to even across the band as shown in figure(4.29) the value coupling balance as maximally flat $S_{31}=m_{36}=-3.610$ dB, its less than 1dB and satisfied the requirements.

Amplitude differences $(S_{13}-S_{12})=(m_{34}-m_{35})=0.080$ dB. its close 0dB.

The measured bandwidth from figure (4.29), will be between the markers m33(1.953GHz) and m35 (2.701GHz) which gives the bandwidth of 784MHz.

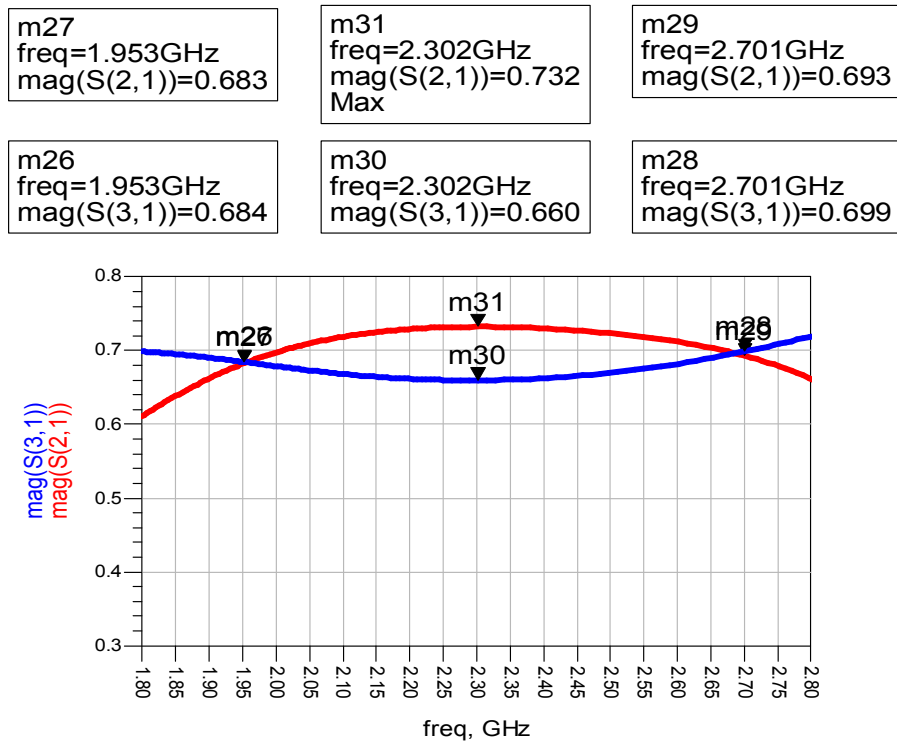


Figure (4.28) Transmission coefficients of cascaded 90° hybrid coupler.

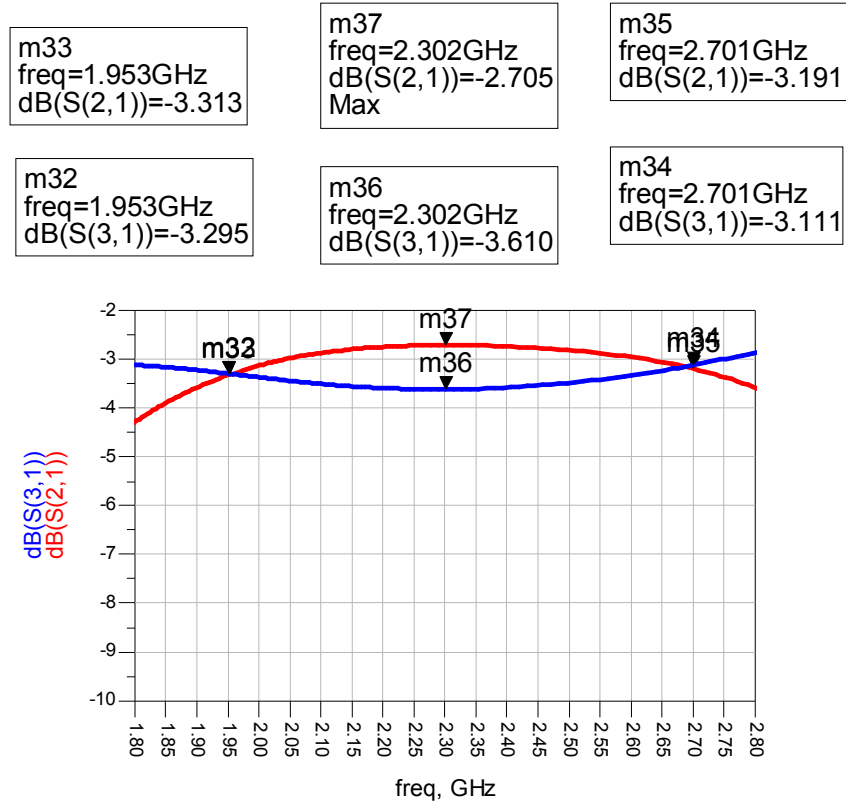


Figure (4.29) Coupling balance of cascaded 90° hybrid coupler.

Figure (4.30), shows phase difference indicated by markers (m38 and m39), then

$$\text{Phase difference } (S_{21}-S_{31})=(m_{38}-m_{39})=89.640^{\circ}.$$

From figure (4.31) the marker m40, shows the phase at the output ports where plot of phase difference between ports 2 and 3 just here, we divide the phase of S(2,1) by phase of S(3,1). The marker m40 gives directly the phase difference between ports 2 and 3 at designed frequency (m40=89.640°).

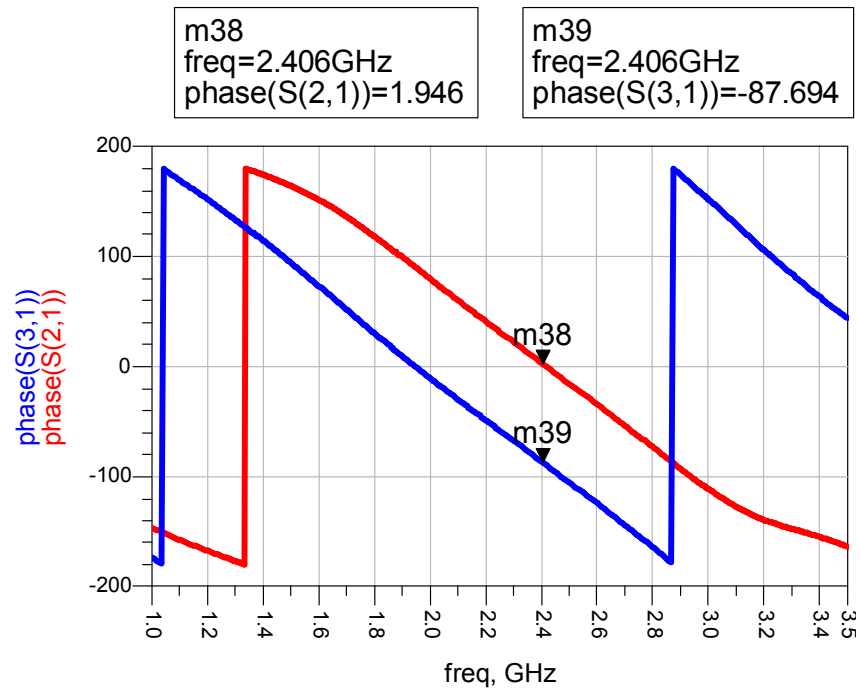


Figure (4.30) Phase difference of cascaded 90° hybrid coupler.

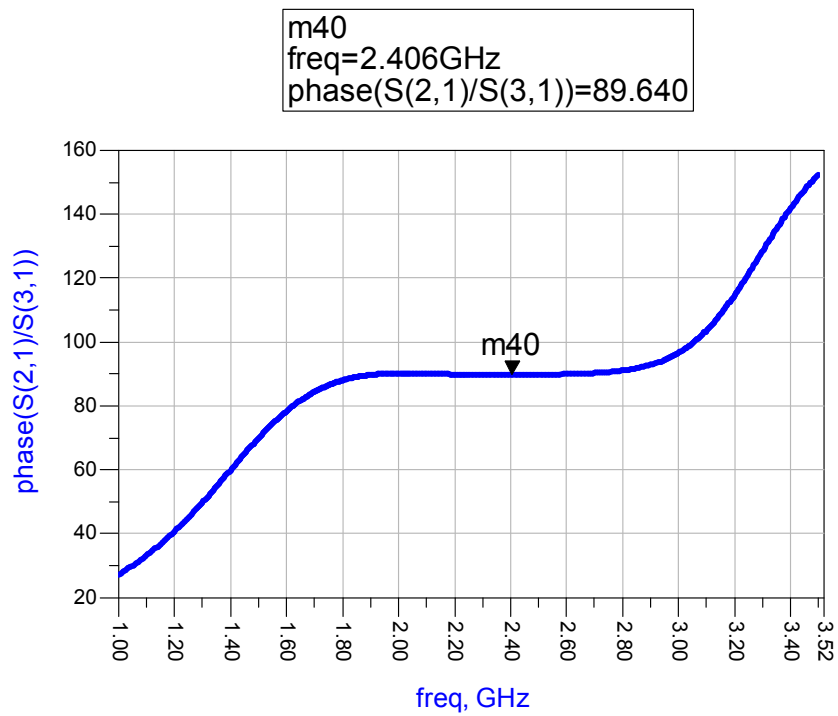


Figure (4.31) Phase of cascaded 90° hybrid coupler.

4.4.2 3D Layout of Cascaded 90° Hybrid

Figure (4.32) shows the layout box of 90° hybrid cascade by X, Y and Z in 3D and the proper highest point of the box is ($h_u=30\text{mm}$). This will help to achieve optimum result for design.

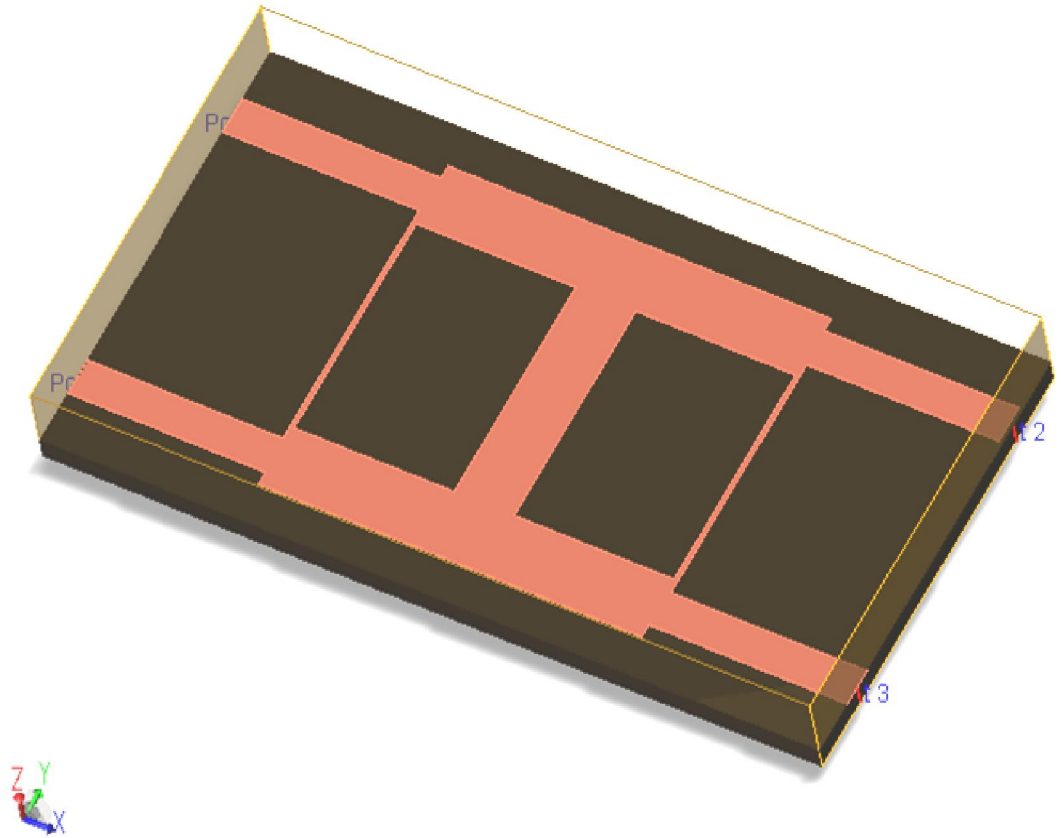


Figure (4.32) 3-Dimensional layout cascaded 90° hybrid coupler.

4.5 Case3. Simulation of 90° Hybrid With Extended TL Output ports

In this case replaced transmission lines of output ports by two extended lengths (a section of open -end transmission line) for enhancement at the phase. The extended ports design maintained in chapter 3 (section 3.4.2).

The design expected delivering high isolation, minimal insertion loss and exceptional phase and amplitude balance. The results obtained from this test case demonstrate the accuracy, efficiency, and scalability of the proposed method.

To reach the best phase simulation to get optimum result for 2.4GHz hybrid coupler, by using the tuning techniques in ADS software in order to tune the extended transmission line lengths. The best results of the lengths and phases are listed in Table(4.1).

Table (4.1) The tuned lengths and phases of 90° hybrid with extended TL output ports.

L	$\Delta\Phi$	Φ
L1= 34.1194mm	170.314 - 80.285	$90^\circ.029$
L2= 34.46059mm	169.417 - 79.415	$90^\circ.005$
L3 = 34.80178mm	168.518 – 78.542	$89^\circ.976$

Figure(4.33), shows 90° hybrid with best extended lengths of output ports which also corresponds to the best phase(L2):-

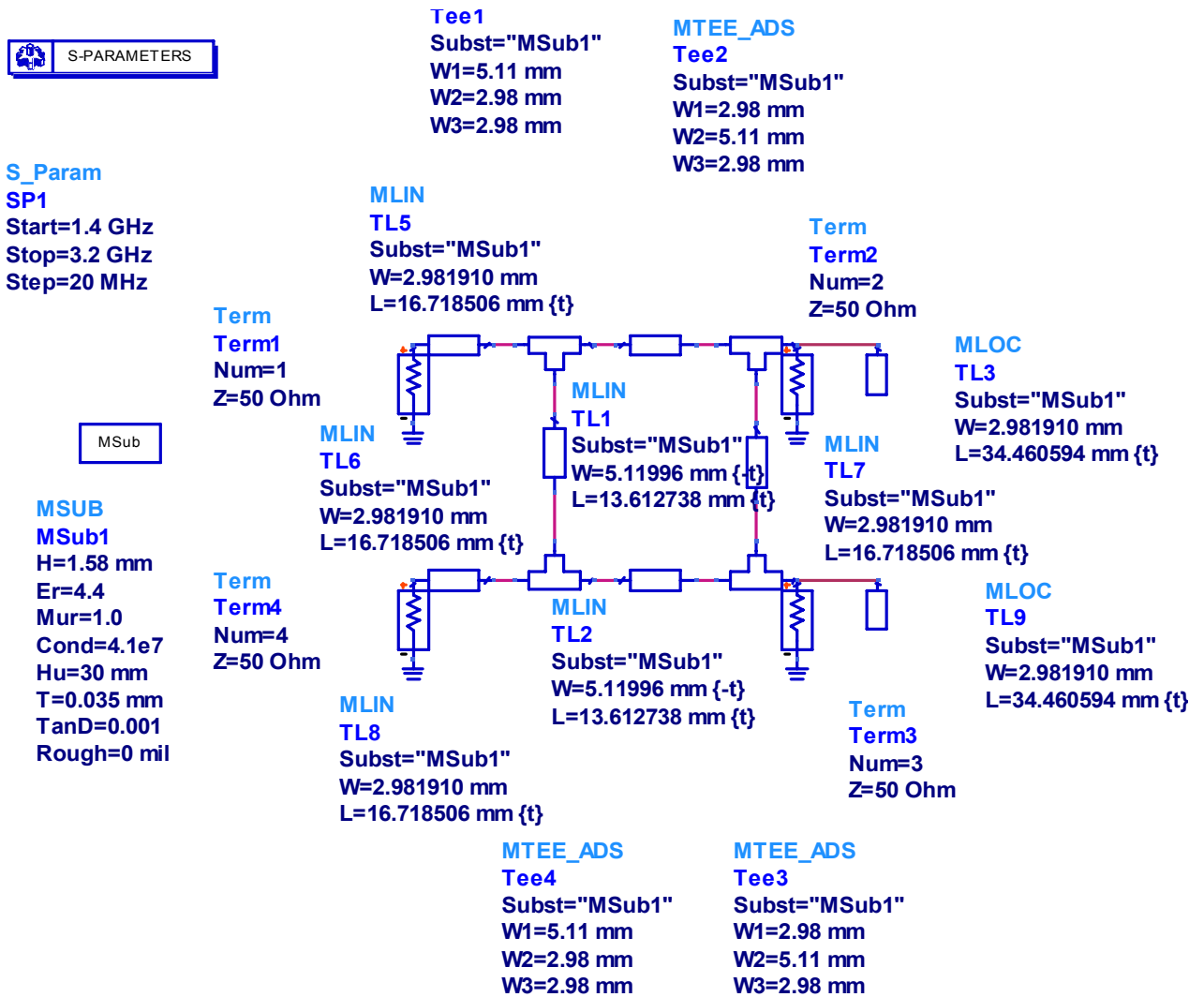
$$L2 = 34.46059\text{mm}$$

The variations of lengths is a good method in order to obtain good phase result in this case and must consider changing in transmission line lengths at the output ports will affect the resonance frequency.

Then figure (4.33) shows, the schematic design of 90° hybrid with extended TL output ports.

At the Open-End of a microstrip line with a width of w , the fields do not stop abruptly, but extend slightly further because of the effect of fringing field.

This effect can be modeled either with an equivalent shunt capacitance C , or with an equivalent length of transmission line L .



Figure(4.33) Schematic diagram of 90° hybrid with extended TL output ports.

Figure (4.34) and figure (4.35) shows, return losses (S11,S22) at markers m1 and m2 respectively, both figures have the same value with little difference (Two ports has low return losses approximately -34dB). In these figures we have a deep null at 2.4GHz, which means that a very good matching to 50Ω at the desired frequency.

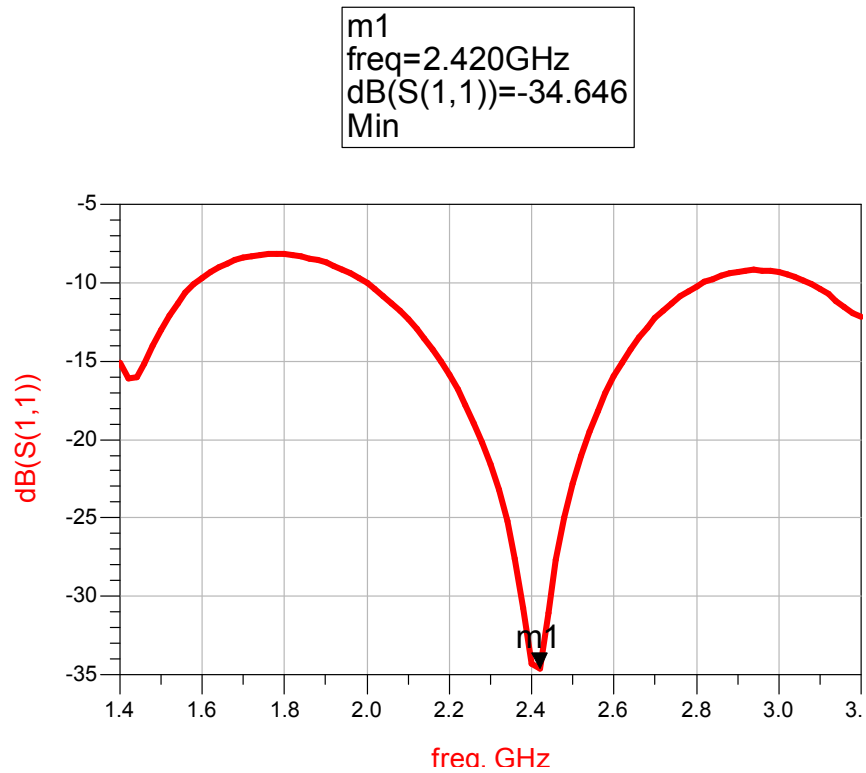


Figure (4.34) Simulation return losses of 90°hybrid with extended TL output ports.

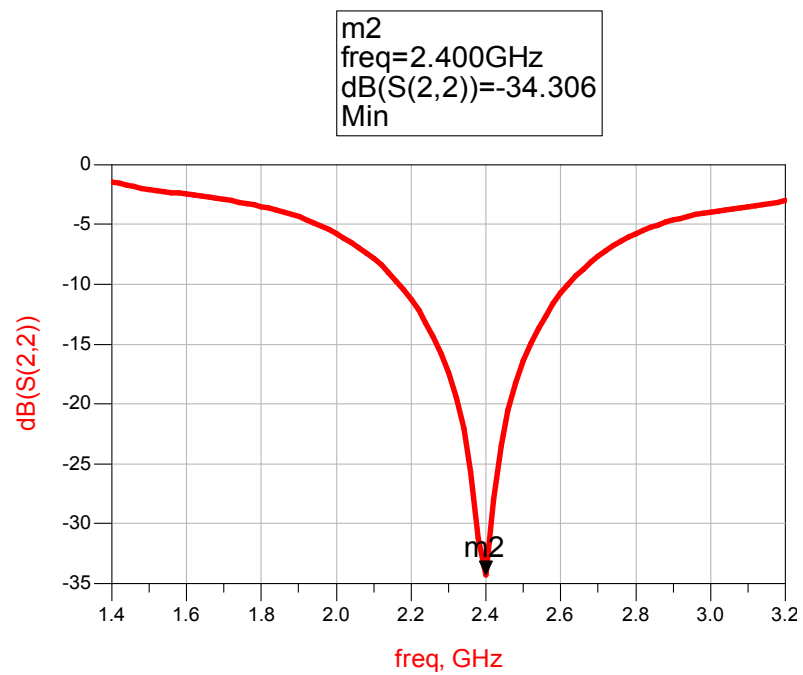


Figure (4.35) Simulation return losses of 90°hybrid extended TL output ports.

Figure (4.36), shows simulation the isolation of 90°hybrid with extended output ports. Isolated output at m3 of $S_{41} = -38.366$ dB, it does not depend on losses. This is generally characteristic of full symmetry couplers. By observing the improvement of isolation of the 90-degree realization provides about 38.366 dB isolation. It has better isolation than the conventional 90°hybrid.

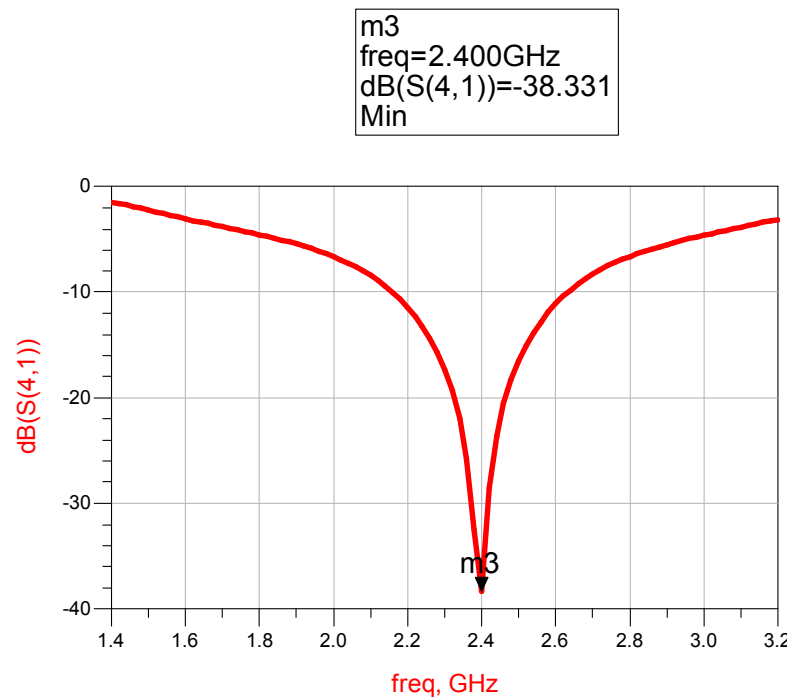


Figure (4.36) Simulation isolation of 90°hybrid with extended TL output ports.

Figure (4.37) shows reciprocal design when the input at port 4 ($S_{ij} = S_{ji}$). That it does not make any difference which port is the input port as long as the relationship between ports remains.

A magnitude transmission coefficient of 90°Hybrid with extended TL output ports at bandwidth frequency starting at 2.240 GHz until 2.520 GHz, from port 1 to ports 2 and 3 is plotted in figure (4.38), this magnitude of coupling voltage at markers m14 and m15 in the range of $\pm 0.707 < 1$ which is close to the ideal case.

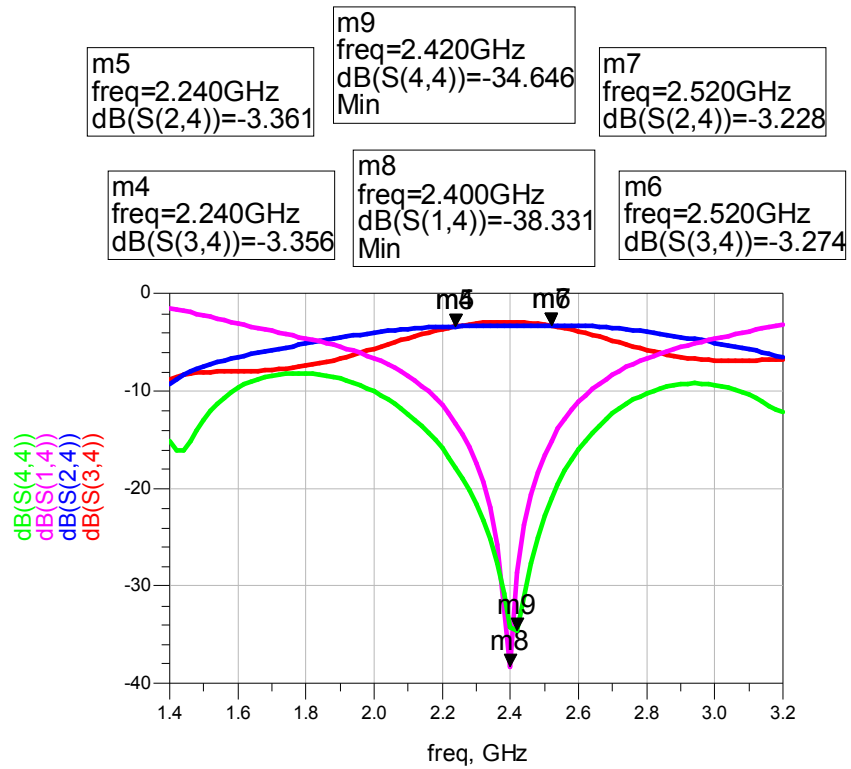


Figure (4.37) Simulation reciprocal of 90° hybrid with extended TL output ports.

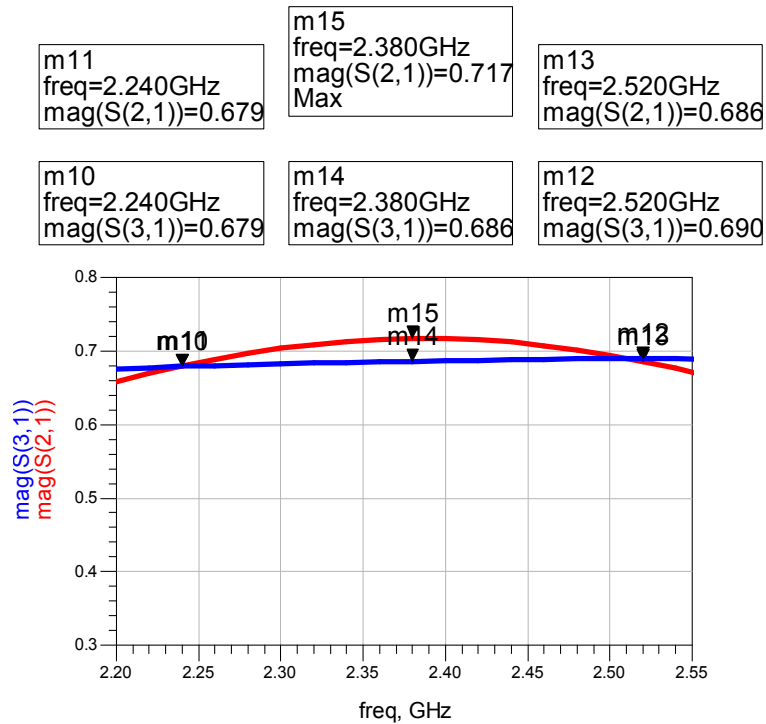


Figure (4.38) Simulation transmission coefficients of 90° hybrid with extended TL outputs

Figure(4.39) the value coupling balance as flat maximally $S_{31}=m_{20}=-3.277$ dB , its less than 1 dB which is as required.

Amplitude differences which are indicated by markers m18 and m19 ($S_{31}-S_{21})=(m_{18}-m_{19})=0.000$ dB, its exactly 0dB as in ideally case, and the bandwidth BW =275 MHZ , which is less than conventional hybrid bandwidth.

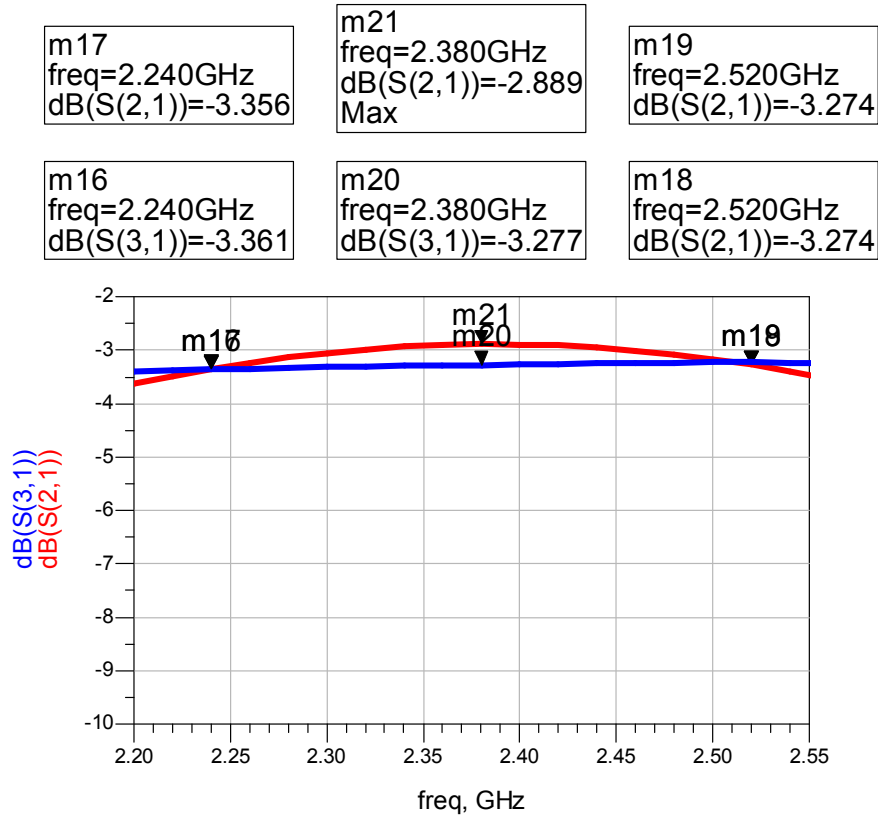


Figure (4.39) Simulation coupling balance of 90°hybrid with extended TL .

As has been illustrated in figure (4.40), the phase difference is indicated by markers (m22 and m23) , which is given by:-

$$\text{Phase difference } (S_{21}-S_{31})=(m_{22}-m_{23})=90^{\circ}.005$$

Figure (4.41) plots the phase of output ports where plot of phase difference between ports 2 and 3 just as divided the phase of $S(2,1)$ by phase of $S(3,1)$.The marker m40 gives directly the phase difference between ports 2 and 3 at designed frequency (m24 =90° .005).

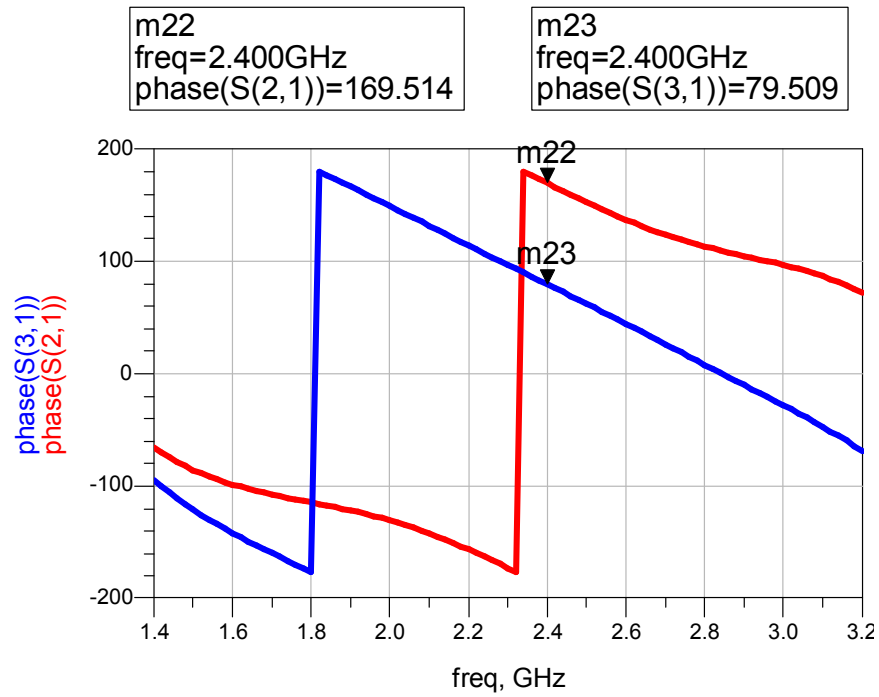


Figure (4.40) Simulation phase difference of 90° hybrid with extended TL .

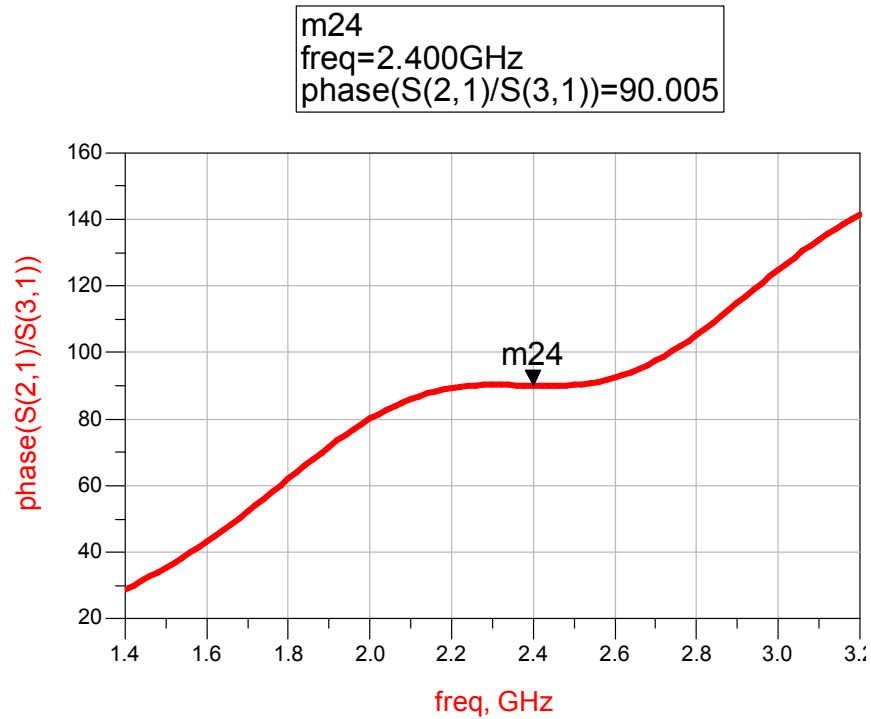


Figure (4.41) Simulation phase of 90° hybrid with extended TL output ports.

4.5.1 Layout of 90° Hybrid With Extended TL Output Ports

By using a command generate layout in ADS software the schematic design of figure (4.33) is transformed to the layout design with all length and width dimensions and substrate specifications as shown in figure (4.42).

Tuning process is used at the layout design in order to obtain the target specifications of the design.

Figure(4.42) shows, the layout design with all dimension after using the tuning process which is satisfies our specification after momentum simulation is run .

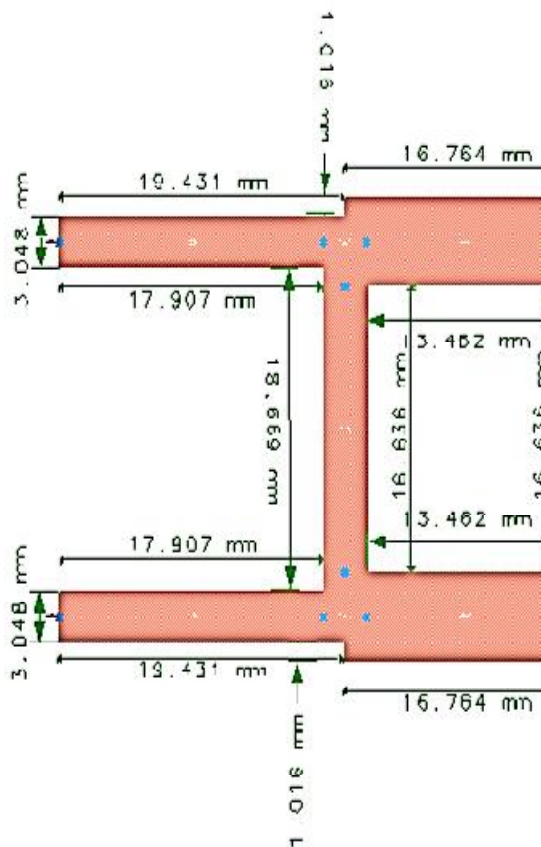


Figure (4.42) Layout of 90°hybrid with extended TL with dimensions .

Figure (4.43) shows, return loss at marker m25=-30.975dB whereas figure (4.44), shows return loss at m26=-24.472dB.

Both figures illustrate that minimum return losses at S11 and S22 which means that minimum reflection coefficient at that ports have good matching (closed to 50Ω at 2.4GHz).

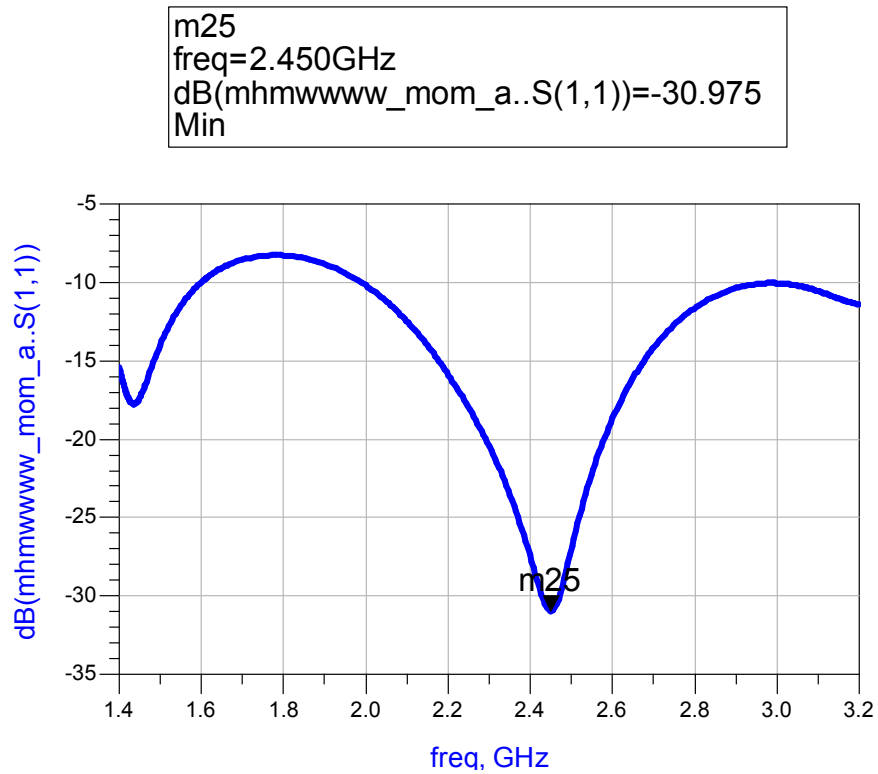


Figure (4.43) Return losses of 90°hybrid with extended TL output ports.

Figure (4.45), shows the simulation isolation of 90°Hybrid with extended outputs.

Isolated output at marker m27 is $S_{41} = -25.588\text{dB}$.

Figure (4.46) shows the magnitude transmission coefficient from port 1 to ports 2 and 3 across the same frequency band as magnitude of coupling voltage at range of ± 0.707 which is less than 1V as indicated by the marker m32=0.662V.

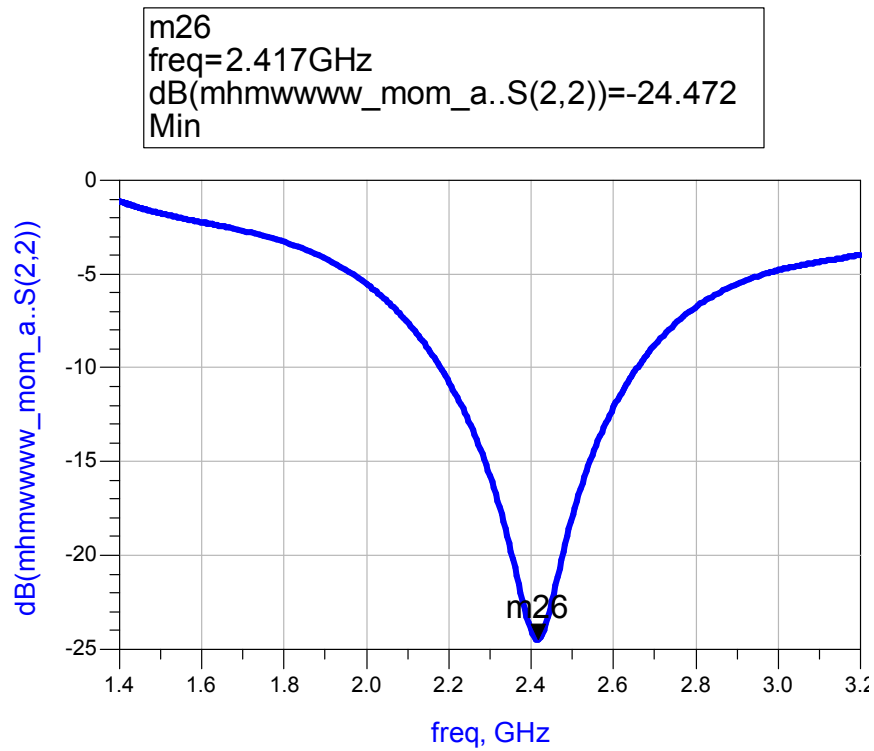


Figure (4.44) Return losses of 90°hybrid with extended TL output ports.

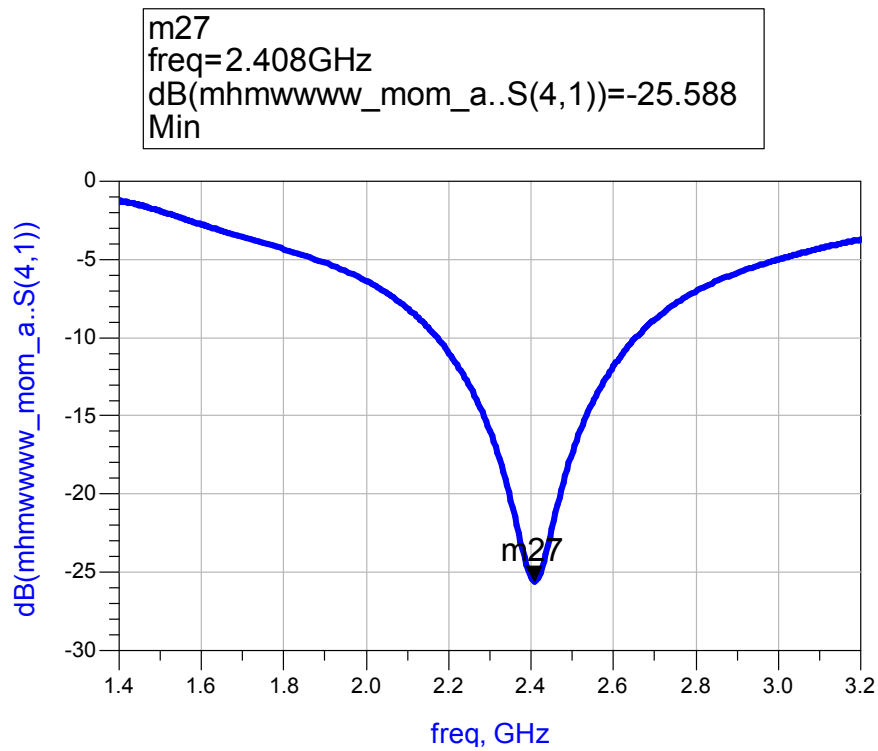


Figure (4.45) Isolation of 90°hybrid extended TL output ports.

m29 freq=2.188GHz mag(S(2,1))=0.653	m33 freq=2.403GHz mag(S(2,1))=0.727 Max	m31 freq=2.619GHz mag(S(2,1))=0.658
m28 freq=2.188GHz mag(S(3,1))=0.654	m32 freq=2.403GHz mag(S(3,1))=0.662	m30 freq=2.619GHz mag(S(3,1))=0.664

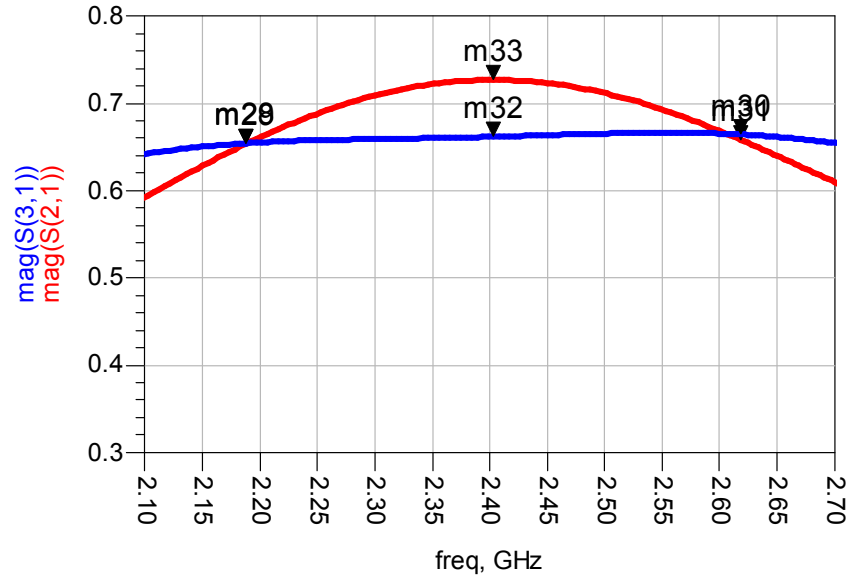


Figure (4.46) Transmission coefficients of 90°hybrid with extended TL .

As been illustrated in Figure(4.47), the value of coupling balance as flat maximally $S_{31}=m_{38}=-3.584$ dB , its less than 1dB as close as ideally case, amplitude differences $(S_{31}-S_{21})=(m_{34}-m_{35})=0.013$ dB, it's close to 0 dB.

The bandwidth of 90°Hybrid with extended TL output ports starting from frequency 2.188 GHZ to 2.619 GHZ, from port 1 to ports 2 and 3 across the same frequency band as shown in figure (4.46) and figure (4.47).

Then the calculated bandwidth is 431 MHZ.

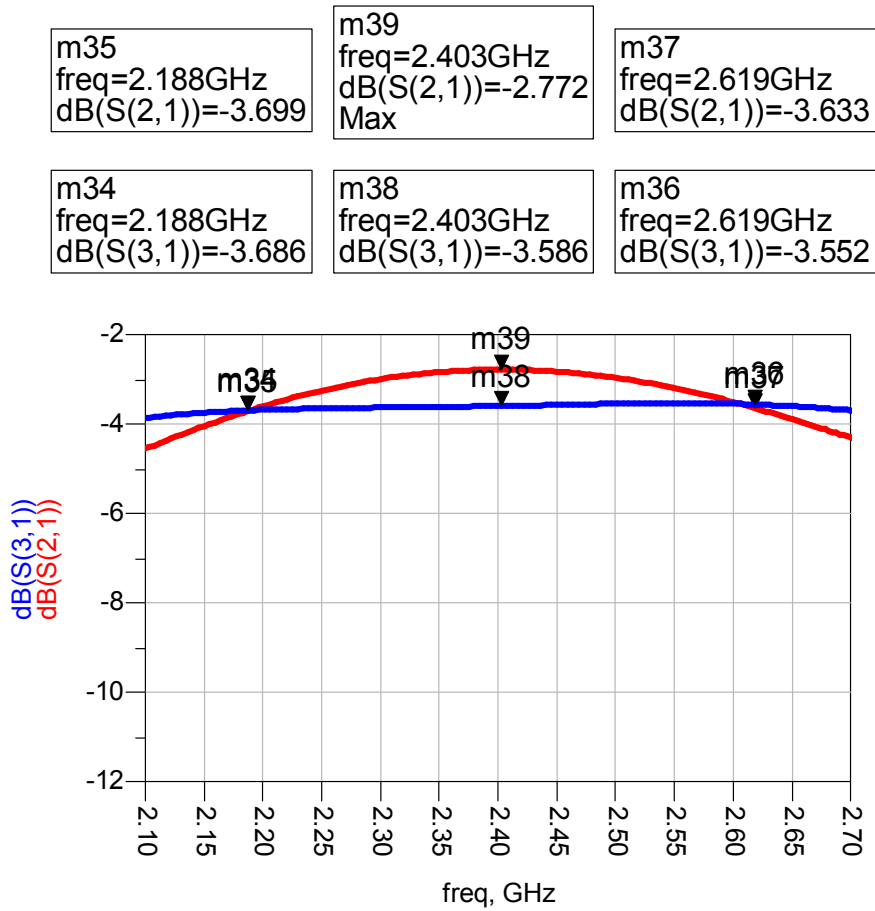


Figure (4.47) Coupling balance of 90°hybrid with extended TL .

We observe improvement of bandwidth in layout diagram than bandwidth in the schematic diagram and also the phase improved as shown in figures (4.48) (4.49).

Figure (4.48) shows phase difference indicated by markers (m40 and m41), then

$$\text{Phase difference (S21-S31)}=(m40-m41) = 90.008^\circ$$

Figure (4.49) record the phase at output ports where the marker m42 gives directly the phase difference between ports 2 and 3 at designed frequency (m42=90.008°).

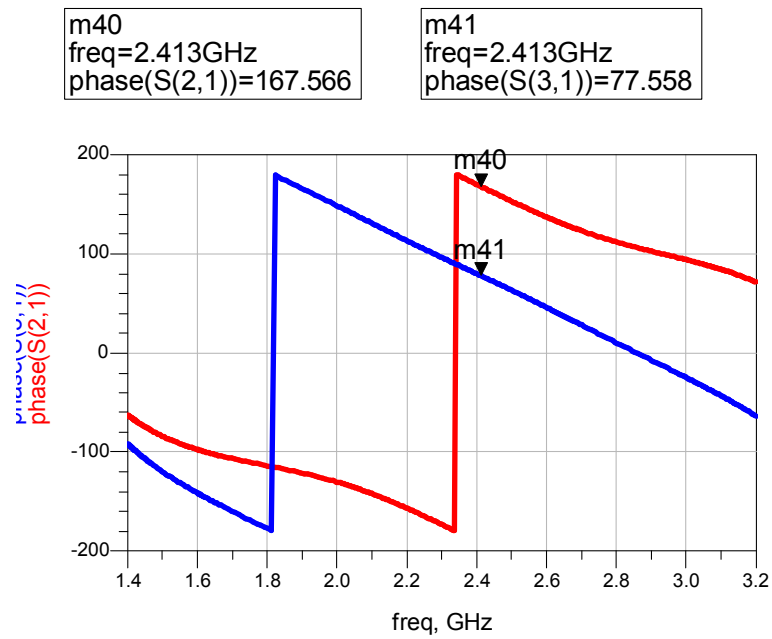


Figure (4.48) Phase difference of 90° hybrid with extended TL output ports .

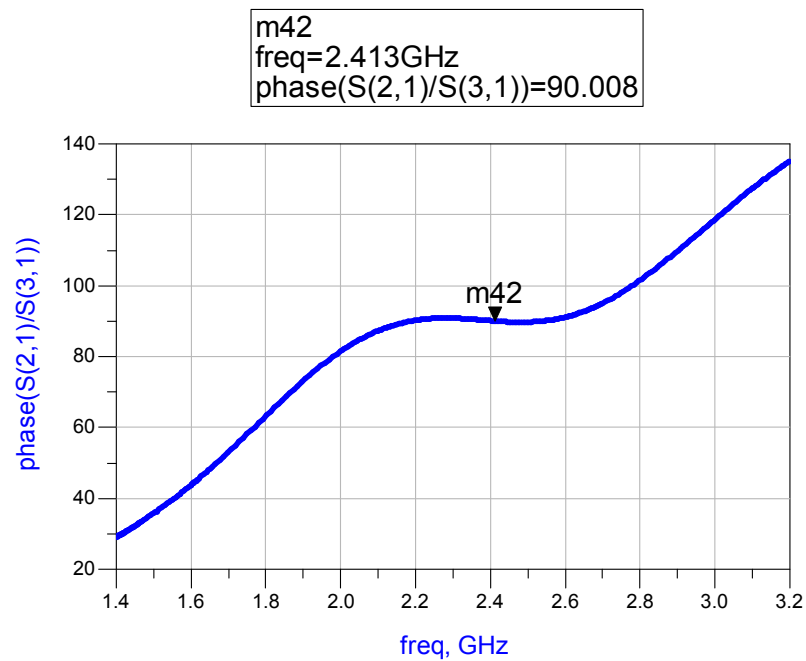
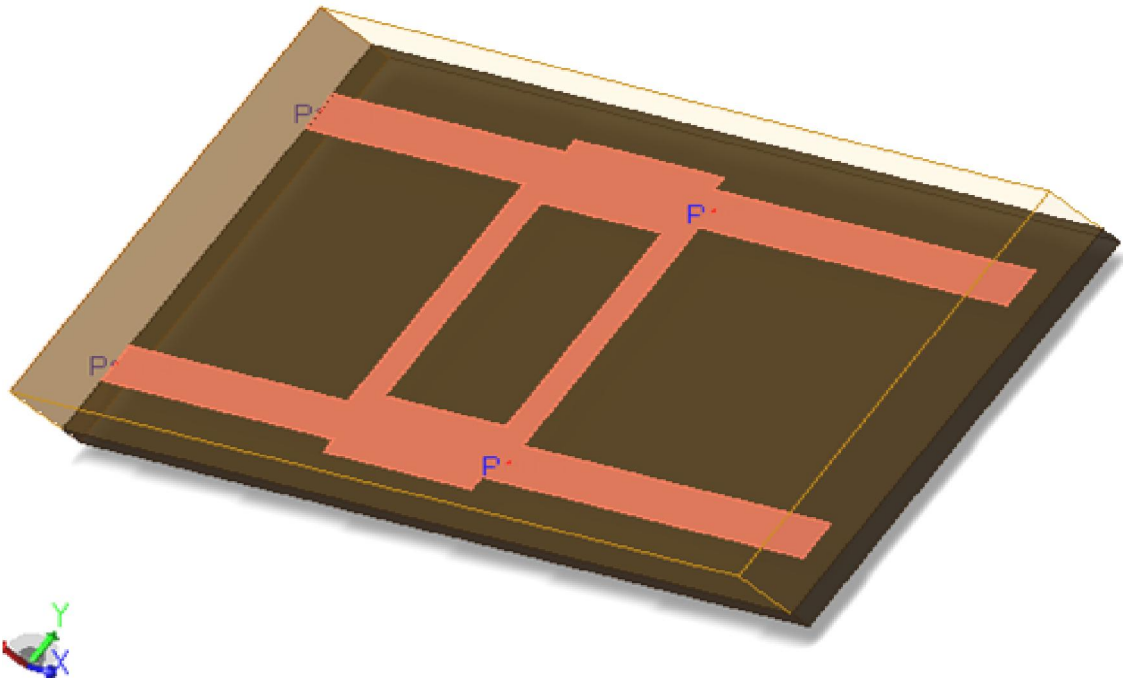


Figure (4.49) Phase of 90° hybrid with extended TL output ports.

4.5.2 3D Layout of 90° Hybrid With Extended TL Output ports

Figure (4.50) shows the layout box of 90° hybrid extended outputs by X, Y, Z 3D and the proper highest point of box is ($h_u=30\text{mm}$). This will help to achieve optimum result for design.



Figure(4.50) 3-Dimensional layout of 90°hybrid with extended TL output ports.

4.6 Case4. Simulation of 90° Hybrid With Single Stub at the Output ports

In this case by fixed two single stubs of output ports to enhance the bandwidth and phase. Useful matching stubs appropriate can be made adjustable so that matching can be corrected on test .

The single stub design maintained in chapter 3. It is ideally suited for hybrids where good matching are required and will benefit for high isolation.

The implementation of stubs are applied at the branch line coupler to increase the frequency bandwidth and obtained phase different between outputs as ideally as 90°. However, this technique is applied for single frequency band which is 2.4GHz. As for conventional design, all branch lines are quarter-wavelength long.

Figure (4.51) display the schematic diagram of hybrid with stubs and RC series load.

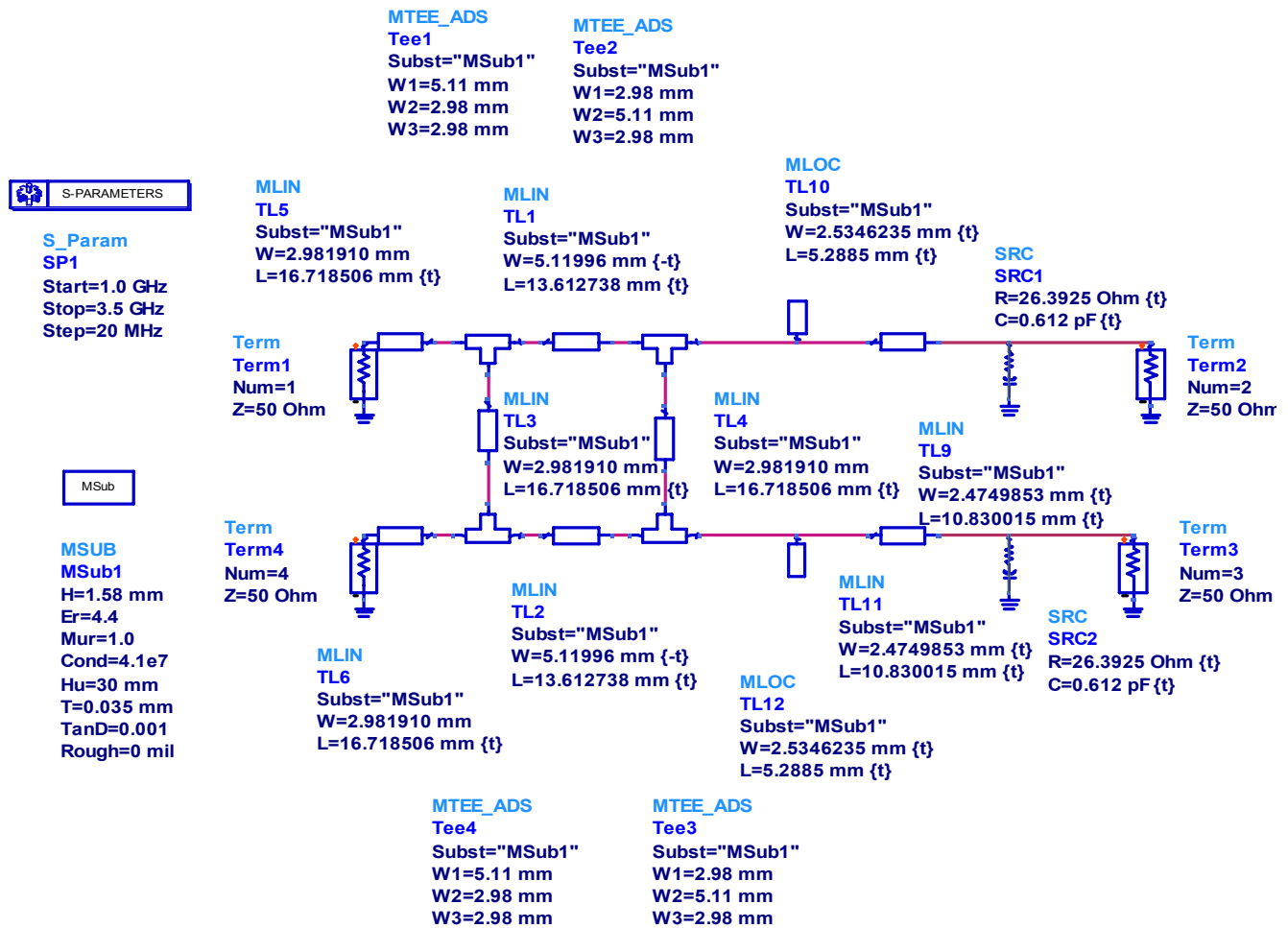


Figure (4.51) Schematic diagram 90° hybrid with two single stub at the output ports.

Figure (4.52) shows return loss at marker (m1) and figure (4.53) shows return loss at marker (m2), both figures of hybrid have different values of resonance at frequency 2.4GHz which indicated by marker (m1) of $S_{11} = -34.586$ dB where $S_{11} = S_{44}$ with deep null of resonance frequency and indicated by marker (m2) of $S_{22} = -20.570$ dB where $S_{22} = S_{33}$, because the stub on output ports obvious increased wide band at output ports; by this configurations obtained good matching for all ports (1, 2, 3 and 4) without any effect to different resonant frequency of the output ports. Observe that improving on Figure (4.54), where the Simulation of isolation of hybrid with single stub output get the best isolation indicated by marker at (m3) of $S_{41} = -43.489$ dB.

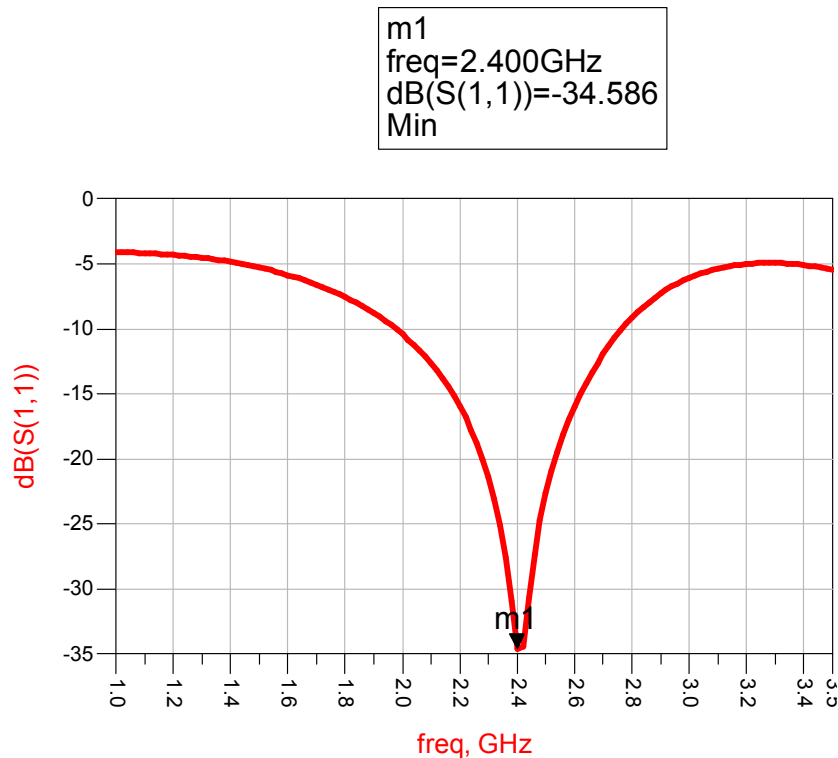


Figure (4.52) Simulation return losses of 90°hybrid with two single stub at the output ports.

This is S41 value very desired, because there will be maximum power delivered to the device at this frequency and then will be provides a best phase balance as shown as figure (4.59). The matching circuits, which incorporate mixed lumped and transmission line elements, are widely used both in hybrid and monolithic design technique.

This results relative by effects the series capacitance and resistance (SRC) with single stub design and transmission lines lengths of stub. The use of the series resistance with a capacitance will give the stability of the phase at the output ports.

The device behavior at reciprocal system ($S_{ij}=S_{ji}$) as shown in figure (4.55).

It also makes no difference which port is the input, this configuration ensures a high degree of isolation between the two output ports and the input port without unwanted interaction between them.

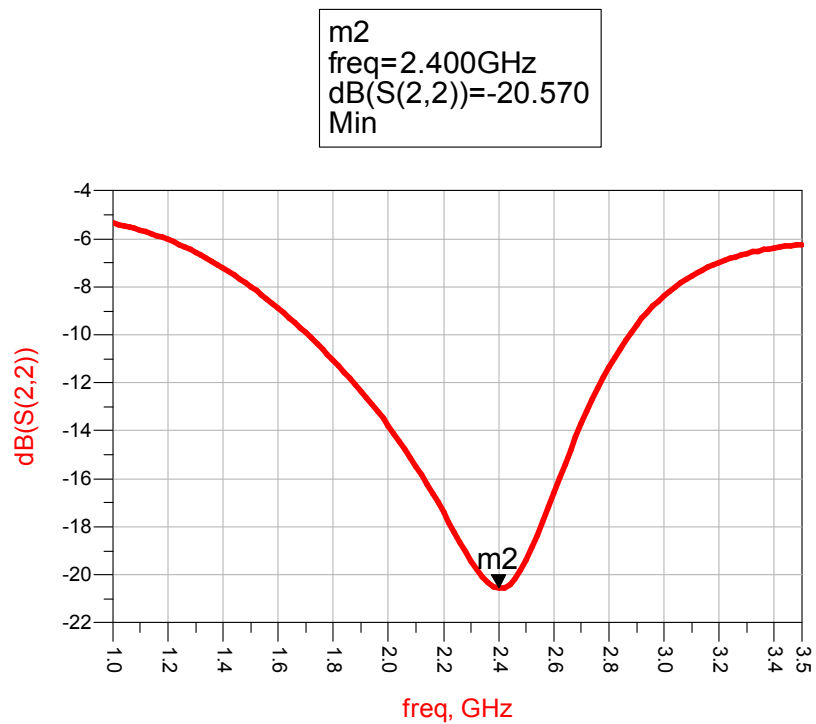


Figure (4.53) Simulation return losses of 90°hybrid with two single stubs at output ports.

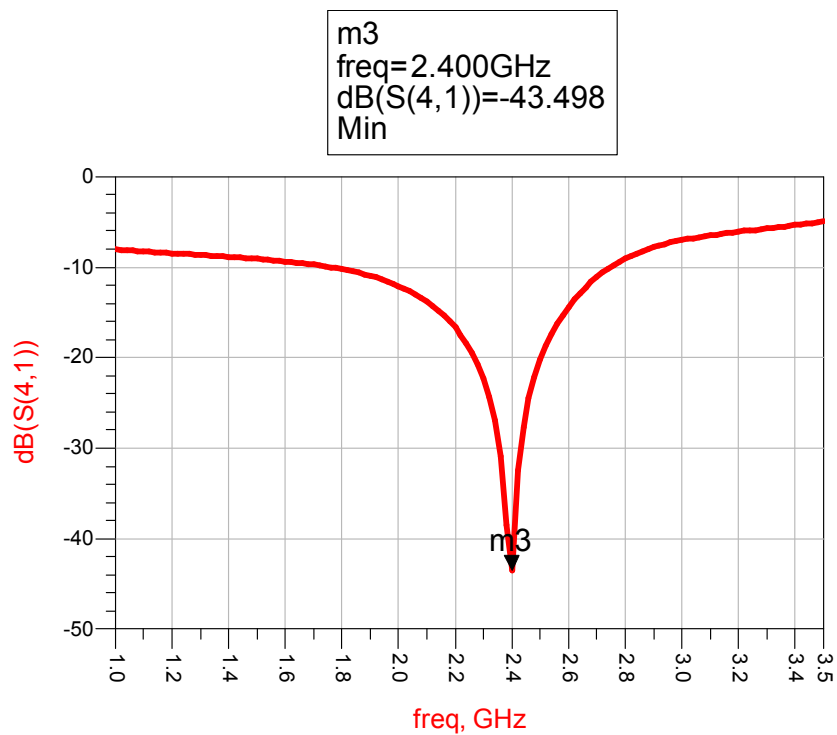


Figure (4.54) Simulation isolation of 90°hybrid with two single stub at the output ports.

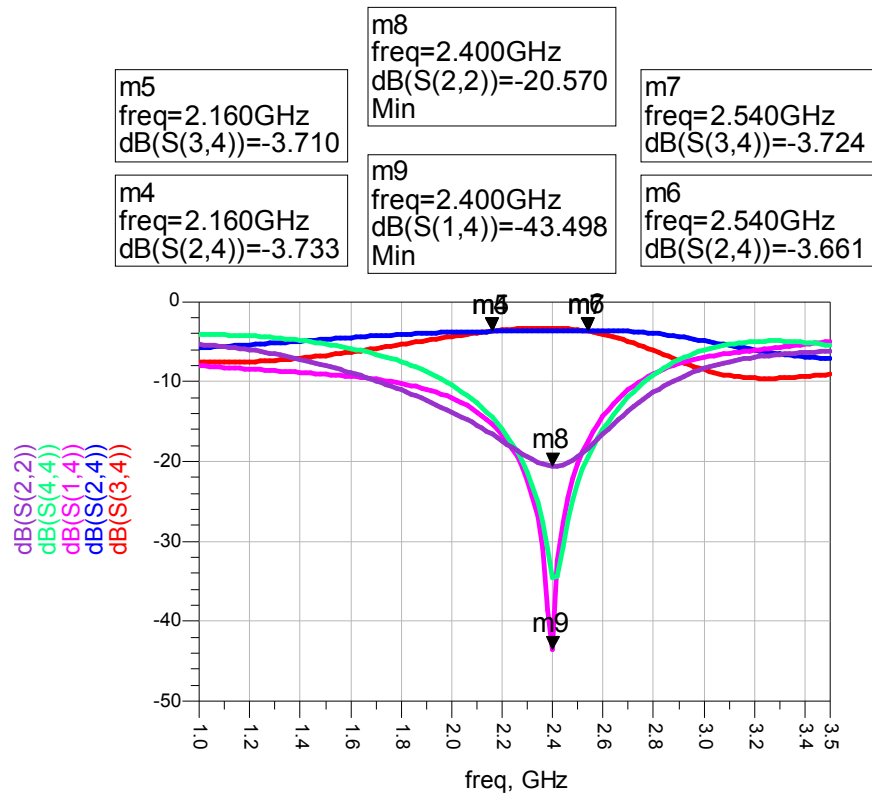


Figure (4.55) Simulation reciprocal of 90°hybrid with two single stub at the output ports.

The 90°Hybridwith single stub output ports shows wide band of frequency starting at 2.160GHz until 2.540 GHz as shown as Figure (4.56), display the transmission coefficient from port 1 to ports 2 and 3 (in dB) across the same frequency band as magnitude of coupling voltage at markers m14 and m15 which have 0.653V and 0.683V respectively, its less than 1 as close as possible to ideal case .

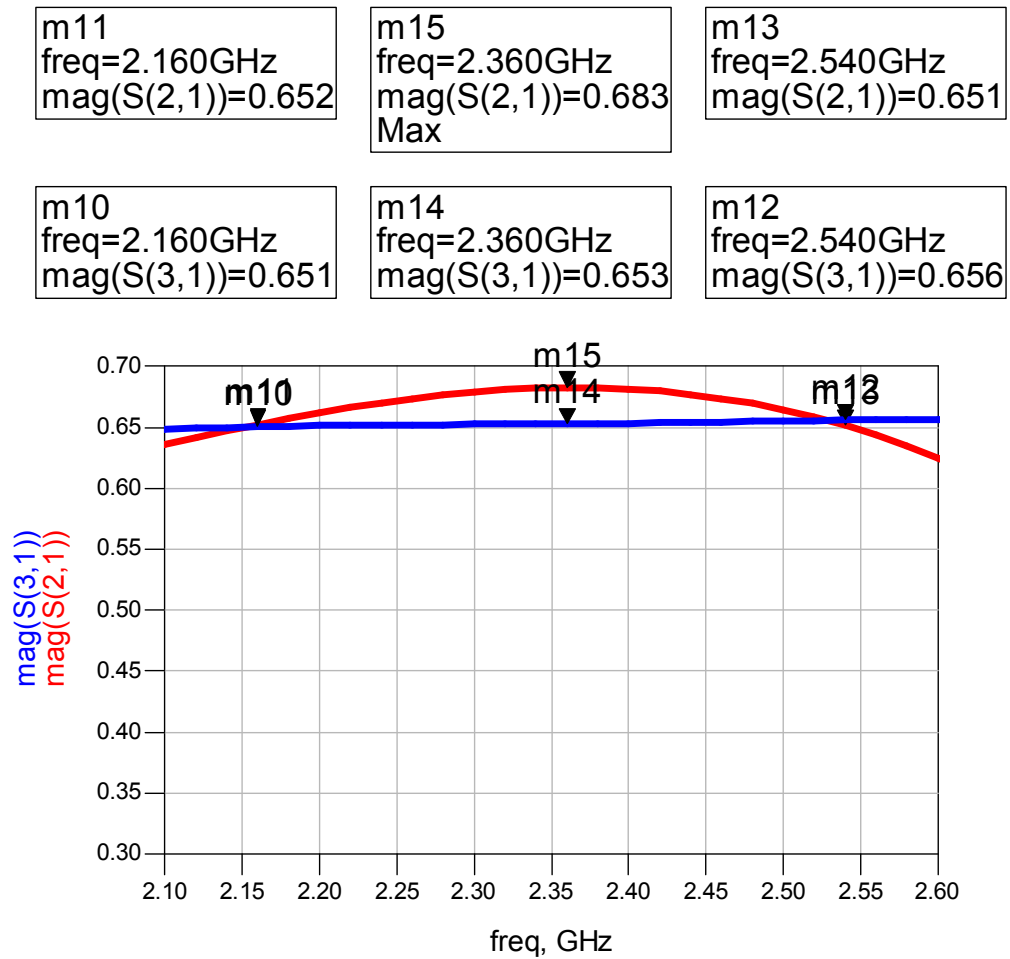


Figure (4.56) Simulation transmission coefficients of 90°hybrid with two Single stub at the output ports.

As been illustrated in Figure (4. 57) the value coupling balance as maximally flat pointed by marker (m20) of $S_{31}=-3.704$ dB, its less than 1dB as close as ideally case .

Amplitude differences pointed at markers $(m18 - m19) = (S_{31}-S_{21})= 0.063$ dB. It's close 0 dB as in ideal case. If result in two equal amplitude signals that are quadrant (90° apart). It also makes extend frequency coverage 2.160 GHz until 2.540 GHz with flat response. Then $BW=420$ MHz .

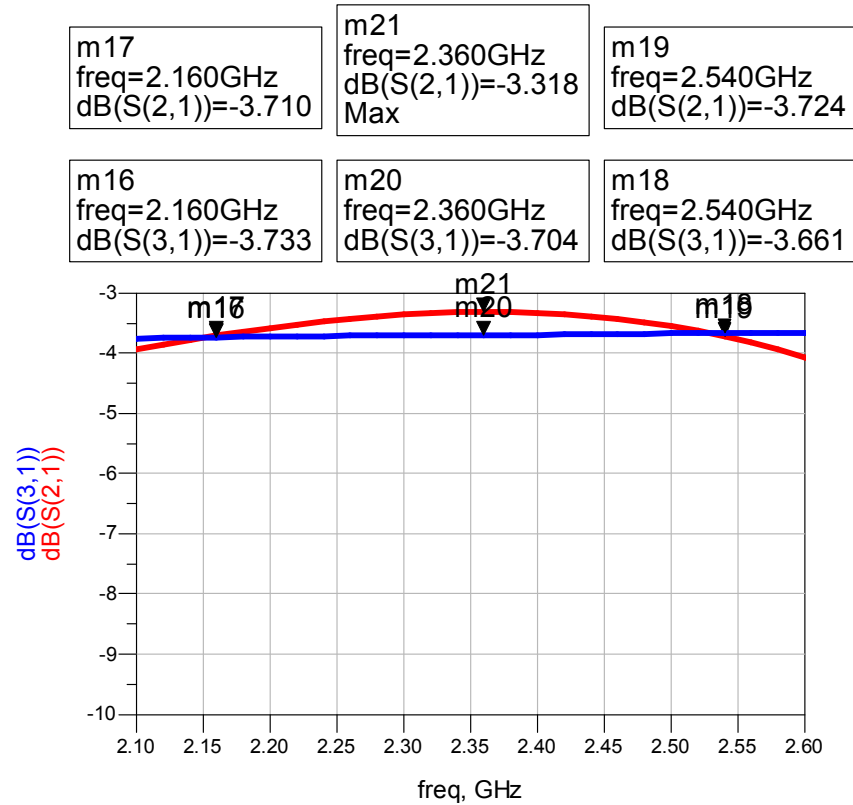


Figure (4.57) Simulation coupling balance of 90° hybrid with two single stub at the output ports.

Figure (4.58), shows the simulated phase at port2 and port 3. The phase difference between port2 and port1 which marketed by (m22) at the frequency 2.4 GHz is 87.469° where as the marker(m23) gives the difference -2.531°, then the phase difference between ports 2 and 3 is

$$\text{Phase difference (m22-m23)} = 90^\circ.000.$$

The device obvious provides best improvement in phase than all last cases..

This ideal phase difference between these two ports. Figure (4.59) is also a plot of phase difference between ports 2 and 3 just have we divided the phase of S(2,1) by phase of S(3,1).The marker (m24) gives directly the phase difference between ports 2 and 3 at designed frequency (m24 = 90°.000).

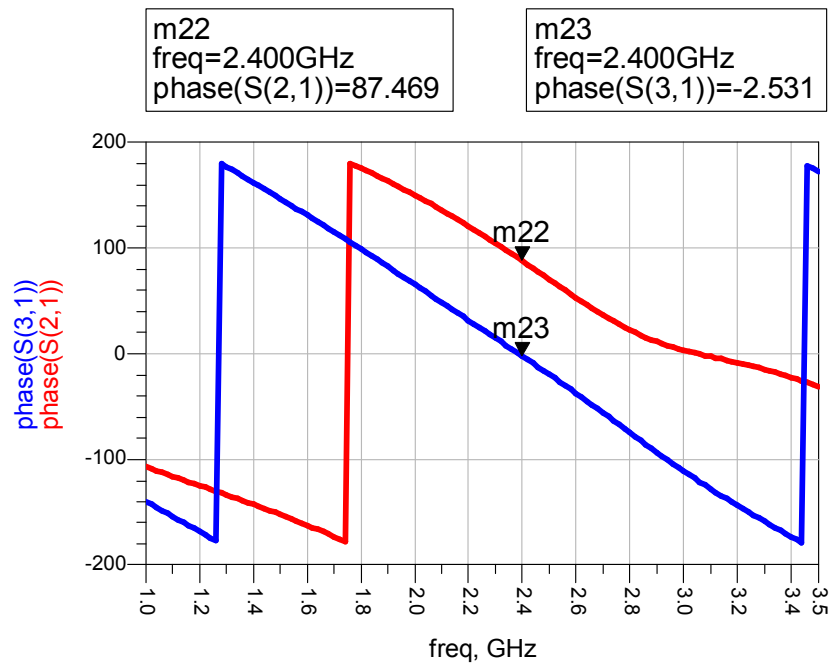


Figure (4.58)) Simulation of phase difference of 90° hybrid with two single stub at the output ports.

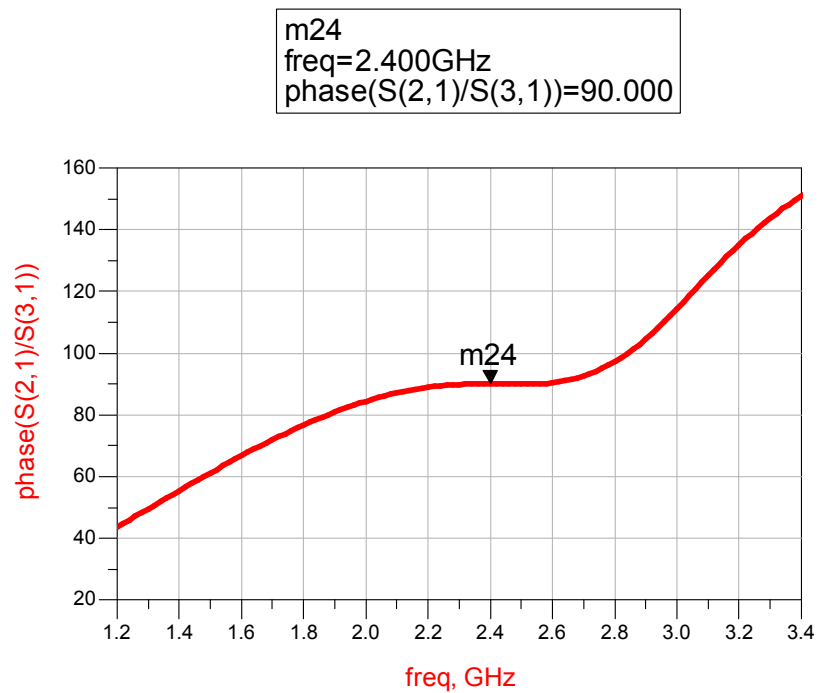


Figure (4.59) Simulation phase of 90° hybrid with two single stub at the output ports.

4.7.1 Layout of 90° Hybrid With Two Single Stubs at the Output ports

By running generate layout in ADS software the schematic design of figure (4.51) is transformed to the layout design with all length and width dimensions and substrate specifications as shown in figure (4.60).

Tuning process is used at the layout design in order to obtain the target specifications of the design.

Figure(4.60), shows the layout design with all dimensions after using the tuning process which satisfies our specification after momentum simulation is run .



Figure (4.60) Layout of 90°hybrid with two single stub at the output ports with dimensions.

Figure(4.61), shows return loss at marker (m25) of $S_{11} = -22.586$ dB where $S_{11} = S_{44}$ and figure (4.62), shows return loss at marker (m26) of $S_{22} = -37.570$ dB where $S_{22} = S_{33}$. Although the difference in resonance frequency is observed, it is obvious all ports matching and the device behavior absorbed maximum power.

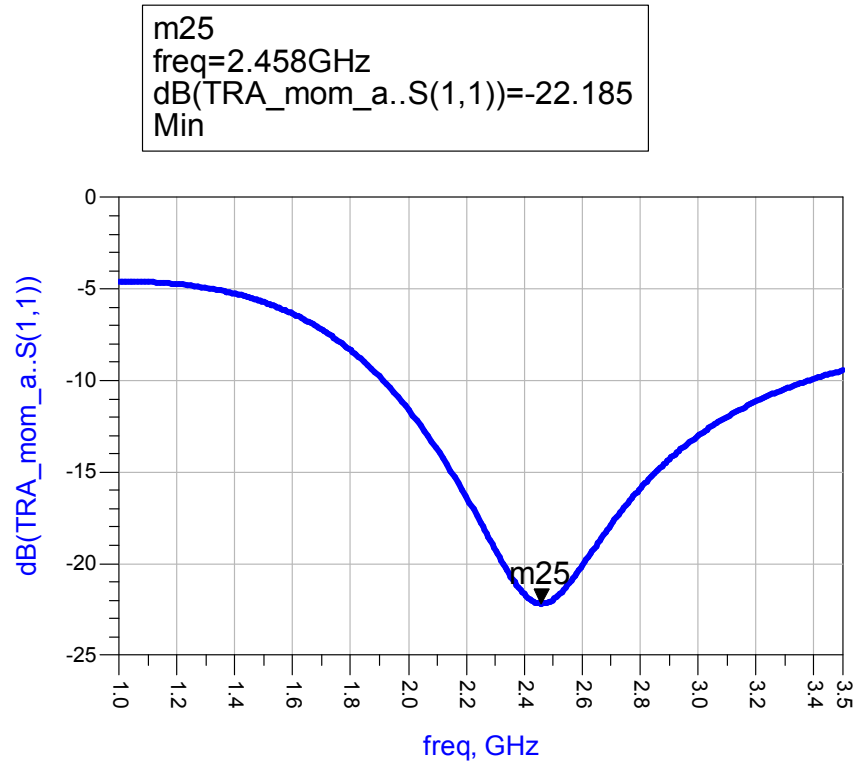


Figure (4.61)Return losses of 90°hybrid with two single stub at the output ports.

The device demonstrated high degree of isolation between the two output ports as shown in figure (4.63), isolated output pointed at marker (m27) of $S_{41} = -38.815$ dB. This behavior ensures the device works as hybrid without changing the layout design .

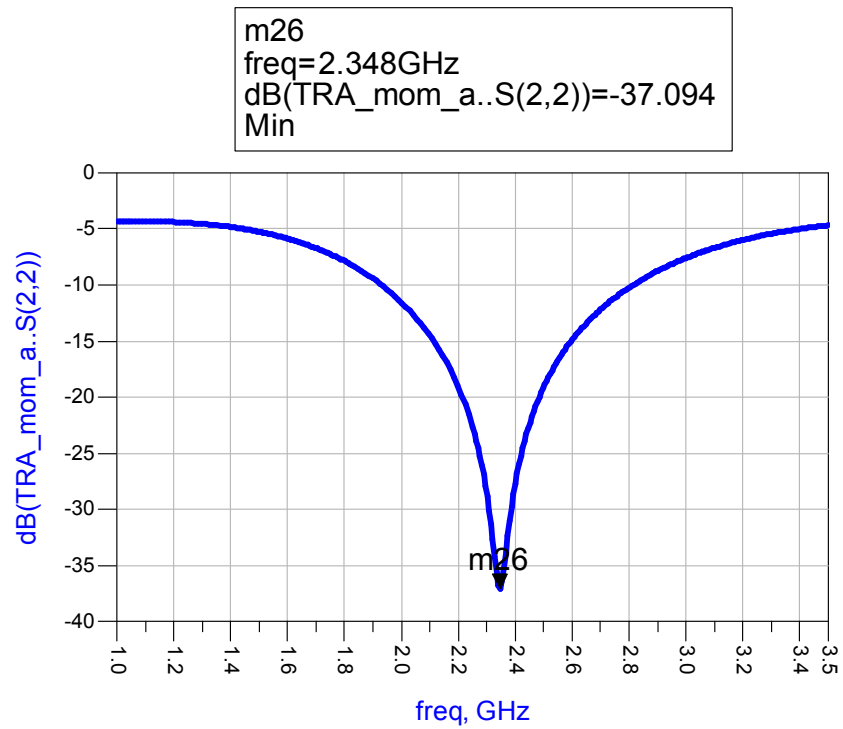


Figure (4.62) Return losses of 90° hybrid with two single stub at the output ports.

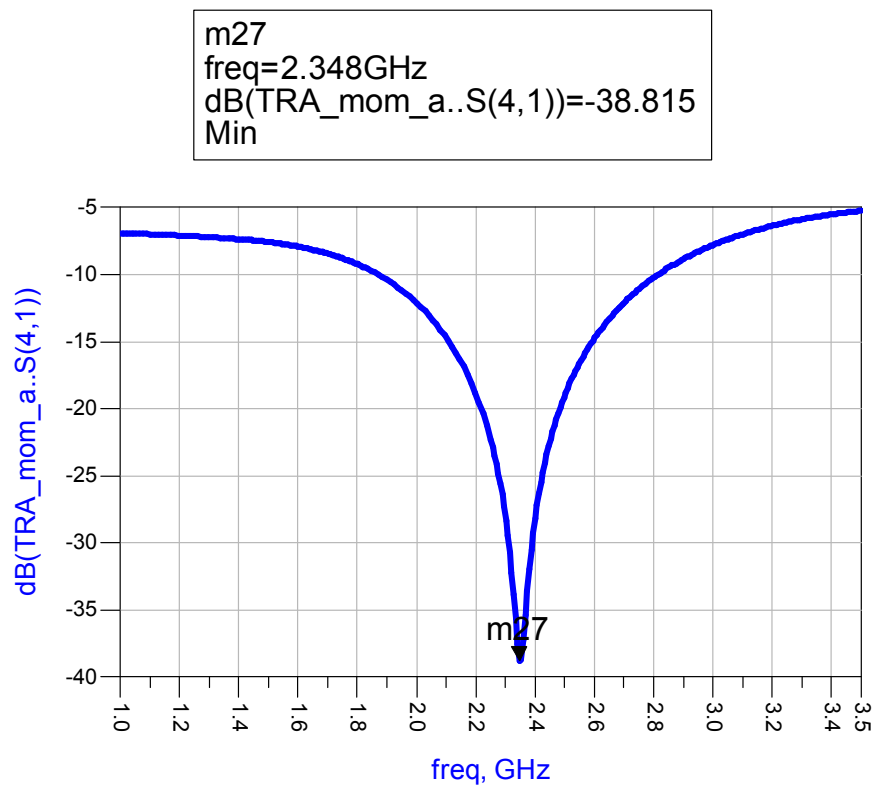


Figure (4.63) Isolation of 90° hybrid with two single stub at the output ports .

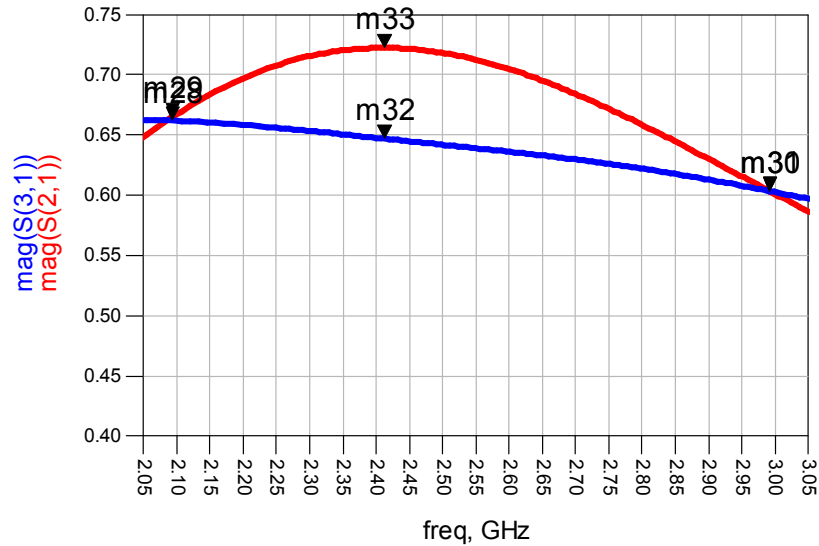
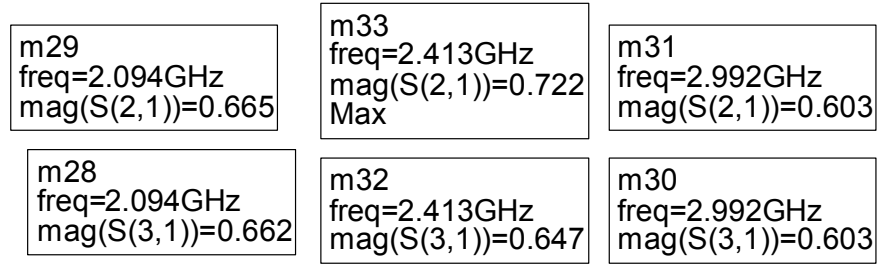


Figure (4.64) Transmission coefficient of 90°hybrid with two single stub at the output ports.

Layout transmission coefficients of 90°Hybrid single stub outputs, exhibit a highest band of frequency starting 2.094 GHz until 2.992 GHz as shown as Figure (4.64) and figure (4.65). Then the calculated bandwidth (BW)=888MHZ.

Figure (4.64) shows the transmission coefficient from port 1 to ports 2 and 3 (in dB) across the same frequency band as magnitude of coupling voltage at $\pm 0.707 < 1$ as close as possible as ideal case , both curves indicated by markers (m32=0.647V) and (m33=0.722V) respectively.

As has been illustrated in Figure (4.65) the value coupling balance as maximally flat pointed by marker (m38) of $S_{31} = -3.783$ dB , its less than 1 dB as close as ideally case , then amplitude differences pointed by markers (m36–m37)=(S_{13} - S_{12})= 0.002 dB, its close 0 dB as in ideally case .

One of the most important challenges in the coupler design is equal power dividing and accurate phase difference .

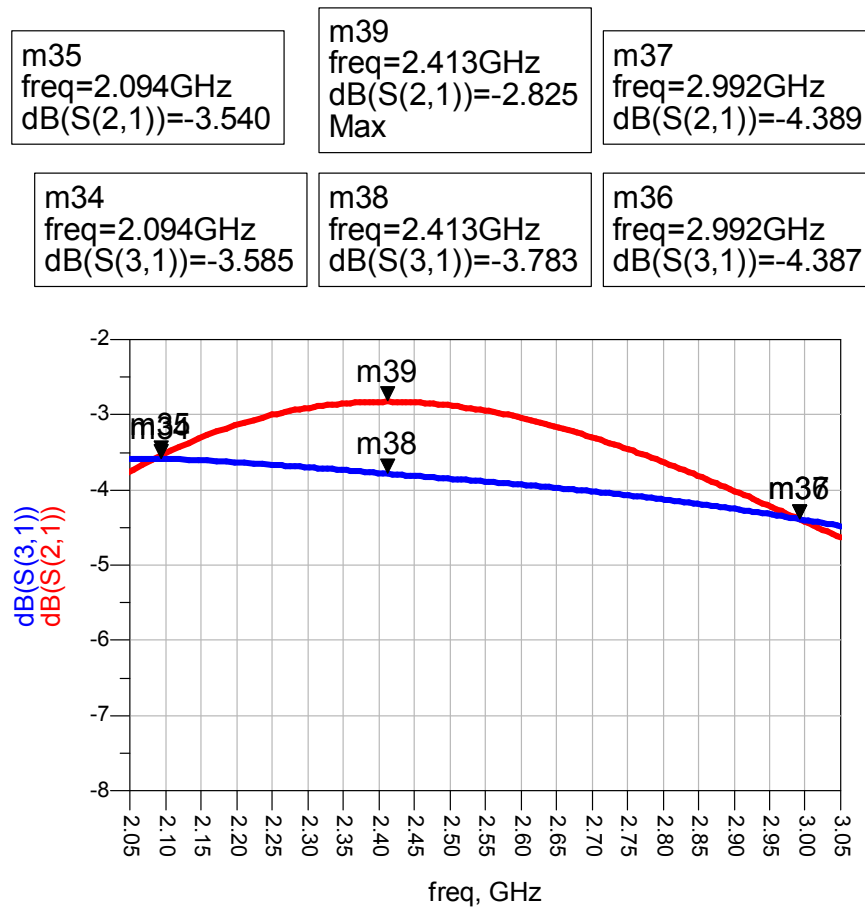


Figure (4.65) Coupling balance of 90°hybrid with two single stub at the output ports.

Figure (4.66) shows difference in phase indicated by marker (m40 and m41) :-

$$\text{Difference phase} = (m40 - m41) = 90^\circ.001$$

Whereas figure (4.67), shows at marker (m42) the layout transformation phase (m42=90°.001) of 90°Hybrid with single stub outputs.

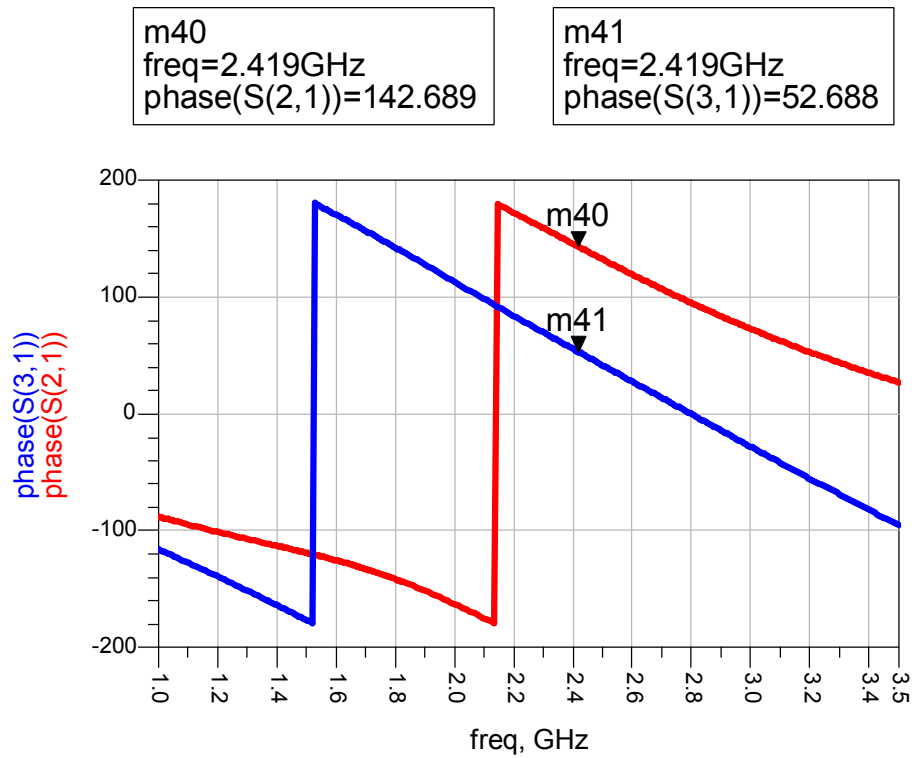


Figure (4.66) Phase difference of 90° hybrid with two single stub at the output ports.

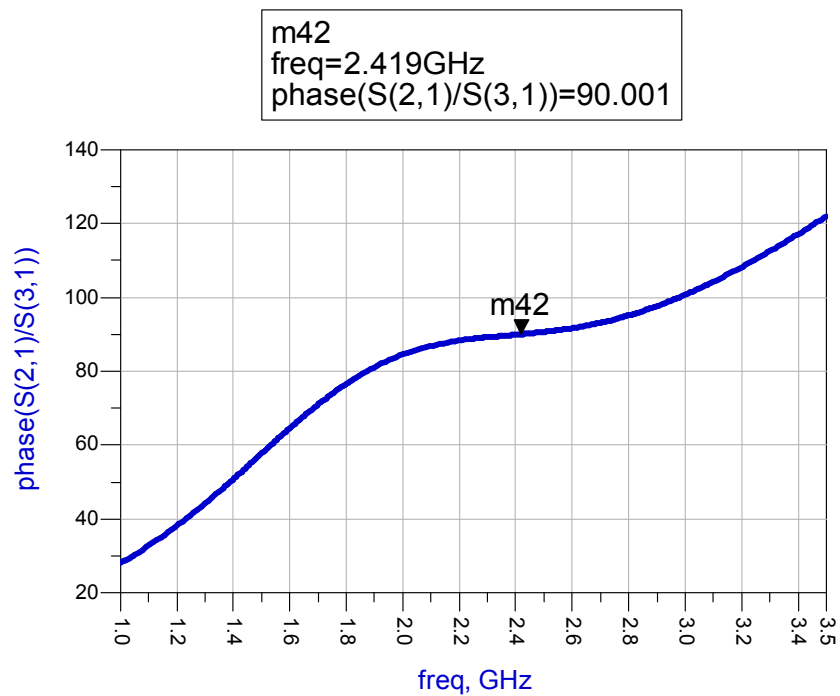


Figure (4.67) Phase of 90° hybrid with two single stub at the output ports.

4.6.2 3D Layout of 90° Hybrid With Two Stubs at the Output Ports

Figure(4.68) shows the layout box of 90° hybrid with two stubs by X ,Y , Z 3D and the best height ($h_u=30\text{mm}$). This will help to achieve optimum result for design.

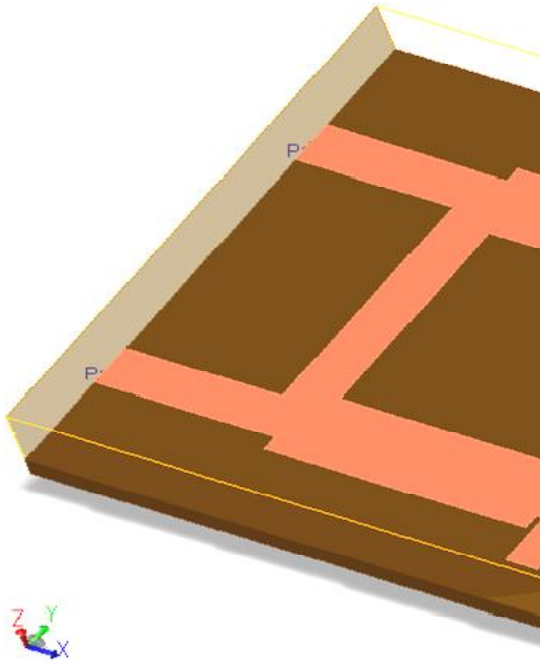


Figure (4.68) 3-Dimensional layout of 90° hybrid with two stub at the output ports.

4.7 Case5. Simulation of 90° Hybrid With Four Stub at the Ports

By observing in case4 (90° hybrid with two single stub at the output Ports) results of the bandwidth and phase for two stubs ports only ,we will use the same design of single stub for all ports to obtain enhancement results of bandwidth and phase too.

Methods have been studied to modify the conventional coupler structure to become a wide-band coupler (maintained before in chapter 3).

Putting a matched network, i.e., a series transmission line and a parallel open stub, in the junction of the coupler can improve the bandwidth of a branch-line coupler.

Figure (4.69) shows schematic design of 90° hybrid four stub ports. This design replaces the single stub into all ports.

The entire design was run tested by ADS software behaved as hybrid and obtained all parameters simulation and layout simulation results.

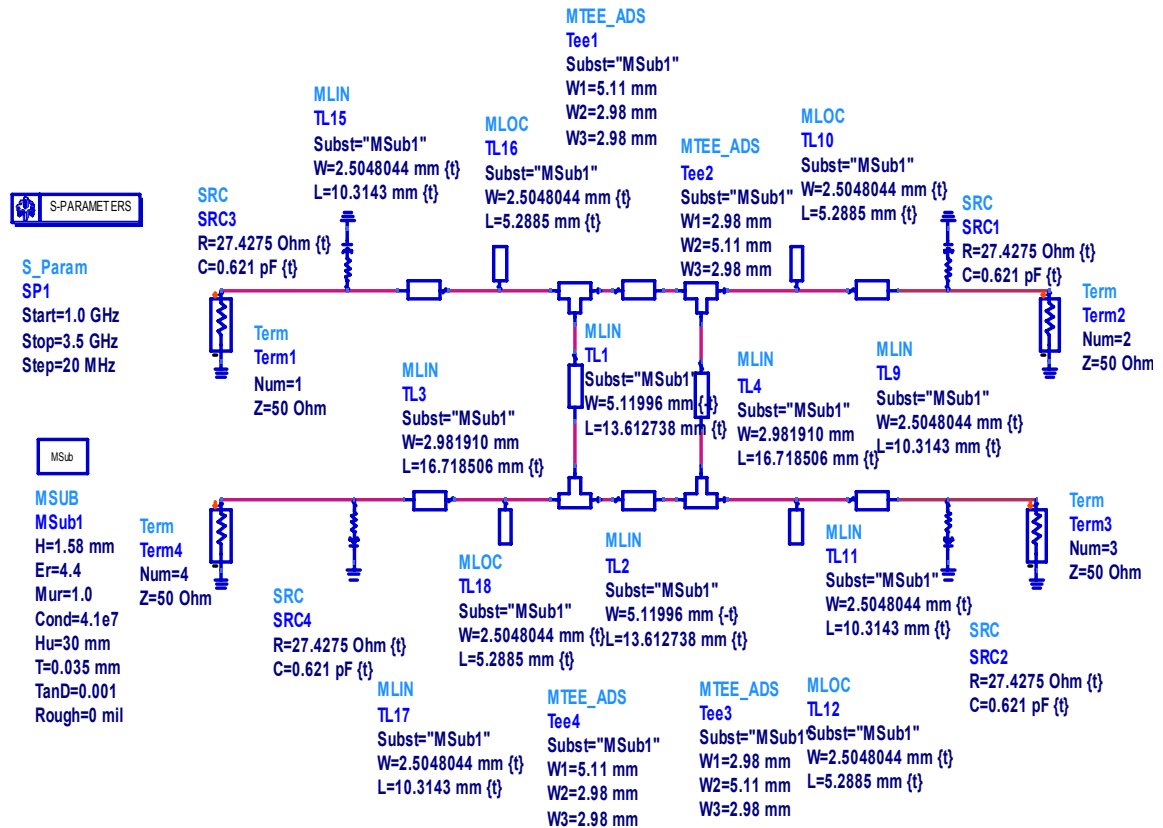


Figure (4.69) Schematic diagram of 90° hybrid with four stub at the ports.

Figure (4.70) shows Simulation of the return losses where the resonant frequency at 2.4GHz a marker (m1) of $S_{11} = -18.186$ dB for 90°hybrid four stub ports also at all ports where ($S_{11}=S_{22}=S_{33}=S_{44}$). Note the return loss has very low value (in dB). This is very desirable, because there will be maximum power delivered to the device at this frequency and all ports are matching and provides wide band improvement.

Figure (4.71) shows simulation a deep null isolation of 90°Hybrid four stub ports where isolated output marker at (m2) of $S_{41} = -32.134$ dB. This value ensures a high degree of isolation between the two output ports and the input port without unwanted interaction between them.

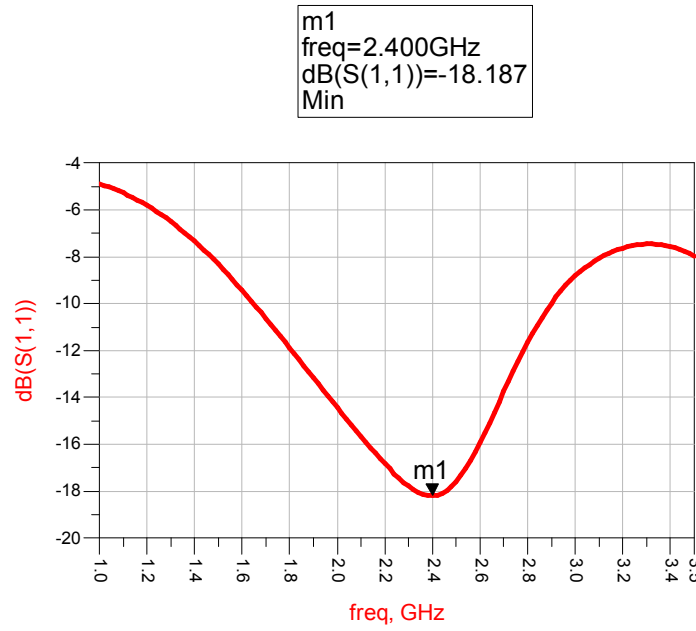


Figure (4.70)Simulation return losses of 90° hybrid with four stub at the ports.

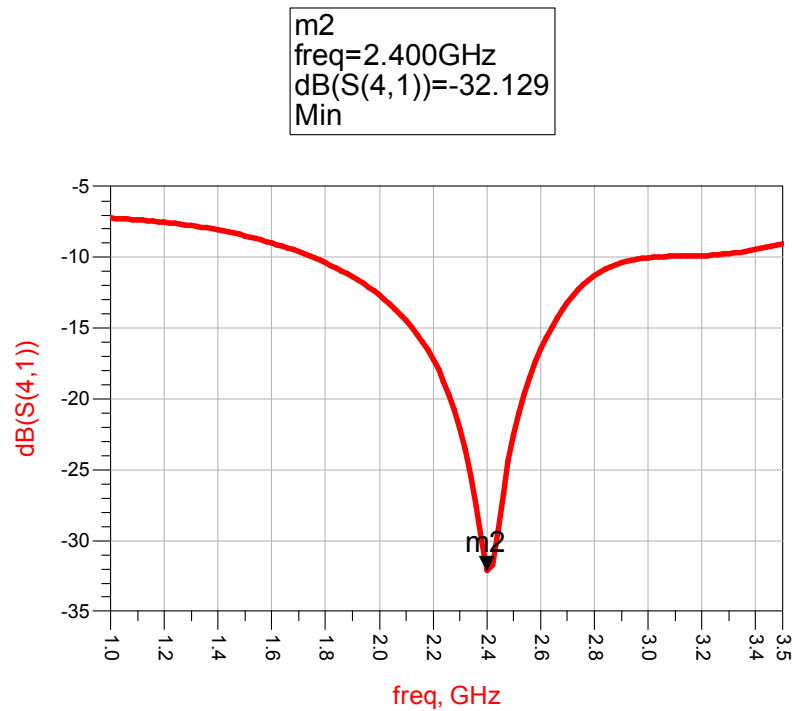


Figure (4.71) Simulation isolation of 90° hybrid with four stub at the ports.

It is obvious the device has matching conditions and reciprocal behavior as shown as figure (4.72). Where any port works as input port as long as the relationship between

ports is remained. This is because the 90° hybrids are electrically and mechanically symmetrical as shown as figure (4.72), then ($S_{ij} = S_{ji}$).

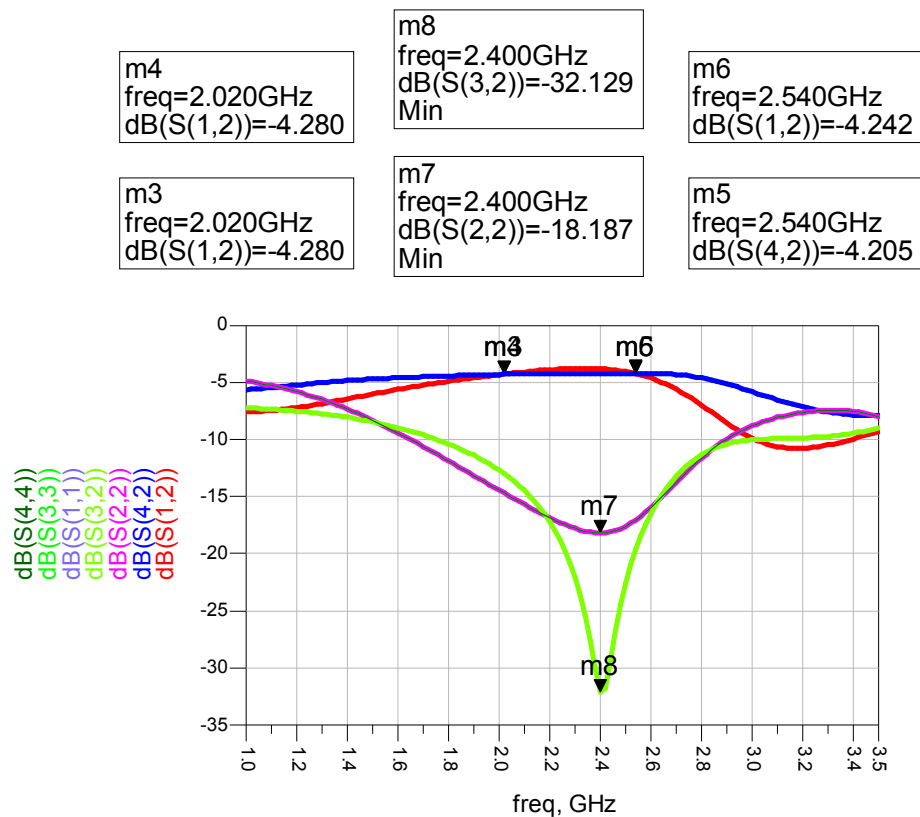


Figure (4.72) Simulation reciprocal of 90° hybrid with four stub at the ports.

Figure (4.73) shows transmission coefficient magnitudes (voltage coupling factor) at the designed frequency 2.4 GHz as indicated by the markers m13 and m14 which are 0.617V and 0.647V respectively. Then values are less than 1 and closed to the typical values.

m10 freq=2.020GHz mag(S(2,1))=0.611	m14 freq=2.340GHz mag(S(2,1))=0.647 Max	m12 freq=2.540GHz mag(S(2,1))=0.614
m9 freq=2.020GHz mag(S(3,1))=0.610	m13 freq=2.340GHz mag(S(3,1))=0.617	m11 freq=2.540GHz mag(S(3,1))=0.616

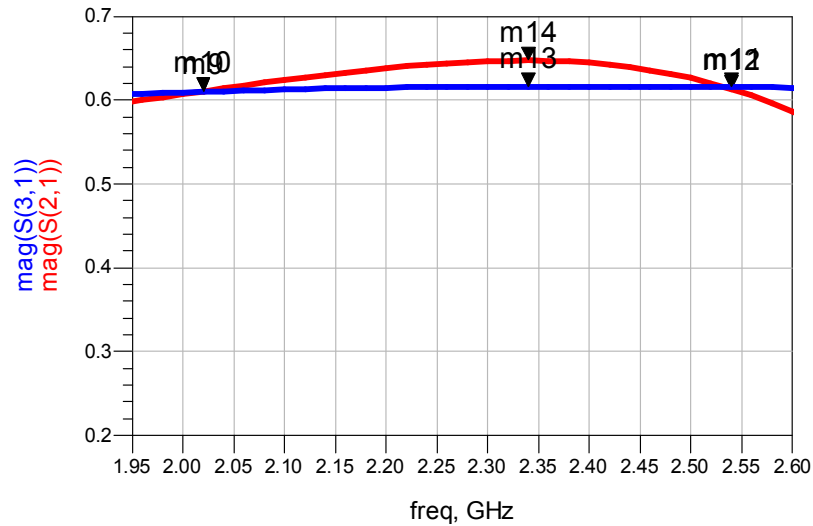


Figure (4.73) Simulation transmission coefficient of 90° hybrid with four stub at the ports.

As has been illustrated in Figure (4.74) the value coupling balance as flat maximally marker m19=S31=-4.201 dB, its less than 1dB as close as ideally case. Amplitude differences (S31-S21)=(m17-m18)=0.037 dB. Its close to 0 dB as in ideally case . will result in two equal amplitude signals that are quadrant (90° apart).

It also makes extended frequency coverage begin from 2.020 GHz until 2.540 GHz with flat response. Then BW=520 MHz.

The coupling of the two arms outputs is shown an equal power split of 3 dB occurs at only the center frequency.

m16 freq=2.020GHz dB(S(2,1))=-4.280	m20 freq=2.340GHz dB(S(2,1))=-3.778 Max	m18 freq=2.540GHz dB(S(2,1))=-4.242
m15 freq=2.020GHz dB(S(3,1))=-4.292	m19 freq=2.340GHz dB(S(3,1))=-4.201	m17 freq=2.540GHz dB(S(3,1))=-4.205

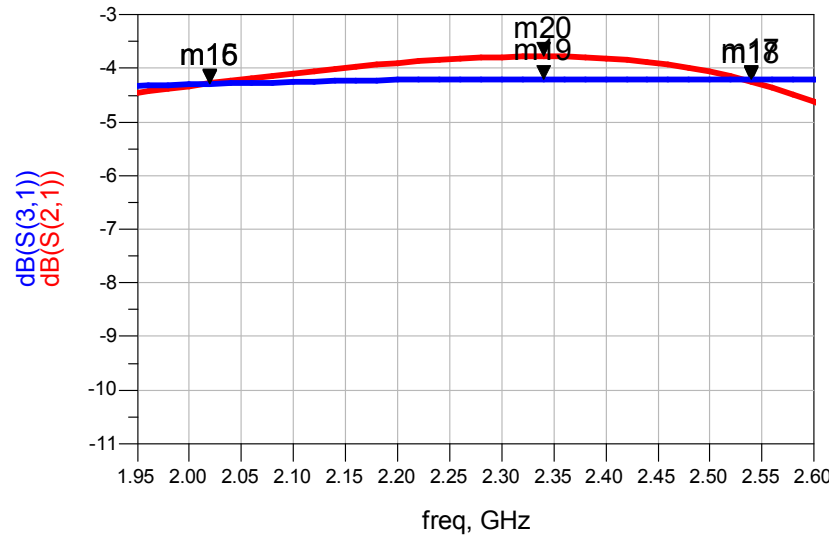


Figure (4.74) Simulation coupling balance of 90° hybrid with four stub at the ports.

Figure (4.75) shows (m21-m22) and figure (4.76) shows m23 as phase balance where

$$\text{Phase difference} = (m21-m22) = 90^{\circ}.006$$

The ideal phase difference between these two ports should be 90 degrees. There is a large bandwidth and coupler has a 90.006 degree phase shift between the two output ports.

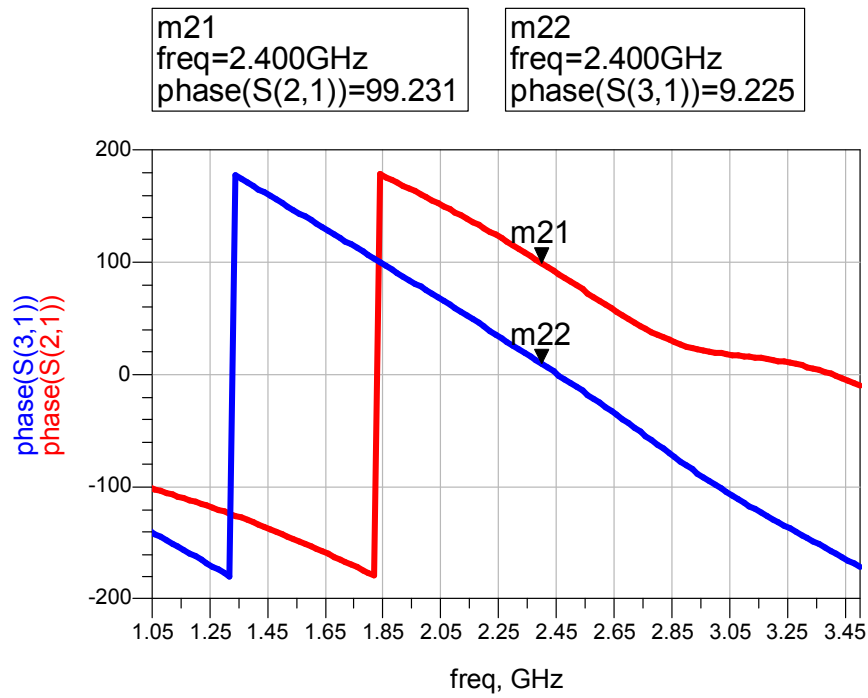


Figure (4.75) Simulation phase difference of 90° hybrid with four stub at the ports.

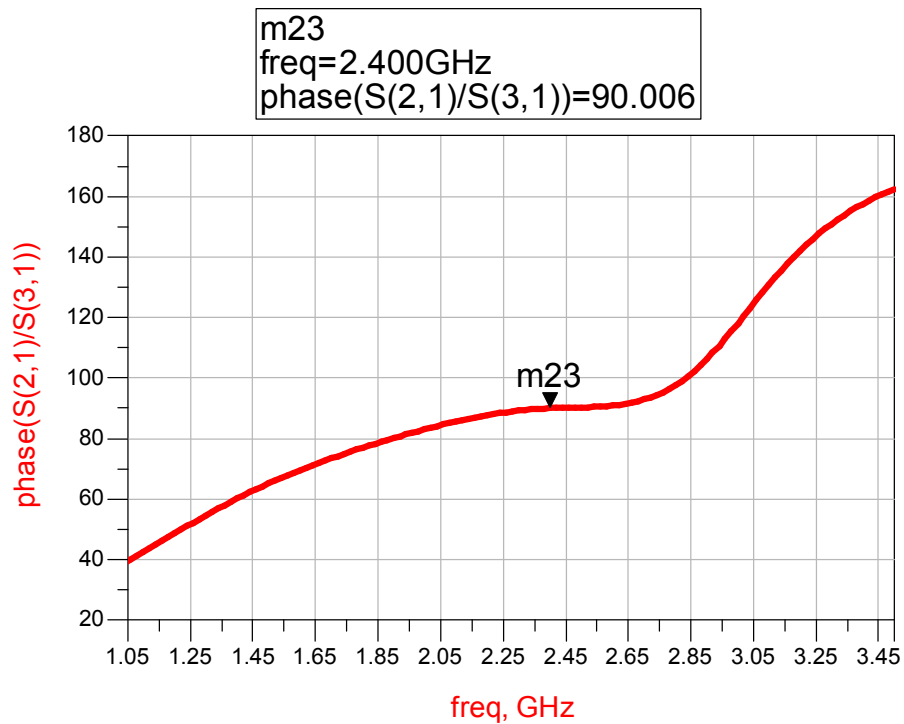


Figure (4.76) Simulation phase of 90° hybrid with four stub at the ports.

4.7.1 The Layout (A) of 90° Hybrid With Four Stub at the Ports

The dimensions of 90° hybrid four stub ports after optimization are shown in figure (4.77).

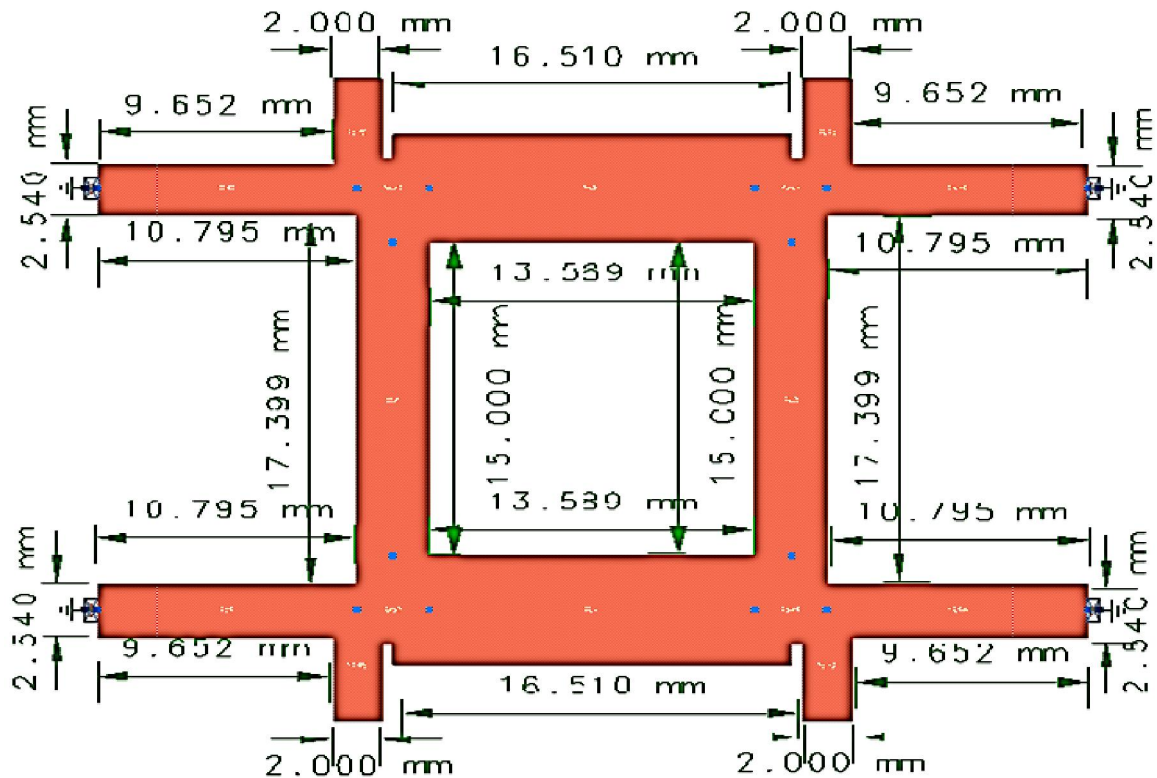


Figure (4.77) Layout (A) of 90° hybrid with four stub at the ports with dimensions.

Simulation results are shown in Figure (4.78) at marker (m24) of $S_{11} = -19.262$ dB of return losses and shows results to previous port assigning, S_{11} , S_{22} , S_{33} , and S_{44} are coupler return losses with the same value because of the symmetry and the reciprocity characteristics. and figure (4.79) shows plot of isolated output at marker (m25) of $S_{41} = -21.495$ dB.

The results display of 90 hybrid provides wide band width, but provides poor isolation which is due to power loss with effect on the phase result.

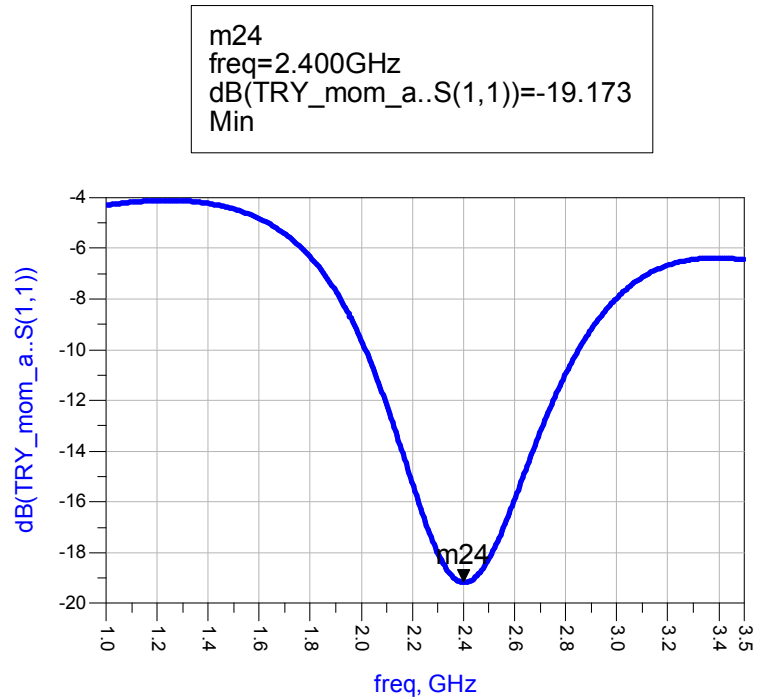


Figure (4.78) Return losses of 90° hybrid with four stub at the ports (A).

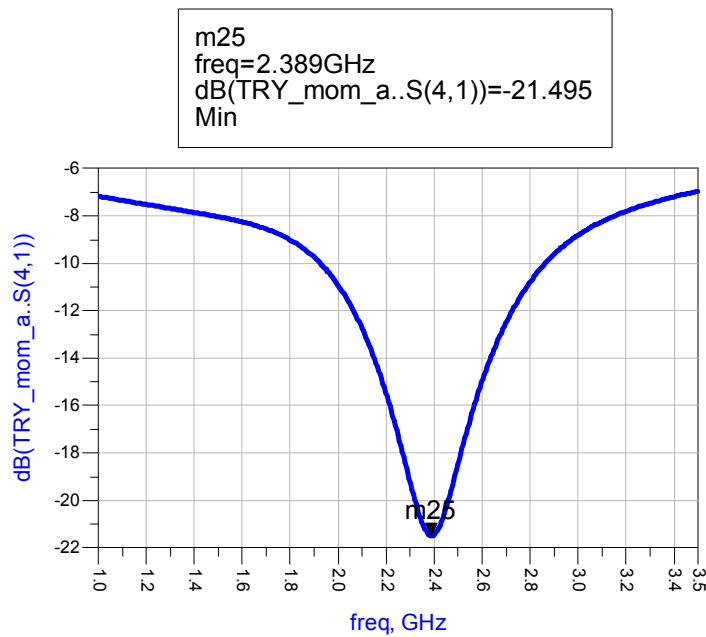


Figure (4.79) Isolation of 90° hybrid with four stub at the ports (A).

Figure (4.80) shows bandwidth frequency starting 2.102 GHZ until 2.823 GHZ and shows the transmission coefficient where marker m30=0.628V and marker m31=0.751V across the same frequency band as magnitude of coupling voltage at $\pm 0.707 < 1$.

m27 freq=2.102GHz mag(S(2,1))=0.655	m31 freq=2.418GHz mag(S(2,1))=0.751 Max	m29 freq=2.823GHz mag(S(2,1))=0.631
m26 freq=2.102GHz mag(S(3,1))=0.661	m30 freq=2.418GHz mag(S(3,1))=0.628	m28 freq=2.823GHz mag(S(3,1))=0.633

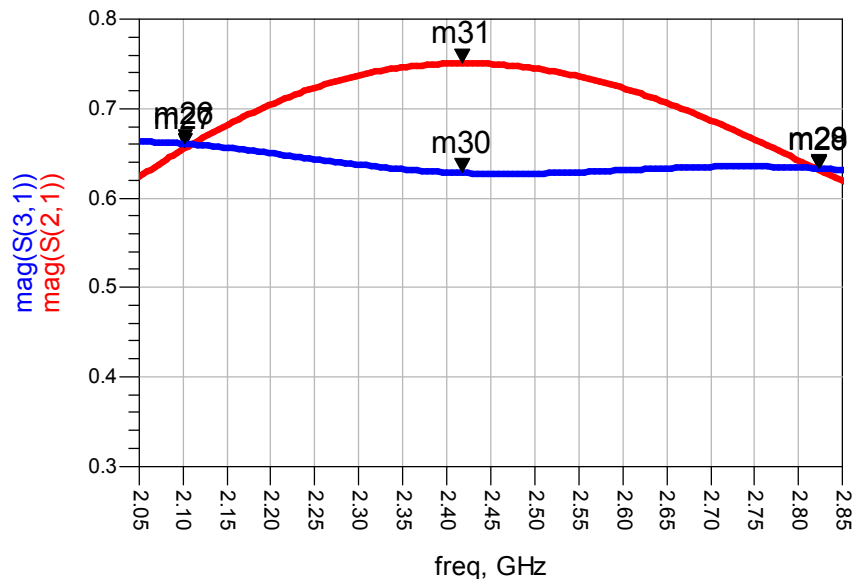


Figure (4.80) Transmission coefficient of 90° hybrid with four stub at the ports (A).

As has been illustrated in Figure(4. 81) the value coupling balance as flat maximally at marker m36 = S31= -4.041 dB , its less than 1 dB .

Amplitude differences (m34 –m35) = (S31-S21) =0.020 dB. its close 0 dB .

Figure (4.80) and Figure (4.81)shows bandwidth frequency starting 2.102 GHz until 2.823 GHz .Then BW = 721 MHz .

m33
freq=2.102GHz
dB(S(2,1))=-3.669

m37
freq=2.418GHz
dB(S(2,1))=-2.491
Max

m35
freq=2.823GHz
dB(S(2,1))=-3.993

m32
freq=2.102GHz
dB(S(3,1))=-3.601

m36
freq=2.418GHz
dB(S(3,1))=-4.041

m34
freq=2.823GHz
dB(S(3,1))=-3.973

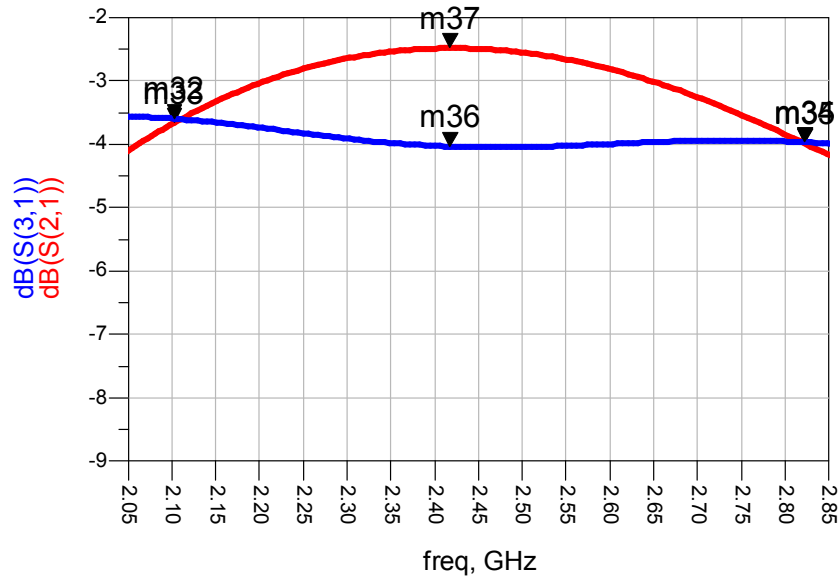


Figure (4.81) Coupling balance of 90° hybrid with four stub at the ports(A).

Figure (4.82) shows both markers (m38, m39) and figure (4.83) shows, the marker (m40) as phase balance where :-

$$\text{Phase difference} = (m38 - m39) = 89.567^\circ$$

The ideal phase difference between these two ports should be 90 degrees.

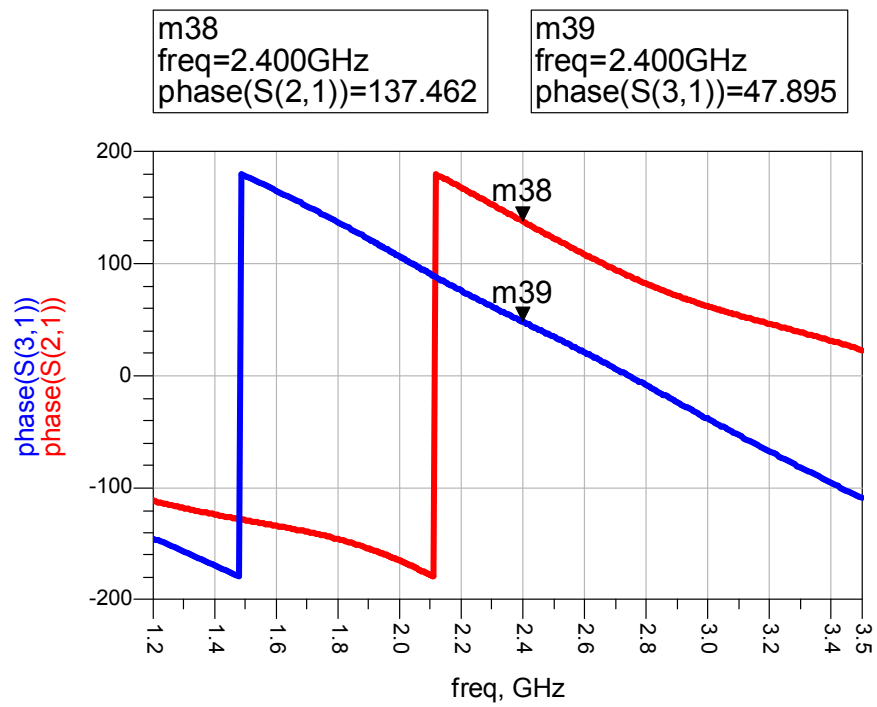


Figure (4.82) Phase difference of 90° hybrid with four stub at the ports (A).

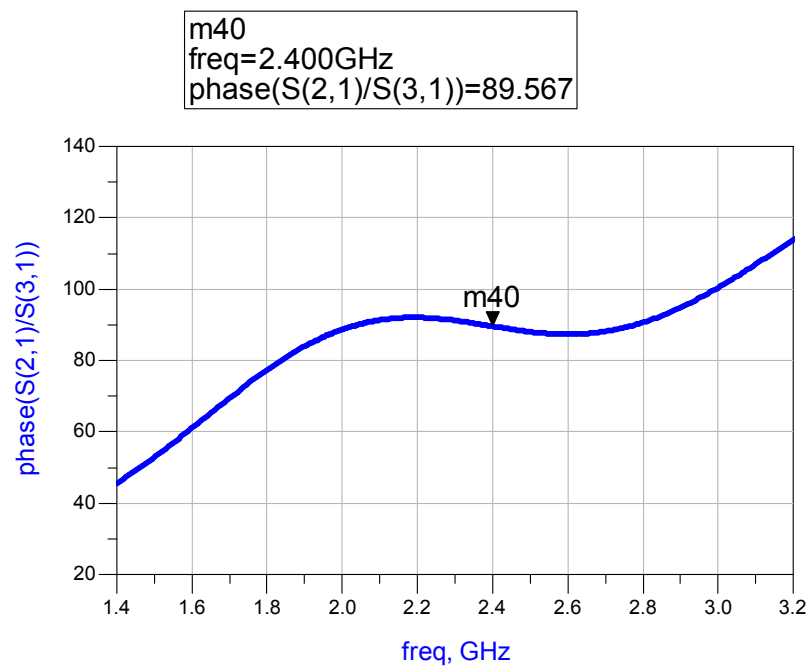


Figure (4.83) Phase of 90° hybrid with four stub at the ports (A) .

4.7.2 3D Layout (A) of 90° Hybrid With Four Stub at the Ports

From figure (4.84) the layout box of 90° hybrid with four stubs ports by X ,Y and Z in 3D and the best Height ($h_u=30\text{mm}$). This will help to achieve optimum result for design.

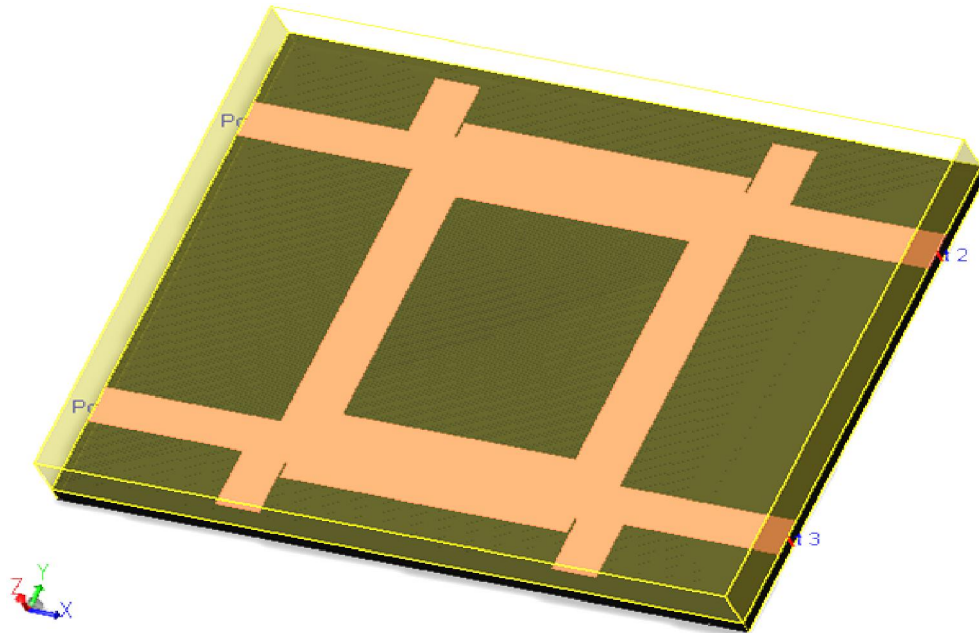


Figure (4.84) 3-Dimensional layout of 90° hybrid with four stub at the ports(A).

4.7.3 Layout (B) of 90° Hybrid With Four Stub at the Ports

The simulation was done in case5 layout (A) is used just here changing the lengths and widths in order to get a good wideband coupler which can be done by varying the widths and lengths of stubs, using tuning process.

In this case reducing size of hybrid was carried out by reducing the lengths of the stubs to the half of the length and simulation proved were performed and the simulation results shows a good results compare to the case (A) .

The entire hybrid circuit layout was then simulated to confirm hybrid behavior, account for the addition of stub ports, and facilitate fine tuning of the circuit layout , it can be seen that the simulation result is closed to the ideal case.

Where the bandwidth, isolation increase and phase is improved .

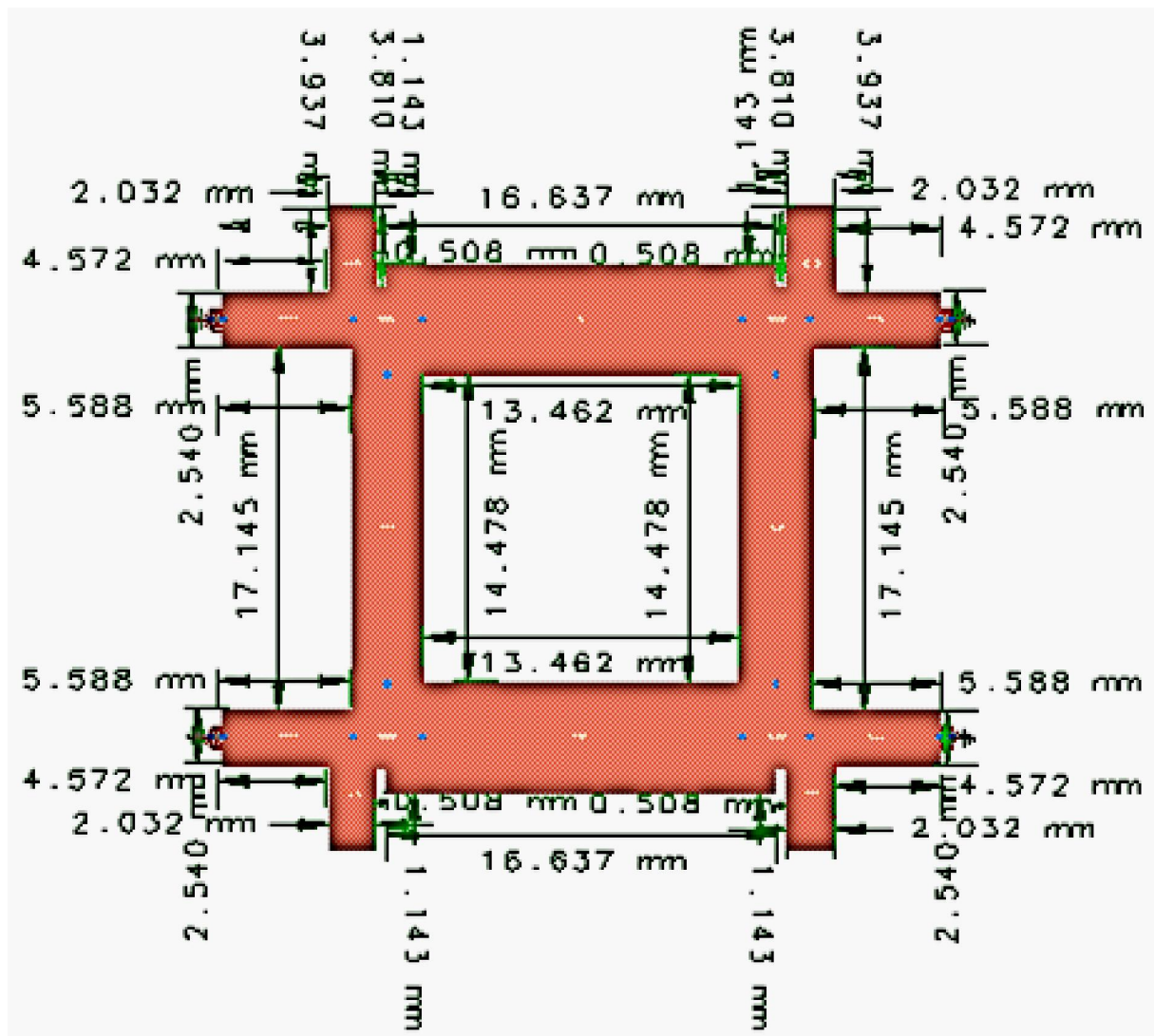


Figure (4.85) Layout(B) of 90° hybrid with four stub at the ports with dimensions.

Simulation results are shown in figure(4.86) and figure(4.87), the plots of return losses and isolation results, the marker (m41) of $S_{11} = -25.5$ dB as all as ports , and isolated output at marker (m42) of $S_{41} = -28.495$ dB. The results display of 90 hybrid provides wide width band, and improve of isolation, which due to maximum power delivered were effect on the phase result.

Also by observing Stubs create reflection zeros at equal distance on both sides of the original perfectly matched points .

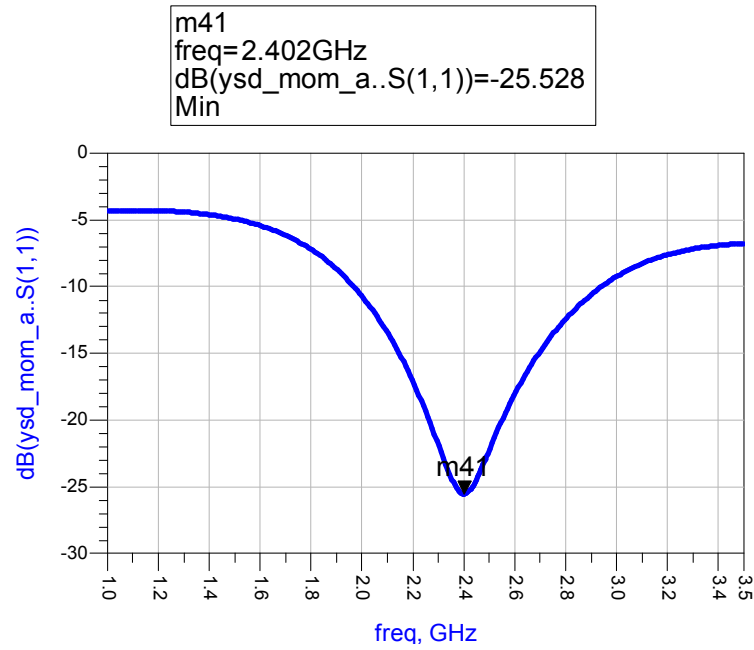


Figure (4.86) Return losses of 90° hybrid with four stub at the ports (B)

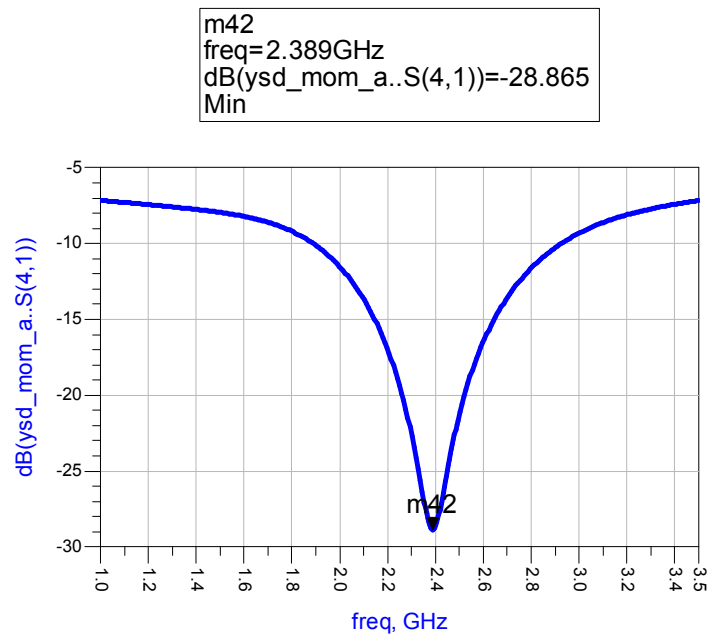


Figure (4.87) Isolation of 90° hybrid with four stub at the ports (B).

Figure (4.88) shows transmission coefficient magnitudes (voltage coupling factor) at the designed frequency 2.4 GHz as indicated by the markers m48 and m47 which are 0.756V and 0.635V respectively ; then values magnitude of coupling voltage at $\pm 0.707 < 1$ as close as in ideally case .

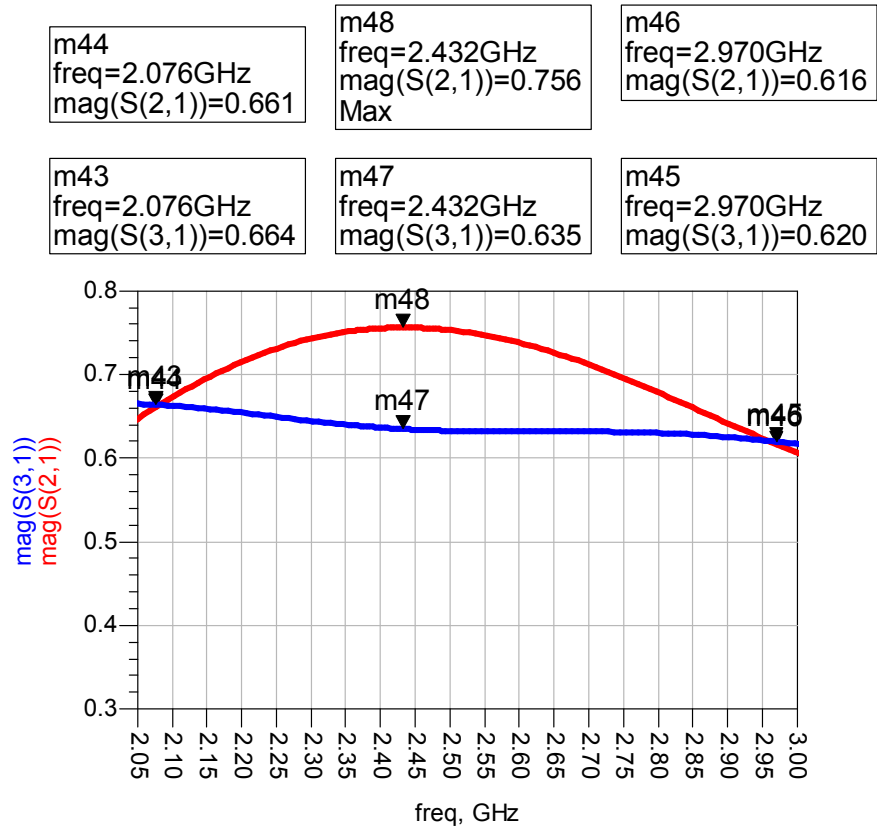


Figure (4.88) Transmission coefficient of 90° hybrid with four stub at the ports (B) .

As has been illustrated in Figure(4.89) the value coupling balance as flat maximally, the marker m53=S31=-3.946 dB, its less than 1dB as close as ideally case. Amplitude differences(S31-S21)=(m49-m50)=0.042 dB. its close 0 dB as in ideally case. will result in two equal amplitude signals that are quadrant (90° apart). It also makes extend frequency coverage 2.076 GHz until 2.970 GHz with flat response.

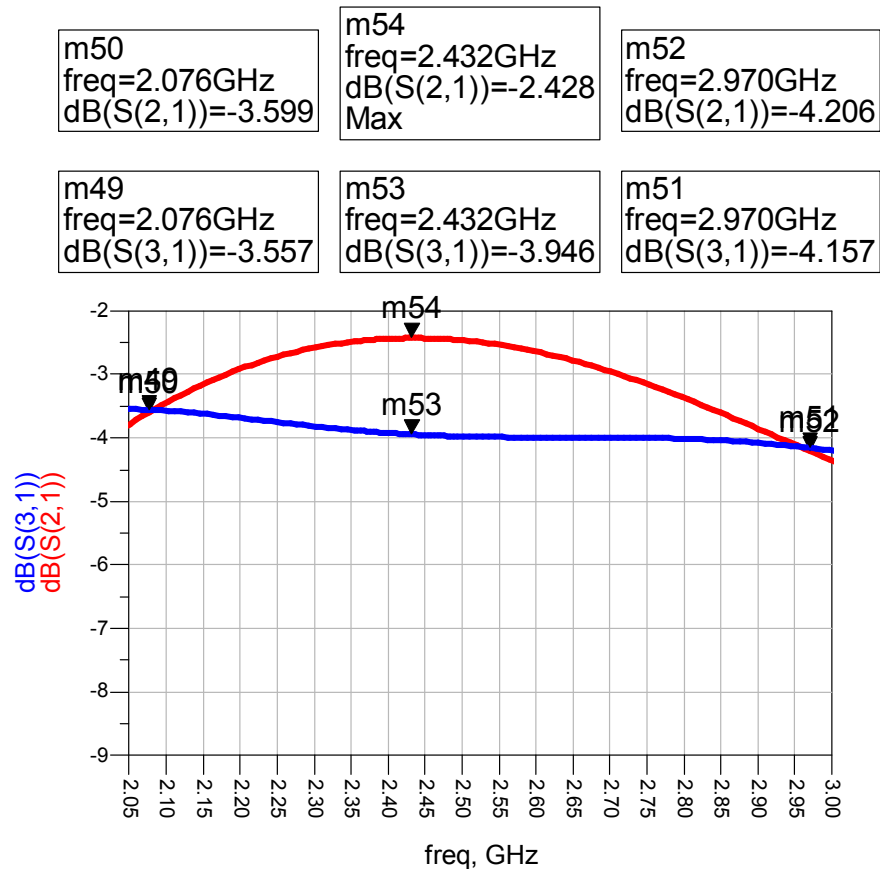


Figure (4.89) Coupling balance 90° hybrid with four stub at the ports (B).

Figure (4.88) and Figure (4.89) show bandwidth frequency starting 2.076 GHz until 2.970 GHz. Then BW = 894MHz.

Figure (4.90) shows (m55–m56) and figure(4.91) shows m57 as phase balance where :-

$$\text{Difference phase} = (m55 - m56) = -270.592^\circ.$$

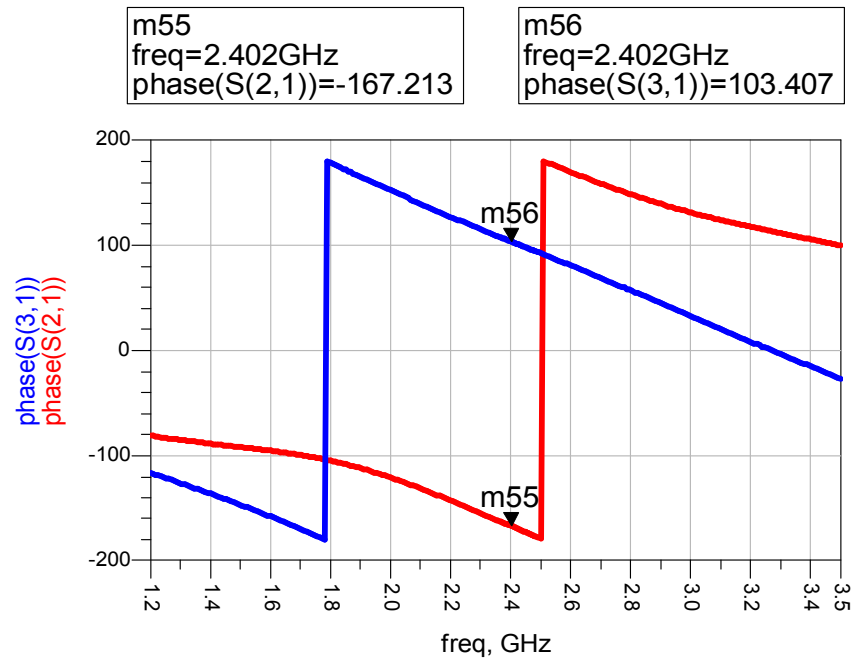


Figure (4.90) Phase difference of 90° hybrid with four stub at the ports (B) .

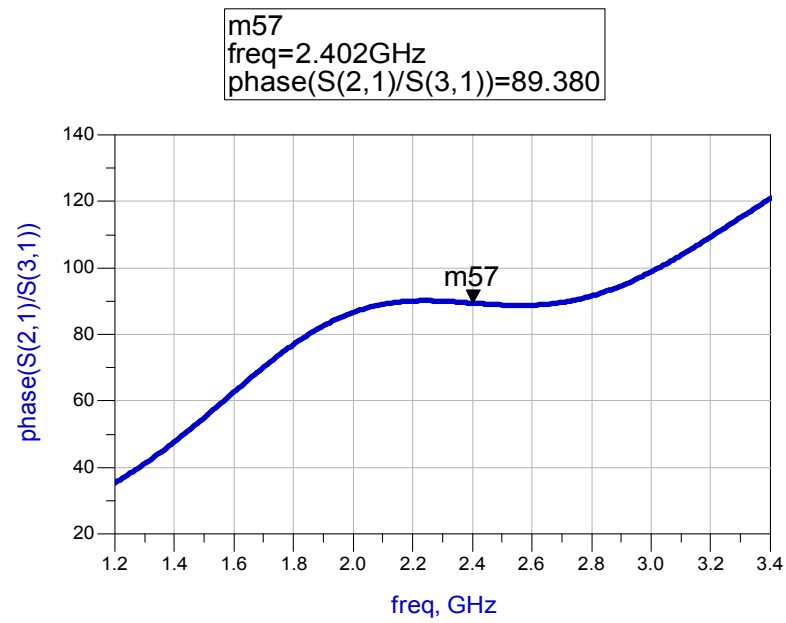


Figure (4.91) Phase of 90° hybrid with four stub at the ports (B).

4.7.4 3D Layout (B) Of 90° Hybrid With Four Stub at the Ports

figure(4.92) shows the layout box of 90° hybrid with four stubs ports by X ,Y , Z in 3D and the best height ($h_u=30\text{mm}$). This will help to achieve optimum result for design.

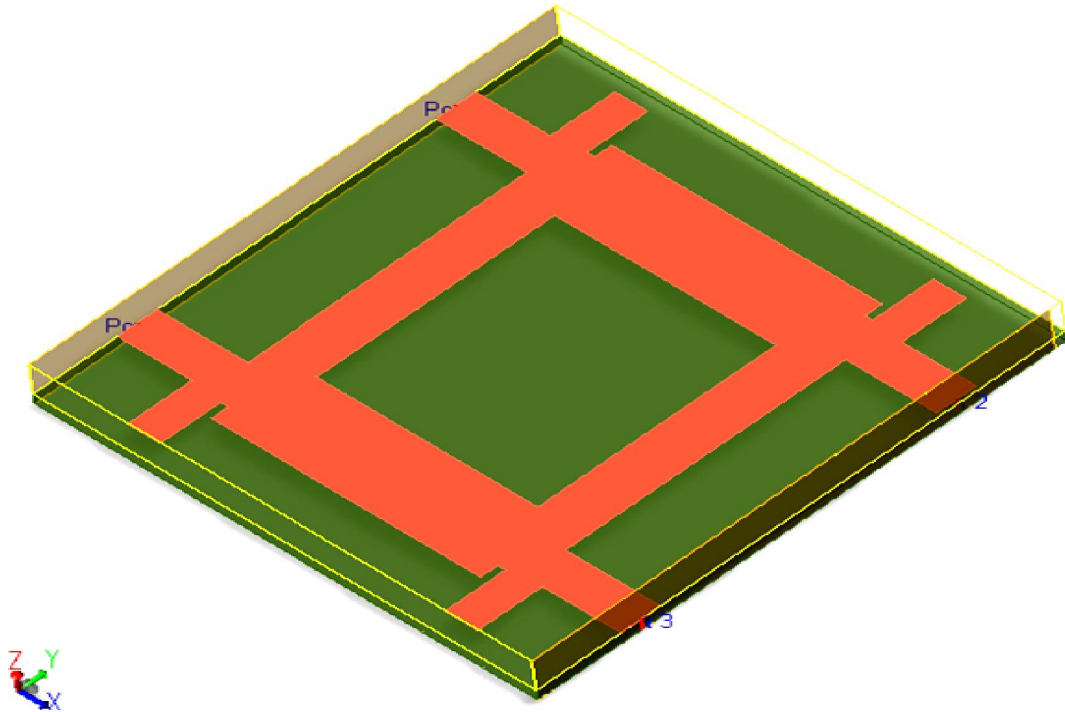


Figure (4.92) 3-Dimensional layout of 90° hybrid with four stub at the ports (B).

4.8 Fabrication

when the simulation result satisfies the objective requirements , fabrication for the design can be performed. Normally, for design implementation microstrip use Flame Retardant 4 (FR4) substrate with dielectric around 3 to 5. A SMA connector is used as the feeding structure four ports coupler via a low profile FR-4 substrate (Dielectric constant being 4.4 is used in our design which available in the Organization of Applied Research and Development in Tripoli.

The photograph of the fabricated printed circuit board of the actual 90° hybrid by four ports attached to 50Ω SMA connectors is shown in figure (4.93).

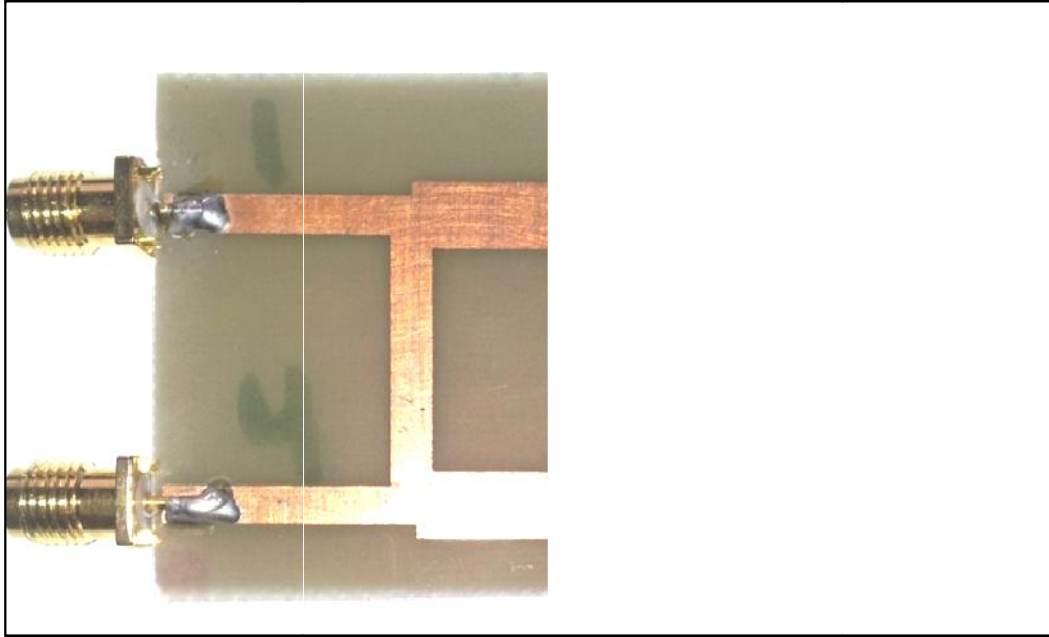


Figure (4.93) Photograph of the fabricated 90° hybrid circuit.

In figures (4.94) and (4.95) the photographs of the fabricated printed circuit board of the actual 90° hybrid cascaded and 90 °hybrid extended TL output ports respectively . both desings with four ports attached to 50Ω SMA connectors.

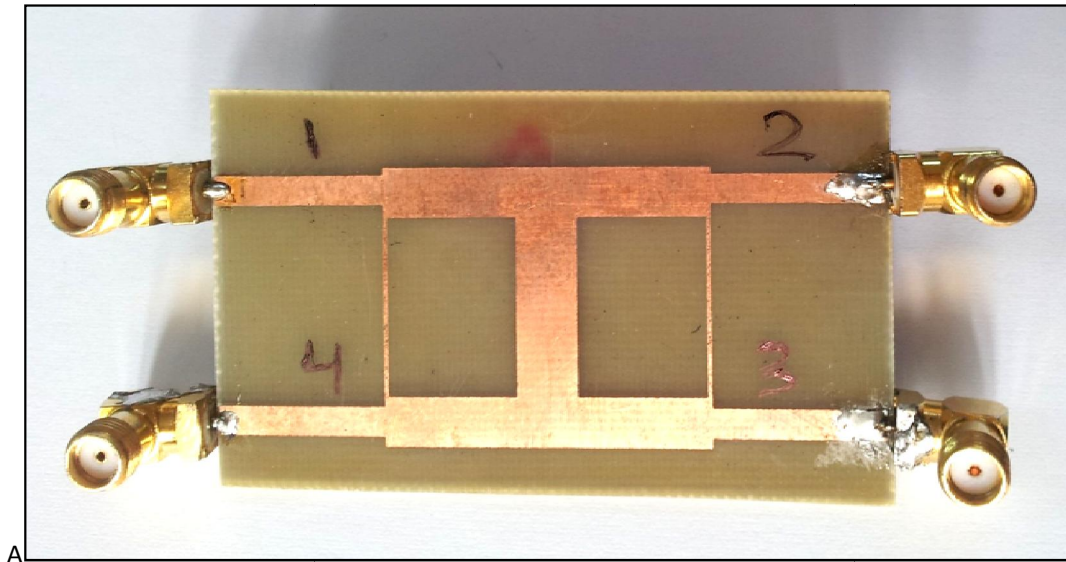


Figure (4.94) Photograph of the fabricated cascaded 90° hybrid circuit.



Figure (4.95) Photograph of the fabricated 90° hybrid with extended TL output ports.

All Those designs fabricated was done in "AL-W'ER" for electronic services (Local shop in Tripoli) for manufacture integrated circuit using planar technology.

4.9 Measurement of Hybrid characteristic parameters

In this part the measurement of the characteristics (amplitude and phase response , return losses, coupling and isolation) of the implemented hybrids on microstrip will be measured. The measurement will be performed using a Hewlett -Packard 8510C vector network analyzer (VNA).

The network analyzer was calibrated across the entire relevant frequency range and the scan resolution (number of sweep points) was set to 112. Also , full two-ports calibration is required to compensate for (calibrate out) the losses and phase delays introduced by cables and connectors/adapters used in the setup.

The HP8510C vector network analyzer (VNA) used in our a versatile , two-port radio frequency (RF) measurements device which allows the user to measure the magnitude and phase of the signal at the input and/or output port of any device under test (DUT).

Based on these measurement a variety of results can be displayed, via the scattering parameters.

Based on parameters, the HP8510C (VNA) provides the user with a host of options to investigate, such as amplitude response, return losses, the input/output impedance and voltage standing wave ratio (VSWR) over the desired frequency range.

The following set of equipments are needed for scattering parameters measurement:-

- 1- HP8510C vector network analyzer (VNA).
- 2- Two 50 Ω – coaxial cables.
- 3- Device under test (DUT our case is 90° hybrid).
- 4- SMA adapter (male to male).
- 5- SMA connectors .
- 6- Matched loads (Open, Short Loads and broadband Loads). .

4.9.1 90° Hybrid Measurements

The experimental set-up for measurements is shown in figure (4.96), its photograph display the HP8510C vector network analyzer (VNA) and the 90° hybrid as Device under test (DUT). All measurements will be carried out will be the same setup only the changing the ports according to the measurement case.



Figure (4.96) The experimental set-up of 90° hybrid measurements.

4.9.1.1 Return Losses Measurements of 90° Hybrid

The experimental set-up in figure (4.96) is used to measure the transmission response and return losses, phases and input impedance.

Measurement result of the return loss is shown in figure (4.97.a), the marker 1 shows -30.466 dB as return loss at the center frequency of 2.4 GHz which is within the operating range.

In the simulation case 1 (figure(4.10)) as it was presented earlier on this chapter, it was evaluated to $S_{11} = -29.26$ dB whereas in practical -30.466 dB which are closed to each other.

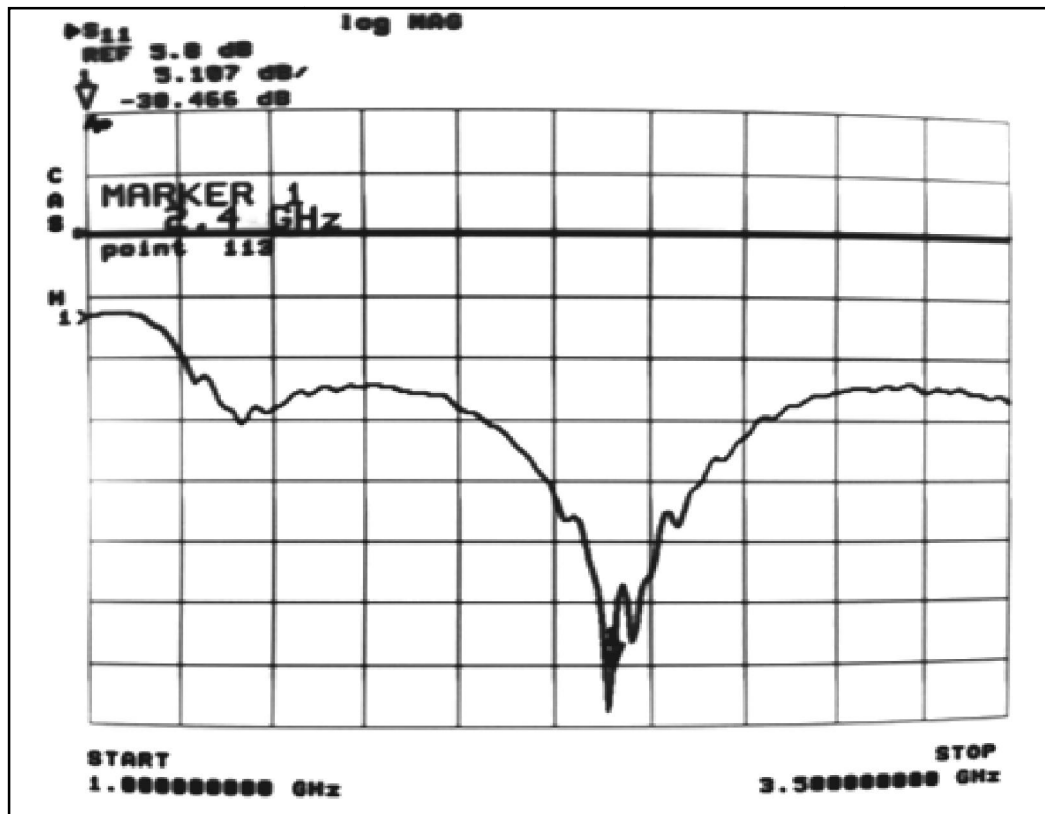
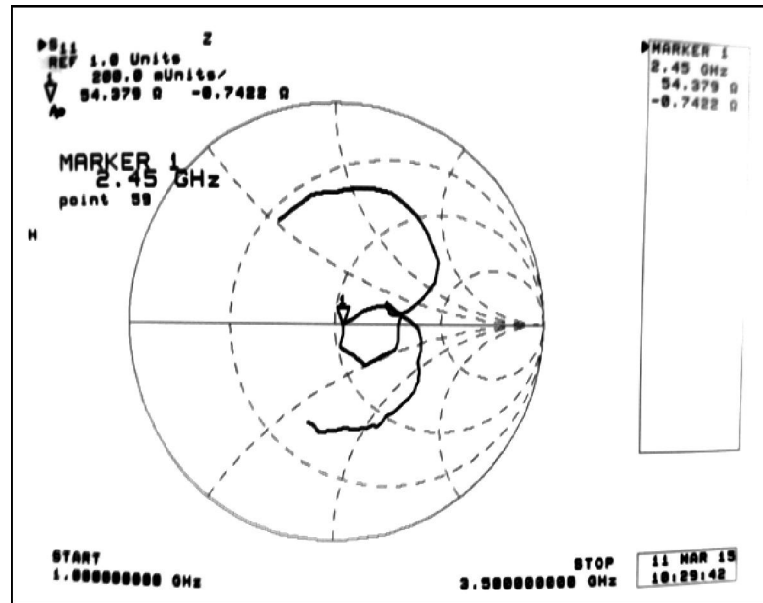


Figure (4.97.a) Measured return losses (S11) of 90° hybrid

4.9.1.2 Measurements of Input Impedance of 90° Hybrid

Measurement result of the input impedance is shown in figure (4.97.b), where marker 1 record the input impedance at the center frequency of 2.4GHz is $(54.379 - j 0.7422)\Omega$.



Figure(4.97.b) Measured of the input impedance of 90° hybrid.

4.9.1.3 Isolation Measurements of 90° Hybrid

The measured of S41 isolation of 90° hybrid is shown in figure(4.98), where marker 1 pointed to the isolation of -26.329 dB at the center frequency of 2.4GHz, whereas in the simulation case 1 (figure(4.11)), it was presented earlier, it was evaluated as S41 = -32.420dB.

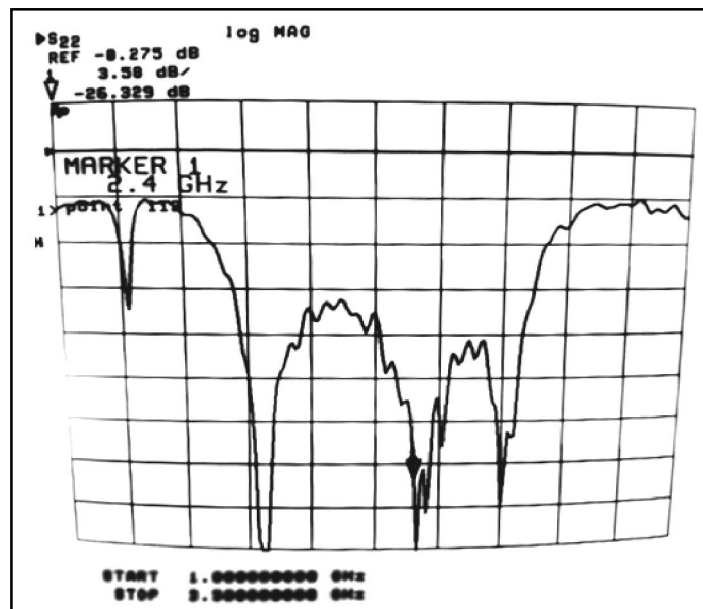


Figure (4.98) Measured of isolation (S41) for 90° hybrid coupler.

4.9.1.4 Phase of 90° Hybrid

Figure (4.99) shows the phases of S21 and S31 outputs of 90° hybrid. From the measurement results of these phases .Marker (1) pointed to the phase of S21 which is($\angle = -90.12^\circ$) whereas the marker (1) in the right side of the figure pointed to the phase of S31($\angle = -179.36^\circ$). Then phase difference (S21-S31) = 89.24° between the port 2 and 3. Comparing these results with simulated phase of 90 hybrid coupler it was (S21-S31) = 89.65° as shown as in figure (4.14) and figure (4.15).

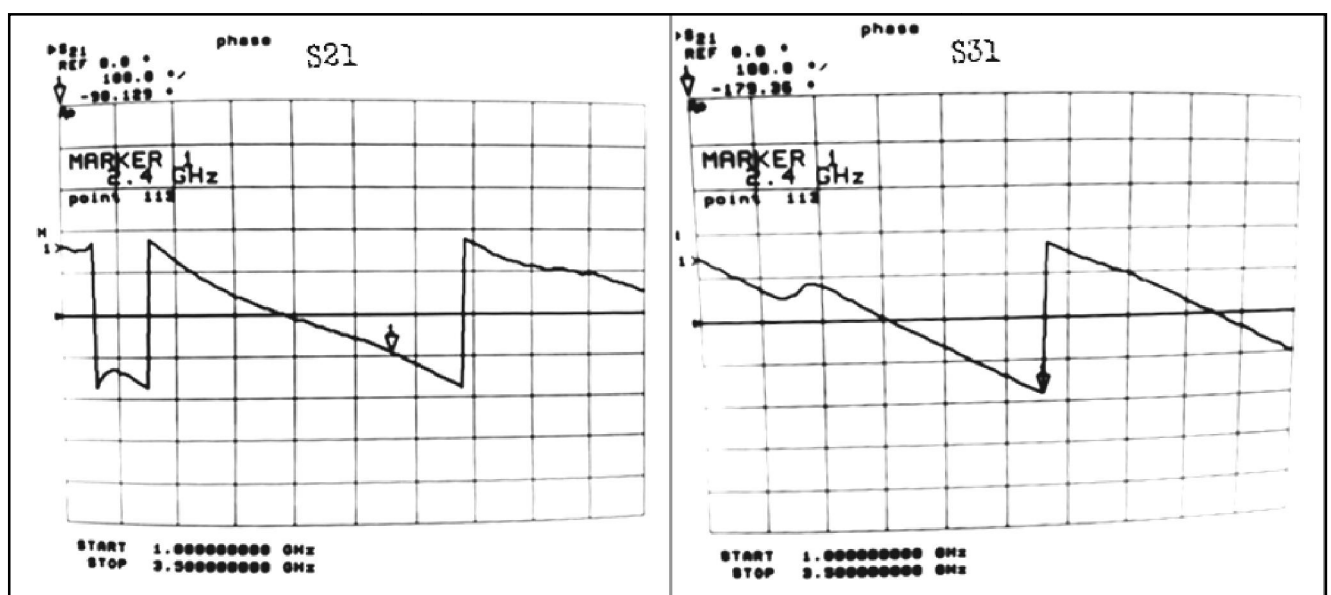


Figure (4.99) Phase measurement of S21 and S31 of 90° hybrid.

4.9.1.5 Bandwidth of 90° Hybrid

Figure (4.100) shows the measured bandwidth between the two markers 3 and 2 which are (2.525 GHz-1.925 GHz). Then the calculated bandwidth BW= 600 MHz which is close to the simulated result, BW= 612 MHz.

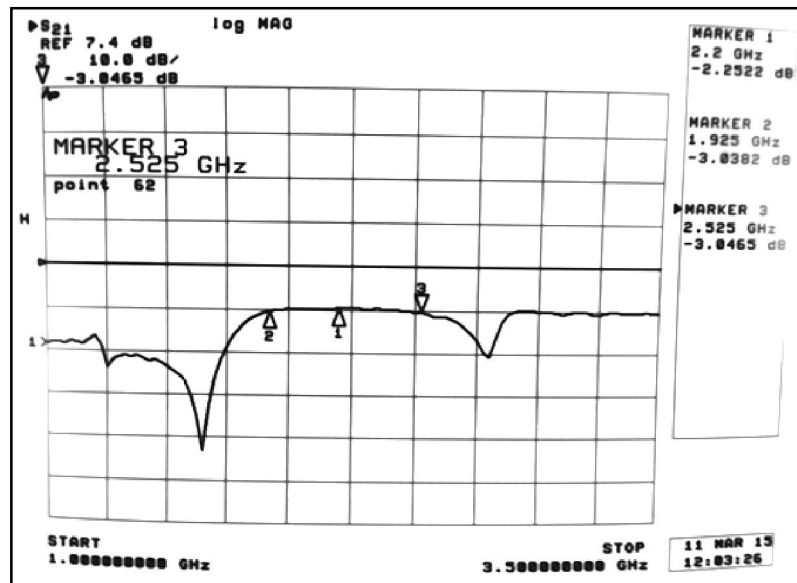


Figure (4.100) Measured bandwidth of 90° hybrid

4.9.2 Measurements of Cascaded 90° Hybrid

The experimental set-up for measurements is shown in figure (4.101), its photograph display the 90° hybrid cascade as Device Under Test (DUT), with terminated loads at ports.



Figure (4.101)The experimental set-up of cascaded 90° hybrid.

4.9.2.1 Return losses Measurements of Cascaded 90° Hybrid

Actual measurement result of the Return losses is shown in figure (4.102.a), where marker 1 shows return losses is $S_{11} = -29.982$ dB at the center frequency of 2.4 GHz .

In the simulation case 1 (figure (4.26)) as it was presented earlier , it was evaluated to $S_{11} = -27.165$ dB which are in a good agreement with measured results.

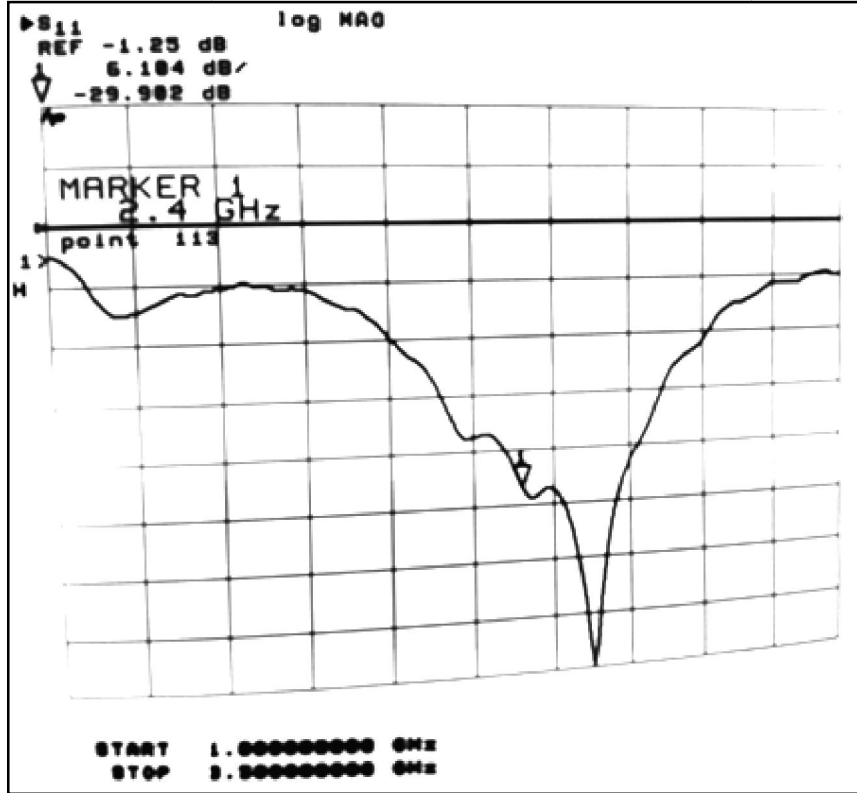
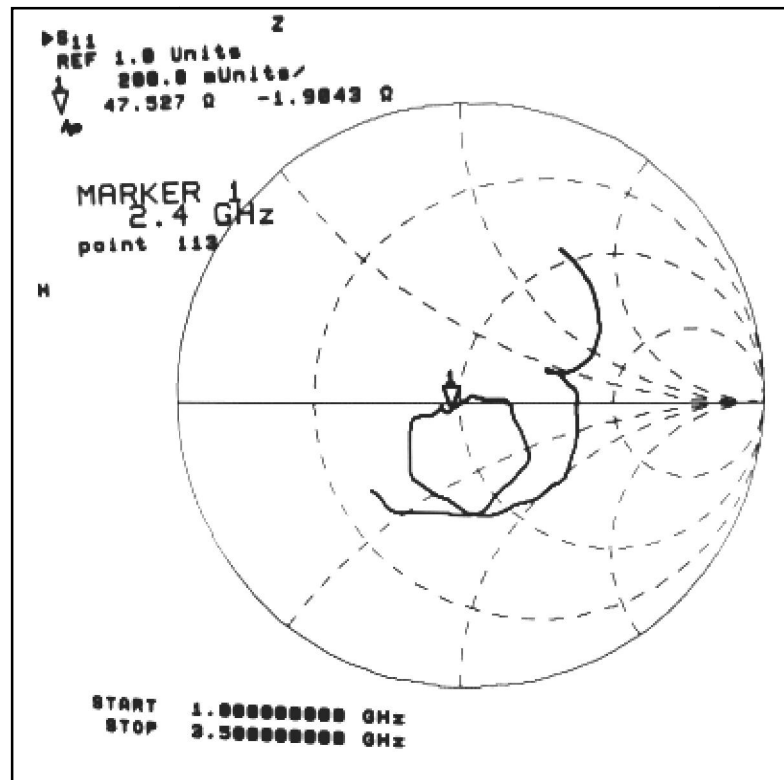


Figure (4.102.a) Measurement return losses of cascaded 90° hybrid .

4.9.2.2 Measurements of Input Impedance of Cascaded 90° Hybrid

Measurement result of the input impedance is shown in figure (4.102. b), where marker 1 pointed to the input impedance at the center frequency of 2.4GHz is $(47.527 - j1.9843)\Omega$.



Figure(4.102.b) Measurement of the input impedance of cascaded 90° hybrid.

4.9.2.3 Isolation Measurements of Cascaded 90° Hybrid

The measured isolation S_{41} of 90° hybrid cascade is shown in figure(4.103), where marker 1 shows the isolation of -31.595 dB at the center frequency of 2.4 GHz. whereas in the simulation case 1 (figure(4.27)), it was presented earlier, it was evaluated as $S_{41} = -28.736$ dB, which are closed to each other.

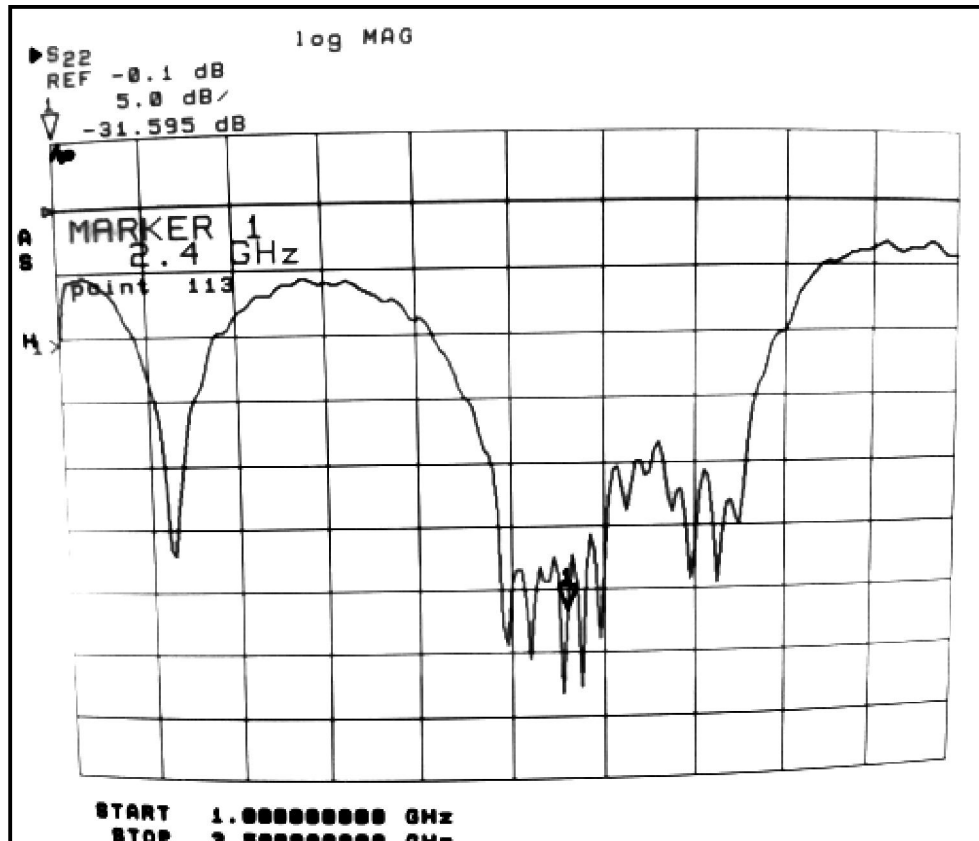


Figure (4.103) Measurement isolation S41 of cascaded 90° hybrid.

4.9.2.4 Phase of Cascaded 90° Hybrid

Figure (4.104) shows the phases of S21 and S31 of 90° hybrid cascade. From the measurement results of these phases .Marker (1) pointed to the phase of S21 which is ($\angle = 137.2^\circ$) whereas the marker (1) in the right side of the figure pointed to the phase of S31 ($\angle = 43.172^\circ$).

Then phase difference (S21-S31) $= 94.17^\circ$ between the port 2 and 3.

Comparing these results with simulated phase of 90 hybrid cascade coupler it was (S21-S31) $= 89.640^\circ$ as shown as in figure (4.30) and figure(4.31). There are discrepancies of phase which is may be explained by there are errors due to the fabrication and measurements.

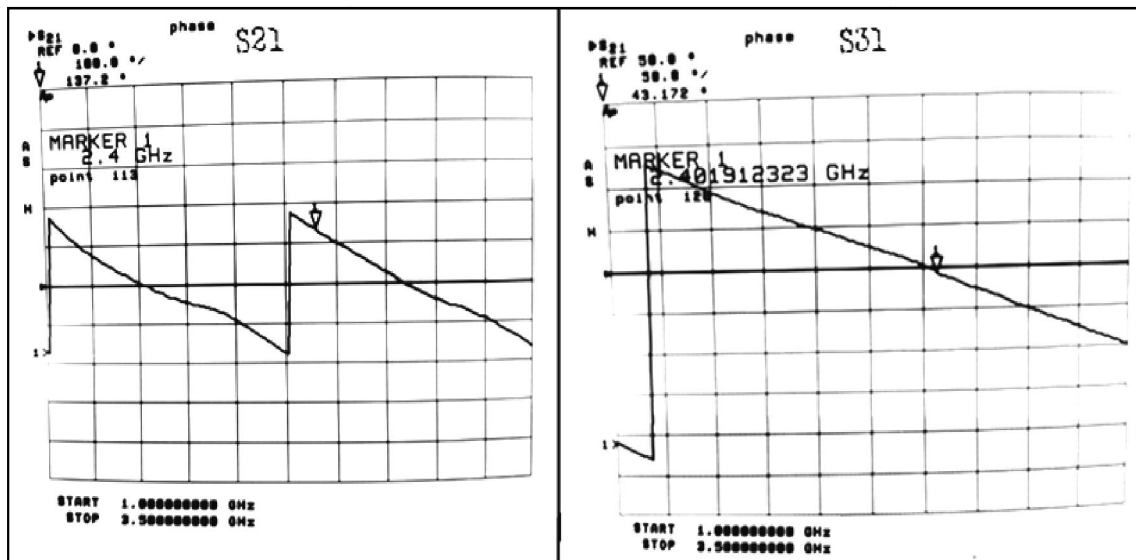


Figure (4.104) The phase measurement of S21 and S31 of cascaded 90° hybrid .

4.9.2.5 Bandwidth of Cascaded 90° Hybrid

Figure (4.105) records, the measured bandwidth between the two markers 1 and 3 which are (2.703 GHz - 2.124 GHz) and then gives the bandwidth BW= 579 MHz which less than the simulated bandwidth BW = 749MHz.

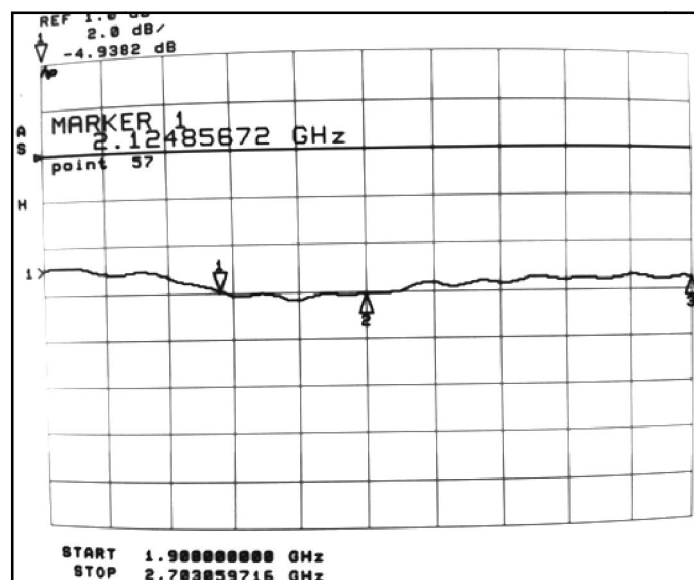


Figure (4.105) Measured bandwidth of cascade 90° hybrid

4.9.3 Measurements 90° Hybrid with Extended TL Output Ports

The experimental set-up for measurements is shown in figure (4.106) its photograph display the HP8510C vector network analyzer (VNA) and the 90 hybrid extended TL output ports as Device Under Test (DUT). In each case of measurement.



Figure (4.106) The experimental set-up of 90° hybrid with extended TL output ports.

4.9.3.1 Return losses Measurements of 90° Hybrid with Extended TL

The measurement result of the return losses is shown in figure (4.107.a), where marker 1 shows return losses is $S_{11} = -24.339$ dB at the center frequency of 2.49 GHz.

In the simulation case 1 (figure (4.43)) as it was presented earlier, it was simulated and given by $S_{11} = -30.975$ dB which are in a good agreement .

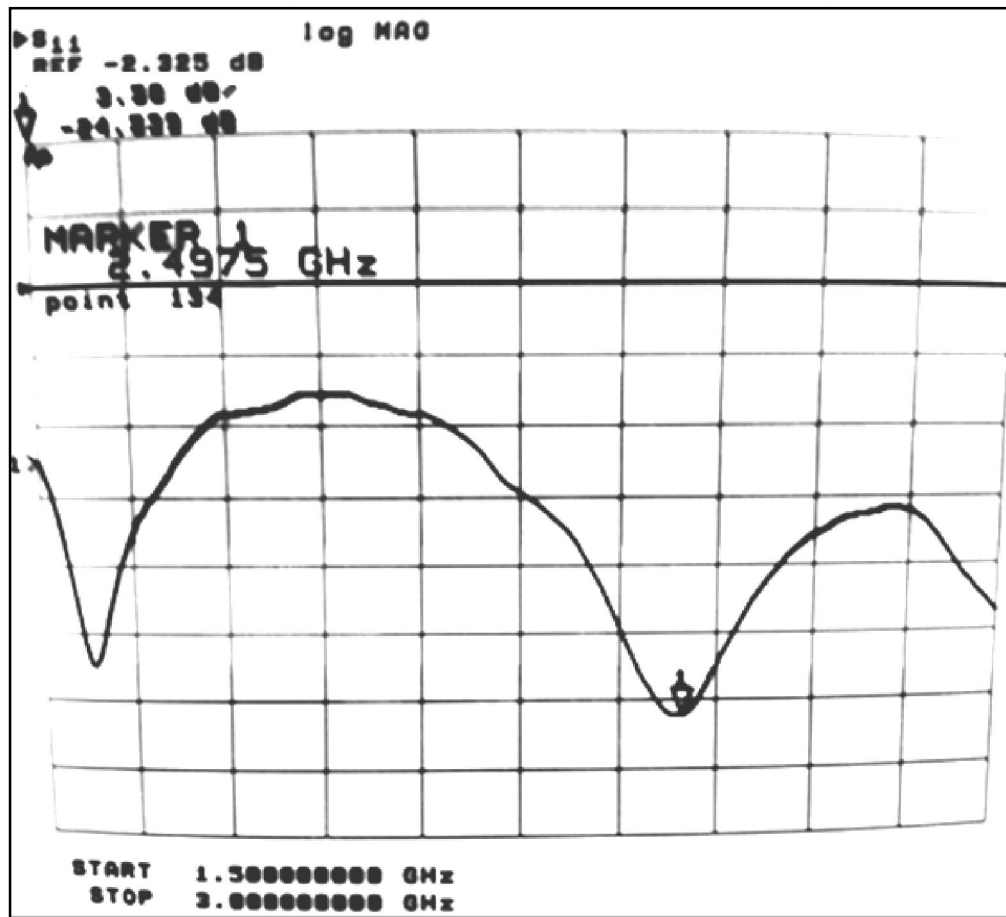
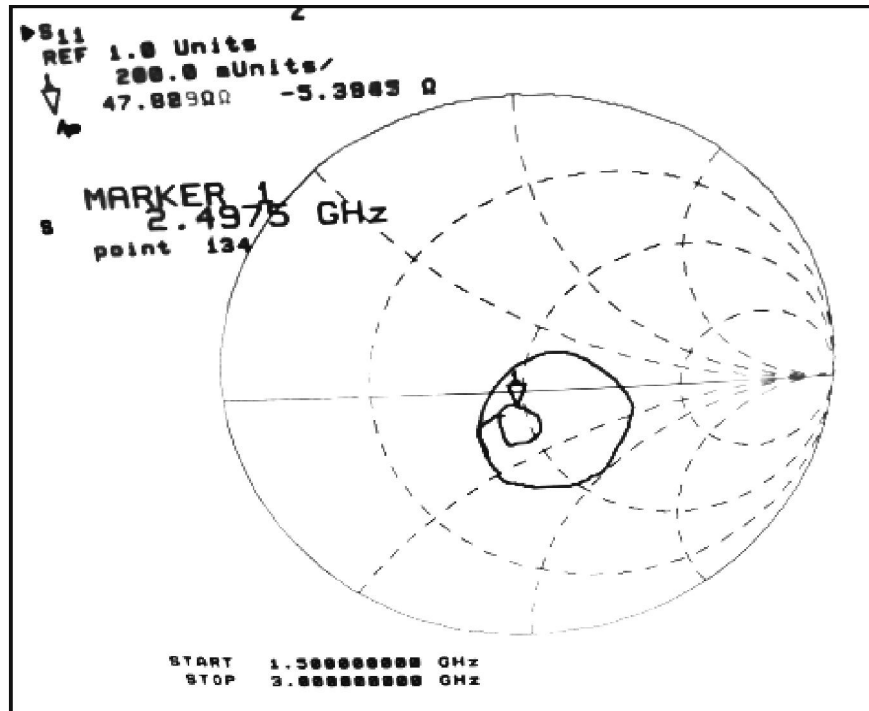


Figure (4.107.a) Measurement return losses of S_{11} of 90° hybrid with extended TL .

4.9.3.2 Measurements of Input Impedance of 90° Hybrid with Extended TL

Measurement result of the input impedance is shown in figure (4.107. b), where marker 1 shows the input impedance at the center frequency of 2.49GHz ($47.829 - j 5.3845 \Omega$).



Figure(4.107.b) Measurement of the input impedance of 90° hybrid with extended TL output ports.

4.9.3.3 Isolation Measurements of 90° Hybrid With Extended TL

The measured of S41 isolation of 90° hybrid extended TL outputs is shown in figure(4.108), where marker 1 shows the isolation of is -19.201dB at the center frequency of 2.5GHz.

Where as in the simulation case 1 (figure(4.45)), it was presented earlier, it was evaluated as $S_{41} = -25.588\text{dB}$.

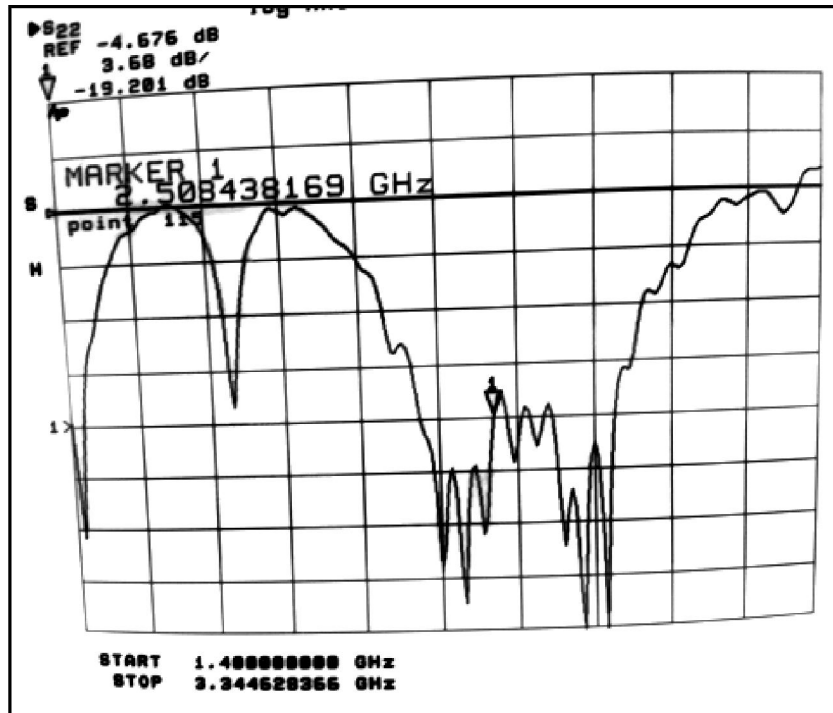


Figure (4.108) Measurement isolation S41 for 90°hybrid with extended TL .

4.9.3.4 Phase of 90° Hybrid With Extended TL

Figure (4.109) shows the phases of S21 and S31 of 90° hybrid extended TL outputs. From the measurement results of these phases. Marker (1) pointed to the phase of S21 which is ($\angle = -74.98^\circ$) whereas the marker (1) in the right side of the figure pointed to the phase of S31 ($\angle = -155.73^\circ$).

Then phase difference ($S21-S31$) = 80.55° between the port 2 and 3.

Comparing these results with simulated phase of 90 hybrid extended TL outputs coupler it was ($S21-S31$)= 90.008° as shown as in figure (4.48) and figure(4.49).

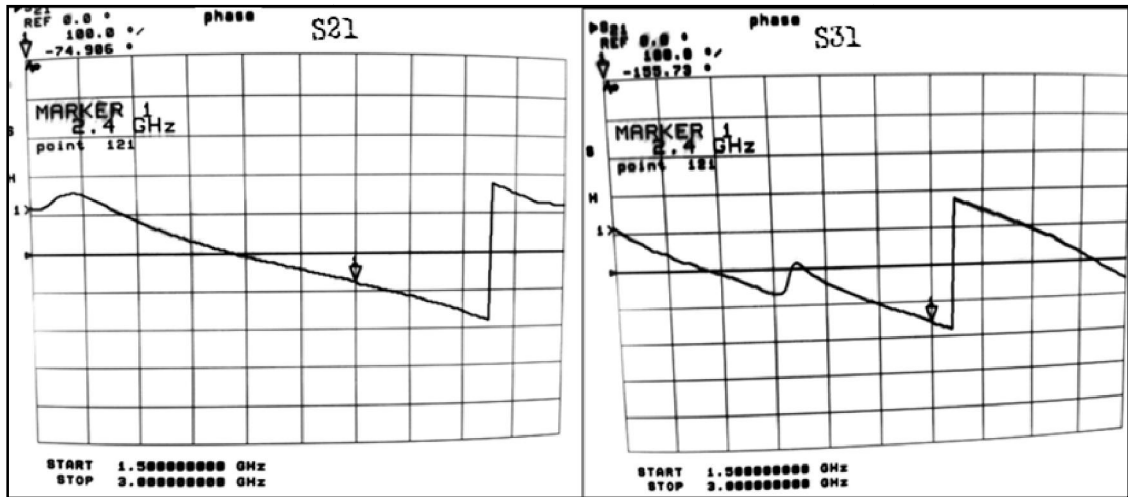


Figure (4.109) Measurement of the phase S21 and S31 of 90°hybrid with extended TL output ports.

4.9.3.5 Bandwidth of 90° Hybrid With Extended TL

Figure (4.110) shows the measured bandwidth between the two markers 1 and 2 which are (2.703 GHz-1.900 GHz). Then BW=1035 MHz which larger than the layout simulated BW= 431 MHz.

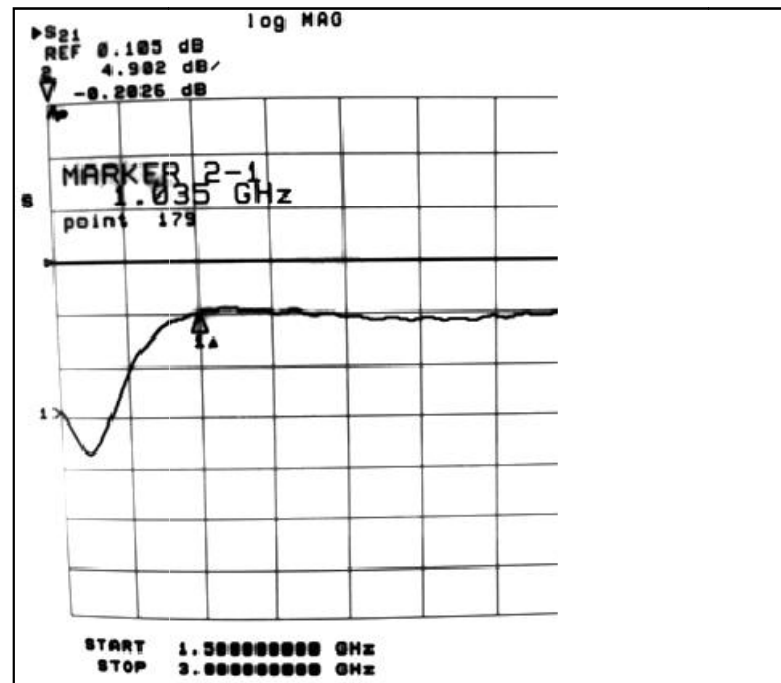


Figure (4.110) Bandwidth measurement of 90°hybrid with extended TL output ports.

Chapter 5

Results and Analysis

5.1 Introduction

This chapter will give the comparison of simulated and measured results in chapter 4, which are related to the improvement of phase and bandwidth of the 90° hybrid. These comparisons will also present the improved bandwidth in percent which are calculated as

$$BW\% = \frac{(BW \text{ of curent case}) - BW \text{ of case1}(90^\circ \text{ hybrid as reference})}{BW \text{ of case1}(90^\circ \text{ hybrid as reference})} \times 100\%$$

5.2 Comparison of simulated and measured parameters

Table (5.1), presents the comparison of the important parameter of all the cases stated in chapter 4 which is gives a good agreements between most of simulated and measured cases.

In Table (5.2), the comparison of bandwidth and phases for all the cases stated in chapter 4, we can observe that the phases are improved by using extended method and stubs and fixed it exactly 90° at the outputs which allows to use the developed Hybrid as 90° phase shifter.

Also the bandwidth is investigated and improved as shown in Table (5.2).

More comparison of parameters are shown in appendix (A).

Table (5.1) Comparison of simulated and measured parameters

90°Hybrid				
Parameter type	Cases			Measurement
	Schematic	Layout		
S11	-34.386 dB	-29.26 dB		-30.466 dB
S41	-36.458 dB	-32.420 dB		-26.329 dB
Coupling Voltage $C = 0.707 < 1$	0.686V	0.661V		-----
Coupling Factor S31(less than 1 dB ideally)	-3.273 dB	-3.593 dB		-----
Amplitude Difference = (S31-S21) = 0dB(ideally)	0.008 dB	0.003 dB		-----
90°HybridCascade				
Parameter type	Cases			Measurement
	Schematic	Layout		
S11	-24.008 dB	-27.165 dB		-29.982 dB
S41	-24.223 dB	-28.736 dB		-31.595 dB
Coupling Voltage $C=0.707 < 1$	0.684V	0.660V		-----
Coupling Factor S31(less than 1 dB ideally)	-3.297 dB	-3.610 dB		-----
Amplitude Difference = (S31-S21) = 0 dB(ideally)	0.026 dB	0.080 dB		-----
90°Hybridwith extended TL outputs ports				
Parameter type	Cases			Measurement
	Schematic	Layout		
S11	-34.646 dB	-30.975 dB		-24.339 dB
S41	-38.366 dB	-25.588dB		-19.201 dB
Coupling Voltage $C = 0.707 < 1$	0.685V	0.662V		-----
Coupling Factor S31(less than 1 dB ideally)	-3.280 dB	-3.584 dB		-----
Amplitude Difference = (S31-S21) = 0dB (ideally)	0.002 dB	0.013 dB		-----
90°Hybrid with two single stubs at outputs ports				
Parameter type	Cases			Measurement
	Schematic	Layout		
S11	-34.586 dB	-22.586 dB		-----
S41	-43.489dB	-38.815 dB		-----
Coupling Voltage $C=0.707 < 1$	0.653V	0.647V		-----
Coupling Factor S31(less than 1 dB ideally)	-3.704 dB	-3.783 dB		-----
Amplitude Difference = (S31-S21)= 0dB (ideally)	0.063 dB	0.002 dB		-----
90°Hybrid with four stub ports				
Parameter type	Cases			Measurement
	Schematic	Layout (A)	Layout (B)	
S11	-18.186 dB	-19.262 dB	-25.5 dB	-----
S41	-32.134 dB	-21.495 dB	-28.49dB	-----
Coupling Voltage $C=0.707 < 1$	0.617 V	0.628V	0.635V	-----
Coupling Factor S31(less than 1 dB ideally)	-4.201 dB	-4.041 dB	-3.946 dB	-----
Amplitude Difference = (S31-S21) = 0 dB(ideally)	0.037 dB	0.020 dB	0.042 dB	-----

Table (5.2) Comparison phases & bandwidths of all Cases

90°Hybrid				
Parameter type	Cases			
	Schematic	Layout		Measurement
Phase difference outputs ΔØ=(S21-S31)	90°.033	89°.65		89°.24
Bandwidth(BW)	360 MHZ	612 MHZ		600 MHZ
BW%	-----	Refrence1		Refrence2
90°HybridCascade				
Parameter type	Cases			
	Schematic	Layout		Measurement
Phase difference outputs ΔØ=(S21-S31)	89°.923	89°.680		94°
Bandwidth (BW)	420 MHZ	749 MHZ		579MHZ
BW%	-----	22.35%		-3.5%
90°Hybrid with extended TL outputs				
Parameter type	Cases			
	Schematic	Layout		Measurement
Phase difference outputs ΔØ=(S21-S31)	90°.002	90°.008		80°.55
Bandwidth (BW)	275 MHZ	431 MHZ		1035MHZ
BW%	-----	-29.5%		72%
90°Hybrid with two single stubs at outputs				
Parameter type	Cases			
	Schematic	Layout		Measurement
Phase difference outputs ΔØ=(S21-S31)	90°.000	90°.001		-----
Bandwidth(BW)	380MHZ	888 MHZ		-----
BW%	-----	45.098%		-----
90°Hybrid with four stubs ports				
Parameter type	Cases			
	Schematic	Layout (A)	Layout (B)	Measurement
Phase difference outputs ΔØ=(S21-S31)	90°.006	89°.567	-270°.592 OR 89°.409	-----
Bandwidth	540 MHZ	721 MHZ	894 MHZ	-----
BW%	-----	17.81%	46.078%	-----

5.3 Comparison of Significant the Results

This comparison between the results which are obtained from the (simulation & measurements) or (simulation schematic & layout) of the hybrid parameters are Shown in the following figures .

5.3.1 Case1. Design 90° Hybrid

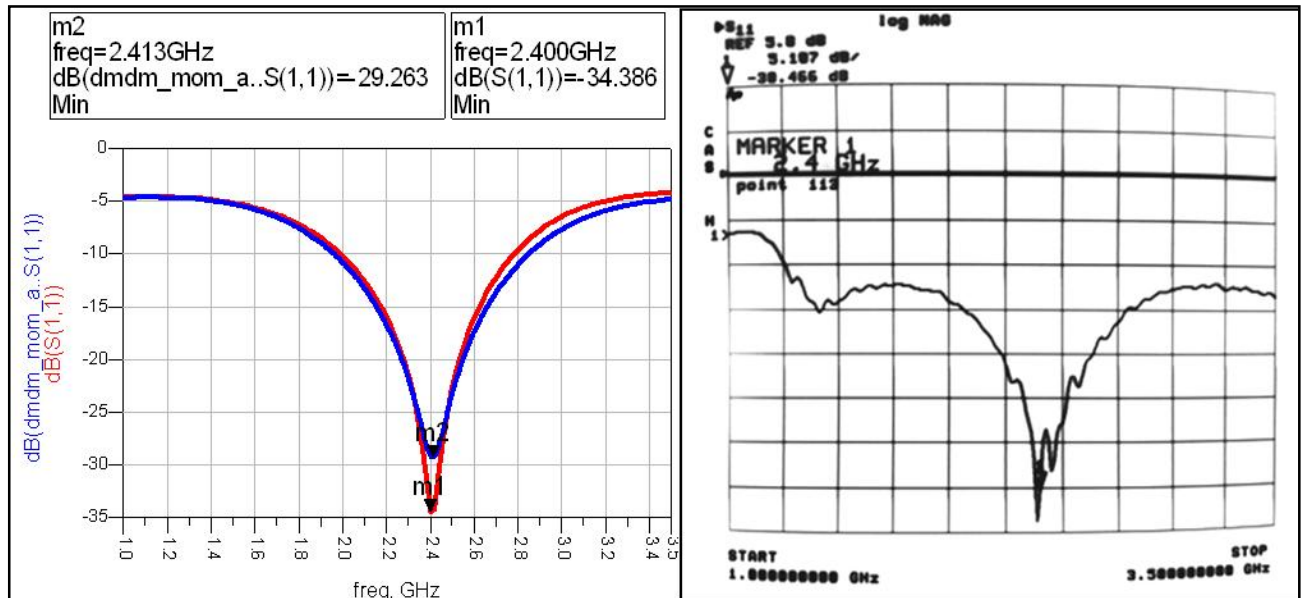
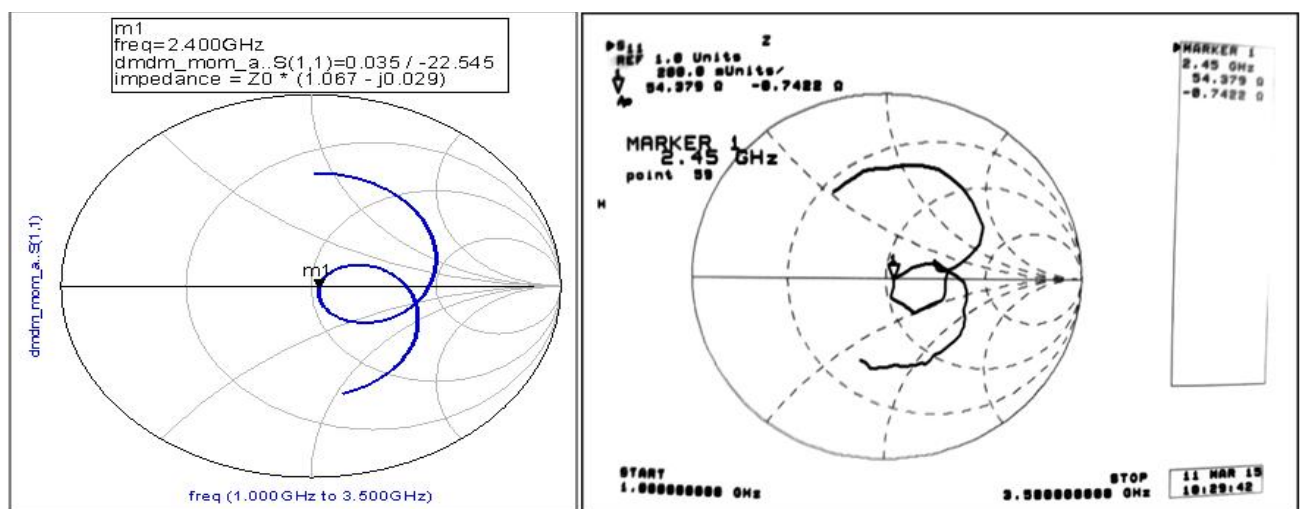


Figure (5.1) Comparison simulation & measurements for return losses of 90° Hybrid



Figure(5.2)Simulation &Measurement of the input impedance of 90° hybrid

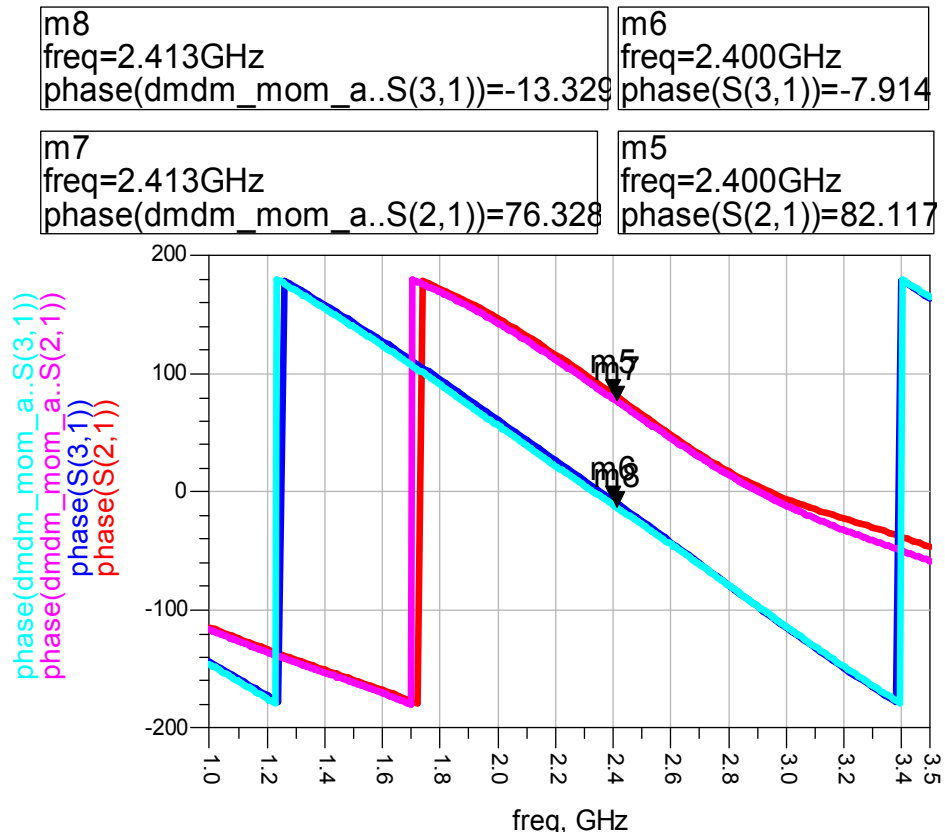


Figure (5.3) Comparison simulation schematic & layout for phase difference of 90° hybrid.

5.3.2 Case 2. Cascaded 90° Hybrid

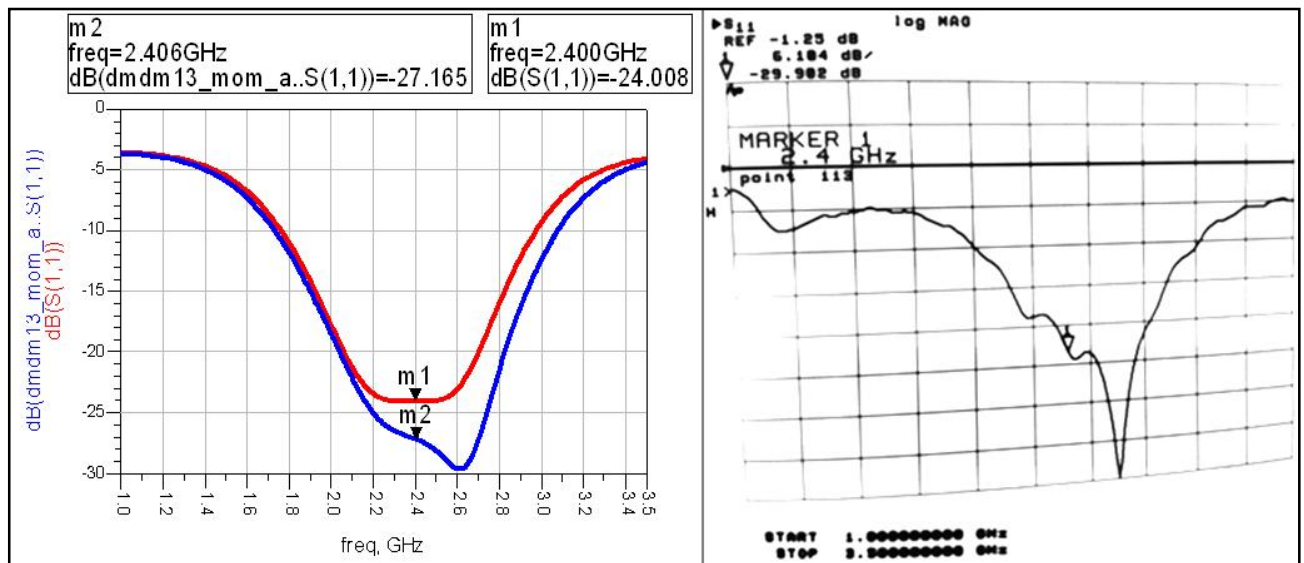


Figure (5.4) Comparison simulation & measurements for return losses of Cascaded 90°Hybrid

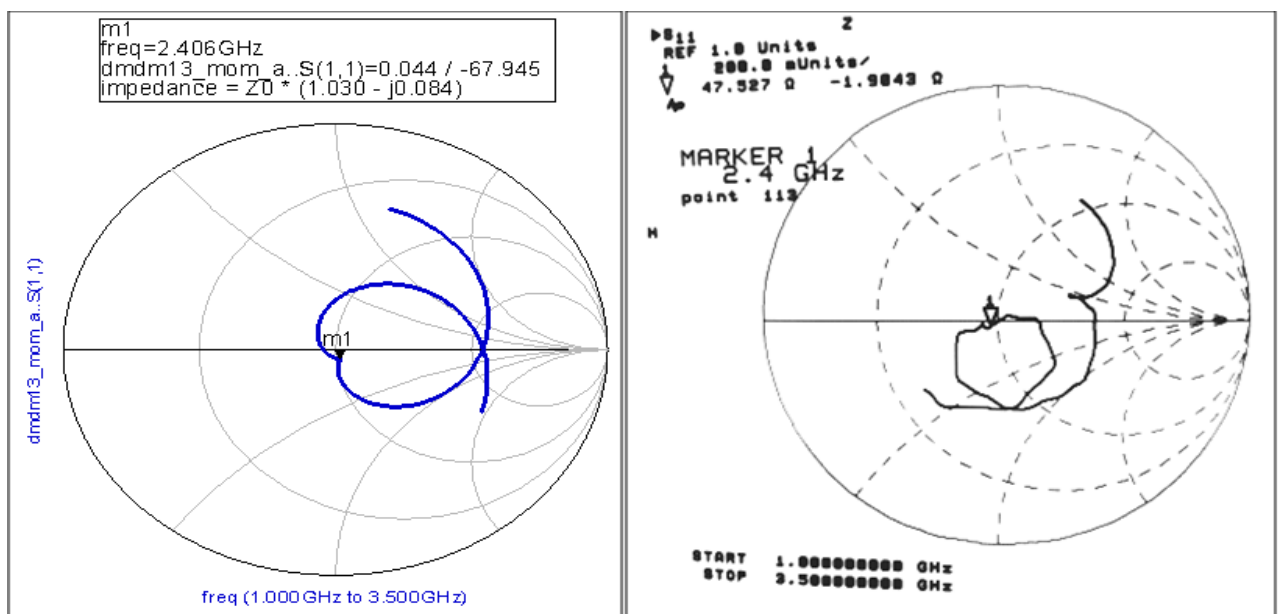


Figure (5.5) Comparison Simulation & Measurement of the input impedance of Cascaded 90° hybrid

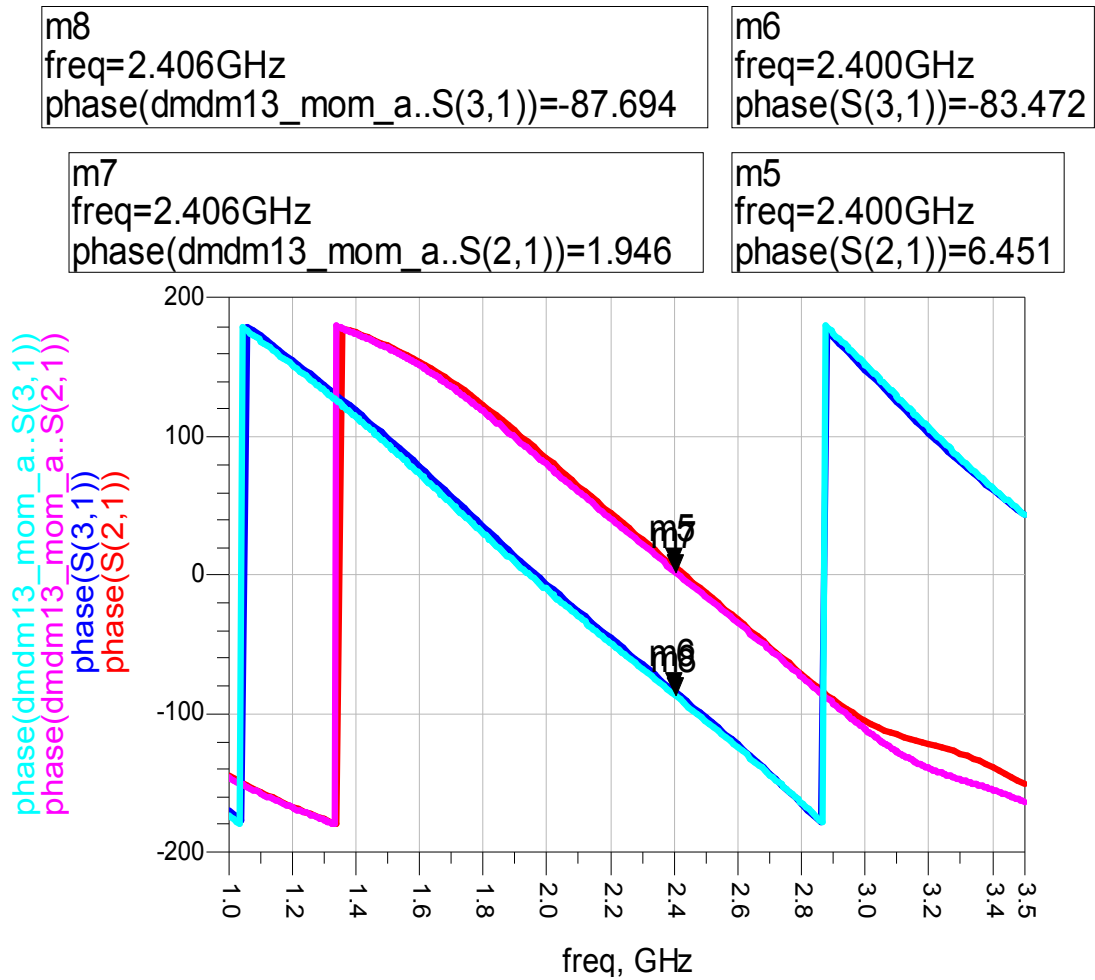


Figure (5.6) Comparison Simulation schematic & layout phase difference of cascaded 90° hybrid

5.3.3 Case3. 90° Hybrid with Extended TL Output Ports

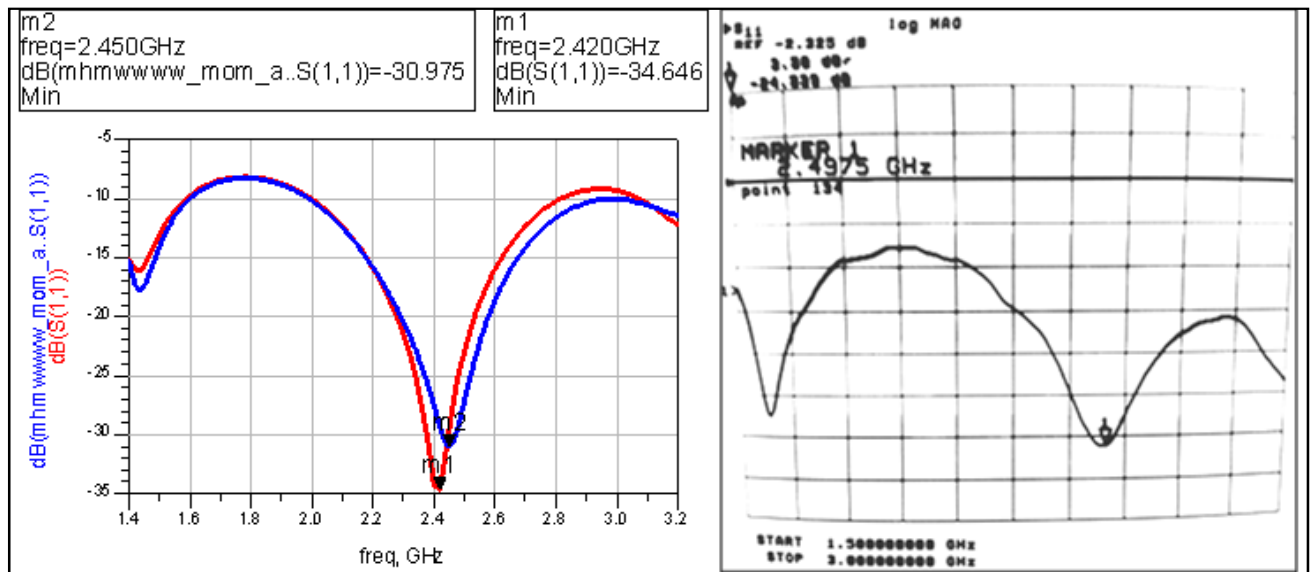
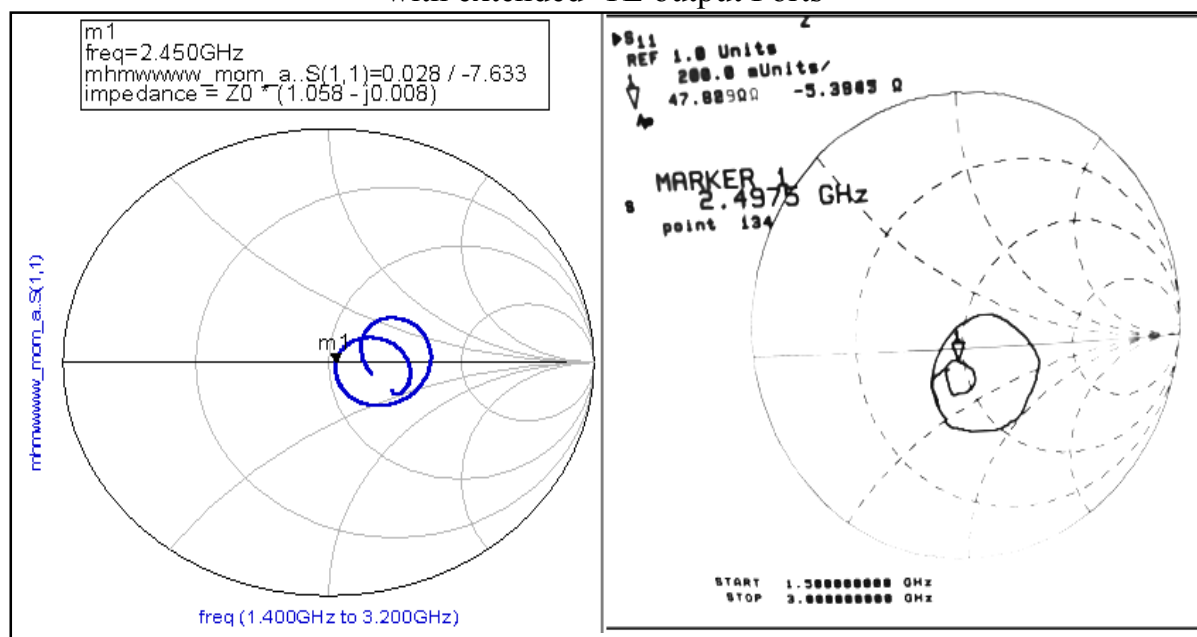


Figure (5.7) Comparison simulation & measurements for return losses of 90° Hybrid with extended TL output Ports



Figure(5.8) Comparison Simulation & Measurement of the input impedance of 90° hybrid with extended TL output Ports

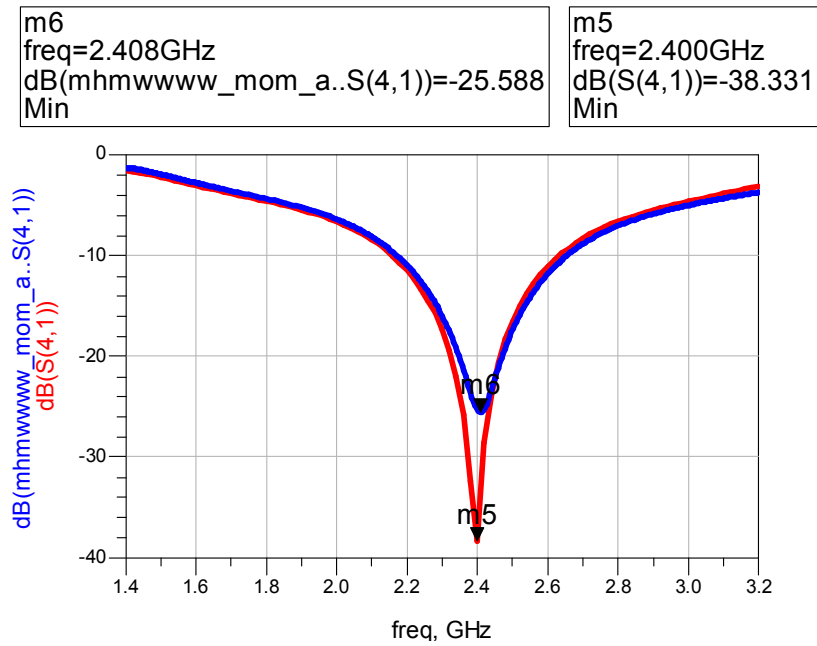


Figure (5.9) Comparison simulation schematic & layout isolation of 90° hybrid with extended TL output Ports .

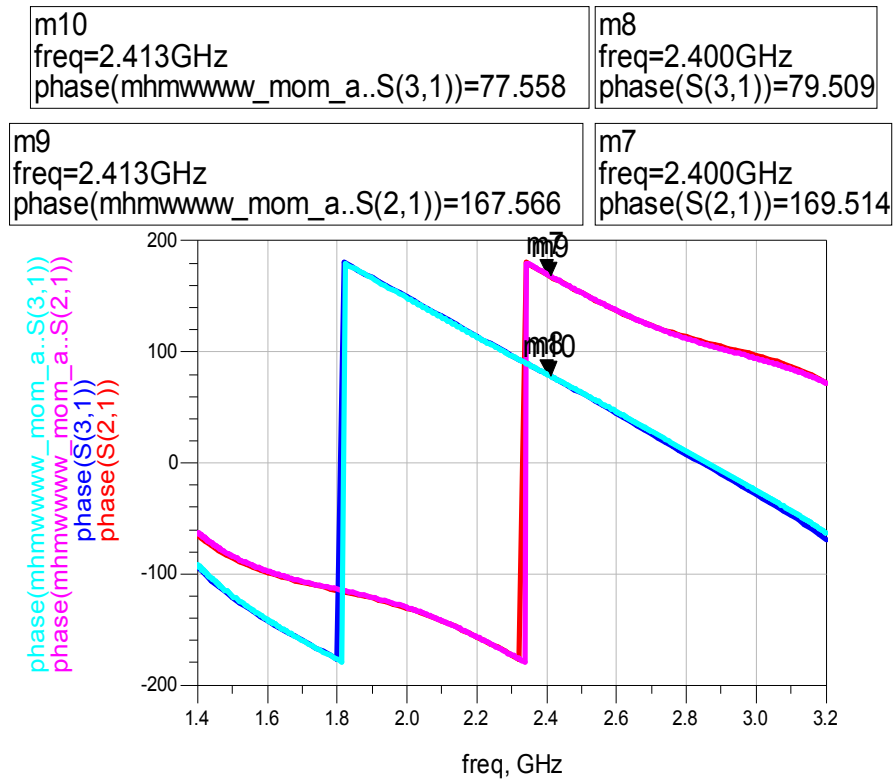


Figure (5.10) Comparison simulation schematic & layout phase difference of 90° hybrid with extended TL output ports .

5.3.4 Case4. Design 90° Hybrid With Two Single Stub at the Output ports

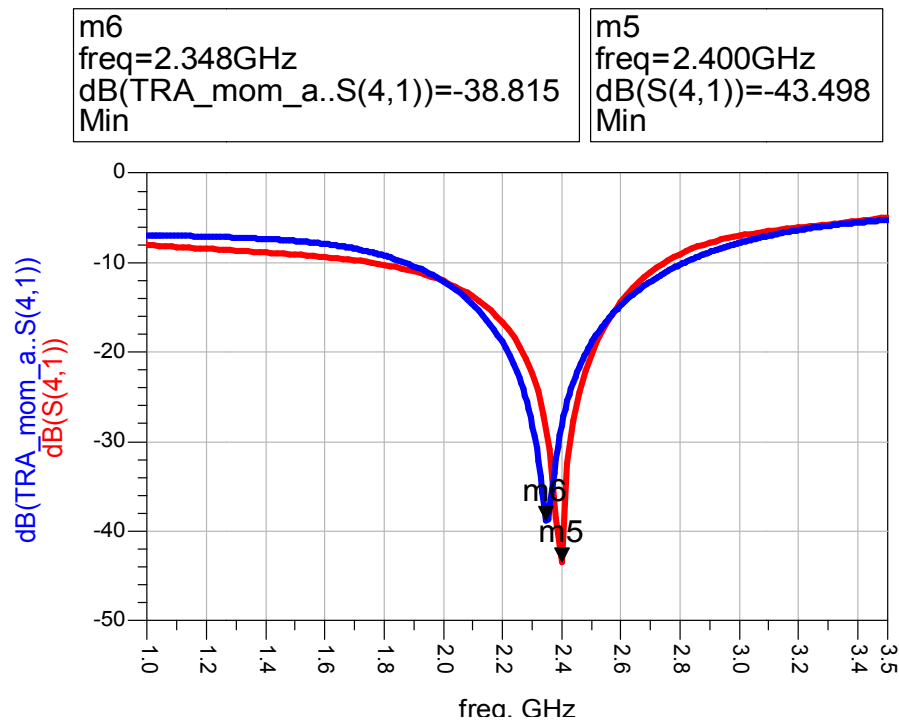


Figure (5.11) Comparison Simulation schematic & layout isolation of 90° Hybrid with two single stub at the output ports

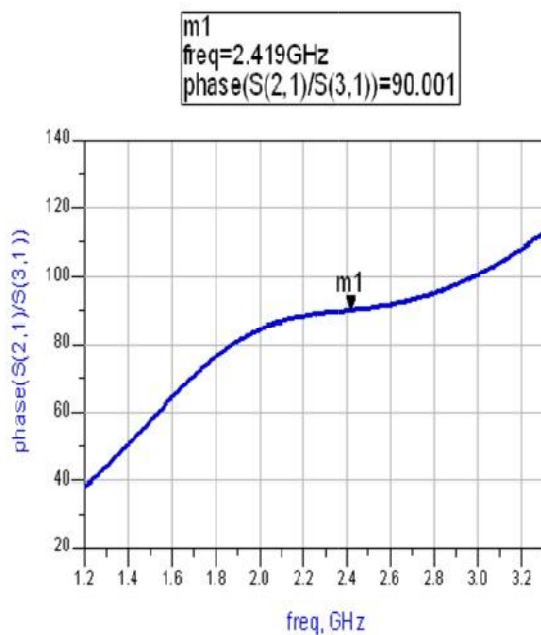


Figure (5.12) Comparison Simulation schematic & layout for phase of 90° Hybrid with two single stub at the output ports

5.3.5 Case5 . Design 90° Hybrid With Four Stub at the Ports

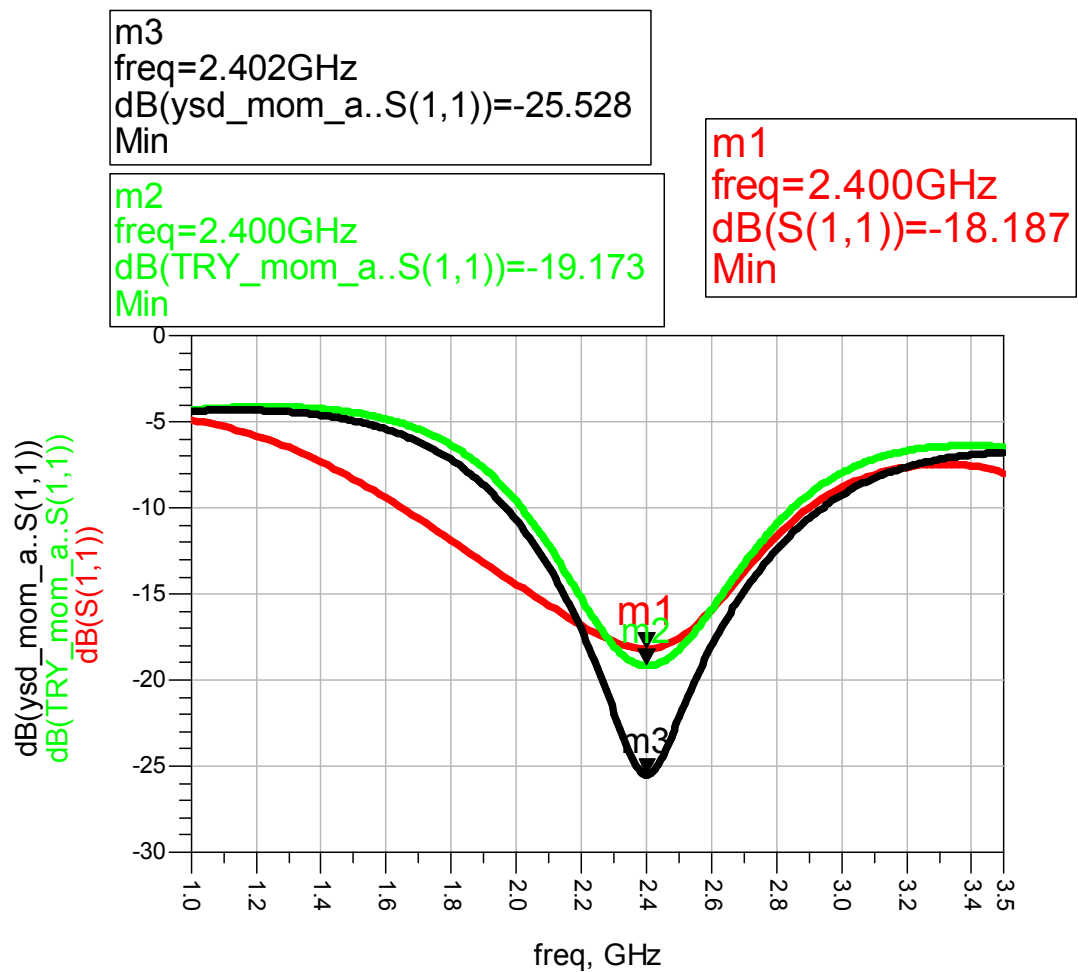


Figure (5.13) Comparison simulation schematic & layout(A and B) return losses of 90°hybrid with four stub at the ports .

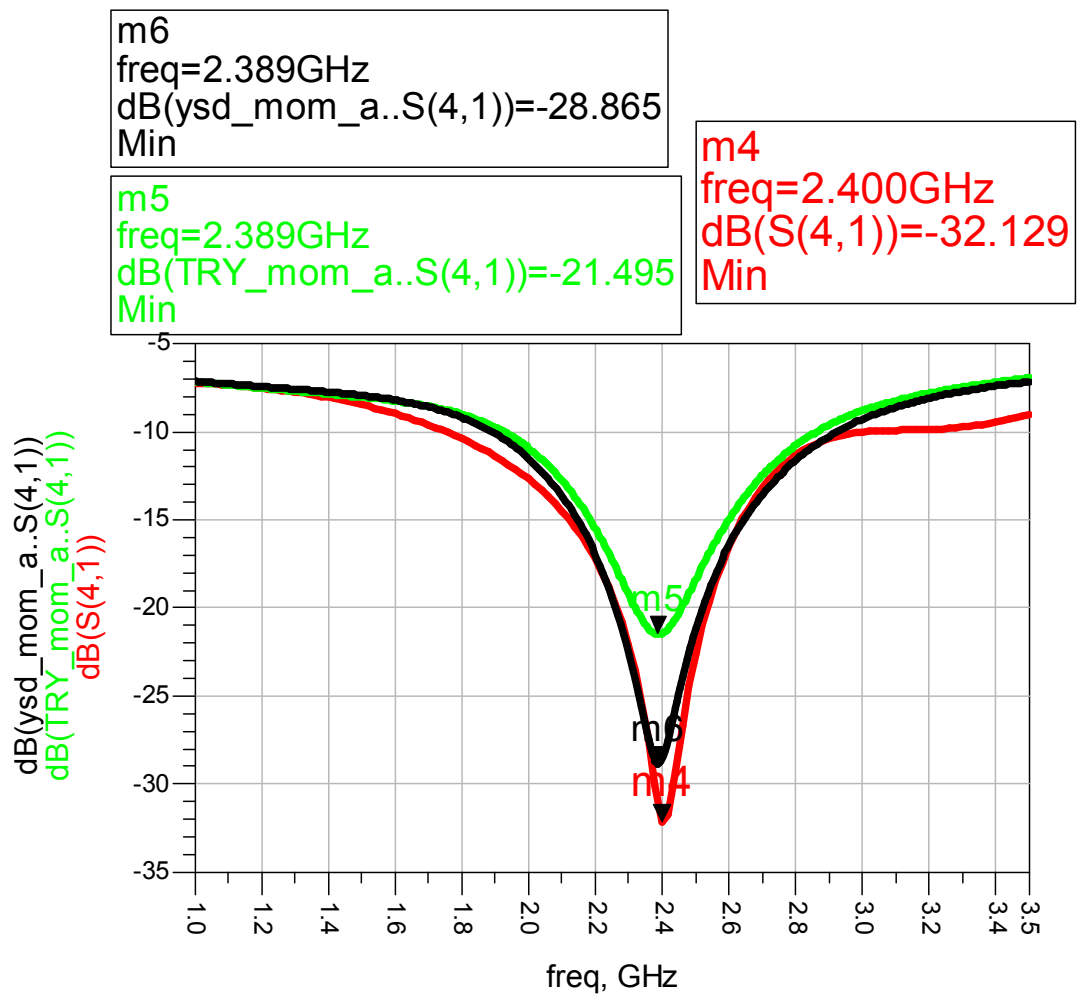


Figure (5.14) Comparison simulation schematic & layout(A and B) isolation of 90° hybrid with four stub at the ports

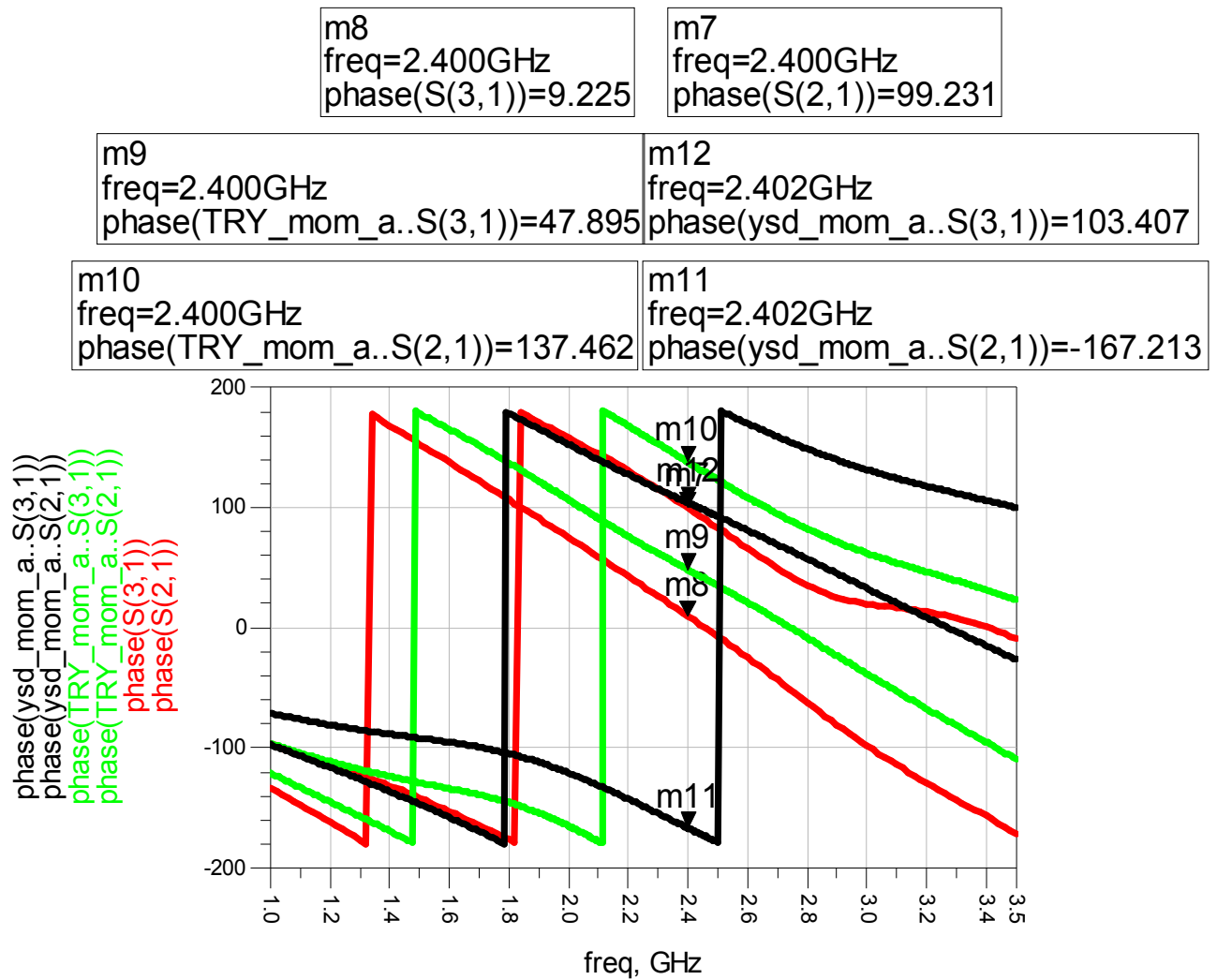


Figure (5.15) Comparison Simulation schematic & layout (A and B) phase difference of 90° Hybrid with four Stub at the Ports .

Chapter 6

Conclusion and Future Work

6.1 Conclusion

The aim of this thesis is to work on 90° hybrid coupler which provides a 90° phase shift between outputs ports over bandwidth centered at 2.4 GHz.

The results of the hybrid coupler very promising. After enhancement methods used the coupler had very good behavior with respect to power split and phase shift.

However, these parameters that would more closely meet these specifications according to the comparison results of simulation layout momentum and experimental measurement .

Summarize the conclusions :

- Our main work is optimization of matching network for a 90° hybrid coupler simulation and experimental measurements for a smaller phase error and greater bandwidth by using improvement methods .
- Investigated the characteristics of five designs of hybrid circuits by using the software which is ADS. Simulation results which are compared with desired results. Three cases of the designs are fabricated on printed circuit board, which are 90 degree hybrid, cascade hybrid and extended 90 hybrid with the specifications listed in table (3.1).
- Investigation of the three fabricated cases were measured including , amplitude, phase response , return losses , coupling and isolation. A good agreement between the simulation and measured results.
- In general, a good agreement between simulated and measured results. However there are several reasons to explain the small discrepancies in the results.
- The discrepancies between simulation and measured results comes from manual fabrication of circuits which gives some errors due to the fabrication process.

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- The measurement process has also some errors that occur because of the DUT in some cases don't have the termination load at the ports due to the lack of the some connectors and terminations in the lap.
 - By using the 90° hybrid with extended TL method the phase was improved and gives us 90° exactly in simulation which allows of using this hybrid as phase shifter.
 - By using 90° hybrid with four single stubs ports , the phase and the bandwidth was enhanced and improved which is encouraging to use this hybrid as phase shifter.
 - By adding series resistance (R) with a capacitance (C) as a load in 90° hybrid with stubs at the ports, the stability of the phase at the outputs are improved.
 - Reduction the size of 90° hybrid with four stubs in case (B) is performed and gives a very good simulation results to obtain the improve the phase and bandwidth.

6.2 Future Work

There are some work should be done in order to complete this work which are considered here by two main points:-

- There are some discrepancy between some simulation and measured results which we expected it from manufacture errors and measurement errors which can be explained in more details as:-
 - ✓ Manufacture errors, the manually procedure was follows in fabricating all the PCB which is not accurate. also the soldering of the SMA connectors causes errors.
 - ✓ Measurement errors, due to the lack of some connectors and $50\ \Omega$ terminations.

for these reasons some Measurements should be repeated and compared with the simulated results.
- Some of the simulated layout circuits such as 90° hybrid with two outputs stubs and four ports stubs (Case 4 and Case 5) should be fabricated, tested on VNA and compared with simulation results.

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