DESIGN OF DIGITAL PHASE SHIFTER WITH VARIOUS ORDERS OF BPF

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Submitted by

AJAY KUMAR
2K09/MOC/02

Under the Supervision of

MR. P. R. Chadha
Associate Professor
Department of Electronics & Communication Engineering

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING
IN ASSOCIATION WITH
DEPARTMENT OF APPLIED PHYSICS

DELHI TECHNOLOGICAL UNIVERSITY

(Formerly Delhi College of Engineering)
Main Bawana Road, Shahabad Daulatapur, New Delhi – 110042
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CERTIFICATE

This is to certify that the dissertation titled **"Design of Digital Phase Shifter with Various Orders of BPF"** is the bonafide work of Ajay Kumar (2K09/MOC/02) under our guidance and supervision in partial fulfillment of requirement towards the degree of Master of Technology in Microwave and Optical Communication Engineering from Delhi Technological University, New Delhi.

Mr. P. R. Chadha	Dr. Rajiv Kapoor
Associate Professor & Supervisor	Head of Department
Electronics and Communication Engineering	Electronics and Communication Engineering
Delhi Technological University	Delhi Technological University
Delhi	Delhi
Date: /06/2011	Date: /06/2011

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Ajay Kumar
2K09/MOC/02
M.Tech (Microwave and Optical Communication Engineering)
Department of Electronics and Communication Engineering
Delhi Technological University

ABSTRACT

This thesis presents the theory and a design method for distributed digital phase shifters, where both the phase-error bandwidth and the return-loss bandwidth are considered simultaneously. The proposed topology of each phase bit consists of a transmission-line (TL) branch and a bandpass filter (BPF) branch. The BPF branch uses grounded shunt quarter wavelength stubs to achieve phase alignment with the insertion phase of the TL branch. By increasing the number of transmission poles of the BPF branch, the returnloss bandwidth can be increased. Analysis of the BPF topology with one, two, and three transmission poles is provided. The design parameters for 22.5 , 45 , 90 , are provided for bandwidths of 30%, 50%**.**

The three bit digital phase shifter is designed with minimum phase shift of 22.5^0 and maximum phase provided is 157.5^0. Results of all three bit phase shifts are produced and their respective phase errors and return losses are compared.

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Motivation

Digital phase shifters are key components in digitally controlled phased-array antennas and other microwave control modules. Usually each bit of a switched phase shifter contains two branches. When switching between the two branches, a differential phase shift is generated while the insertion loss remains the same. In the design of digital phase shifters, both phase error and return loss should be considered simultaneously, and the overall bandwidth of the phase shifter is the intersection of the phase-error bandwidth and return-loss bandwidth.

Phase shifting can be realized by using lumped-element networks and distributed networks . At high frequencies, the parasitics of the lumped elements usually degrade the phaseshifter bandwidth. Therefore, topologies with transmission lines (TLs) have been investigated to increase the bandwidth.

Background

Before discussing topologies of phase shifter design and their comparison we need to know- basics of phase shifter, their types and conventional phase shifters. Basics of transmission lines and bandpass filters are required.

Organization of thesis

Chapter1: "basics of phase shifter" includes the basic theory required to understand phase shifter design and conventional phase shifter structures.

Chapter2: "Topologies for three bit phase shifter design" contains basic design of digital phase shifter having two branches i.e. TL branch and BPF branch. BPF can be implemented with different topologies with different orders of filter. Their limitations and specific cases are discussed in this chapter.

Chapter3: "simulation results" contains simulated output results of digital phase shifter bits and their comparison for different bandwidths.

Chapter4: "conclusion and future work" contains final summary of what result we have achieved and the future research work possible in extension to this this thesis work.

Chapter 1

Basics of phase shifter

1.1 Definition

Phase Shifters are devices, in which the phase of an electromagnetic wave of a given frequency can be shifted when propagating through a transmission line. In many fields of electronics, it is often necessary to change the phase of signals. RF and microwave Phase Shifters have many applications in various equipments such as phase discriminators, beam forming networks, power dividers, linearization of power amplifiers, and phase array antennas.

The major parameters which define the RF and microwave Phase Shifters are:

- frequency range,
- bandwidth (*BW*),
- total phase variance (Δj),
- insertion loss (*IL*),
- switching speed,
- power handling (*P*),
- accuracy and resolution,
- input/output matching (*VSWR*) or return loss (*RL*),
- harmonics level.

1.2 Relation between Propagation Constant, Phase Shift, Delay, and Wavelength

In a transmission line the **Propagation Constant** is a complex number having two parts:

the real portion is the **attenuation constant** (**α**, neper per unit length) and the imaginary portion **βx** is called the **phase constant** (**β**, radians per unit length).

The **attenuation constant** α determines the way a signal is reduced in amplitude as it propagates down the line, while the **phase constant β** shows the difference in phase between the voltage at the sending end of the line and at a **distance x**.

The phase constant **βx** shows the phase shift of the voltage (or current) at a point located at a distance **x** along a transmission line with respect to the sending voltage (or current).

A phase shift of 360° (or **2π** radians) equals one wavelength and, as shown in figure below, marks the distance between successive points on the waveform (such as zero crossings).

The wavelength **λ** is the distance **x** required to make the phase angle **βx** increase by **2π** radians.

$$\lambda = 2\pi / \beta \qquad (1.1)$$

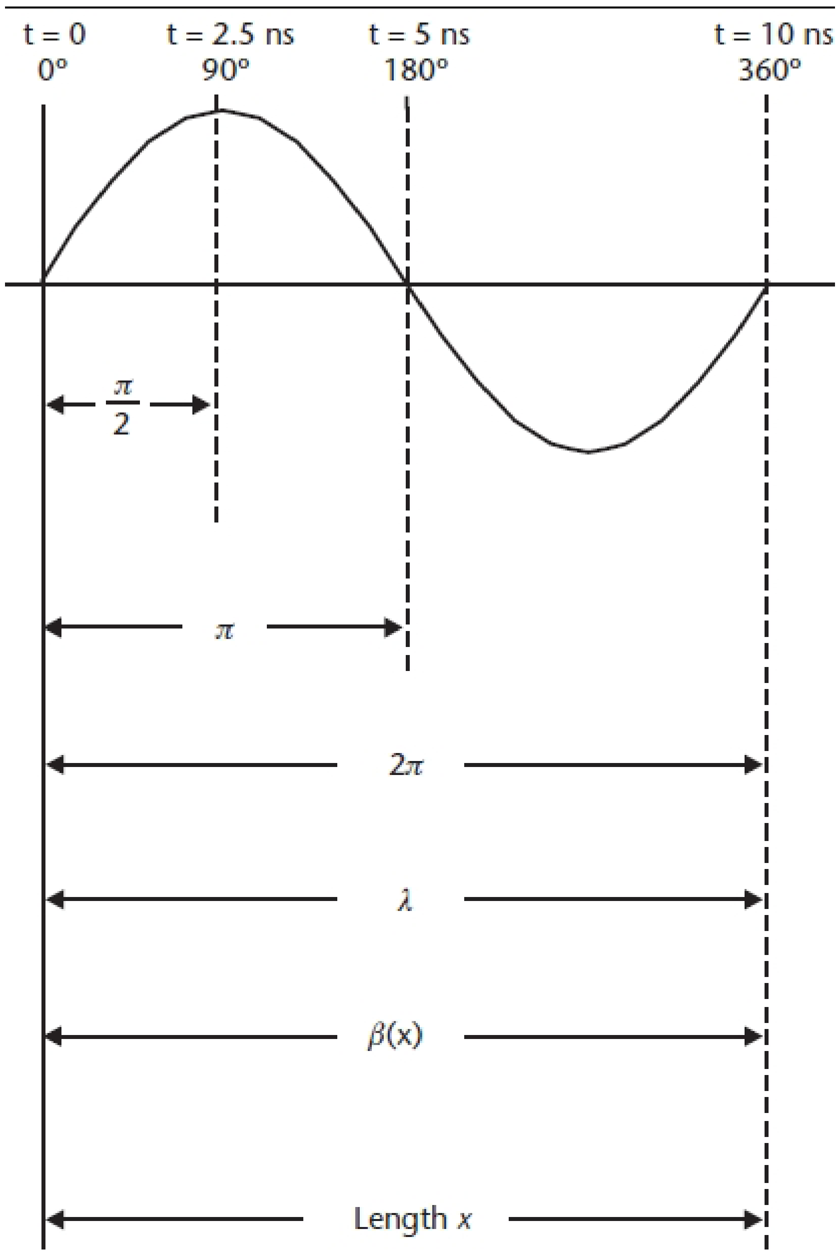


Figure1.1 Relationship between degrees (°), radians (2π), phase shift (β), and wavelength (λ)

From the figure above can be stated that a phase shift may also be seen as a delay. The relation between the phase shift and the time delay is given by:

Time delay (seconds) = [Phase Shift (°)] / [360 x frequency (Hz)]

The time delay is proportional to the inverse of velocity **Vp**.

The amount of delay a transmission line introduces per distance **x** is:

Time delay = x / Vp = (βx / ω)

1.3 Classification of Phase Shifters

The function of a phase-shifter is to control the phase characteristics in the processing of an input signal via a digital or analogue command and can be accomplished in a passive device. There are basically two general methods of producing phase-shifters: mechanical and electronic.

1.3.1 Mechanical and Electronic Phase Shifter

Most of the mechanical phase shifters incorporate a lumped element quadrature hybrid, together with a matched pair of L-C networks, to realize variable phaseshifts. Variable L-C networks linked to output ports act as sliding short circuits. Placed at the output ports of the hybrid, these short circuits reflect incident energy back towards the source.

Electronically-controlled, continuously variable phase-shifters operate similarly to the mechanical phase-shifters described above. The principal difference is that variable-controlled capacitors (varactors), such as controlled by voltage, are used instead of manually adjusted capacitors.

Larger mechanical devices are expensive and cannot take advantage of the economies of scale needed to make affordable phase-shifters. Therefore electronic phase-shifters have broad applicability for both commercial and military applications, including advanced military radars, cellular base stations, satellite communications, and automotive anti-collision radar.

1.3.2 Analogue and Digital Phase Shifters

Additionally there are two broad types of variable electronic phase-shifters: analogue and digital. Analogue phase-shifters change the output phase by means of an analogue signal (e.g. voltage) to provide a continuously variable phase. Digital phase-shifters use a digital signal to change the output phase in quantized steps, providing a discrete set of phase states that are controlled by two-state phase bits.

1.4 Basic block diagram of phase shifter

Basic block diagram of phase shifter is as given below. The phase introduced by it is expressed in terms of S parameters.

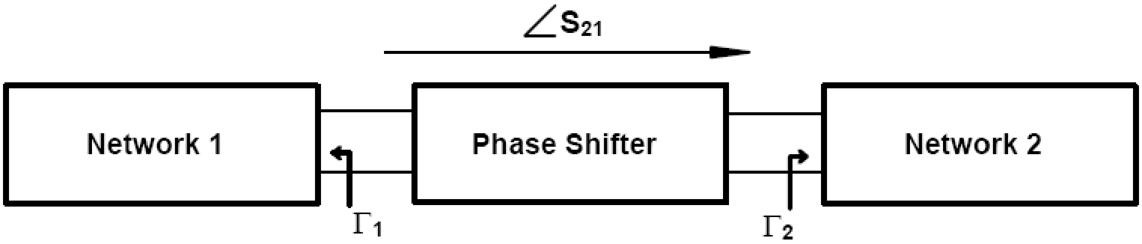


Figure 1.1 A phase shifter and its loads at two ends

In above block diagram Γ1 is reflection coefficient of network1 and Γ2 is reflection coefficient of network2.

1.5 Review of conventional phase shifters

Phase shifters have been developed for phased array antennas for more than half a century [17]. Generally, there are three types of phase shifters: mechanical phase shifters, ferrite phase shifters and semiconductor device phase shifters. We are concerned with only electrical planar phase shifters switched by semiconductor devices are covered.

1.5.1 Topologies to Achieve Phase Shifts

An ideal phase shifter is a two-port device whose insertion phase can be changed while its insertion loss remains the same. Assume the reference state has an insertion phase of ϕ_1, and the phase shifting state has an insertion phase of ϕ_2, the phase shift $\Delta\phi$ is the phase difference between the two states, which is given by

$$\Delta\phi = \phi_2 - \phi_1 \qquad (1.1)$$

Normally, for phase shifters with multiple phase states the phase shifts all refer to one reference phase state. In the following, four commonly used types of phase shifter topologies are reviewed: loaded-line phase shifters, switched network phase shifters, reflection type phase shifters and vector summation phase shifters.

1.5.2 Loaded-line Phase Shifter

The concept of the loaded-line phase shifters is to use the loads to change the electrical length of a fixed transmission line. And therefore, it is a transmission type phase shifter. The prototype is shown in Figure 1.2.

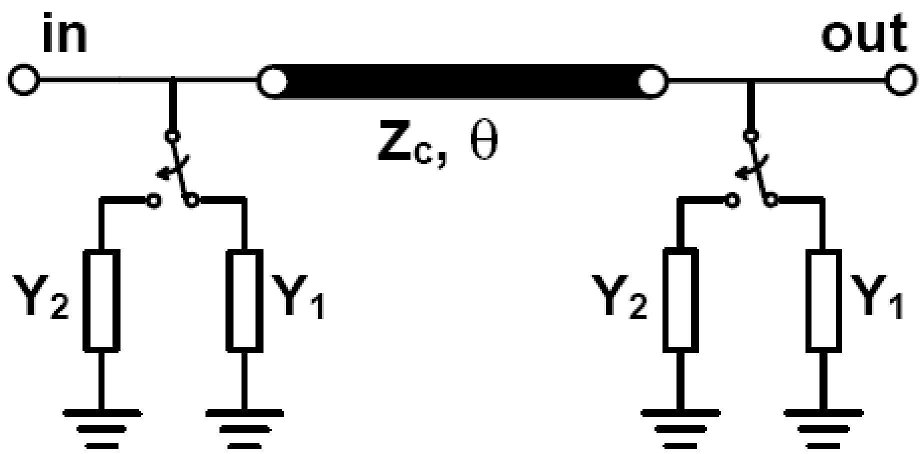


Figure 1.2 Loaded-line phase shifter prototype

The loads inserted at the two ends of the transmission line can be controlled digitally to change the electrical length of the centre line with impedance of Z_c and original electrical length of θ. In an analog phase shifter, the loads are controlled continuously. However, the perfect matching may not be always met. Here we are discussing a binary phase shifter with two possible load states: Y1 and Y2. When the loss of the loads is neglected, the insertion phase ϕ_i(i=1, 2) can be obtained from given equation

$$\phi_i = \arccos(\cos\theta - Y_i/Z_c \sin\theta) \qquad (1.2)$$

Therefore, when Y1 changes to Y2, the phase shift can be obtained from (2-1). Based on this prototype, the analysis for different loads and switching components can be done. Three different classes with nonzero and unequal loads (Class I), load-unload loads (Class II) and complex-conjugate loads (Class III) are concluded. It is found that the loaded line phase shifter has the largest bandwidth and perfect matching at the center frequency for Class III when the original electrical length of θ is around 90^0. The design parameters can be obtained from given equations.

$$Z_c = \cos(\Delta\phi/2)\, Z_0 \qquad (1.3)$$

$$Y_i = \pm\tan(\Delta\phi/2) Y_0 \qquad (1.4)$$

It is noted that this topology cannot be used for 180º phase shifters due to the extreme values from (2-4). When the design parameters are obtained, the return loss and phase error can be written as a function of the normalized frequency. Therefore, the return loss bandwidth (BW_{RL}) and phase error bandwidth (BW_{PE}) are related to the phase shift values. The relations for Class III loaded-line phase shifters with discrete loads are plotted in Figure 1.3 for phase shift ranges from 5^0 to 120^0.

From Figure 1.3, the overall bandwidth of the loaded-line phase shifter is mainly limited by the return loss when the required phase shift increases. For example, when the phase shift is 90^0, the 4^0 ($\pm2^0$) phase error bandwidth BW_{PE} is 35% while the 15 dB BW_{RL} is 14%, and the overall

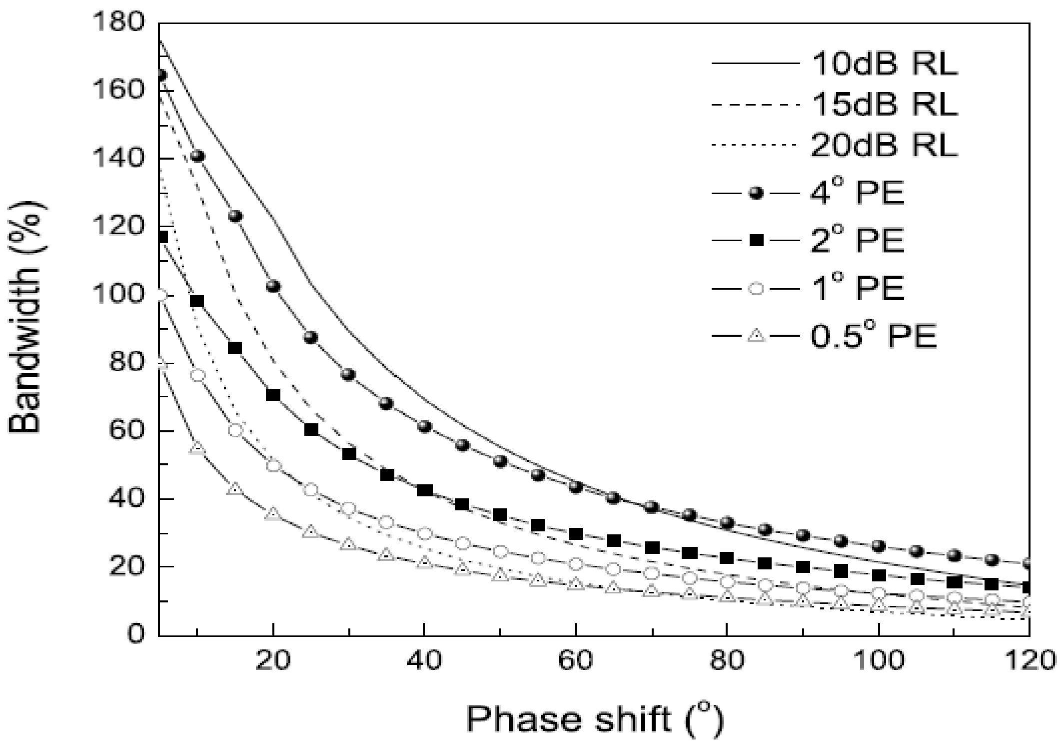


Figure 1.3 Bandwidth versus the phase shifts for different phase errors and return losses of the lumped element loaded Class III phase shifters.

bandwidth is limited to 14% when the required return loss is 15 dB. Besides high power applications, this prototype is also preferred for high frequency designs due to its simple structure.

1.5.3 Switched Network Phase Shifter

The switched network phase shifter is another type of transmission phase shifter. The development in this type of phase shifter is very active because the number of network combination is numerous.

Common switched network phase shifters switch between different pass bands of different networks to obtain the phase difference as shown in Figure 1.4.

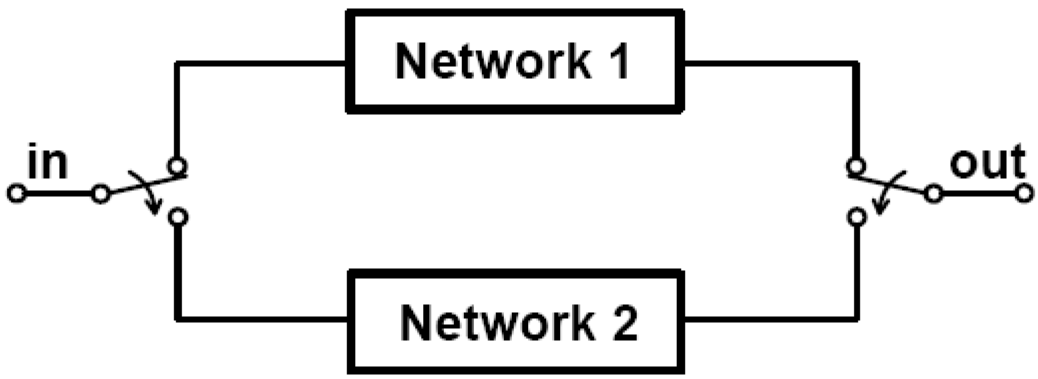


Figure 1.4 Schematic of the switched network phase shifters

Here, two popular used topologies are reviewed: the high-pass/low-pass phase shifter and the all-pass network phase shifter, which includes the Schiffman phase shifter.

1.5.4 High-pass/low-pass (HP/LP) phase shifter

The HP/LP phase shifter switches between high-pass filters and low-pass filters. A typical third order HP/LP phase shift bit is shown in Figure 1.5. The T-network is chosen for the HP filter and the π-network is chosen for the LP filter to minimize the number of inductors used.

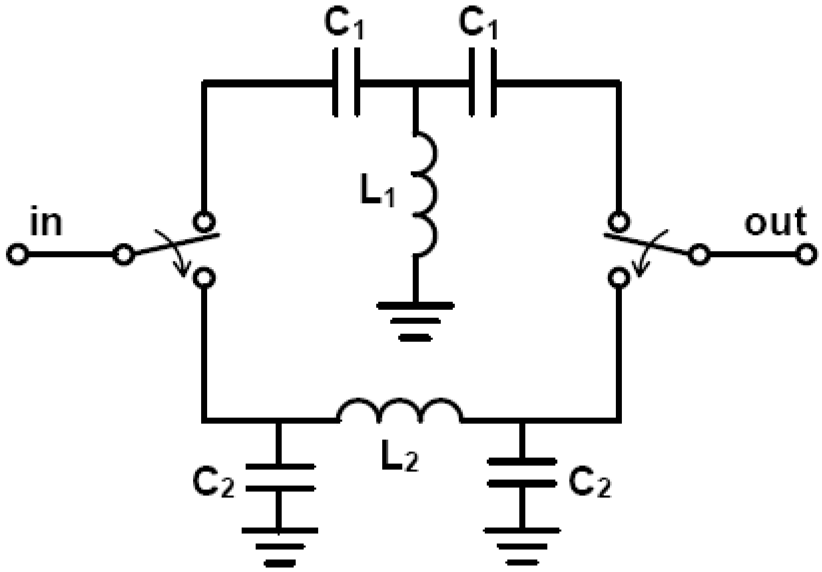


Figure 1.5 Third order high-pass/low-pass phase shifter.

Design parameters can be obtained from for the T-network and π-network for a given center frequency ω_0 and desired phase shift $\Delta\phi$ at ω_0. For smaller phase shifters, larger inductors are needed for the high-pass filter using this topology. The insertion loss is affected by the parasitic, the matching from the adjacent phase bits and the two SPDT switches used in one phase bit. For millimetre wave designs, semiconductor based SPDT switches suffer from a high insertion loss and poor isolation. The travelling wave concept can be adopted in the design of SPDT switches but the circuit size can be significant.

To minimize the circuit area and to extend the bandwidth, the parasitics of the switching components are absorbed rather than avoided in the FET-integrated designs. The ON state of the FET can be modeled by a small resistor and the OFF state of the FET can be modeled by a small capacitor as shown in Figure 1.6.

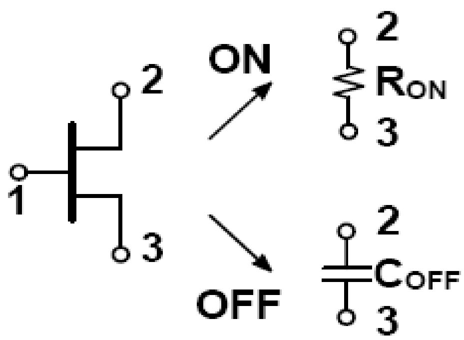


Figure 1.6 Modelled ON and OFF state of the FET.

This methodology is used for phase shifter implementation and a bandwidth of nearly an octave is achieved. When more stages of HP/LP networks are used, larger bandwidth can be realized.

This topology is also popular in microwave systems where large phase shifts are required. Examples using the HP/LP circuits to provide the required phase difference include the TX-RX switches with leakage cancelation and the broadband multi-port direct receiver.

1.5.5 All-pass network phase shifter

Another popular phase shifter switches between all-pass networks. In a single stage four-element all-pass network to achieve a bandwidth ratio of 1.6:1 is presented. The two topologies of the all-pass networks shown in Figure 1.7 share the same performance when the values satisfy following relation.

$$LC = 1/\omega_0^2, \quad L/C = Z_0^2 \qquad (1.5)$$

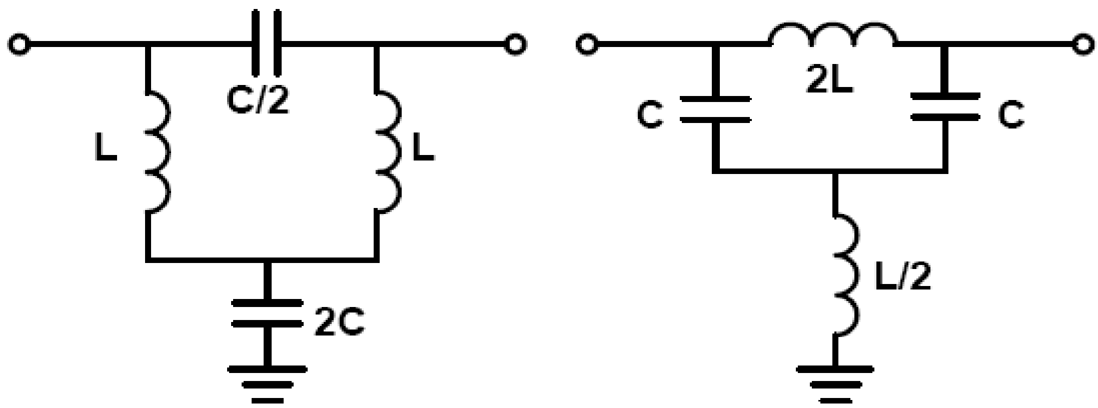


Figure 1.7 Two four-element all-pass networks.

Where ω_0 is the transition frequency where the insertion phase is -180^0 and Z_0 is the characteristic impedance of the system. The phase shift can be obtained by changing ω_0 of two all-pass networks and a phase shift peak will be obtained for each setting. The advantage of the all-pass network is that the insertion phase can be treated separately due to the very high return loss. Therefore, when two stages are cascaded, two peaks in the phase response can be obtained to get a phase shift ripple and the bandwidth can be significantly improved. For the two-section phase shifter, the bandwidth ratio increases to 4:1 for the same performance of the single-section phase shifter.

Besides the large bandwidth applications, the all-pass network is also used in compact integrated circuit designs. Like the HP/LP topologies, FETs can be integrated into the circuits to minimize the circuit area. The transition frequency can be shifted much higher than the designed frequency and therefore the parameter values are smaller.

Furthermore, the capacitors in both the HP/LP phase shifters and the lumped all-pass network phase shifters can be changed to varactors to realize continuous phase tuning.

1.5.6 The Schiffman phase shifter

The all-pass network based phase shifter can also be realized in distributed form. Schiffman presented a topology based on a carefully designed coupled line and a transmission line. The design theory is synthesized and a phase error of $\pm 2^0$ is achieved for a 90^0 phase shift over a bandwidth of 60%. However, it is assumed that the even- and odd-mode velocities of the RF signal are the same in the original design. Therefore, the topology was implemented in striplines. For microstrip lines, the Schiffman phase shifter needs to be modified to maintain the bandwidth.

1.5.7 Reflection Type Phase Shifter

The reflection type phase shifters consist of two parts: a 3-dB directional coupler and two tunable terminations. As shown in Fig. 2.7, the input signal is first split into two parts, and then reflected by the two loads. When the loads are the same, the voltage at the output port can be written as

$$V_{out} = j\Gamma V_{in} \qquad (1.6)$$

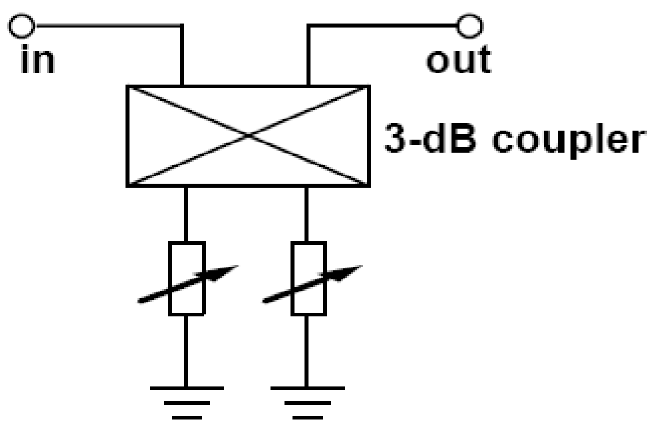


Figure 1.8 Reflection type phase shifter prototype.

Compared with the transmission phase used in the transmission type phase shifters, reflection type phase shifters transform the reflection coefficient phase to the transmission phase first, and the phase shift is calculated from (2-1). When the power is fully reflected from the loads, the insertion loss is minimized.

The bandwidth of the reflection type phase shifter is determined by both the coupler and the terminations. For example , broadband terminations are used, but the slot fin-line coupler limits the bandwidth to 15% only. The reported bandwidth using a branch line coupler is from 20% to 25% .it is claimed that the isolation and the return loss of the coupler influence the phase shifter performance, and ±3% to ±4% phase errors are achieved for a bandwidth ratio larger than 5 by using Lange couplers. Lange couplers can provide a large bandwidth and the bandwidth of the phase shifter is then limited by the terminations.

Because couplers are used, the circuit size increases especially for the digital designs . Therefore, to reduce the number of cascaded phase bits, continuous tuning is proposed to cover small phase shifts to reduce the insertion loss and circuit size. In[56], parallel series resonators are used as the reflective terminations and , cascaded reflection type phase shifters with a phase shift peak summation methodology are used to achieve an analog phase shift over 100% bandwidth.

The problem for broadband analog reflective phase shifters is achieving a large phase shift range. To achieve 360^o phase coverage, more stages are required which immediately degrades the performance. Research has been done to increase the phase coverage for a single section reflective phase shifter. However, the bandwidth is reduced. An impedance-transforming branch line coupler is used together with transformer integrated terminations and a full 360^0 phase shift is achieved. The circuit size is further reduced by implementing the coupler using lumped elements. As a tradeoff, the bandwidth reduces to 10% to 15%.

1.5.8 Vector Based Phase Shifter

The vector based phase shifter first generates base vectors, and then combines the vectors with different amplitude weights to achieve different phase shifts. The concept is shown in Figure 1.9.

The commonly used phase difference between the vectors can be around 90^0 where four vectors are needed, or around 120^0 where three vectors are needed . The 90^0 phase difference is mainly realized using combiners, directional couplers, polyphase filters, and all-pass networks , and the 120^0 phase difference is mainly realized using HP/LP filters and all-pass networks.

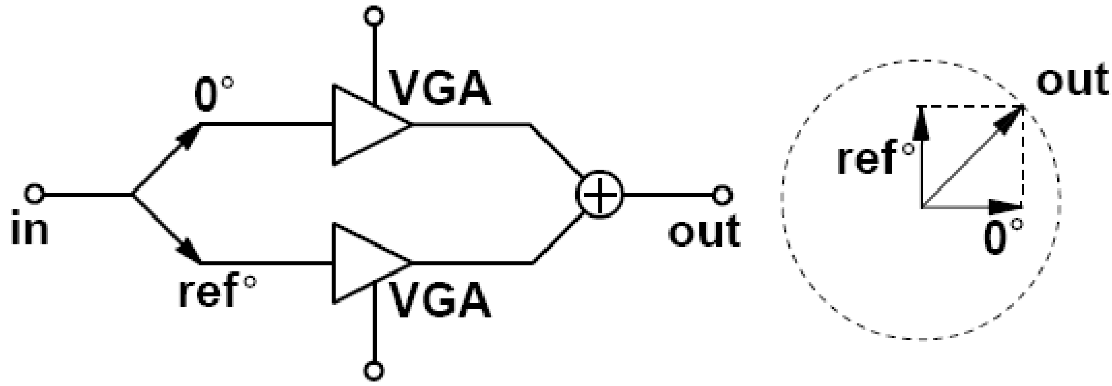


Figure 1.9 Basic concept of the vector based phase shifters.

The main advantage of the vector based phase shifter is the potential for a small circuit size. For most vectors generated by polyphase filters, the amplitude variation is a concern. Usually the phase bandwidth is much larger than the return loss bandwidth. The phase shift response is quite smooth over a multi-octave band while the return loss is around an octave or less.

Another concern is the power consumption, power handling capability, and linearity of this type of phase shifters. Compared to the conventional phase shifter designs, the vector based phase shifter uses VGAs and DACs to control the weight of the vectors, which consume extra power of the circuits.

Chapter 2

Topologies for three bit phase shifter design

In most conventional phase shifter design methods, when the center frequency and the required phase shift at the center frequency are fixed, the performance is then fixed. There is no room for tradeoffs to obtain an optimum performance for a given bandwidth. In the previous chapter, improvements are possible in terms of return loss and phase error. However, more substantial tradeoffs between the performance and the bandwidth are needed, so that designers may synthesize phase shifters according to the requirement, similar to the expert filter design methods.

This chapter introduces more tradeoffs between the performance and the bandwidth of several novel phase shifter topologies, where the optimum phase error for a given bandwidth is realized.

Basic block diagram of three bit phase shifter consists of four blocks each representing individual bit. Smallest possible phase shift is 22.5^0 and largest possible phase shift is 157.5^0. It is clear from figure 2.1.

Figure 2.1 Block diagram of the four-bit phase shifter.

2.1 Phase Slope Alignment Using Quarter Wavelength Stubs

In Figure 2.2, the topology for a phase shift bit consisting of a transmission line branch, a bandpass filter branch and two SPDT switches to switch between the two branches is shown. In the analysis that follows, the insertion phase of the transmission line branch at the center frequency ω_c is chosen as the sum of the desired phase shift and the insertion phase of the bandpass filter at $\omega_{c..}$

$$\Phi_{TL}(\omega_c) = \Delta\phi(\omega_c) + \Phi_{BPF}(\omega_c) \qquad (2.1)$$

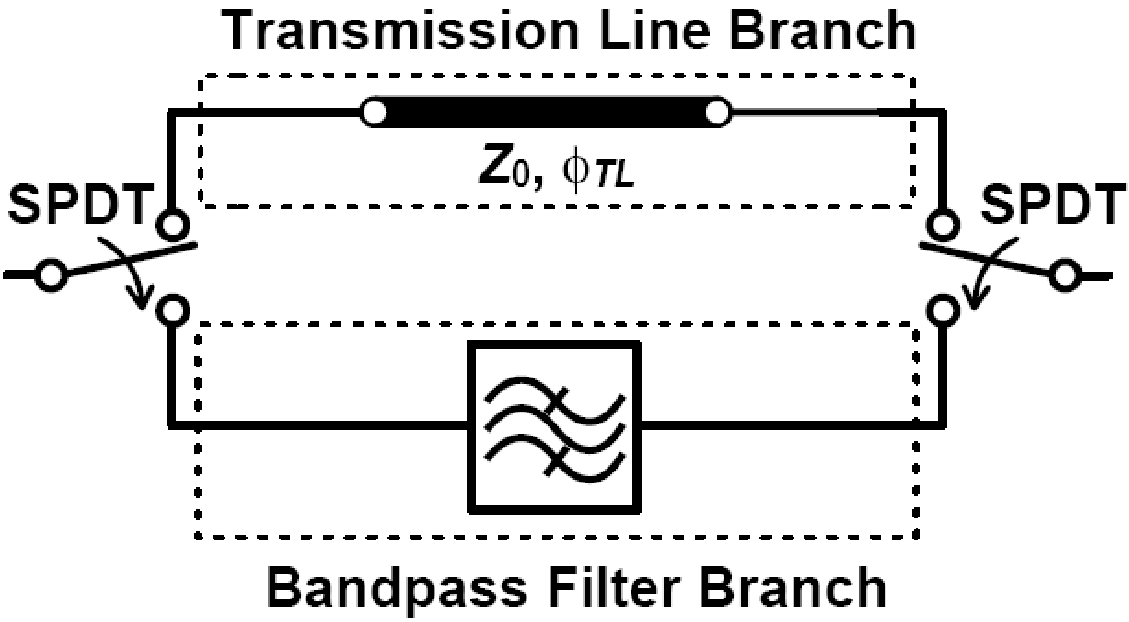


Figure 2.2 topology for phase shifter bit.

In the following, bandpass filters with one, two and three transmission poles for the bandpass filter branch are analyzed. The bandwidth of the phase shifter increases with the number of poles. The return loss, phase error and the achievable phase shift for each case are discussed in detail.

2.1.1 Topology 1: BPF with one pole

Figure 2.3 shows the topology of the single pole bandpass filter which consists of a λ/4 short circuited shunt stub resonant at ω_c. At an arbitrary frequency, the insertion phase of the band-pass filter is given by

$$\phi_{BPF}\left(\bar{\omega}\right) = \arctan\left[\frac{1}{2\bar{Z}_1}\tan\left(\frac{\pi}{2}-\frac{\pi}{2}\bar{\omega}\right)\right] \qquad (2.2)$$

Where $\bar{Z}_1 = Z_1/Z_0$ is the normalized impedance and $\bar{\omega} = \omega/\omega_c$ Is the normalized frequency.

The topology for single pole bandpass filter using quarter wavelength transmission line is give in figure as follows.

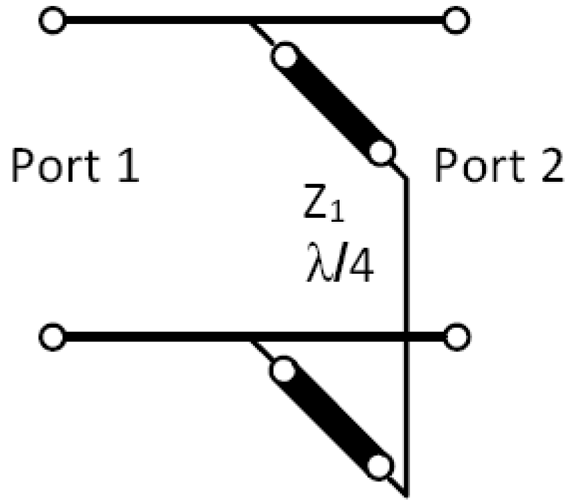


Figure 2.3 Topology 1: Single pole bandpass filter.

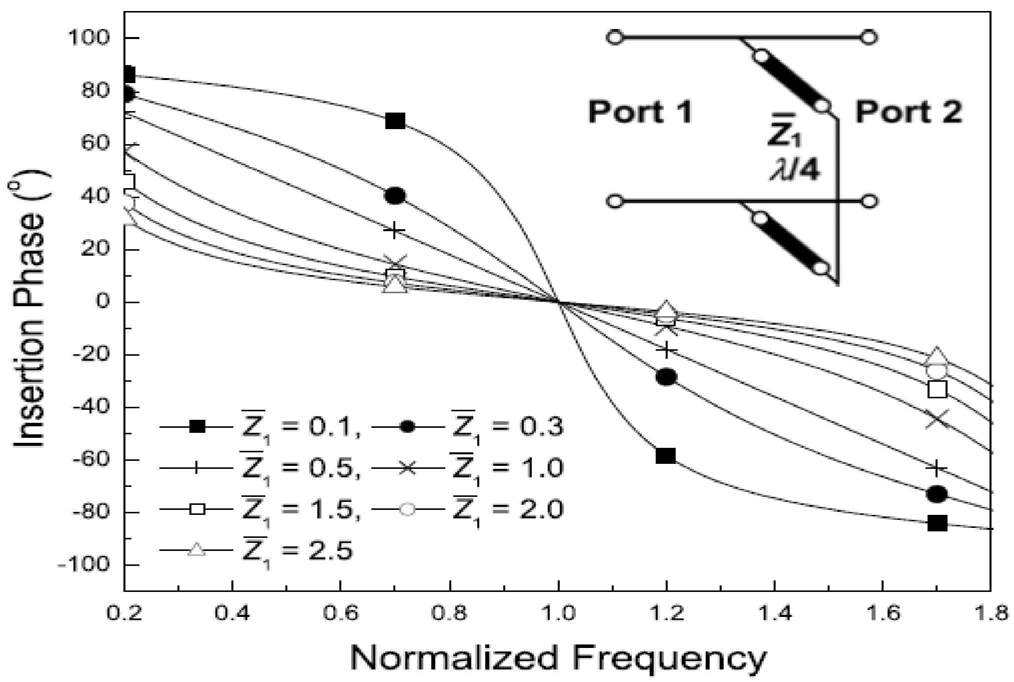


Figure 2.4 Insertion phase versus normalized frequency of a grounded shunt stub.

In Figure 2.4 the insertion phase is shown versus the normalized frequency for different normalized impedance . It is observed that when normalized impedance is changed, the phase slope of the branch can be controlled without changing the insertion phase at the center frequency.

In Fig. 2.5(a) and (b) two possible phase shift responses Δϕ are shown. In these figures four characteristic frequencies are defined.

$\overline{\omega}_L$ and $\overline{\omega}_H$ Are the lower bound and upper bound of the desired bandwidth.

$\overline{\omega}_{r1}$ and $\overline{\omega}_{r2}$ are the frequencies where the phase slope of TL branch and BPF branch are equal.

The maximum phase error in the desired bandwidth is denoted by *PE* and the phase shift within the desired bandwidth satisfies

$$\Delta\phi(\overline{\omega}_C) - PE \le \Delta\phi(\overline{\omega}) \le \Delta\phi(\overline{\omega}_C) + PE \qquad (2.3)$$

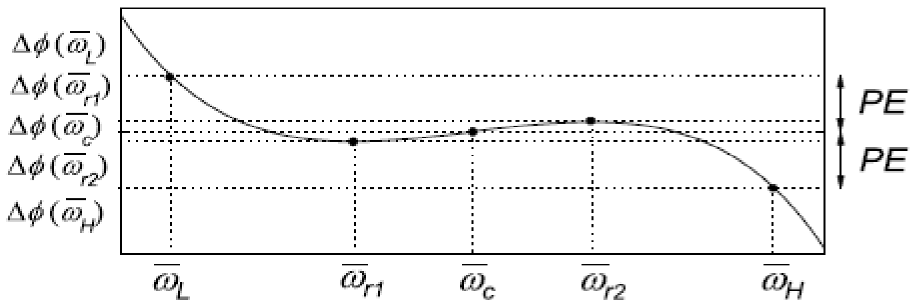
As shown in figure 2.5(c), the optimum phase error PE_{opt} is achieved when

$$\Delta\phi(\overline{\omega}_L) = \Delta\phi(\overline{\omega}_{r2}) \text{ and } \Delta\phi(\overline{\omega}_H) = \Delta\phi(\overline{\omega}_{r1}) \qquad (2.4)$$

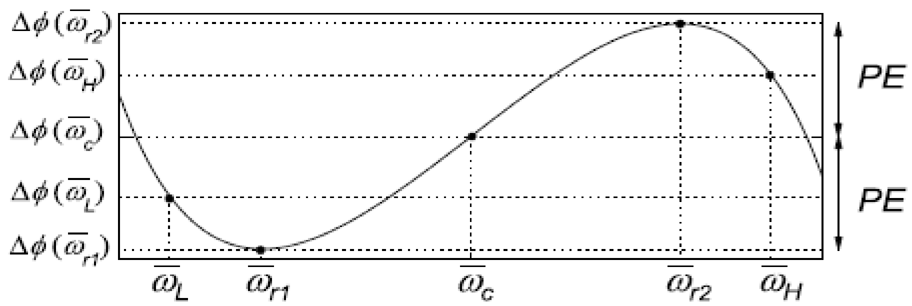
Note that if one of these conditions is satisfied, the other is automatically satisfied too. Because the phase shift is anti-symmetric around central frequency , only frequencies below central frequency are considered in the following discussion. To find the optimum phase error PE_{opt}, w_{r1} must be determined first. The phase shift is given by

$$\Delta\phi(\overline{\omega}) = \phi_{BPF}(\overline{\omega}) - \phi_{TL}(\overline{\omega}) \qquad (2.5)$$

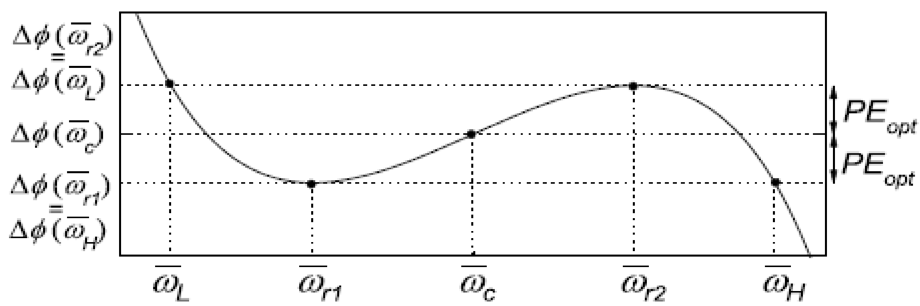
Typical phase shift responses are given as follows.



(a)



(b)



(c)

Figure 2. 5. (a), (b) Typical phase shift responses (c) optimum phase shift response.

Differentiating equation 2.5 with respect to normalized frequency and equating to zero to obtain optimum case, we get

$$\frac{d\Delta\phi(\overline{\omega})}{d\overline{\omega}} = \frac{d\phi_{BPF}(\overline{\omega})}{d\overline{\omega}} - \frac{d\phi_{TL}(\overline{\omega})}{d\overline{\omega}} = 0 \qquad (2.6)$$

The phase slope of bandpass filter is found using equation 2.2

$$\frac{d\phi_{BPF}(\overline{\omega})}{d\overline{\omega}} = -\frac{\pi\left[1+\tan^2\left(\frac{\pi\overline{\omega}}{2}\right)\right]\overline{Z}_1}{1+\overline{Z}_1^2\tan^2\left(\frac{\pi\overline{\omega}}{2}\right)} \qquad (2.7)$$

And phase slope of transmission line is constant

$$\frac{d\phi_{TL}(\overline{\omega})}{d\overline{\omega}} = -\Delta\phi(\overline{\omega}_c) \qquad (2.8)$$

Substituting (2.7) and (2.8) in (2.6), we can solve for $\overline{\omega}_{r1}$:

$$\overline{\omega}_{r1} = \frac{2}{\pi}\arctan\sqrt{\frac{\pi\overline{Z}_1-\Delta\phi(\overline{\omega}_c)}{\overline{Z}_1\left[4\overline{Z}_1\Delta\phi(\overline{\omega}_c)-\pi\right]}} \qquad (2.9)$$

With the condition:

$$\frac{\pi\overline{Z}_1-\Delta\phi(\overline{\omega}_c)}{\overline{Z}_1\left[4\overline{Z}_1\Delta\phi(\overline{\omega}_c)-\pi\right]} \geq 0 \qquad (2.10)$$

By substituting (2-9) into (4-5), the phase shift at $\overline{\omega}_{r1}$: is

$$\Delta\phi(\bar{\omega}_{r1}) = \arctan\frac{\sqrt{\bar{Z}_1\left[4\bar{Z}_1\Delta\phi(\bar{\omega}_c)-\pi\right]}}{2\bar{Z}_1\sqrt{\pi\bar{Z}_1-\Delta\phi(\bar{\omega}_c)}}+\frac{2\Delta\phi(\bar{\omega}_c)}{\pi}\sqrt{\frac{\pi\bar{Z}_1-\Delta\phi(\bar{\omega}_c)}{\bar{Z}_1\left[4\bar{Z}_1\Delta\phi(\bar{\omega}_c)-\pi\right]}} \tag{2.11}$$

The optimum phase error PE_{opt} is then obtained by numerically solving

$$\Delta\phi(\bar{\omega}_L)-\Delta\phi(\bar{\omega}_c)=\Delta\phi(\bar{\omega}_c)-\Delta\phi(\bar{\omega}_{r1}) \tag{2.12}$$

In Fig. 4.5 PE_{opt} and characteristic impedance are shown versus the phase shift for bandwidths of 30%, 50%, 67% (one octave) and 100% (3:1). When the bandwidth and phase shift are specified, the characteristic impedance is obtained from the left axis of Fig. 2.6, and the corresponding optimum phase error is found from the right axis.

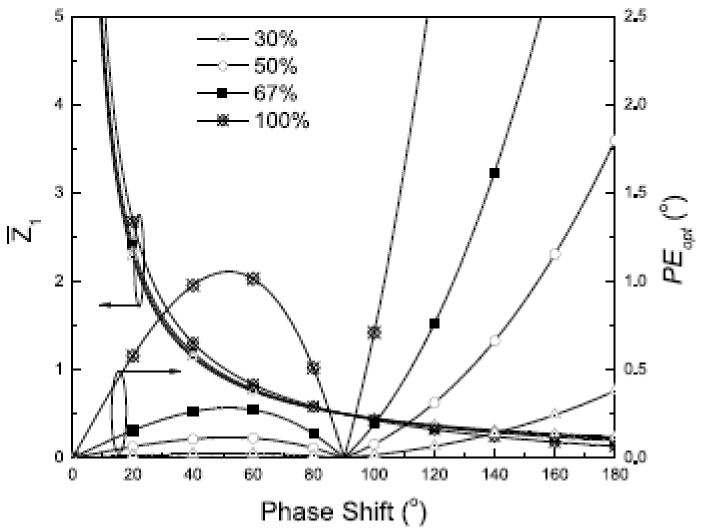


Figure 2.6 PE_{opt} and $\bar{Z}_1$ of the grounded shunt stub versus the phase shift.

Note that so far only the bandwidth for the phase error is considered. However, for phase shifter design the bandwidth of a specified return loss should also be considered. Because the return loss bandwidth of the TL branch ideally is infinite, the bandwidth is limited by the return loss bandwidth of the BPF branch. In the case of the grounded shunt $\lambda/4$ stub, the return loss is determined by the characteristic impedance. For an given impedance and a desired return loss RL, the bandwidth is given by:

$$BW_{RL} = 2\left(1 - \frac{2}{\pi}\arctan\frac{\sqrt{10^{RL/10}-1}}{2\overline{Z}_1}\right)\times 100\% \qquad (2.13)$$

In Fig. 2.7 the return loss bandwidth BWRL is shown versus the impedance $\overline{Z}_1$ For return losses of 10, 15, and 20 dB. Because the return loss and the phase shift are both determined by $\overline{Z}_1$ Their relation is shown in figure 2.8

As an example, consider a phase shifter with an octave bandwidth (67%) and a required return loss of 15 dB. From Figure 2.8 it is found that only phase shifts up to 30^0 can be achieved. The associated phase error is found in Figure 2.6.

In conclusion, when large bandwidths are required with return losses of 15 dB or larger, the achievable phase shift with the single grounded shunt λ/4 stub is limited to small phase shifts. Therefore, other topologies must be considered to improve the bandwidth of the return loss of the BPF branch. The single pole BPF is used for the 22.5° phase bit of the realized four-bit phase shifter discussed later.

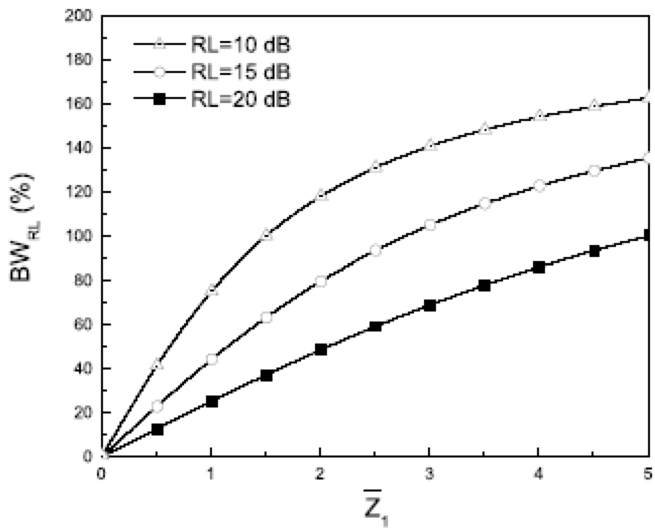


Figure 2.7 return loss bandwidth versus $\overline{Z}_1$

The achievable phase shift with optimum phase error can be shown as given in following figure.

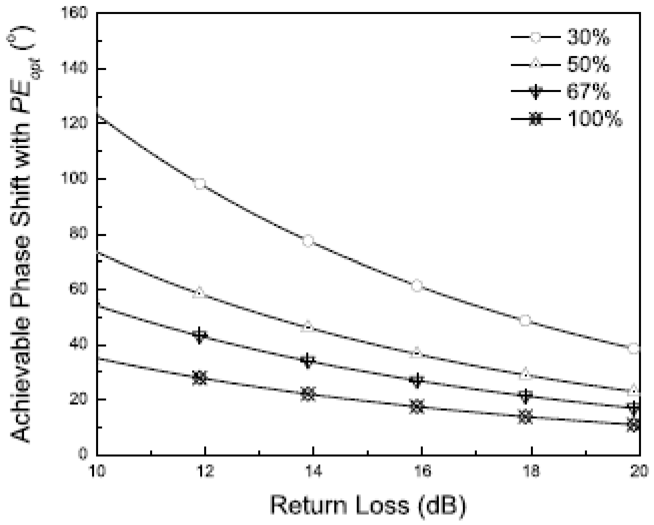


Figure 2.8 Relation between the return loss and the achievable phase shift with optimum phase error for different bandwidths.

2.1.2 Topology 2: BPF with two poles

To improve the bandwidth of the return loss for the larger phase shifting bits, transmission poles are added to the single stub bandpass filter. To achieve two poles a series $\lambda/4$ transmission line is inserted between two short circuited shunt $\lambda/4$ stubs as shown in Figure 2.9.

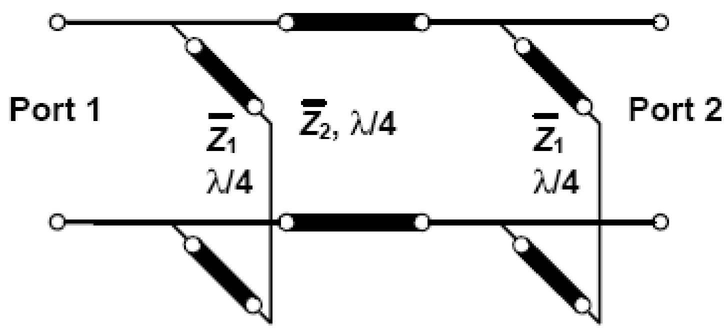


Figure 2.9 topology 2: double pole bandpass filter.

The insertion phase of double pole bandpass filter is given by

$$\phi_{BPF}(\overline{\omega}) = \arctan\left[\frac{2\overline{Z}_1\overline{Z}_2 + \Omega\overline{Z}_2^2 - \Omega^2\overline{Z}_2^2\left(\overline{Z}_2^2+1\right)}{2\Omega\overline{Z}_1\overline{Z}_2\left(\overline{Z}_1+\overline{Z}_2\right)}\right] \tag{2.14}$$

where $\Omega = \tan(0.5\pi\overline{\omega})$
when $\overline{Z}_2 < 1$, Two transmission poles are found:

$$\overline{\omega}_{1,2} = 1 \pm \left[1 - \frac{2}{\pi}\arctan\sqrt{\frac{\overline{Z}_2}{\overline{Z}_1^2}\cdot\frac{2\overline{Z}_1+\overline{Z}_2}{1-\overline{Z}_2^2}}\right] \tag{2.15}$$

Note that when $\overline{Z}_2 = 1$, Only one transmission pole is found at $\overline{\omega}_c$ And when $\overline{Z}_2 > 1$, No trans--mission pole are found.

Optimum phase error is found again by solving equation (2.12) and the solution is a set of com--bination of $\overline{Z}_1$ and $\overline{Z}_2$. Therefore, the combination of $\overline{Z}_1$ and $\overline{Z}_2$. That satisfies the ret--urn loss requirement within the required bandwidth must be determined.

There are two extreme cases that meet the return loss requirement. This is illustrated in Fig. 2.10 for a return loss requirement of 15 dB. In case 1, the return loss requirement is met at the band edges $\overline{\omega}_L$ and $\overline{\omega}_H$ and the requirement is exceeded at $\overline{\omega}_c$. in case 2 requirement is met at $\overline{\omega}_c$ And exceeded at $\overline{\omega}_L$ and $\overline{\omega}_H$. Each case has an optimum phase error and the case with the smaller optimum phase error should be selected.

As a further illustration, in Fig. 4.10 the optimum phase error is shown versus the required return loss for an octave bandwidth of 22.5°, 45° and 90° phase bits. For all three phase bits, case 1 and case 2 intersect at the point when the return loss at the band edges is equal to the return loss at central frequency. The inset of Fig. 2.11 shows that in case 1 there can be zero, one or two poles when characteristic impedance is larger than, equal to, or smaller than 1 respectively.

For the three phase shifts, it is observed that case 1 provides the smaller optimum phase error. For a return loss requirement of 15 dB the optimum phase error for the 22.5° and 45° phase bit is 0.24° and 0.55°, respectively. It is also observed that it is not possible to achieve a return loss of 15 dB for the 90° phase bit.

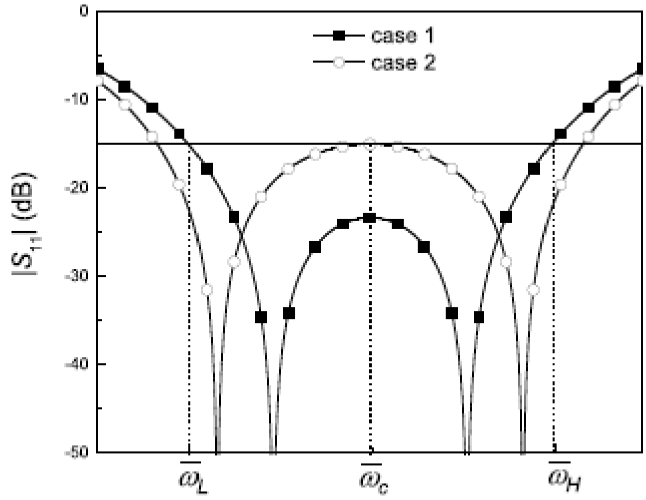


Figure 2.10 return loss of BPF with two poles

Case 1: return loss requirement is met at $\bar{\omega}_L$ and $\bar{\omega}_H$.

Case 2: return loss requirement is met at $\bar{\omega}_c$.

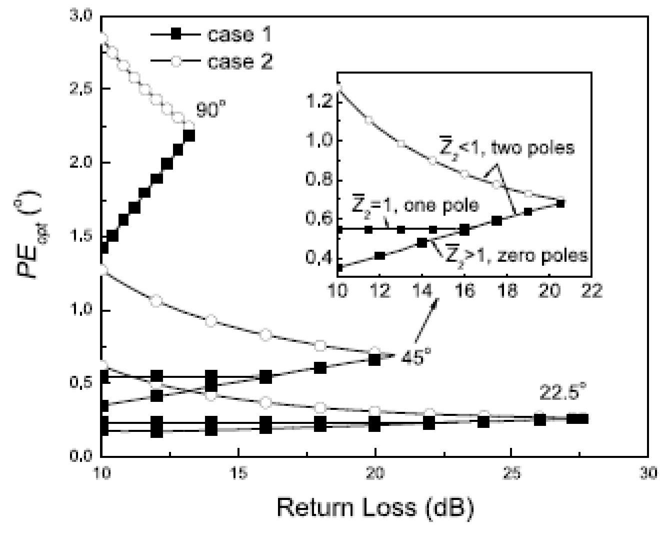


Figure 2.10 Two cases of *PEopt* versus return loss for 22.5º, 45º and 45º phase shifts in an octave bandwidth.

Case 1: return loss requirement is met at $\bar{\omega}_L$ and $\bar{\omega}_H$.

Case 2: return loss requirement is met at $\bar{\omega}_c$.

For the three phase shifts, it is observed that case 1 provides the smaller optimum phase error. For a return loss requirement of 15 dB the optimum phase error for the 22.5° and 45° phase bit is 0.24° and 0.55°, respectively. It is also observed that it is not possible to achieve a return loss of 15 dB for the 90° phase bit.

In short procedure to calculate $\overline{Z}_1$ and $\overline{Z}_2$ is as follows

1. set the requirements for the frequencies $\overline{\omega}_L$ and $\overline{\omega}_H$, Return loss RL and phase shift $\Delta\phi(\overline{\omega}_c)$

2. use equation (2.8) and (2.14) in (2.6) to find the expression for PE_{opt} with $\overline{Z}_1$ and $\overline{Z}_2$ as vari- -able

3. substitute $\overline{\omega}_L$ and $\overline{\omega}_c$ in the expression of $|S_{11}|$ Of the BPF branch.

4. using the expressions found in step 2 and step 3, determine $\overline{Z}_1$ and $\overline{Z}_2$ for the two cases

And select the case with the smallest PE_{opt}

The above procedure is applied to determine the relation between bandwidth and achievable phase shift for the three return loss cases of 10, 15, and 20 dB. The results are shown in Fig. 2.11. It is observed that the achievable phase shift decreases when the required return loss increases.

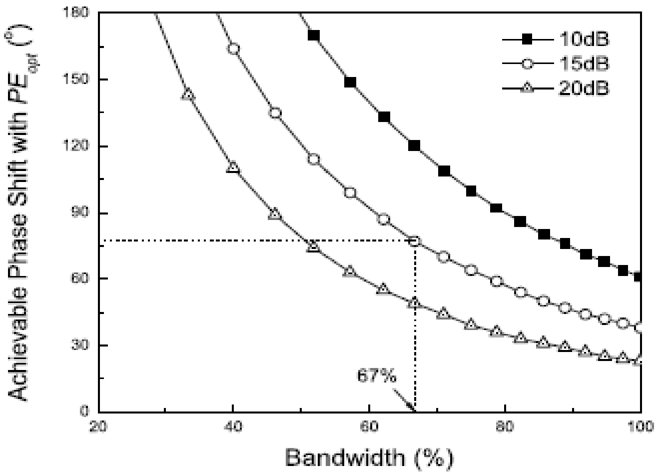


Figure 2.11 Relation between the achievable phase shift with *PEopt* and bandwidth for different return losses.

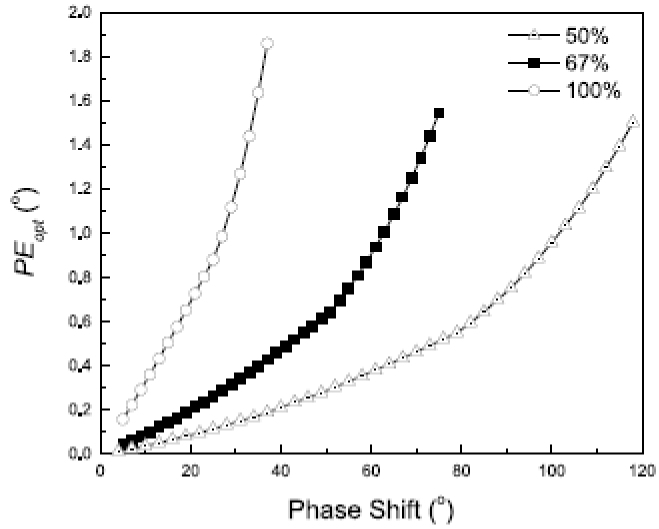


Figure 2.12 *PEopt* versus phase shift for a bandwidth of 50%, 67%, and 100% with a return loss of 15 dB.

The specific case of an octave bandwidth is also indicated in the figure. For a return loss of 15 dB the maximum achievable phase shift is 77°.

For a return loss requirement of 15 dB the relation between the optimum phase error and the required phase shift is shown in Fig. 4.12 for bandwidths of 50%, 67%, and 100%. Note that all three traces end at the point where the return loss requirement can no longer be met. For the design of an octave bandwidth four bit phase shifter, as discussed later, the topology with two poles is used for the 45° phase bit. As observed from Fig. 4.10, the topology with two poles still can not achieve a return loss of 15 dB for the 90° and 180° phase bits with an octave bandwidth.

2.2 Phase Slope Alignment Using LC Resonators

Resonators employing inductors and capacitors can also be used to control the phase slope. Fig. 2.13 shows series and parallel LC resonators and their combination. The combination of one series and two shunt parallel LC resonators shown in Fig. 2.13(c) is actually a BPF. Both of the two types of resonators can be used to control the phase slope. Therefore, the insertion phase slope of the BPF can be controlled to achieve the best alignment with the insertion phase of the reference state over a large bandwidth.

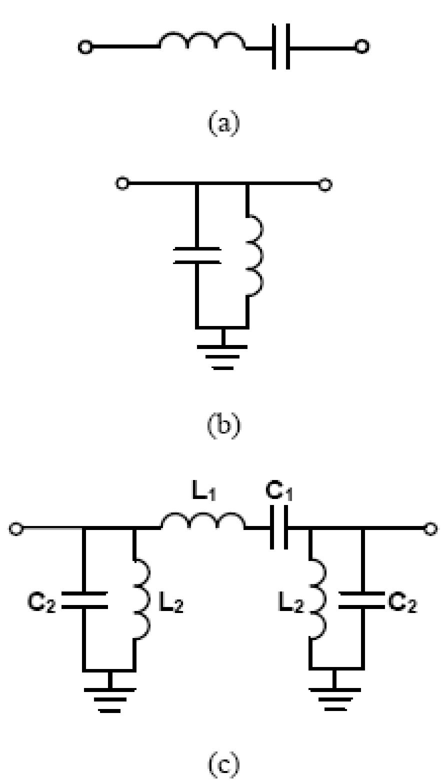


Fig. 2.13. The (a) series (b) parallel LC resonators and their (c) combination.

How to control the phase slope using a single LC resonator cab be described as given below. For a series LC resonator resonating at ω_0 , the *S*-parameters are found as

$$S_{11} = \frac{j\overline{X}\left(\overline{\omega}-\overline{\omega}^{-1}\right)}{2+j\overline{X}\left(\overline{\omega}-\overline{\omega}^{-1}\right)}$$
$$S_{21} = \frac{2}{2+j\overline{X}\left(\overline{\omega}-\overline{\omega}^{-1}\right)} \qquad (2.18)$$

where $\overline{X} = \omega_0 L / Z_0 = 1/\omega_0 C Z_0$ and $\overline{\omega} = \omega/\omega_0$.

For a shunt parallel LC resonator resonating at ω_0, the *S*-parameters are found as

$$S_{11} = \frac{-j\overline{B}\left(\overline{\omega} - \overline{\omega}^{-1}\right)}{2 + j\overline{B}\left(\overline{\omega} - \overline{\omega}^{-1}\right)}$$
$$S_{21} = \frac{2}{2 + j\overline{B}\left(\overline{\omega} - \overline{\omega}^{-1}\right)} \qquad (2.19)$$

where $\overline{B} = \omega_0 C Z_0 = Z_0 / \omega_0 L$.

Broadband phase shifter topology given below is used to implement 90 degree phase bit.

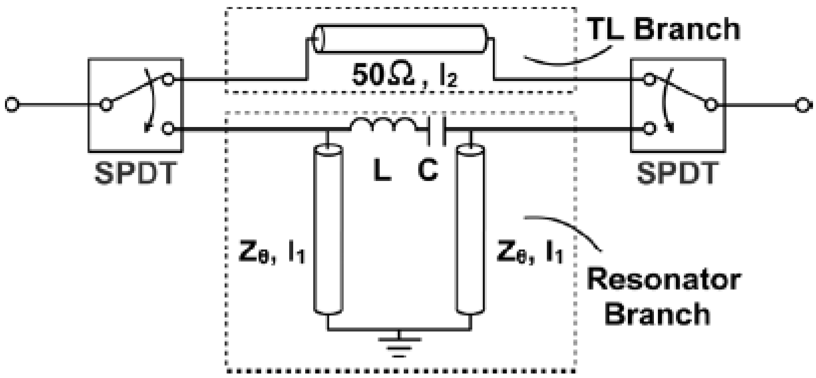


Figure 2.14 Broadband phase shifter topology.

Chapter 3

Simulation results

3.1 Simulated results for 30% bandwidth phase shifter

Both the branches of phase shifter bit are simulated for different phase shift bits and their results are plotted as give below.

3.1.1 Results for 22.5 degree phase bit

First branch of 22.5 phase bit is a transmission line providing a phase angle of 22.5 degree. Its s -parameter results are as given below.

3.1.1(a) TL branch

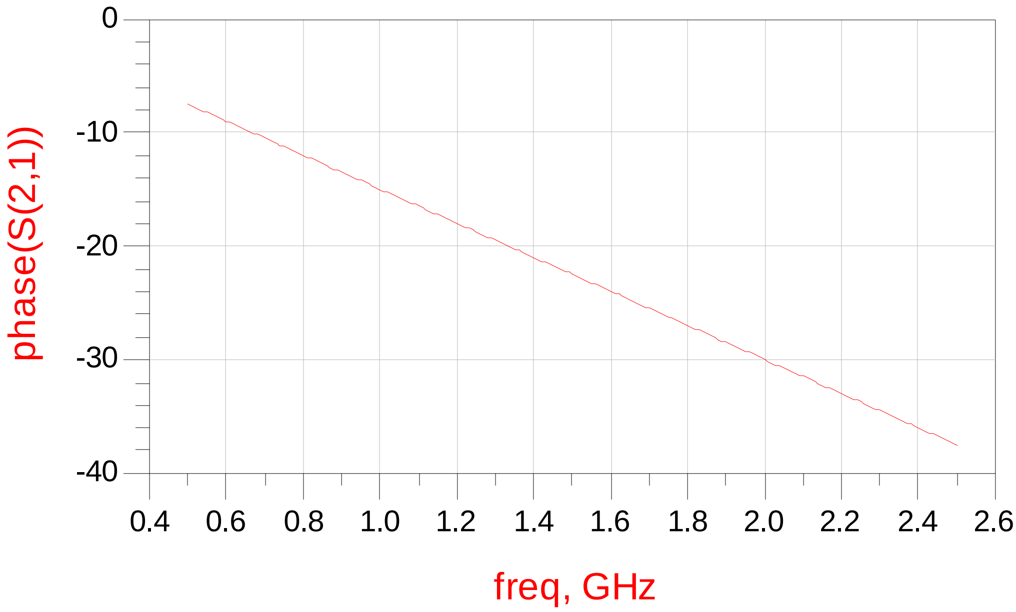


Figure 3.1 S(2,1) parameter of 22.5 degree TL branch.

Above figure show the phase of s(2,1) parameter variation with frequency. It is clear that at central frequency 1.5 GHz, the phase is 22.5 degree. To get the same phase in frequency range of bandwidth the phase slope alignment is used, which is provided by BPF branch.

3.1.1(b) BPF branch

Simulation results of BPF branch of 22.5 degree phase bit are as given below.

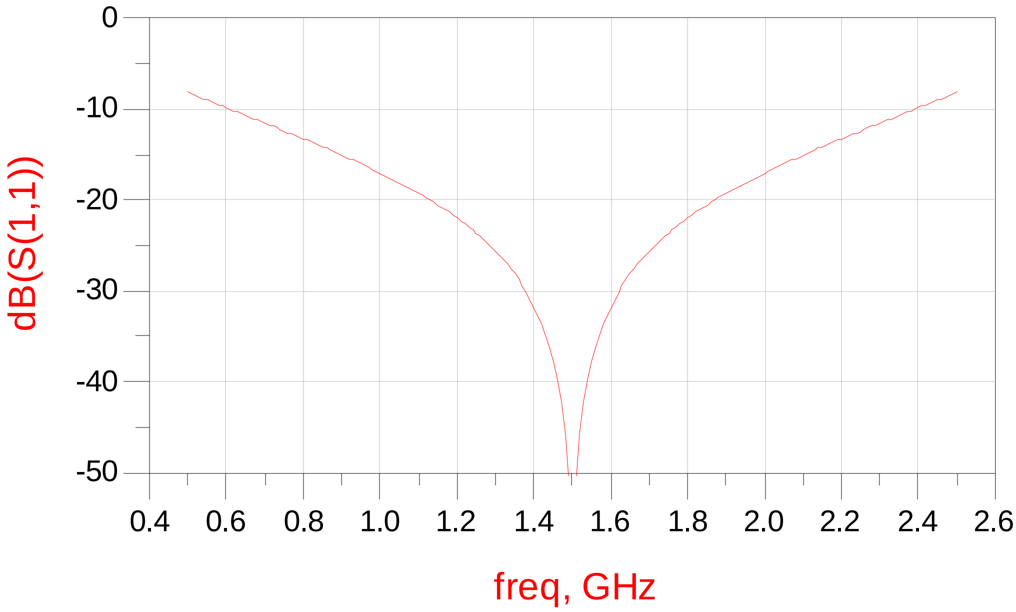


Figure 3.2 S(1,1) parameter of BPF branch versus frequency.

The s(1,1) parameter represents return loss due to BPF branch. From the figure 3.2 it is clear that at central frequency 1.5GHz, return loss is below –50 dB. At the corner frequencies of bandwidth the **return loss** is **–24.56** dB.

Figure 3.3 represents S(2,1) parameter of BPF branch of 22.5 degree phase bit. It is clear from the figure that at central frequency phase provided by BPF branch is zero and positive for frequencies below central frequency and negative for frequencies above central frequency. In this way the phase slope of TL branch is aligned by BPF branch.

Figure 3.5 shows the net phase of 22.5 degree phase bit. From figure it is clear that at central frequency net phase is 22.5 degree and at corner frequencies net phase is 22.52 and 22.48 degrees. Hence **phase error** of 22.5 degree phase bit is **0.02** degree.

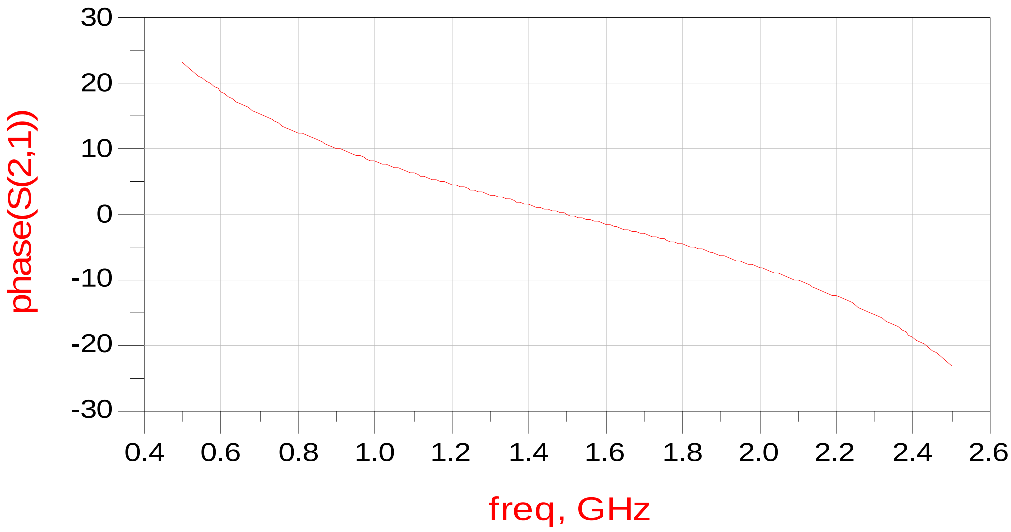


Figure 3.3 S(2,1) parameter of BPF branch of 22.5 degree phase bit

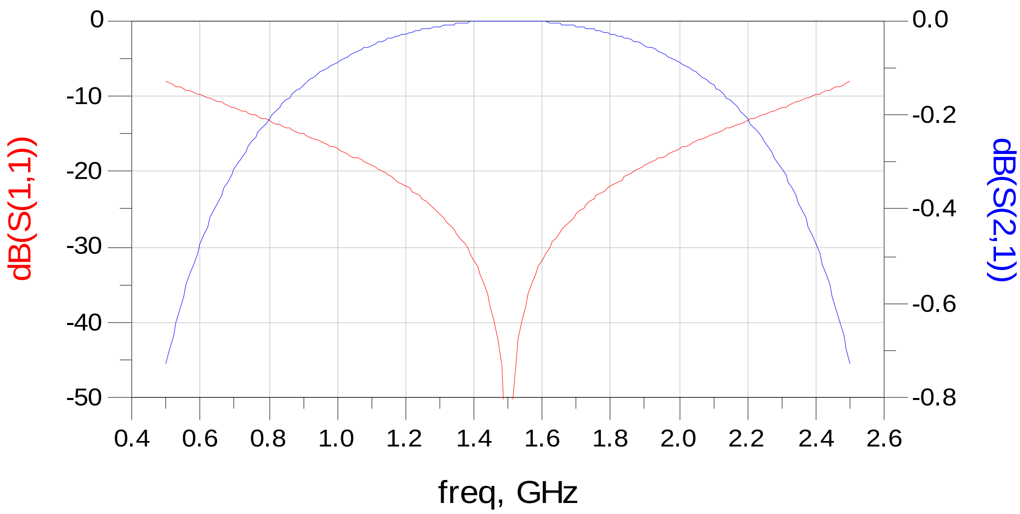


Figure 3.4 S(1,1) and S(2,1) magnitude versus frequency

3.1.1(c) Net phase of 22.5 phase bit

The net phase of 22.5 degree phase bit is as given in figure 3.5

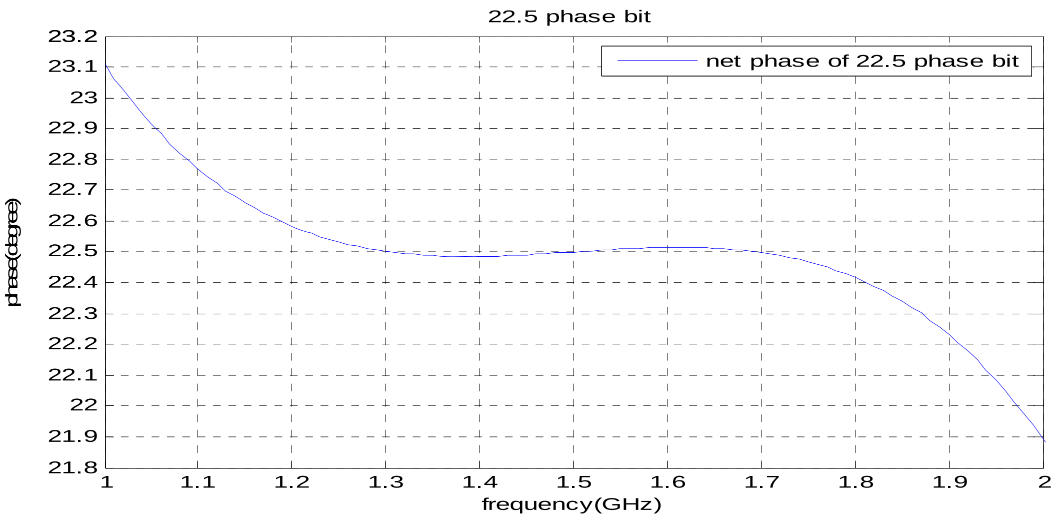


Figure 3.5 net phase versus frequency plot of 22.5 degree phase bit

From above figure it is clear that at central frequency 1.5 GHz the net phase is 22.5 degree and at corner frequencies it is 22.52 and 22.48. Hence the **phase error** of 22.5 degree phase bit is **0.02 degree**.

3.1.2 results for 45 degree phase bit

First branch of 45 phase bit is a transmission line providing a phase angle of 45 degree. Its s-parameter results are as given below.

3.1.2(a) TL branch

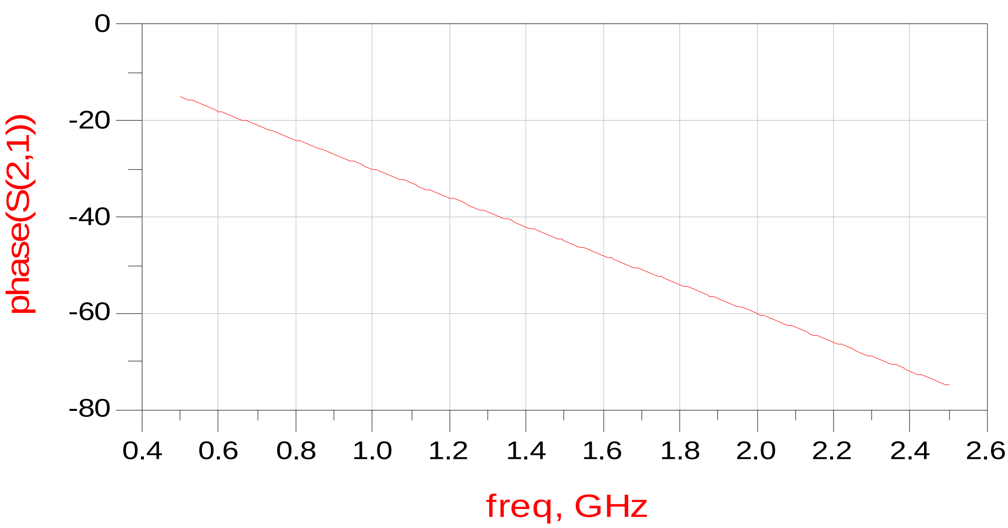


Figure 3.6 S(2,1) parameter of TL branch of 45 degree phase bit

figure 3.6 show the phase of s(2,1) parameter variation with frequency. It is clear that at central frequency 1.5 GHz, the phase is 45 degree. To get the same phase in frequency range of bandwidth the phase slope alignment is used, which is provided by BPF branch.

3.1.2(b) BPF branch

Simulation results of BPF branch of 45 degree phase bit are as given below.

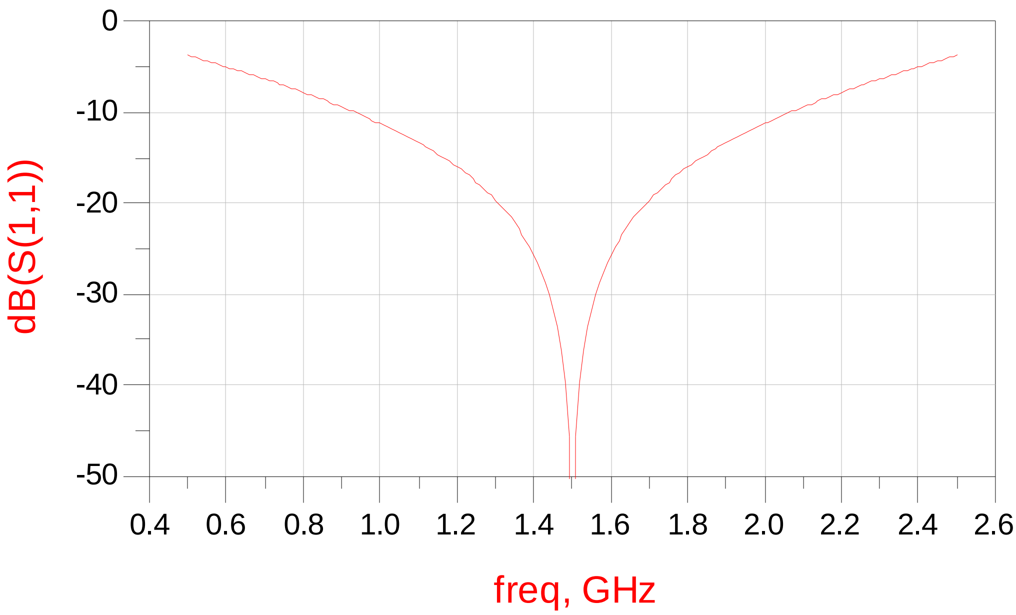


Figure 3.7 S(1,1) parameter of BPF branch versus frequency.

The s(1,1) parameter represents return loss due to BPF branch. From the figure 3.7 it is clear that at central frequency 1.5GHz, return loss is below –50 dB. At the corner frequencies of bandwidth the **return loss –18.52** dB.

Figure 3.8 represents S(2,1) parameter of BPF branch of 45 degree phase bit. It is clear from the figure that at central frequency phase provided by BPF branch is zero and positive for frequencies below central frequency and negative for frequencies above central frequency. In this way the phase slope of TL branch is aligned by BPF branch.

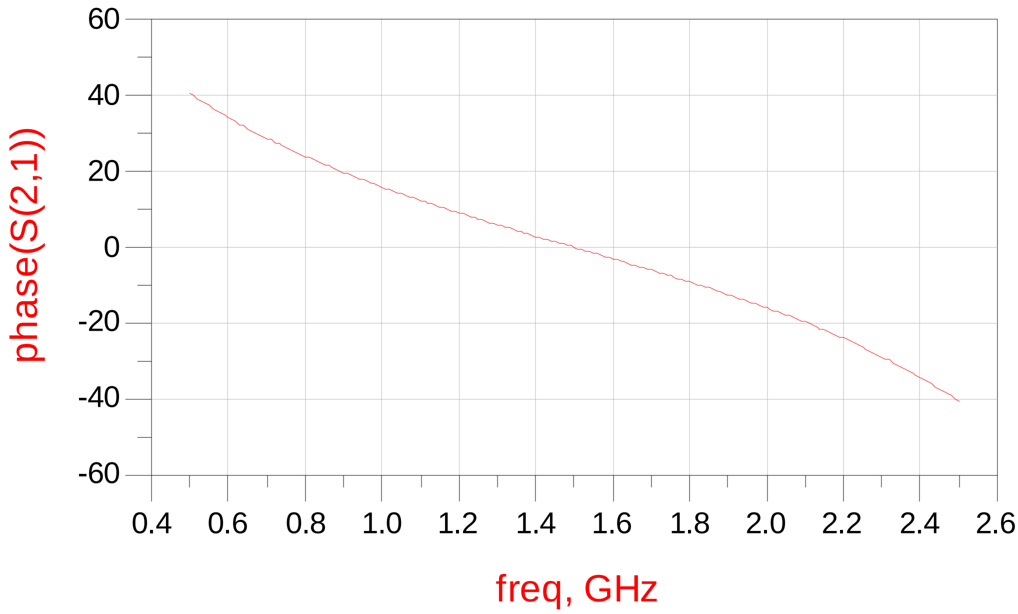


Figure 3.8 S(2,1) parameter of BPF branch of 45 degree phase bit

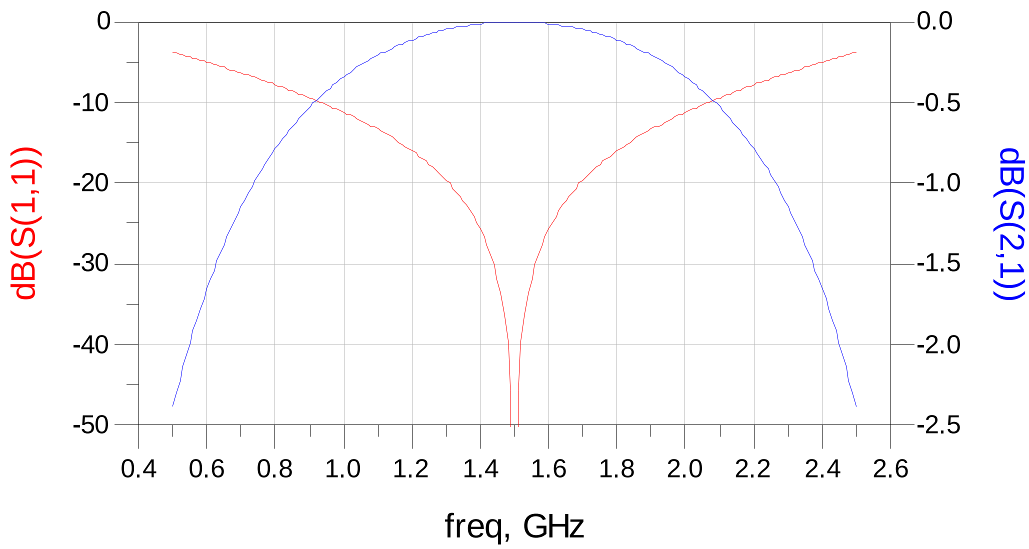


Figure 3.9 S(1,1) and S(2,1) magnitude versus frequency

3.1.2(c) Net phase of 45 degree phase bit

Net phase of 45 degree phase bit is difference of phase provided by transmission line and BPF branch. It can be plotted versus frequency as given below.

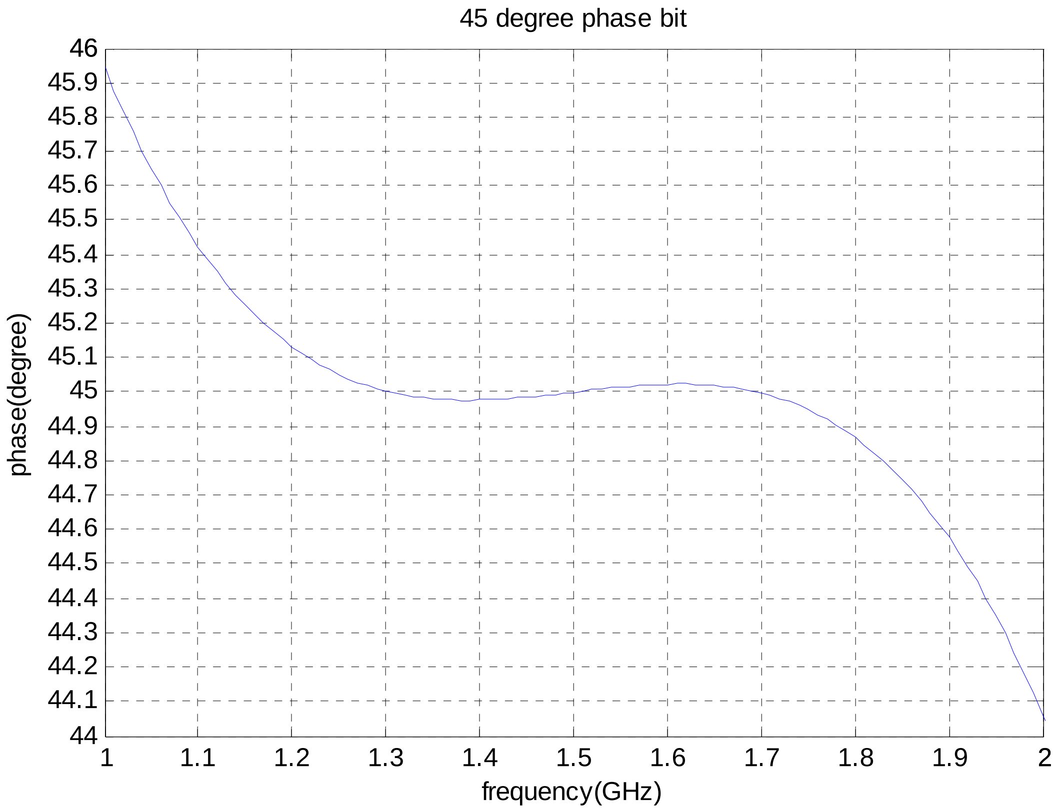


Figure 3.10 net phase versus frequency of 45 degree phase bit

From above figure it is clear that at central frequency 1.5 GHz the net phase is 45 degree and at corner frequencies it is 45.03 and 44.97. Hence the **phase error** of 45 degree phase bit is **0.03 degree**.

3.1.3 results for 90 degree phase bit

First branch of 90 phase bit is a transmission line providing a phase angle of 90 degree. Its s-parameter results are as given below.

3.1.3(a) TL branch

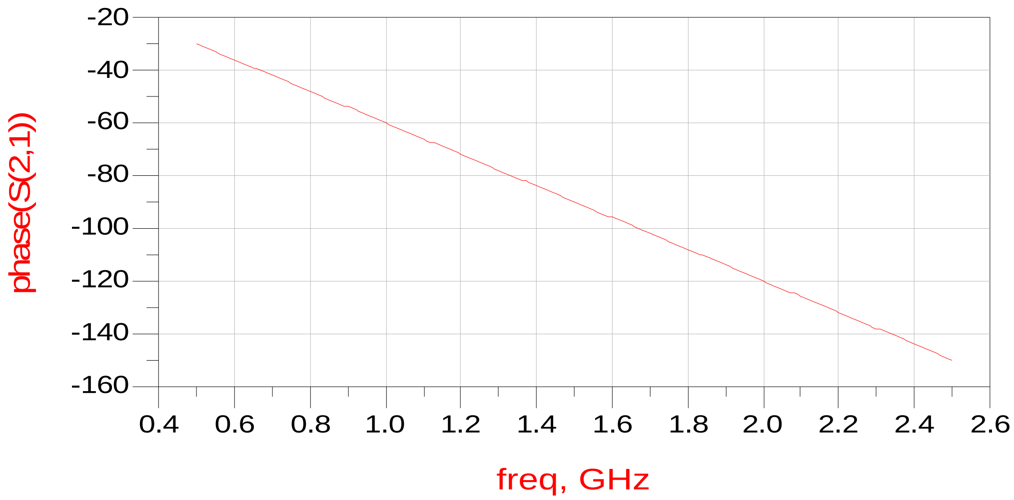


Figure 3.11 S(2,1) parameter of 90 degree TL branch.

Above figure show the phase of s(2,1) parameter variation with frequency. It is clear that at central frequency 1.5 GHz, the phase is 90 degree. To get the same phase in frequency range of bandwidth the phase slope alignment is used, which is provided by BPF branch. BPF is implemented using L and C components.

3.1.3(b) BPF branch

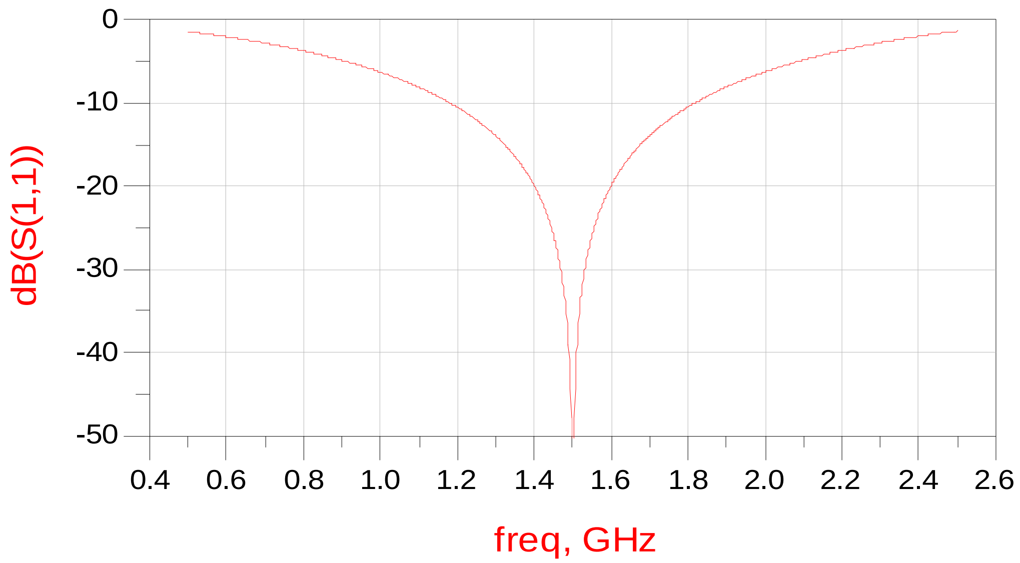


Figure 3.12 S(1,1) parameter of BPF branch versus frequency.

The s(1,1) parameter represents return loss due to BPF branch. From the figure 3.2 it is clear that at central frequency 1.5GHz, return loss is below –50 dB. At the corner frequencies of bandwidth the **return loss** is **–13.12 dB.**

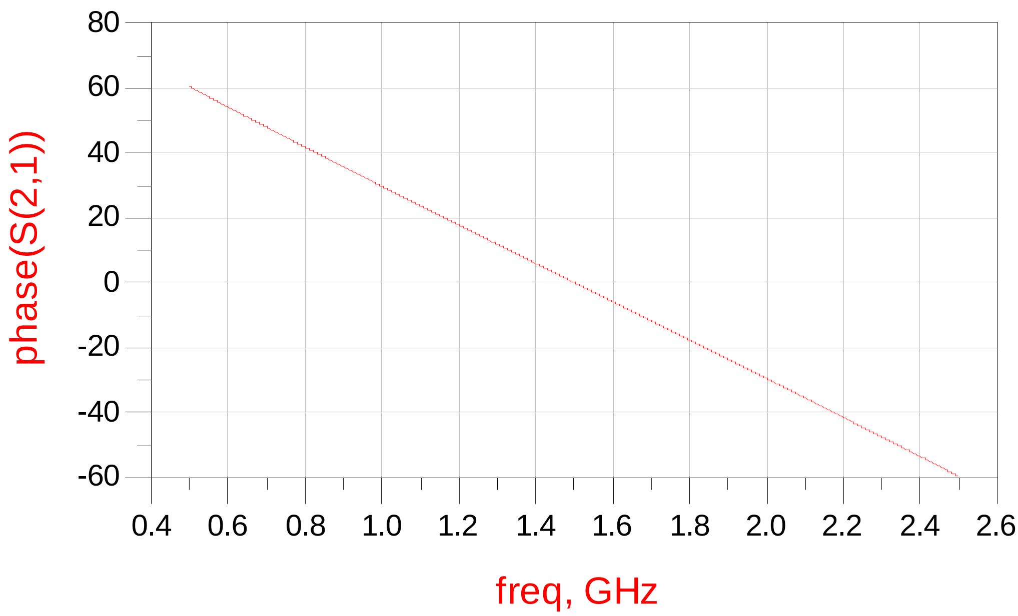


Figure 3.13 S(2,1) parameter of BPF branch of 90 degree phase bit

Figure 3.13 represents S(2,1) parameter of BPF branch of 90 degree phase bit. It is clear from the figure that at central frequency phase provided by BPF branch is zero and positive for frequencies below central frequency and negative for frequencies above central frequency. In this way the phase slope of TL branch is aligned by BPF branch.

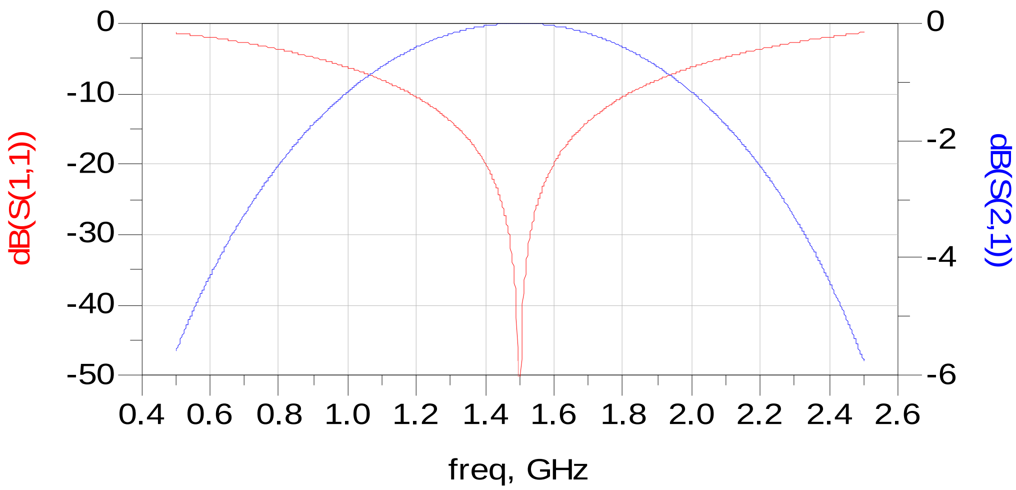


Figure 3.14 S(1,1) and S(2,1) magnitude versus frequency

3.1.3(c) Net phase of 90 phase bit

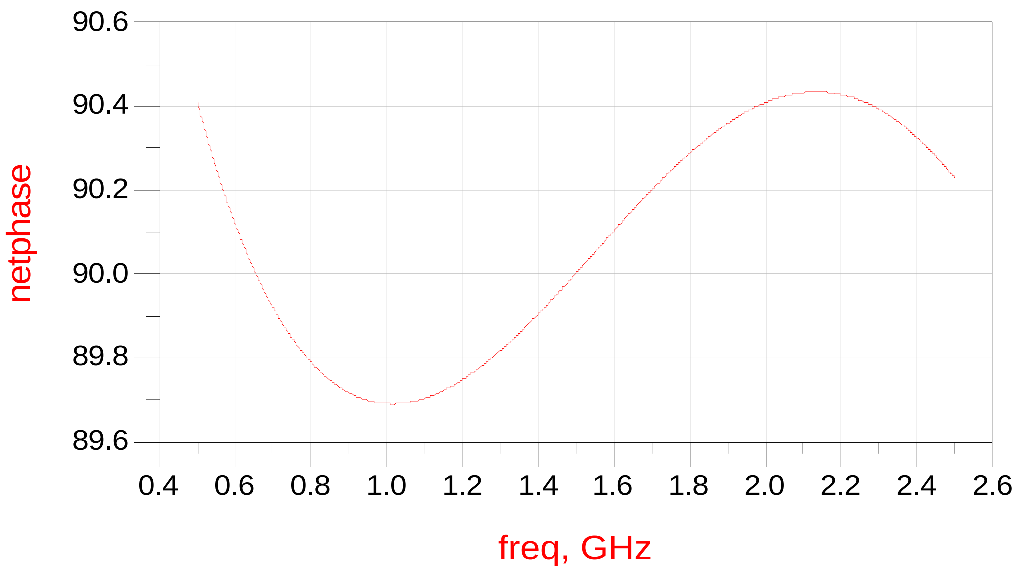


Figure 3.15 Net phase versus frequency plot of 90 degree phase bit

From above figure it is clear that at central frequency 1.5 GHz the net phase is 90 degree and at corner frequencies it is 89.79 and 90.22. Hence the **phase error** of 45 degree phase bit is **0.17 degree**.

3.2 Simulated results for 50% bandwidth phase shifter

Both the branches of phase shifter bit are simulated for different phase shift bits and their results are plotted as give below.

3.2.1 results for 22.5 degree phase bit

First branch of 22.5 phase bit is a transmission line providing a phase angle of 22.5 degree. Its s -parameter results are as given below.

3.2.1(a) TL branch

The figure 3.16 show the phase of s(2,1) parameter variation with frequency. It is clear that at central frequency 1.5 GHz, the phase is 22.5 degree. To get the same phase in frequency range of bandwidth the phase slope alignment is used, which is provided by BPF branch.

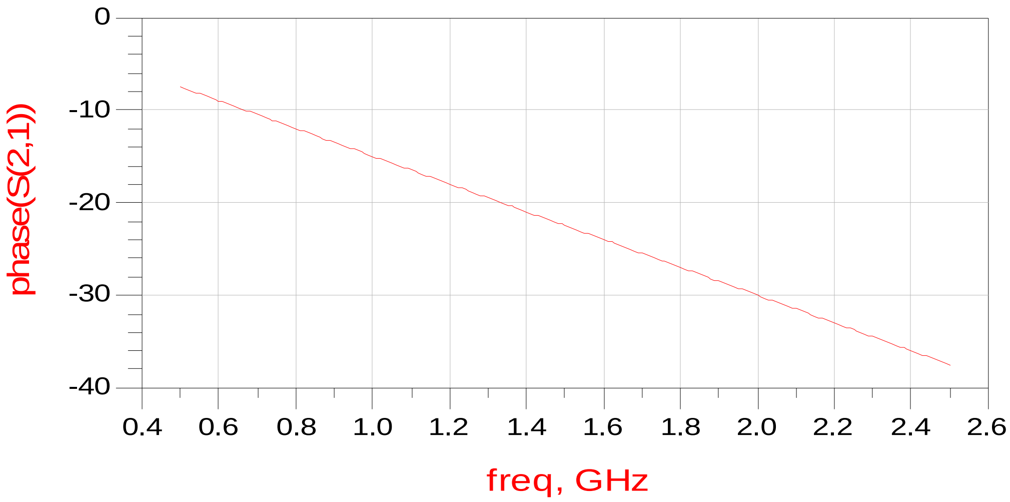


Figure 3.17 S(2,1) parameter of 22.5 degree TL branch.

3.2.1(b) BPF branch

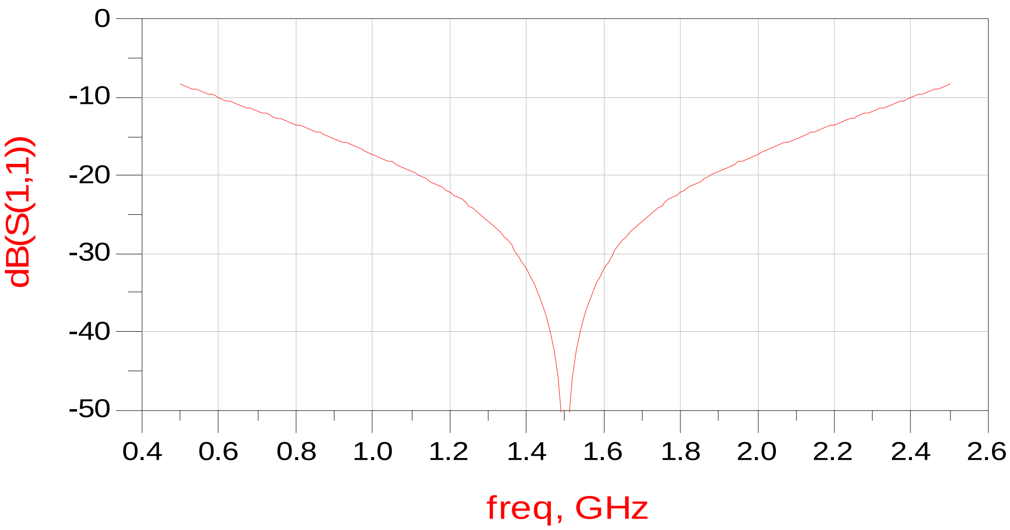


Figure 3.18 S(1,1) parameter of BPF branch versus frequency.

The s(1,1) parameter represents return loss due to BPF branch. From the figure 3.18 it is clear that at central frequency 1.5GHz, return loss is below –50 dB. At the corner frequencies of bandwidth the **return loss** is **–20.19** dB.

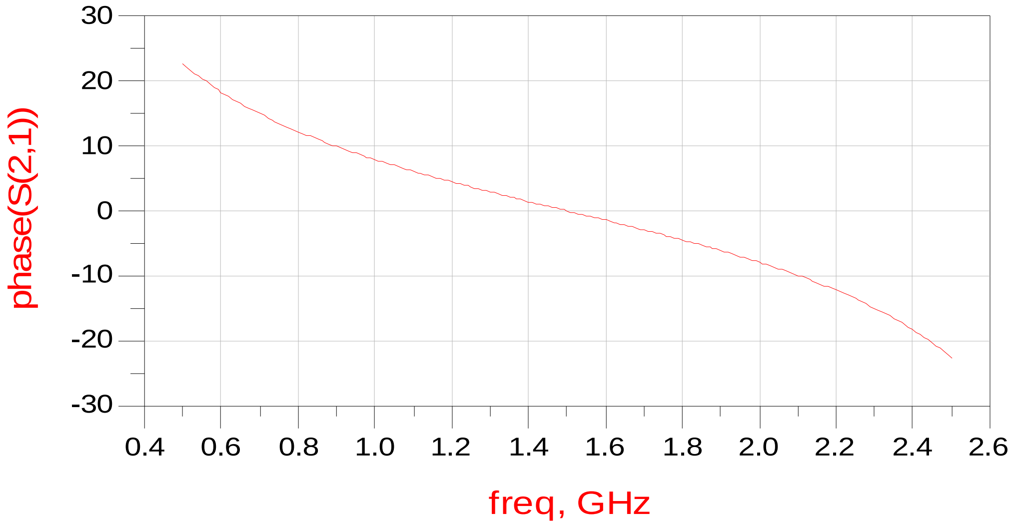


Figure 3.19 S(2,1) parameter of BPF branch versus frequency.

Figure 3.19 represents S(2,1) parameter of BPF branch of 22.5 degree phase bit. It is clear from the figure that at central frequency phase provided by BPF branch is zero and positive for frequencies below central frequency and negative for frequencies above central frequency. In this way the phase slope of TL branch is aligned by BPF branch.

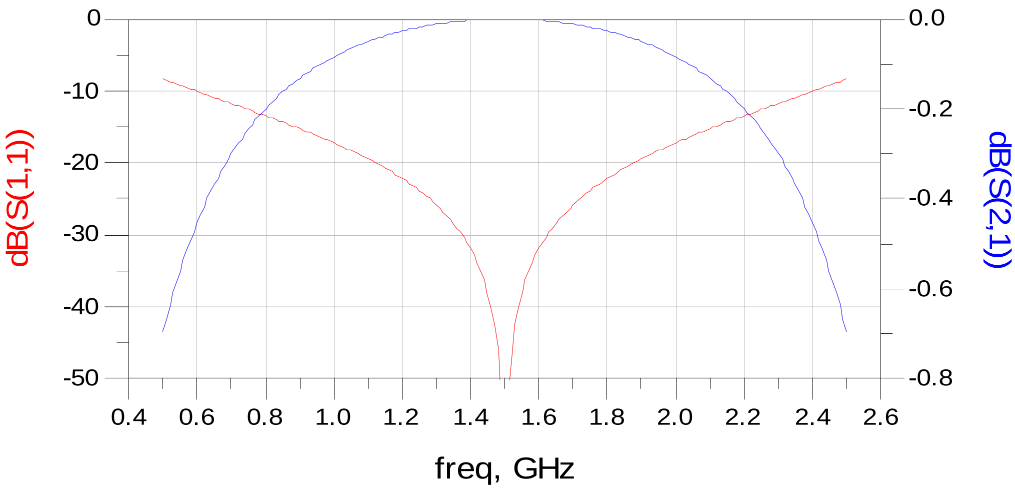


Figure 3.20 S(1,1) and S(2,1) magnitude versus frequency

3.2.1(c) Net phase of 22.5 phase bit

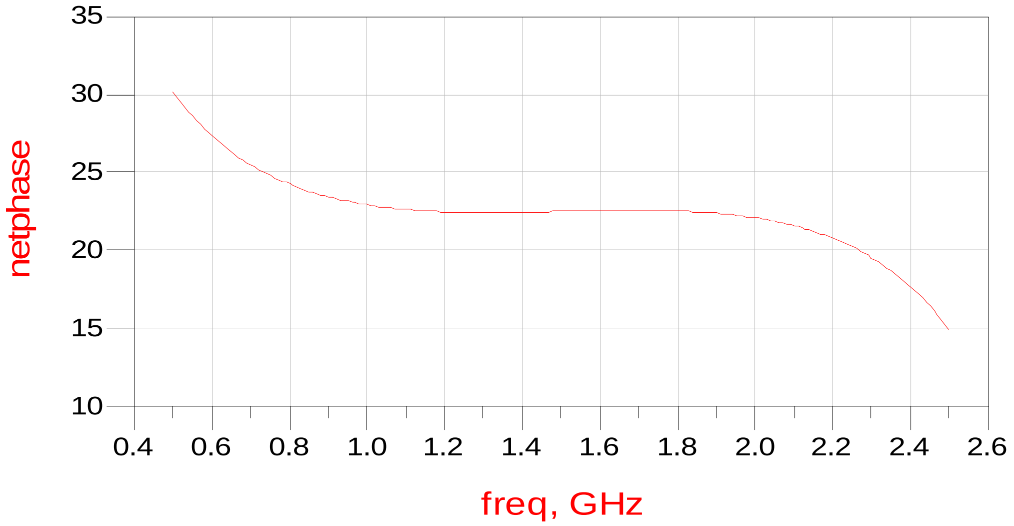


Figure 3.21 Net phase of 22.5 degree phase bit versus frequency

From above figure it is clear that at central frequency 1.5 GHz the net phase is 22.5 degree and at corner frequencies it is 22.44 and 22.44. Hence the **phase error** of 22.5 degree phase bit is **0.06 degree**.

3.2.2 results for 45 degree phase bit

First branch of 45 phase bit is a transmission line providing a phase angle of 45 degree. Its s-parameter results are as given below.

3.2.2(a) TL branch

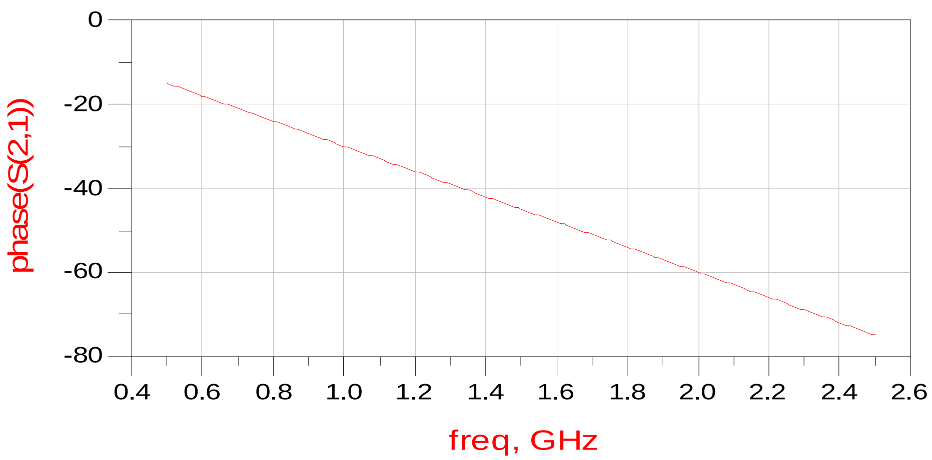


Figure 3.22 S(2,1) parameter of TL branch of 45 degree phase bit

3.2.2(b) BPF branch

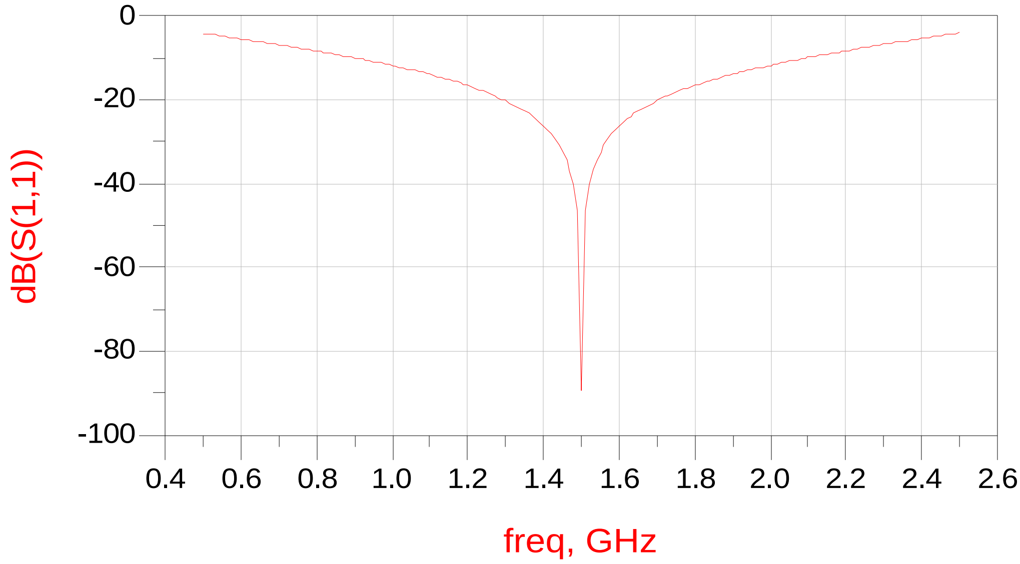


Figure 3.23 S(1,1) parameter of BPF versus frequency

The s(1,1) parameter represents return loss due to BPF branch. From the figure 3.23 it is clear that at central frequency 1.5GHz, return loss is below –50 dB. At the corner frequencies of bandwidth the **return loss** is **–14.52** dB.

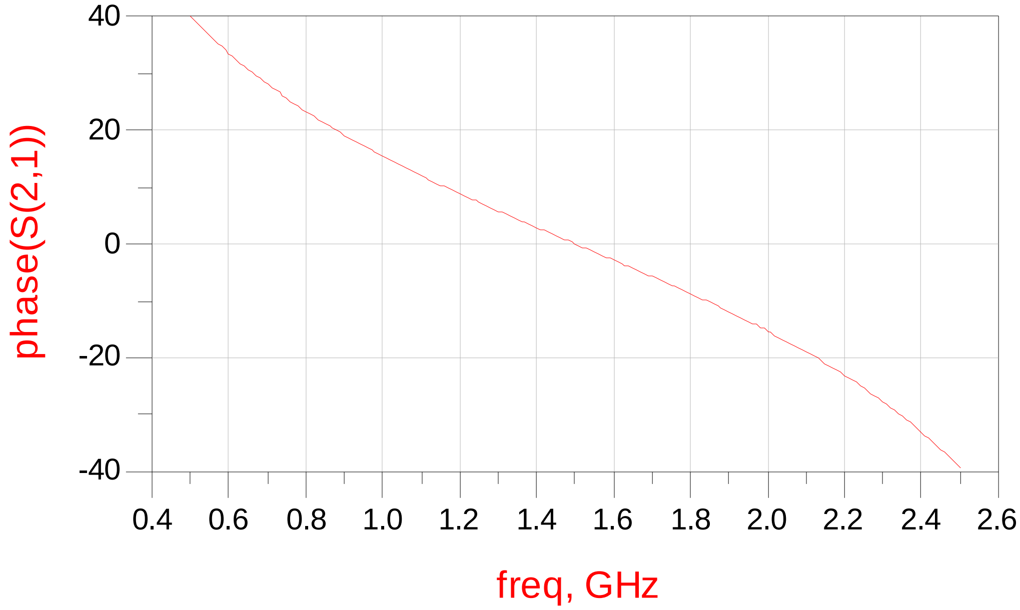


Figure 3.24 S(2,1) parameter of BPF versus frequency

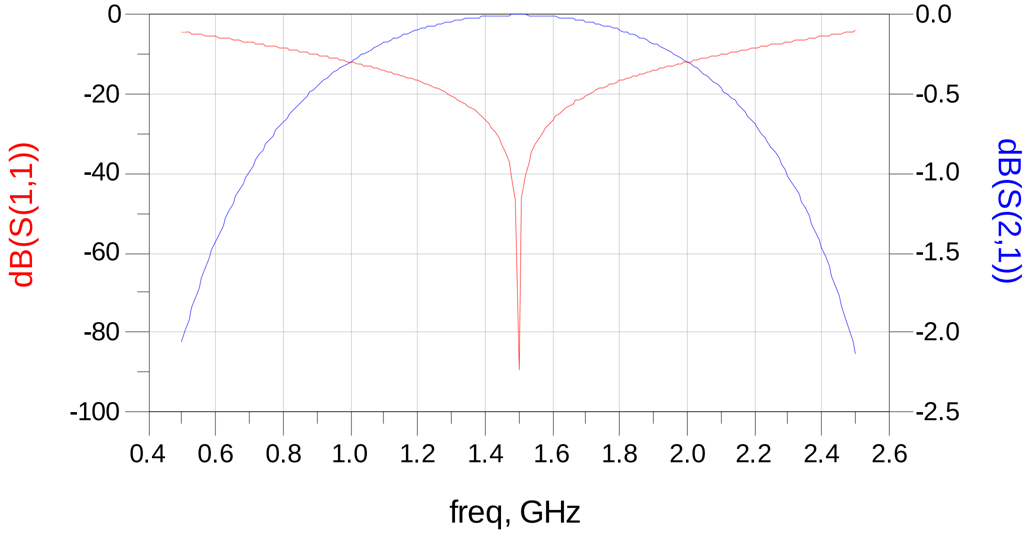


Figure 3.25 S(1,1) and S(2,1) magnitude versus frequency

3.2.2(c) Net phase of 45 phase bit

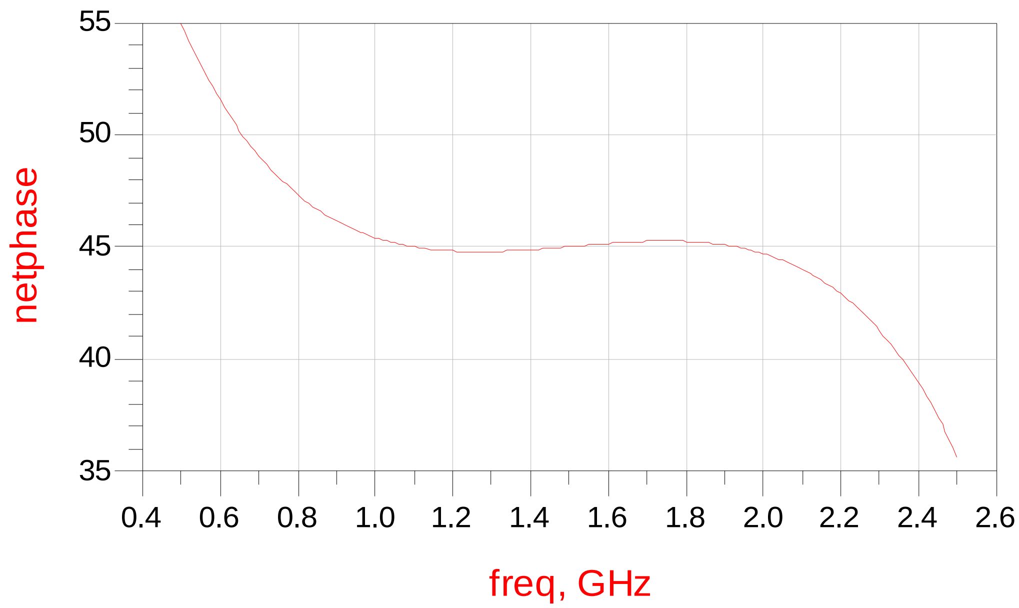


Figure 3.26 Net phase of 45 degree phase bit versus frequency

From above figure it is clear that at central frequency 1.5 GHz the net phase is 45 degree and at corner frequencies it is 44.93 and 45.08. Hence the **phase error** of degree phase bit is **0.07 degree**.

3.2.3 results for 90 degree phase bit

First branch of 90 phase bit is a transmission line providing a phase angle of 90 degree. Its s-parameter results are as given below.

3.2.3(a) TL branch

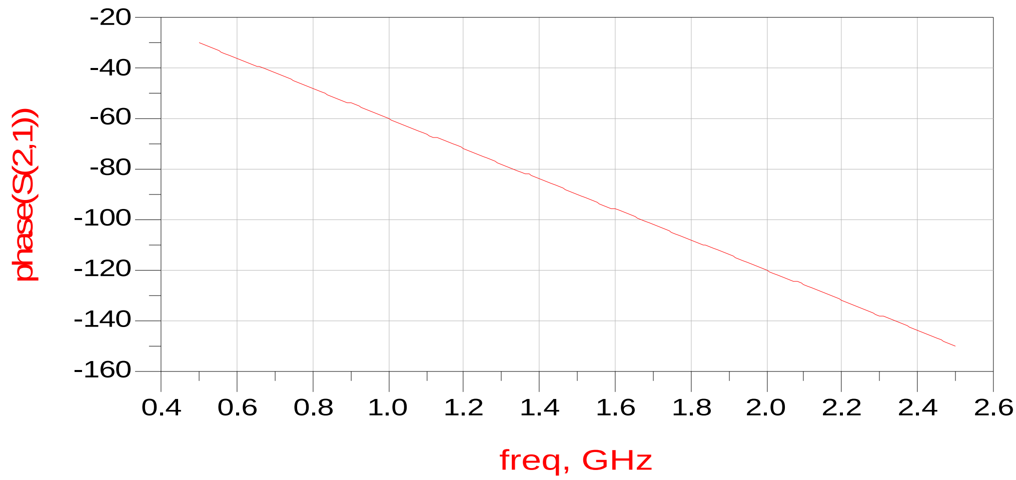


Figure 3.27 S(2,1) parameter of 90 degree TL branch.

Above figure show the phase of s(2,1) parameter variation with frequency. It is clear that at central frequency 1.5 GHz, the phase is 90 degree. To get the same phase in frequency range of bandwidth the phase slope alignment is used, which is provided by BPF branch. BPF is implemented using L and C components.

3.2.3(b) BPF branch

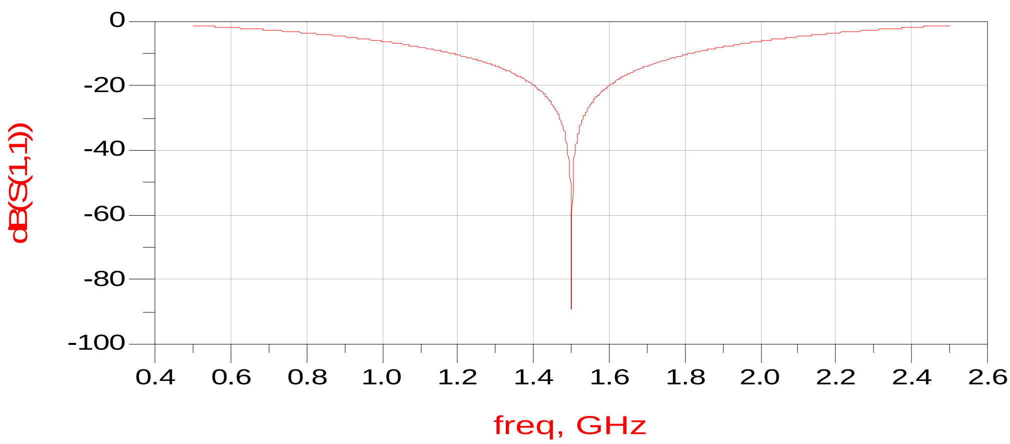


Figure 3.27 S(1,1) parameter of BPF branch versus frequency

The s(1,1) parameter represents return loss due to BPF branch. From the figure 3.27 it is clear that at central frequency 1.5GHz, return loss is below –50 dB. At the corner frequencies of bandwidth the **return loss** is **–9.12 dB.**

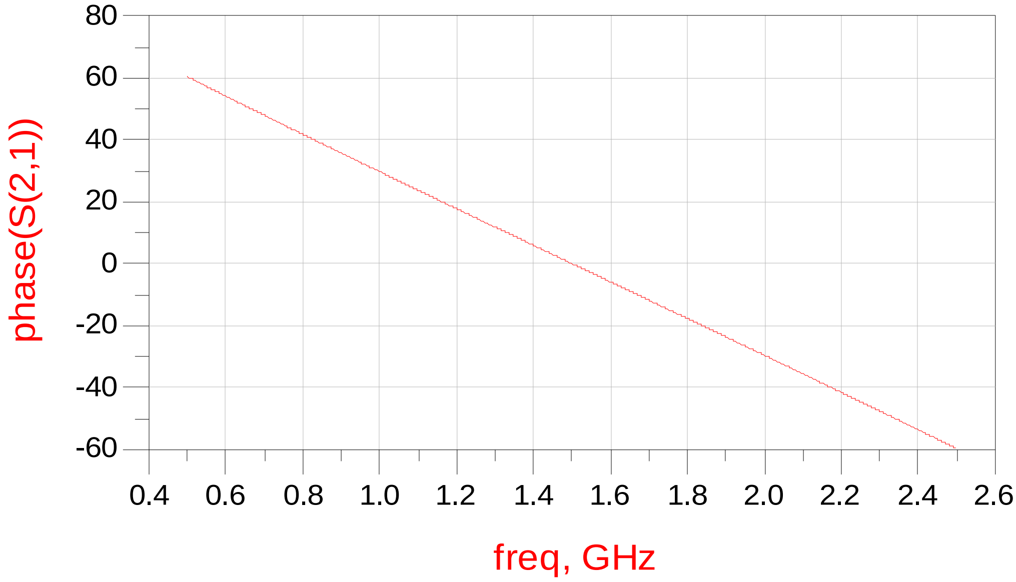


Figure 3.28 S(2,1) parameter of BPF branch versus frequency

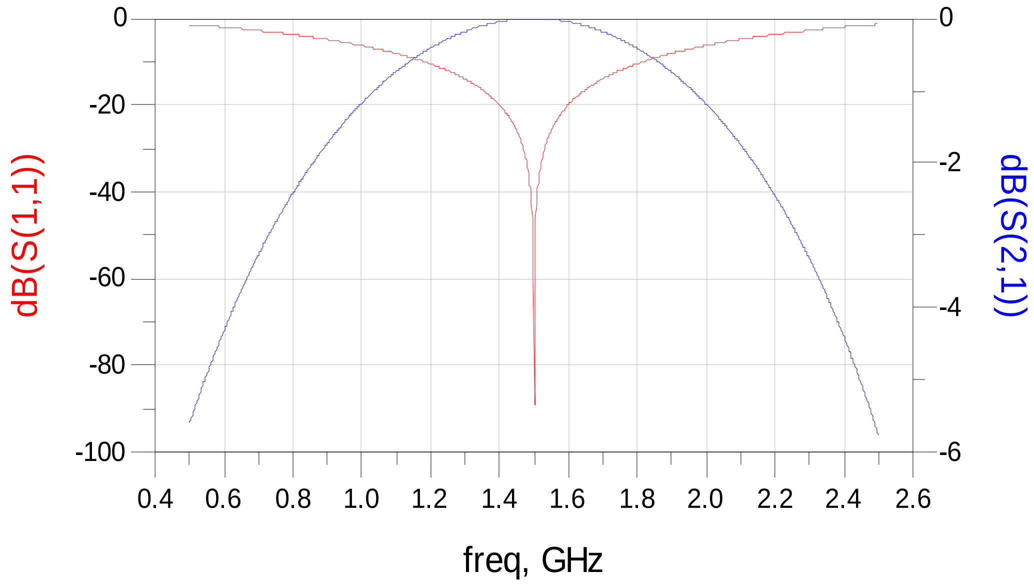


Figure 3.29 S(2,1) and S(1,1) parameter magnitude versus frequency

3.2.3(c) Net phase of 90 degree phase bit

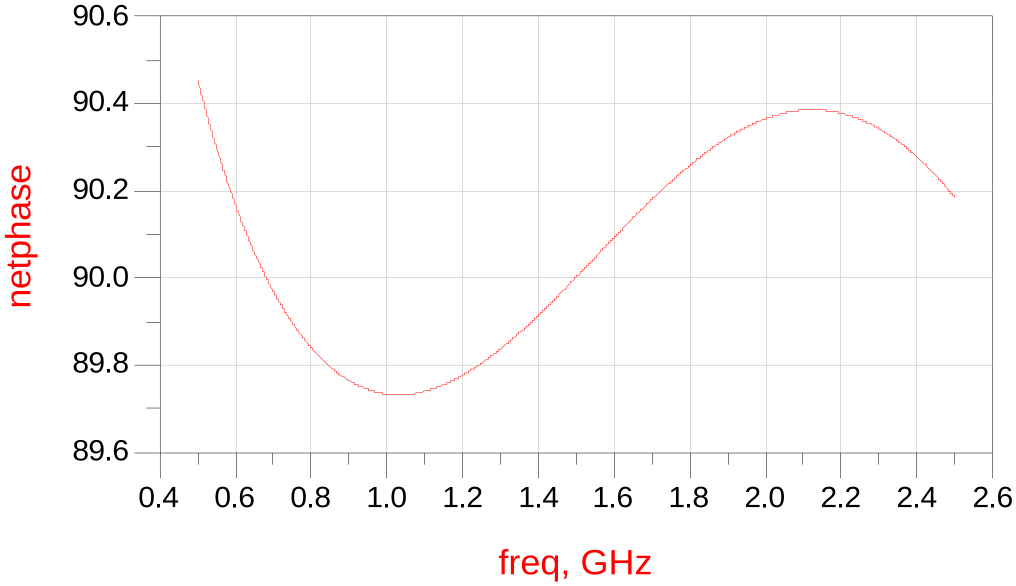


Figure 3.30 Net phase of 90 degree phase bit versus frequency

From above figure it is clear that at central frequency 1.5 GHz the net phase is 90 degree and at corner frequencies it is 89.72 and 90.21. Hence the **phase error** of 45 degree phase bit is **0.28 degree**.

3.3 Three bit phase shifter design and simulation results

In the design of three bit phase shifter LSB represents 22.5 degree phase bit and MSB represents 90 degree phase bit. There are three bits in design so eight possible phase shifts are there. Basic design of three bit phase shifter is given in figure 3.31.

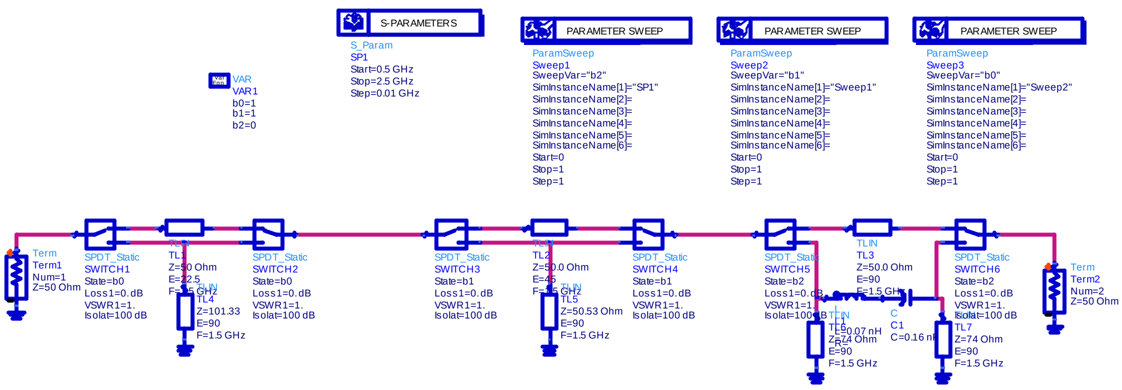


Figure 3.31 Basic design of three bit phase shifter using ADS simulator

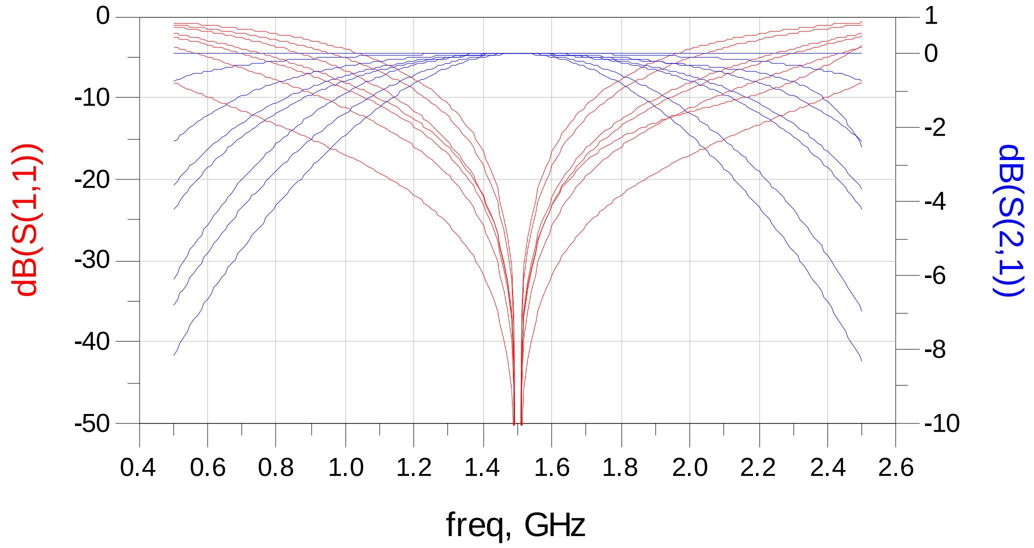


Figure 3.31 return loss variation of three bit phase shifter versus frequency

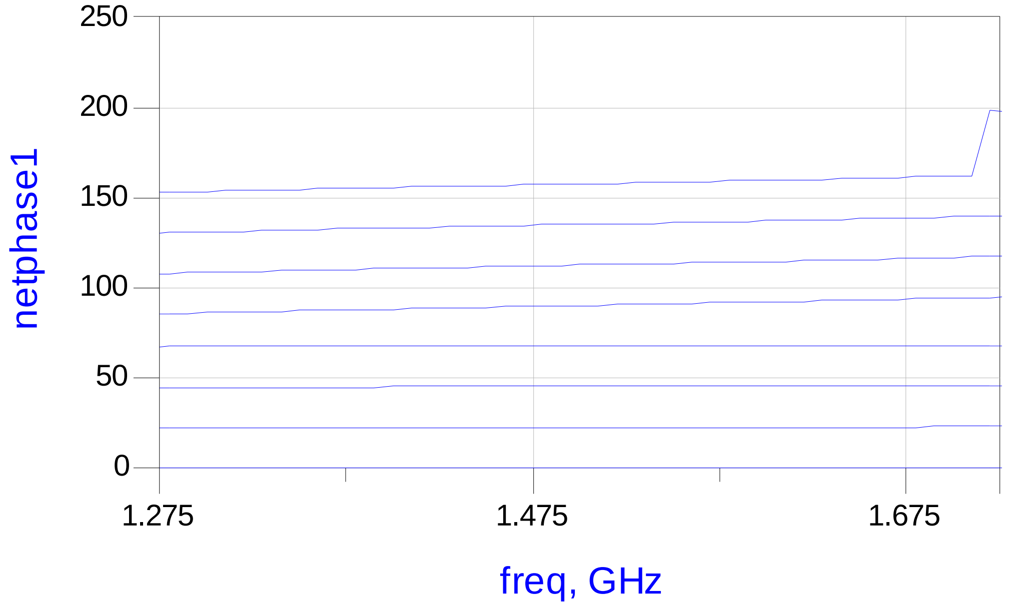


Figure 3.33 Net phase of three bit phase shifter versus frequency

Chapter 4

Conclusion and future work

4.1 Conclusion

The main objective of this project was to design digital phase shifter which works in microwave frequency range. The phase error and return loss should be within limits.

Three bits are designed separately then a three bit phase shifter is designed whose central frequency is 1.5GHz and works in 30% bandwidth. The eight possible phase shifts are then produced with phase error within limits.

4.2 Summary of results

Table given below shows the summary of different phase bits designed.

Table 4.1 results comparison

Phase bit (degree)	Bandwidth %	Phase error (degree)	Return loss (dB)
22.5	30	0.02	–24.56
45	30	0.03	–18.52
90	30	0.17	–13.12
22.5	50	0.06	–20.19
45	50	0.07	–14.52
90	50	0.28	-9.12

4.2 future work

The three bit phase shifter design can be fabricated and the results can be compared with simulated results.

180 degree phase bit can also be designed and four bit phase shifter can be designed.

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