ZADÁNÍ DIPLOMOVÉ PRÁCE

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obor

Název tématu:
Fázový posouvací s vedením periodicky zatíženým varaktory
(Phase Shifter Based on Varactor-Loaded Transmission Line)

Zásady pro vypracování:

Seznamte se s různými druhy fázových posouvacích prvků pro mikrovlnná pásmena a porovnejte jejich vlastnosti z hlediska funkce i náročnosti realizace.

Navrhněte a realizujte diferenciální fázový posouvací na bázi vedení periodicky zatíženého varaktory. Obvod by měl vykazovat fázový posuv alespoň 90 stupňů v pásmu jedné oktávy se středním kmitočtem 1 GHz či vyšším. Snažte se dosáhnout útlumu odrazu alespoň 10dB.

Preferujte jednoduché a levné řešení.

Práce by měla obsahovat teoretický rozebor, návrh obvodu vč. simulace reálných vlastností použitých prvků (varaktory, vedení, diskontinuita) a chování při velkosignálovém vyzvučení. Výsledky simulace porovnejte s měřenými na realizovaném vzorku.
Seznam doporučené odborné literatury:


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V Praze, květen 2009
I hereby declare that I have written this thesis without any help from others and without the use of documents and aids other than those stated below, that I have mentioned all used sources and that I have cited them correctly according to established academic citation rules.

Matěj Vokáč
May 22, 2009
Prague
Abstract

The following master’s thesis describes the function principles, design and optimization of a phase shifter based on a varactor-loaded transmission line. The phase shifter consists of a microstrip transmission line that is periodically loaded with varactor diodes. Since electrical length of such a synthetic transmission line is a function of reverse bias voltage applied to the diodes, the phase of the signal passing through the line can be shifted according to the control voltage.

In order to simulate the designed phase-shifter circuit accurately, a BB857 varactor-diode nonlinear model has been developed based on own measurements of the diode. The model is also used to analyze the large-signal behavior of the resulting circuit. The designed and subsequently fabricated phase shifter meets the objective of the maximum differential phase shift of no less than 90° within a one-octave frequency band for the center frequency of 1 GHz.

Index Terms – Phase shifter, microstrip transmission line, BB857 varactor diode, equivalent circuit, modeling, large-signal behavior
Abstract
in the Czech Language

Následující diplomová práce popisuje principy funkce, návrh a optimalizaci fázového posouvače založeného na vedení zatíženém varaktory. Tento fázový posouvač sestává z mikropáskového vedení, které je varaktory periodicky zatížené. Jelikož elektrická délka takto vzniklého syntetického vedení je funkci záporného předpětí těchto kapacitních diod, může být fáze signálu procházejícího vedením posouvána v závislosti na řídícím napětí.

Aby bylo možné obvod navrženého fázového posouvače přesně simulovat, byl na základě vlastních měření varaktoru BB857 vytvořen jeho nelineární model. Tento model je také použit k analýze velkosignálového chování výsledného obvodu. Navržený a následně vyrobený fázový posouvač splňuje cíl maximálního rozdílového fázového posuva o hodnotě alespoň 90° ve frekvenčním pásmu jedné oktávy pro střední frekvenci 1 GHz.

Klíčové pojmy – Fázový posouvač, mikropáskové vedení, varaktor BB857, náhradní obvod, modelování, velkosignálové chování
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<td>$[A]$</td>
<td>General chain matrix</td>
</tr>
<tr>
<td>$[A_u]$</td>
<td>Chain matrix of a distributed phase-shifter unit cell</td>
</tr>
<tr>
<td>$[A_p]$</td>
<td>Chain matrix of a homogenous transmission line section</td>
</tr>
<tr>
<td>$[A_{var}]$</td>
<td>Chain matrix of a shunt varactor diode to ground</td>
</tr>
<tr>
<td>$C_i$</td>
<td>Capacitance per unit length of a homogenous transmission line</td>
</tr>
<tr>
<td>$C_j$</td>
<td>PN junction capacitance</td>
</tr>
<tr>
<td>$C_{j0}$</td>
<td>Zero-bias PN junction capacitance</td>
</tr>
<tr>
<td>$C_{L}$</td>
<td>Capacitance per unit length of a synthetic transmission line</td>
</tr>
<tr>
<td>$C_p$</td>
<td>Varactor-diode package capacitance</td>
</tr>
<tr>
<td>$C_t$</td>
<td>Equivalent lumped capacitance of a transmission-line section</td>
</tr>
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<td>Tunable varactor-diode capacitance</td>
</tr>
<tr>
<td>$f$</td>
<td>Frequency</td>
</tr>
<tr>
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<td>Center frequency of operation</td>
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<tr>
<td>$f_B$</td>
<td>Bragg frequency</td>
</tr>
<tr>
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<td>Maximum frequency of operation</td>
</tr>
<tr>
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<td>Minimum frequency of operation</td>
</tr>
<tr>
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</tr>
<tr>
<td>$h$</td>
<td>Height of a microstrip-line substrate</td>
</tr>
<tr>
<td>$l$</td>
<td>Physical length of a transmission line</td>
</tr>
<tr>
<td>$l_{sp}$</td>
<td>Physical length of varactor-diodes spacing</td>
</tr>
<tr>
<td>$L_i$</td>
<td>Inductance per unit length of a homogenous transmission line</td>
</tr>
<tr>
<td>$L_{L}$</td>
<td>Inductance per unit length of a synthetic transmission line</td>
</tr>
<tr>
<td>$L_s$</td>
<td>Varactor-diode series inductance</td>
</tr>
<tr>
<td>$L_t$</td>
<td>Equivalent lumped inductance of a transmission-line section</td>
</tr>
<tr>
<td>$n$</td>
<td>PN junction profile exponent</td>
</tr>
<tr>
<td>$N$</td>
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<td>Resistance per unit length of a homogenous transmission line</td>
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<td>$R_j$</td>
<td>PN junction resistance</td>
</tr>
<tr>
<td>$R_s$</td>
<td>Varactor-diode series resistance</td>
</tr>
<tr>
<td>$[S]$</td>
<td>General scattering matrix</td>
</tr>
<tr>
<td>$[S_u]$</td>
<td>Scattering matrix of a distributed phase-shifter unit cell</td>
</tr>
<tr>
<td>$t$</td>
<td>Thickness of a microstrip-line metallization</td>
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<tr>
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<td>Phase velocity on an unloaded transmission line</td>
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<tr>
<td>$v_{L}$</td>
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<tr>
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<td>Reverse bias voltage</td>
</tr>
<tr>
<td>$V_0$</td>
<td>PN junction diffusion voltage</td>
</tr>
<tr>
<td>$w$</td>
<td>Width of a microstrip transmission line</td>
</tr>
<tr>
<td>$Y_{var}$</td>
<td>Tunable varactor-diode admittance</td>
</tr>
</tbody>
</table>
$Z_i$ Characteristic impedance of an unloaded transmission line
$Z_{in}$ Reference impedance at an input port
$Z_{out}$ Reference impedance at an output port
$Z_L$ Characteristic impedance of a varactor-loaded transmission line
$\alpha$ Attenuation constant
$\beta$ Phase constant
$\gamma$ Propagation constant of a homogenous transmission line
$\delta$ Loss angle
$\varepsilon_r$ Dielectric constant
$\lambda_g$ Guided wavelength on a transmission line
$\varphi_i$ Electrical length of an unloaded transmission line
$\varphi_L$ Electrical length of a varactor-loaded transmission line
$\varphi_{sp}$ Electrical length of an unloaded transmission-line section between varactor diodes
$\Phi_L$ Maximum differential phase shift of a varactor-loaded transmission-line phase shifter
$\Phi_u$ Maximum differential phase shift of a varactor-loaded transmission-line phase-shifter unit cell
$\omega$ Angular frequency
$\omega_B$ Bragg angular frequency
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Chapter 1

Introduction

1.1 Thesis Objectives

Microwave systems have become an integral part of modern life, particularly in the field of wireless communications, and have found their applications in various other spheres of human activity including industry, navigation, radio-location, transportation, medicine, pharmacology and scientific research [1].

A phase shifter is an example of a microwave device that enhances the features of systems operating at microwave frequencies and enables them to provide the user with new beneficial functions thanks to its ability to adjust the phase of a passing signal as desired.

The following master’s thesis describes different function principles and technologies of implementation of microwave phase-shifter circuits. The objective of this thesis is then to explain the function of a varactor-loaded transmission-line phase shifter in relation to its physical parameters, specify key variables of the structure and discuss an appropriate choice of their values in order to design and optimize the circuit.

Furthermore, the thesis aims to cover development and fabrication of a particular phase shifter that operates within a one-octave frequency band for the center frequency of 1 GHz or higher. A band that spans from 0.7 GHz to 1.4 GHz has been chosen for this purpose. The maximum differential phase shift of the circuit should be no less than 90° within the whole frequency band and return loss of no less than 10 dB should also be achieved. Moreover, it is desirable that the phase shifter be implemented by simple and inexpensive technique.
Inaccurate circuit-element description or missing element parasitics may lead to imprecise conclusions during microwave systems design and to significant differences between the designed circuit simulations and measurements of the manufactured circuits. Care must be taken to rely on as precise element models as possible in order to obtain acceptable agreement between the simulated and the measured data.

For the designed phase-shifter circuit to be simulated accurately, a nonlinear model of a BB857 varactor diode, manufactured by Infineon Technologies AG, needs to be developed. The model should precisely simulate the real element within the frequency range of 0.5 GHz through 3 GHz at several significant diode bias voltages with respect to the operating conditions of the phase-shifter circuit, in which the diodes are to be employed. Obviously, the diode package parasitics and the chip bias-dependent character should also be taken into account.

Since the function of a varactor-loaded transmission-line phase shifter relies on voltage-controlled PN junction capacitance of varactor diodes, precise varactor-diode modeling is crucial to the phase shifter design. In addition, the nonlinear dependence of the PN junction capacitance on the applied bias voltage \([2]\) represents the main source of nonlinearity of the circuit and thus the model is also important for the large-signal behavior analysis of the phase-shifter circuit.

### 1.2 Thesis Outline

Chapter 1 introduces phase shifters as useful components of modern microwave systems, presents the main objectives of the thesis, covers the fundamentals of the phase-shifter function and explains why accurate element modeling is important for the whole system design. It also informs about different function principles and technologies of implementation of phase-shifter circuits intended to operate in microwave frequency bands as well as about various applications that benefit from the phase-shifter function.

Chapter 2 explores in greater detail the structure, key design parameters and the function principle of a varactor-loaded transmission-line phase shifter. It also explains the phenomenon of the Bragg frequency and formulates its restriction on the periodic structure of the distributed phase shifter circuit. In addition, the basics of capacitance diodes are mentioned in the chapter.

Chapter 3 covers design and optimization of the phase-shifter circuit with
ideal elements as well as the design objectives and constraints, provides a nonlinear equivalent circuit of a BB857 varactor diode and employs the model in simulations of the phase-shifter circuit with real elements. Accuracy of both the model provided by the manufacturer and the resulting improved model is also evaluated.

Chapter 4 provides technical details of the circuit fabrication and presents results of small-signal vector measurements of the circuit as well as its measured large-signal behavior. The characteristics obtained from the measurements are compared with the simulated data.

Chapter 5 concludes the thesis by summing up and discussing the results reached and by proposing additional steps to take in order to further improve the approach to this topic.

1.3 Phase-Shifter Fundamentals

A phase shifter is a device that adjusts the phase of an input signal according to a control signal. This function can be achieved by several principles of operation and implemented by different technologies.

This master’s thesis focuses on a phase shifter based on a varactor-loaded transmission line. The circuit employs voltage-dependent PN junction capacitance of varactor diodes to alter electrical length of the synthetic transmission line formed by periodical loading of a microstrip transmission line with the diodes. And because electrical length of such a synthetic transmission line is a function of reverse bias voltage applied to the diodes, the phase of the signal passing through the line can be shifted according to the control voltage.

The phase shifter consists of a microstrip transmission line that is periodically loaded with varactor diodes. Since electrical length of such a synthetic transmission line is a function of reverse bias voltage applied to the diodes, the phase of the signal passing through the line can be shifted according to the control voltage.

Based on the nature of the control signal, phase shifters can be basically divided into two groups—analog and digital phase shifters.
1.4 Function Principles

There are several means how to achieve the phase-shifting function. Based on their function principle, phase shifters can be divided into transmission-type loaded-line phase shifters [3], reflective-type phase shifters [4], high-pass phase shifters [5], tunable electromagnetic crystal (EMXT) surfaces [6] etc.

For example, reflective-type phase shifters (RTPS) provide better return loss and larger phase tuning range than, for instance, a transmission-type loaded-line phase shifter. On the other hand, the RTPS typically take up more space [4].

1.5 Technologies of Implementation

Phase shifters can be implemented by the microwave monolithic integrated circuits (MMICs) [8], the pseudomorphic high electron-mobility transistors (pHEMTs) [15], silicon varactor diodes—which is the technology employed in this thesis—, ferroelectric varactors [14], ferrites, micro-electro-mechanical systems (MEMS) etc.

For example, the disadvantage of the MMIC technology are the high costs of implementation and technological challenges [8].

1.6 Applications

Various modern electronical and microwave systems benefit from the ability of phase shifters to adjust the phase of a passing signal as desired, e. g., beam scanning antenna arrays, phase modulators [15], [16], jitter and phase error compensation circuits and frequency tunable active microwave filters.

Many radar [10], [11], [8], [6] and wireless communication [9], [10], [6] systems are based on phased-array antennas for achieving electronic beam control and fast beam scanning. Such systems can be used for military, e. g., early-warning and missile defense systems [10], [8], and commercial applications, e. g., flight control [10], automotive collision-avoidance [9], [10], [8] and cruise-control [8] systems, blind-spot indicators [8], multipoint communications [8], the global positioning system (GPS) [10], compact scanning arrays [8] etc.
Chapter 2

Varactor-Loaded Transmission-Line Phase Shifter

2.1 Basic Structure Description

A phase shifter that is in the center of attention of this thesis consists of a high-impedance transmission line with characteristic impedance $Z_i$ periodically loaded with varactor diodes of capacitance $C_{\text{var}}$. The spacing between the varactor diodes can be characterized by its physical and electrical length, $l_{sp}$ and $\varphi_{sp}$, respectively. Such a structure along with corresponding parameters is shown in Fig. 2.1.

![Varactor-loaded transmission-line circuit](image)

Fig. 2.1: Varactor-loaded transmission-line circuit

Key parameters of a transmission line and the means how these parameters are affected by varactor loading need to be explained in order to gain further knowledge of the phase shift principle. Before proceeding to the
phase shifter design itself, theoretical foundations for certain fields of the design should also be laid. Namely, the Bragg frequency restriction, the cascade structure description and capacitance diodes basics are discussed in the following sections.

2.2 Key Transmission-Line Parameters

Analysis of an elementary transmission-line segment which is excited by a harmonic voltage source of angular frequency $\omega$ leads to formula (2.1). It expresses propagation constant of a homogenous transmission line $\gamma$ as follows [20]:

$$
\gamma = \sqrt{(R_i + j\omega L_i)(G_i + j\omega C_i)} = \alpha + j\beta,
$$

(2.1)

where $R_i$ and $G_i$ stand for resistance and conductance per unit length of a homogenous transmission line, respectively, and $L_i$ and $C_i$ are inductance and capacitance per unit length of a homogenous transmission line, respectively. Factors $\alpha = \Re\{\gamma\}$ and $\beta = \Im\{\gamma\}$ denote attenuation constant and phase constant, respectively.

Characteristic impedance of a homogenous transmission line $Z_i$ is given by equation (2.2) derived from the same analysis [20]:

$$
Z_i = \sqrt{\frac{R_i + j\omega L_i}{G_i + j\omega C_i}}.
$$

(2.2)

Primary transmission-line parameters—$R_i, G_i, L_i, C_i$—dependencies on the line characteristic impedance $Z_i$ and the propagation constant $\gamma$ can then be determined using formulas (2.1) and (2.2) as:

$$
R_i = \Re\{\gamma Z_i\},
$$

(2.3)

$$
L_i = \frac{1}{\omega} \Im\{\gamma Z_i\},
$$

(2.4)

$$
G_i = \Re\{\frac{\gamma}{Z_i}\},
$$

(2.5)
\[ C_i = \frac{1}{\omega} \Im\{\frac{\gamma}{Z_i}\}, \quad (2.6) \]

because

\[ \gamma Z_i = R_i + j\omega L_i, \quad (2.7) \]
\[ \frac{\gamma}{Z_i} = G_i + j\omega C_i. \quad (2.8) \]

The speed at which wavefront points of a given phase value propagate down a transmission line is called phase velocity. Phase velocity on a transmission line is expressed [21] by equation (2.9).

\[ v_i = \frac{\omega}{\beta} \quad (2.9) \]

Supposing the considered homogenous transmission line is lossless, i.e., its resistance \( R_i \) and conductance \( G_i \) per unit length equal to zero, formulas (2.1) and (2.2) can be simplified to

\[ \gamma = \sqrt{j\omega L_i j\omega C_i} = j\omega \sqrt{L_i C_i} = j\beta, \quad (2.10) \]
\[ Z_i = \sqrt{\frac{L_i}{C_i}}. \quad (2.11) \]

Based on (2.10), it is evident that a lossless transmission line is characterized by the following relations:

\[ \alpha = \Re\{\gamma\} = 0, \quad (2.12) \]
\[ \beta = \Im\{\gamma\} = \omega \sqrt{L_i C_i}. \quad (2.13) \]

Relations (2.9) and (2.13) further result in the expression of lossless transmission-line phase velocity:

\[ v_i = \frac{\omega}{\omega \sqrt{L_i C_i}} = \frac{1}{\sqrt{L_i C_i}}. \quad (2.14) \]
Dependences of inductance $L_i$ and capacitance $C_i$ per unit length on characteristic impedance $Z_i$ and phase velocity $v_i$ can be obtained by modification of (2.4), (2.6) and (2.9) as follows:

$$L_i = \frac{1}{\omega} \Im\left\{ \frac{Z_i}{v_i \beta} \Im\{ \gamma \} \right\} = \frac{Z_i}{v_i \beta} \Im\{ j \beta \} = \frac{Z_i}{v_i \beta} = \frac{Z_i}{v_i}, \quad (2.15)$$

$$C_i = \frac{1}{\omega} \Im\left\{ \frac{\gamma}{Z_i} \right\} = \frac{1}{v_i \beta Z_i} \Im\{ \gamma \} = \frac{1}{v_i \beta Z_i} \Im\{ j \beta \} = \frac{\beta}{v_i \beta Z_i} = \frac{1}{Z_i v_i}, \quad (2.16)$$

These dependences are employed in section 2.4 in order to express Bragg frequency $f_B$ by means of varactor diode capacitance $C_{\text{var}}$, characteristic impedance of an unloaded transmission line $Z_i$, electrical length of an unloaded transmission-line section between varactor diodes $\varphi_{sp}$ and angular frequency $\omega$.

### 2.3 Phase Shift Principle

The goal of this section is to demonstrate that phase difference between a signal at the input port and the same signal at the output port of a transmission-line segment can be altered by variable capacitance by which the transmission line is loaded. The concept of a distributed phase shifter that uses a tunable synthetic transmission line to control the phase of the incident signal [3], [22] will be presented.

When a signal propagates down a transmission line, the difference between its respective phases at the input port and at the output port of the line equals to electrical length of the line. Electrical length of a transmission line $\varphi_i$ is defined [23] by equation (2.17):

$$\varphi_i = 2\pi \frac{l}{\lambda_g}, \quad (2.17)$$

where $l$ denotes the transmission-line physical length and $\lambda_g$ stands for signal wavelength on the transmission line, which is given [23] as follows:

$$\lambda_g = \frac{2\pi}{\beta}. \quad (2.18)$$
Using relations (2.18) and (2.13), equation (2.17) can be modified to express electrical length of a lossless homogenous transmission line by its primary parameters, inductance $L_i$ and capacitance $C_i$ per unit length:

$$\varphi_i = \beta l = \omega l \sqrt{L_i C_i} = 2\pi f l \sqrt{L_i C_i}.$$  \hspace{1cm} (2.19)

As can be seen in (2.19), the extent to which the phase of a signal is shifted when passing through a lossless homogenous transmission line is proportional to signal frequency $f$, physical length of the line $l$ and its inductance $L_i$ and capacitance $C_i$ per unit length.

The distributed phase-shifter circuit, shown in Fig. 2.1, comprises a high-impedance transmission line periodically loaded with varactor diodes of capacitance $C_{var}$ while physical spacing between the diodes equals to $l_{sp}$. A unit cell can be defined for this periodic structure. It consists of a section of the transmission line of length $l_{sp}$ and a shunt variable capacitor $C_{var}$ to ground.

Each such a transmission-line section can be approximated as lumped inductance $L_t$ and capacitance $C_t$, as shown in the equivalent circuit in Fig. 2.2, where

$$L_t = L_i l_{sp},$$  \hspace{1cm} (2.20)

$$C_t = C_i l_{sp}.$$  \hspace{1cm} (2.21)

Fig. 2.2: Varactor-loaded transmission-line equivalent circuit

For frequencies $f$ well below the Bragg frequency $f_B$, i.e. $f \ll f_B$, the periodically loaded line may be treated as a synthetic transmission line [22], depicted in Fig. 2.3, whose capacitance per unit length $C_t$ has been increased from the former value of $C_i$ due to the periodic loading with shunt capacitance.
Fig. 2.3: Synthetic transmission line described by voltage-controlled characteristic impedance and electrical length

$C_{\text{var}}$ [3]. The inductance per unit length $L_L$ for this synthetic line remains unchanged from that of the original unloaded line, $L_i$. Total inductance $L_L$ and capacitance $C_L$ per unit length of the designed synthetic transmission line are then found to be

$$L_L = L_i, \quad (2.22)$$

$$C_L = C_i + \frac{C_{\text{var}}}{l_{sp}}. \quad (2.23)$$

Based on relations (2.19), (2.22) and (2.23), the equivalent electrical length $\varphi_L$ of the varactor-loaded transmission line, which forms the synthetic transmission line, is given as

$$\varphi_L = \omega l \sqrt{L_L C_L} = 2\pi f l \sqrt{L_i \left( C_i + \frac{C_{\text{var}}}{l_{sp}} \right)}. \quad (2.24)$$

In equation (2.24), the line inductance $L_i$ and line capacitance $C_i$ are normalized per unit length. In assuming a synthetic transmission line, the discrete variable capacitance is essentially distributed over the length of the cell [3]. This is why all terms involving $C_{\text{var}}$ are divided by the spacing between varactor diodes, $l_{sp}$. This approach breaks down in the vicinity of the Bragg frequency where a more exact analysis must recognize the discrete nature of the loading [3]. The phenomenon of the Bragg frequency is discussed in section 2.4.

Capacitance $C_{\text{var}}$ depends on reverse bias voltage applied to corresponding varactor diodes [2]. Therefore, electrical length of the designed synthetic transmission line $\varphi_L$, which is proportional to $C_{\text{var}}$, can be controlled by tuning this voltage as is evident in equation (2.24). This is the phase shift
principle employed in a varactor-loaded transmission-line phase shifter the thesis focuses on.

To sum it up, phase difference between a signal at the input port and the same signal at the output port of a varactor-loaded transmission-line segment equals to its electrical length $\varphi_L$ that is voltage-controlled. The phase of the signal passing through the distributed phase shifter can thus be shifted according to the control voltage.

When varying the varactor diodes capacitance $C_{\text{var}}$ at any given frequency $f$ that is well below the Bragg frequency $f_B$, i.e. $f \ll f_B$, the maximum possible differential phase shift $\Phi_L$ that can be achieved using a varactor-loaded transmission line of given physical line length $l$ and diodes spacing $l_{sp}$ can be determined, based on (2.24), as follows

$$\Phi_L = \Delta \varphi_L = \varphi_L^{\text{max}} - \varphi_L^{\text{min}} =$$

$$= 2\pi f l \sqrt{L_i \left( \sqrt{C_i + \frac{C_{\text{var}}^{\text{max}}}{l_{sp}}} - \sqrt{C_i + \frac{C_{\text{var}}^{\text{min}}}{l_{sp}}} \right)},$$

(2.25)

where $C_{\text{var}}^{\text{max}}$ and $C_{\text{var}}^{\text{min}}$ indicate the maximum and the minimum capacitance of the varactor diode, respectively. $\varphi_L^{\text{max}}$ and $\varphi_L^{\text{min}}$ mean the maximum and the minimum corresponding equivalent electrical length of the varactor-loaded transmission line, respectively.

Besides its electrical length $\varphi_L$, there are other voltage-dependent properties of the varactor-loaded transmission line, such as its characteristic impedance $Z_L$ and phase velocity $v_L$, due to voltage-controlled capacitance $C_{\text{var}}$. On the basis of formulas (2.22) and (2.23), expressions (2.11) and (2.14) can be modified accordingly to reflect the line loading:

$$Z_L = \sqrt{\frac{L_L}{C_L}} = \sqrt{\frac{L_i}{C_i + \frac{C_{\text{var}}^{\text{max}}}{l_{sp}}}},$$

(2.26)

$$v_L = \frac{1}{\sqrt{L_L C_L}} = \frac{1}{\sqrt{L_i \left( C_i + \frac{C_{\text{var}}^{\text{max}}}{l_{sp}} \right)}}.$$  

(2.27)
2.4 Bragg Frequency

This section discusses the phenomenon of the Bragg frequency that affects the periodic structure of a distributed phase-shifter circuit and that needs to be taken into account in the varactor-loaded transmission-line phase-shifter design.

A result of creating a periodic structure that is shown in Fig. 2.1 is the existence of a cutoff frequency, called Bragg frequency \( f_B \), near the point where the guided wavelength \( \lambda_g \) approaches the periodic spacing of the discrete components \( l_{sp} \) [24].

At the Bragg frequency, the periodic filter structure of the distributed loaded line causes the line impedance \( Z_L \) to become zero [25] and thus almost total reflection occurs [26] so there is no power transfer from one port to the other [25]. For this reason, Bragg frequency has to be considered carefully to determine the upper operational frequency limit of the circuit [26].

Characteristic impedance \( Z_L \) of a loaded line that can be approximated by the lumped-element transmission-line model depicted in Fig. 2.2 is expressed [24], [26] by (2.28) as

\[
Z_L = \sqrt{\frac{L_t}{C_t + C_{var}}} \sqrt{1 - \left(\frac{\omega}{\omega_B}\right)^2},
\]

where \( \omega_B \) denotes the Bragg angular frequency and is given [24], [26] as

\[
\omega_B = \frac{2}{\sqrt{L_t(C_t + C_{var})}}.
\]

By a simple modification of equation (2.29) and by employing relations (2.20) and (2.21), the Bragg frequency \( f_B \) can be written [3], [27] as follows:

\[
f_B = \frac{\omega_B}{2\pi} = \frac{1}{\pi \sqrt{L_t(C_t + C_{var})}} = \frac{1}{\pi \sqrt{L_t l_{sp}(C_t l_{sp} + C_{var})}}.
\]

As can be seen in (2.30), the Bragg frequency \( f_B \) is inversely proportional to periodic spacing of the varactor diodes \( l_{sp} \), varactor diode capacitance \( C_{var} \) and to inductance and capacitance per unit length of an unloaded transmission line \( L_t \) and \( C_t \), respectively.
It is evident from (2.28) that for frequencies well below the Bragg frequency, i.e. $f \ll f_{B} \sim \omega \ll \omega_{B}$, expression (2.28) can be simplified to (2.26) using (2.20) and (2.21) as follows:

$$Z_{L} = \sqrt{\frac{L_{i}}{C_{t} + C_{\text{var}}}} \cdot \sqrt{1 - \left(\frac{\omega}{\omega_{B}}\right)^{2} \omega \ll \omega_{B}} = \sqrt{\frac{L_{i}}{C_{t} + C_{\text{var}}} \cdot \sqrt{1 - \left(\frac{\omega}{\omega_{B}}\right)^{2}}} = \sqrt{\frac{L_{i}}{C_{t} l_{sp} + C_{\text{var}}}} = \sqrt{\frac{L_{i}}{C_{t} + C_{\text{var}}}} l_{sp}.$$  \hspace{1cm} (2.31)

On the other hand, when frequency of operation reaches the Bragg frequency, i.e. $f \rightarrow f_{B} \sim \omega \rightarrow \omega_{B}$, characteristic impedance of a loaded transmission line $Z_{L}$ approaches zero as is demonstrated in formula (2.32).

$$Z_{L} = \sqrt{\frac{L_{i}}{C_{t} + C_{\text{var}}}} \cdot \sqrt{1 - \left(\frac{\omega}{\omega_{B}}\right)^{2} \omega \rightarrow \omega_{B}} \rightarrow 0$$  \hspace{1cm} (2.32)

The existence of the Bragg frequency thus imposes certain restrictions on the phase-shifter design. The distributed phase-shifter structure should be optimized to ensure that the Bragg frequency is much higher than the frequency of operation. Based on (2.30), one way to accomplish this is to reduce the periodic spacing of the varactor diodes $l_{sp}$.

In order to evaluate the Bragg frequency constraint for the purpose of the phase-shifter design, it is convenient to express Bragg frequency $f_{B}$ by means of varactor diode capacitance $C_{\text{var}}$, characteristic impedance of an unloaded transmission line $Z_{i}$, electrical length of an unloaded transmission-line section between varactor diodes $\varphi_{sp}$ and angular frequency $\omega$.

Using equations (2.15), (2.16), (2.9) and (2.19), such a dependence of the Bragg frequency $f_{B}$ can be derived from expression (2.30) and is written in formula (2.33).

$$f_{B} = \frac{1}{\pi \sqrt{L_{i} l_{sp}(C_{t} l_{sp} + C_{\text{var}})}} \overset{(2.15),(2.16)}{=} \frac{1}{\pi \sqrt{Z_{i} l_{sp} (l_{sp} Z_{i} + C_{\text{var}})}} \overset{(2.9)}{=} \frac{1}{\pi \sqrt{Z_{i} \varphi_{sp} (\varphi_{sp} Z_{i} + C_{\text{var}})}} \overset{(2.19)}{=} \frac{1}{\pi \sqrt{Z_{i} \varphi_{sp} (\varphi_{sp} Z_{i} + C_{\text{var}})}}$$  \hspace{1cm} (2.33)
The existence of the Bragg frequency, which is related to the periodic structure of a distributed phase-shifter circuit, can be proved by a simulation in a CAD circuit simulation program. Frequency characteristics of the transmission scattering coefficient $s_{21}$ of a phase-shifter circuit with ideal elements have been examined in AWR Microwave Office.

The simulated structure is the same as the one depicted in Fig. 3.1 where for a basic configuration, number of sections $N = 8$, varactor diodes capacitance $C_{\text{var}} = 2.3 \text{ pF}$, characteristic impedance $Z_i = 100 \Omega$ and electrical length $\varphi_{\text{sp}} = 16.8^\circ$ at $f = 1 \text{ GHz}$. Each time, one of these parameters is tuned and its effect on the $s_{21}$ frequency dependence is observed. The respective value of the Bragg frequency $f_B$ can then be determined as a significant fall of the magnitude of $s_{21}$.

Tab. 2.1: Analytically computed values of $f_B$ for different distributed phase-shifter circuits

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value 1</th>
<th>Value 2</th>
<th>Value 3</th>
<th>Value 4</th>
<th>Value 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>$N = 8$, $f = 1 \text{ GHz}$, $\varphi_{\text{sp}} = 16.8^\circ$, $Z_i = 100 \Omega$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$C_{\text{var}} \text{ [pF]}$</td>
<td>1.3</td>
<td>1.8</td>
<td>2.3</td>
<td>2.8</td>
<td>3.3</td>
</tr>
<tr>
<td>$f_B \text{ [GHz]}$</td>
<td>3.51</td>
<td>3.09</td>
<td>2.80</td>
<td>2.58</td>
<td>2.40</td>
</tr>
<tr>
<td>$N = 8$, $f = 1 \text{ GHz}$, $C_{\text{var}} = 2.3 \text{ pF}$, $Z_i = 100 \Omega$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$\varphi_{\text{sp}} \text{ [\degree]}$</td>
<td>12.8</td>
<td>14.8</td>
<td>16.8</td>
<td>18.8</td>
<td>20.8</td>
</tr>
<tr>
<td>$f_B \text{ [GHz]}$</td>
<td>3.28</td>
<td>3.02</td>
<td>2.80</td>
<td>2.62</td>
<td>2.47</td>
</tr>
<tr>
<td>$N = 8$, $f = 1 \text{ GHz}$, $C_{\text{var}} = 2.3 \text{ pF}$, $\varphi_{\text{sp}} = 16.8^\circ$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$Z_i \text{ [\Omega]}$</td>
<td>60</td>
<td>70</td>
<td>80</td>
<td>90</td>
<td>100</td>
</tr>
<tr>
<td>$f_B \text{ [GHz]}$</td>
<td>3.43</td>
<td>3.23</td>
<td>3.07</td>
<td>2.93</td>
<td>2.80</td>
</tr>
</tbody>
</table>

Results of the simulation can be seen in Fig. 2.4 through 2.6 and compared with respective theoretical values of $f_B$ computed analytically from relation (2.33). The computed values are recorded in Tab. 2.1 and indicate very good agreement with the simulated characteristics. Conclusions about the Bragg frequency phenomenon explored in this section are employed in section 3.1 during the phase-shifter design.
Fig. 2.4: The magnitude of $s_{21}$ versus frequency $f$ dependence of a distributed phase shifter with ideal elements for varactor diodes capacitance $C_{var}$ of 1.3 pF, 1.8 pF, 2.3 pF, 2.8 pF and 3.3 pF where number of sections $N = 8$, electrical length $\varphi_{sp} = 16.8^\circ$ at $f = 1$ GHz and characteristic impedance $Z_i = 100 \ \Omega$.

Fig. 2.5: The magnitude of $s_{21}$ versus frequency $f$ dependence of a distributed phase shifter with ideal elements for electrical length $\varphi_{sp}$ of 12.8°, 14.8°, 16.8°, 18.8° and 20.8° at $f = 1$ GHz where number of sections $N = 8$, varactor diodes capacitance $C_{var} = 2.3 \ \text{pF}$ and characteristic impedance $Z_i = 100 \ \Omega$. 
Fig. 2.6: The magnitude of $s_{21}$ versus frequency $f$ dependence of a distributed phase shifter with ideal elements for characteristic impedance $Z_i$ of 60 Ω, 70 Ω, 80 Ω, 90 Ω and 100 Ω where number of sections $N = 8$, varactor diodes capacitance $C_{var} = 2.3$ pF and electrical length $\varphi_{sp} = 16.8^\circ$ at $f = 1$ GHz.
2.5 Unit-Cell Matrix Description

The distributed phase-shifter structure can be described as a cascade of identical two-ports that each corresponds to a varactor-loaded transmission-line unit cell. Each such a unit cell can be further subdivided into a cascade of three elementary two-ports that make up the cell. This reflects the unit-cell configuration since it comprises a transmission-line section of length $l_{sp}$ that is split into two equal parts by a shunt varactor diode to ground of variable capacitance $C_{var}$, as is depicted in Fig. 2.7.

![Fig. 2.7: Varactor-loaded transmission-line unit cell](image)

With respect to the above mentioned structure, it is desirable to represent each two-port in the cascade by the corresponding chain matrix [28]. The resulting chain matrix of the two-port cascade then simply equals to the matrix product of the component chain matrices [29] as is employed in equation (2.37). For the purpose of the distributed phase-shifter design, chain matrix of the distributed phase-shifter unit cell $[A_u]$ will be determined and converted into the corresponding scattering matrix $[S_u]$.

A chain matrix $[A_g]$ of a homogenous transmission-line section of length $\frac{l_{sp}}{2}$ is found [30] to be

$$[A_g] = \begin{bmatrix}
\cosh \frac{\gamma l_{sp}}{2} & Z_i \sinh \frac{\gamma l_{sp}}{2} \\
\frac{1}{Z_i} \sinh \frac{\gamma l_{sp}}{2} & \cosh \frac{\gamma l_{sp}}{2}
\end{bmatrix}, \quad (2.34)$$

where $Z_i$ stands for characteristic impedance of the unloaded transmission line and $\gamma$ denotes propagation constant of the line.

A chain matrix $[A_{var}]$ of a shunt varactor diode to ground of capacitance $C_{var}$, which loads the line, is given [31] as
\[
[A_{\text{var}}] = \begin{bmatrix} 1 & 0 \\ Y_{\text{var}} & 1 \end{bmatrix}, \tag{2.35}
\]

where \(Y_{\text{var}}\) means tunable varactor-diode admittance for given angular frequency \(\omega\), which is expressed \([32]\) by formula (2.36).

\[
Y_{\text{var}} = j\omega C_{\text{var}} \tag{2.36}
\]

A chain matrix \([A_u]\) of the distributed phase-shifter unit cell that consists of the three elementary two-ports—two homogenous transmission-line sections of length \(l_{sp}/2\) and the shunt varactor diode to ground in between—put in a cascade can then be determined by the matrix product of the respective chain matrices \([29]\)—\([A_g]\), \([A_{\text{var}}]\) and \([A_g]\)—as

\[
[A_u] = [A_g] \cdot [A_{\text{var}}] \cdot [A_g] = \\
\begin{bmatrix}
\cosh \gamma l_{sp}^{\text{var}} & Z_s \sinh \gamma l_{sp}^{\text{var}} \\
\frac{1}{Z_s} \sinh \gamma l_{sp}^{\text{var}} & \cosh \gamma l_{sp}^{\text{var}}
\end{bmatrix}
\begin{bmatrix} 1 & 0 \\ Y_{\text{var}} & 1 \end{bmatrix}
\begin{bmatrix}
\frac{1}{Z_s} \sinh \gamma l_{sp}^{\text{var}} & \cosh \gamma l_{sp}^{\text{var}} \\
\cosh \gamma l_{sp}^{\text{var}} & Z_s \sinh \gamma l_{sp}^{\text{var}}
\end{bmatrix} = \begin{bmatrix} a_{u11} & a_{u12} \\ a_{u21} & a_{u22} \end{bmatrix}, \tag{2.37}
\]

where

\[
a_{u11} = \cos^2 \frac{\varphi_{sp}}{2} - \sin^2 \frac{\varphi_{sp}}{2} + j Y_{\text{var}} Z_s \sin \frac{\varphi_{sp}}{2} \cos \frac{\varphi_{sp}}{2}, \tag{2.38}
\]

\[
a_{u12} = -Y_{\text{var}} Z_s^2 \sin^2 \frac{\varphi_{sp}}{2} + 2 j Z_s \sin \frac{\varphi_{sp}}{2} \cos \frac{\varphi_{sp}}{2}, \tag{2.39}
\]

\[
a_{u21} = Y_{\text{var}} \cos^2 \frac{\varphi_{sp}}{2} + \frac{2 j}{Z_s} \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2}, \tag{2.40}
\]

\[
a_{u22} = \cos^2 \frac{\varphi_{sp}}{2} - \sin^2 \frac{\varphi_{sp}}{2} + j Y_{\text{var}} Z_s \sin \frac{\varphi_{sp}}{2} \cos \frac{\varphi_{sp}}{2}. \tag{2.41}
\]

Variable \(\varphi_{sp}\), which is present e. g. in equations (2.38)–(2.41), stands for electrical length of the unloaded transmission-line section between the varactor diodes.
In order to evaluate a contribution of the unit cell to differential phase shift, return loss and insertion loss of the whole phase-shifter circuit, it is useful to convert its chain matrix \([A_u]\) into the corresponding scattering matrix \([S_u]\) by formulas (A.1)–(A.5) that are recorded in Appendix A.

The resulting scattering matrix of the distributed phase-shifter unit cell \([S_u]\) is given by equations (2.42)–(2.52), where \(Z_{in}\) and \(Z_{out}\) denote reference impedance at the input and the output port, respectively.

\[
[S_u] = \begin{bmatrix}
    s_{u11} & s_{u12} \\
    s_{u21} & s_{u22}
\end{bmatrix} \tag{2.42}
\]

\[s_{u11} = \frac{s_{u11a}}{s_{u11b}} \tag{2.43}\]

\[
s_{u11a} = 2Z_{out}Z_i \cos^2 \frac{\varphi_{sp}}{2} + jY_{var}Z_{out}Z_i^3 \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} - Z_{out}Z_i + \\
  + 2jZ_i^2 \sin \frac{\varphi_{sp}}{2} \cos \frac{\varphi_{sp}}{2} - Y_{var}Z_i^3 + Y_{var}Z_i^3 \cos^2 \frac{\varphi_{sp}}{2} - \\
  - 2jZ_{in}^*Z_{out} \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} - Y_{var}Z_{in}^*Z_{out}Z_i \cos^2 \frac{\varphi_{sp}}{2} + Z_{in}^*Z_i - \\
  - 2Z_{in}^*Z_i \cos^2 \frac{\varphi_{sp}}{2} - jY_{var}Z_{in}^*Z_i^2 \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} \tag{2.44}\]

\[s_{u11b} = 2Z_i^2 \sin \frac{\varphi_{sp}}{2} \cos \frac{\varphi_{sp}}{2} + jY_{var}Z_{in}Z_i^2 \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} + \\
  + jY_{var}Z_{out}Z_i^2 \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} + 2jZ_{in}Z_{out} \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} + \\
  + 2Z_{out}Z_i \cos^2 \frac{\varphi_{sp}}{2} + Y_{var}Z_{in}Z_{out}Z_i \cos^2 \frac{\varphi_{sp}}{2} + 2Z_{in}Z_i \cos^2 \frac{\varphi_{sp}}{2} - \\
  - Z_{out}Z_i - Z_{in}Z_i + Y_{var}Z_i^3 \cos^2 \frac{\varphi_{sp}}{2} - Y_{var}Z_i^3 \tag{2.45}\]

\[s_{u12} = \frac{s_{u12a}}{s_{u12b}} \tag{2.46}\]

\[s_{u12a} = 2Z_i \sqrt{\Re{\{Z_{in}\}} \Re{\{Z_{out}\}}} \tag{2.47}\]

\[s_{u12b} = s_{u11b} \tag{2.48}\]
\[ s_{u21} = s_{u12} \quad (2.49) \]
\[ s_{u22} = \frac{s_{u22a}}{s_{u22b}} \quad (2.50) \]
\[ s_{u22a} = -2Z_{\text{out}}^* Z_i \cos^2 \frac{\varphi_{sp}}{2} - jY_{\text{var}} Z_{\text{out}}^* Z_i^2 \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} + Z_{\text{out}}^* Z_i^+ \]
\[ + 2jZ_i^2 \sin \frac{\varphi_{sp}}{2} \cos \frac{\varphi_{sp}}{2} - Y_{\text{var}} Z_i^3 + Y_{\text{var}} Z_i^3 \cos^2 \frac{\varphi_{sp}}{2} - \]
\[ - 2jZ_{\text{in}} Z_{\text{out}}^* \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} - Y_{\text{var}} Z_{\text{in}} Z_{\text{out}}^* Z_i \cos^2 \frac{\varphi_{sp}}{2} + Z_{\text{in}} Z_i^+ \]
\[ + 2Z_{\text{in}} Z_i \cos^2 \frac{\varphi_{sp}}{2} + jY_{\text{var}} Z_{\text{in}} Z_i^2 \cos \frac{\varphi_{sp}}{2} \sin \frac{\varphi_{sp}}{2} \quad (2.51) \]
\[ s_{u22b} = s_{u11b} \quad (2.52) \]

As can be seen in relations (2.42)–(2.52), transmission scattering coefficient of the unit cell \( s_{u21} \) is a function of tunable varactor-diode admittance \( Y_{\text{var}} \), which is proportional to varactor diode capacitance \( C_{\text{var}} \) as is obvious from (2.36). The transition between two \( Y_{\text{var}} \) admittance states thus produces a change in the phase of the transmission scattering coefficient \( s_{u21} \) [33].

Therefore, the maximum differential phase shift of the distributed phase-shifter unit cell \( \Phi_u \) for given angular frequency \( \omega \) is determined as

\[ \Phi_u = \Delta \arg s_{u21} = | \arg s_{u21}(C_{\text{var}}^{\text{max}}) - \arg s_{u21}(C_{\text{var}}^{\text{min}}) |, \quad (2.53) \]

where \( C_{\text{var}}^{\text{max}} \) and \( C_{\text{var}}^{\text{min}} \) indicate the maximum and the minimum capacitance of the varactor diode, respectively.

When the whole phase-shifter circuit is assembled, i.e., \( N \) its unit cells are put in a cascade, the maximum differential phase shift of the varactor-loaded transmission-line phase shifter \( \Phi_L \) is found to be

\[ \Phi_L = N \cdot \Phi_u, \quad (2.54) \]

where the number \( N \) is determined during the phase-shifter design in order to achieve the goal of \( \Phi_L \) by the circuit.

Finally, return loss \( RL_u \) and insertion loss \( IL_u \) of the distributed phase-shifter unit cell can be counted by formulas (2.55) and (2.56), respectively, where \( |s_{u11}| \) and \( |s_{u21}| \) represent respective magnitudes of \( s_{u11} \) and \( s_{u21} \).
\[ RL_u = -20 \log |s_{u11}| \quad (2.55) \]
\[ IL_u = -20 \log |s_{u21}| \quad (2.56) \]

### 2.6 Capacitance Diodes Basics

The main function of capacitance diodes is to provide variable capacitance and there are several means how to implement it. These include a PN junction depletion layer, a metal-semiconductor junction or a MIS structure [2]. Capacitance diodes are known as varicaps or varactors.

In practice, the PN junction depletion layer capacitance is the most used capacitance diode technique. Since the width of a reverse biased PN junction depletion layer is voltage dependent and the capacitance of the depletion layer is inversely proportional to its width, the desired capacitance can thus be adjusted by an appropriate reverse bias voltage applied to a diode [2]. Reverse biased PN junction capacitance is characterized by low temperature dependence, low noise and by frequency independence reaching as high as the millimeter-wave bands [2].

The PN junction depletion layer capacitance is referred to as the barrier capacitance. Equation (2.57) shows the dependence of barrier capacitance \( C_j \) on reverse bias voltage \( V \) applied to the junction [2]:

\[ C_j = C_{j0} \left( 1 + \frac{V}{V_0} \right)^{-n}, \quad (2.57) \]

where \( C_{j0} \) stands for zero-bias PN junction capacitance, \( V_0 \) is the PN junction diffusion voltage and \( n \) is the PN junction profile exponent.

The applications of capacitance diodes benefit from the ability to control the capacitance by the applied voltage. Capacitance diodes are therefore key elements to a number of subsystems, namely harmonic signals generators, parametric amplifiers, mixers, detectors, frequency modulators, phase shifters, tunable resonant circuits with voltage-controlled resonant frequency etc [2].

A typical equivalent circuit of packaged capacitance diodes is shown in Fig. 2.8. It consists of PN junction resistance \( R_j \), PN junction capacitance \( C_j \), series resistance \( R_s \) of the semiconductor material and ohmic contacts, series
inductance $L_s$ of metal contact wires and package capacitance $C_p$. Tab. 2.2 lists typical values of these elements as reported in [2].

The varactor-loaded transmission-line phase shifter is to be implemented using the BB857 varactor diodes, manufactured by Infineon Technologies AG. Typical values of the BB857 varactor diode chip and the related SCD80 package equivalent circuit elements are recorded in Tab. 2.3 based on data provided by the manufacturer [34]. The manufacturer declares the RF-package parasitics equivalent circuit to be valid up to 6 GHz [34]. Detailed characteristics and parameters of the BB857 varactor diode are recorded in Appendix B [34].

Tab. 2.2: Typical values of the packaged capacitance diode equivalent circuit elements as reported in [2]

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_j$</td>
<td>$10^4 \Omega$–$10^8 \Omega$</td>
<td>PN junction resistance</td>
</tr>
<tr>
<td>$C_j$</td>
<td>0.1 pF–1 µF</td>
<td>PN junction capacitance</td>
</tr>
<tr>
<td>$R_s$</td>
<td>0.1 Ω–10 Ω</td>
<td>Series resistance</td>
</tr>
<tr>
<td>$L_s$</td>
<td>1 nH–10 nH</td>
<td>Series inductance</td>
</tr>
<tr>
<td>$C_p$</td>
<td>approx. 1 pF</td>
<td>Package capacitance</td>
</tr>
</tbody>
</table>

Tab. 2.3: Typical values of the BB857 varactor diode and the related SCD80 package equivalent circuit elements based on data provided by the manufacturer [34]

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_j$</td>
<td>0.52 pF–6.6 pF</td>
<td>PN junction capacitance</td>
</tr>
<tr>
<td>$R_s$</td>
<td>1.5 Ω</td>
<td>Series resistance</td>
</tr>
<tr>
<td>$L_s$</td>
<td>0.6 nH</td>
<td>Series inductance</td>
</tr>
<tr>
<td>$C_p$</td>
<td>90 fF</td>
<td>Package capacitance</td>
</tr>
</tbody>
</table>
The series resistance $R_s$ is sometimes assumed to be constant with reverse bias voltage and frequency in order to simplify use of the model. However, additional losses often result in a significantly higher value for $R_s$ at microwave frequencies. Also, $R_s$ decreases with increasing bias voltage as the width of the depletion region around the PN junction increases, reducing the length of the conductive path through the bulk semiconductor material surrounding the junction [37].
Chapter 3

Phase-Shifter Design and Optimization

3.1 Design Objectives and Constraints

This section presents basic objectives of the phase-shifter design as well as constraints imposed by technological parameters of employed design elements. These constraints and trade-offs are then discussed in order to find optimal implementation of a varactor-loaded transmission-line phase shifter.

The phase shifter is to be designed to operate within a one-octave frequency band whose center frequency is 1 GHz or higher. A band that spans from 0.7 GHz to 1.4 GHz has been chosen for this purpose. The maximum differential phase shift of the circuit should be no less than $90^\circ$ within the whole frequency band and return loss of no less than 10 dB should also be achieved. Moreover, it is desirable that the phase shifter be implemented by simple and inexpensive technique. These phase-shifter design objectives are summarized in Tab. 3.1.

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency band of operation</td>
<td>$0.7 \text{ GHz} \leq f \leq 1.4 \text{ GHz}$</td>
</tr>
<tr>
<td>Maximum differential phase shift</td>
<td>$\Phi_L \geq 90^\circ$</td>
</tr>
<tr>
<td>Return loss</td>
<td>$RL \geq 10 \text{ dB}$</td>
</tr>
<tr>
<td>Implementation technique</td>
<td>simple and inexpensive</td>
</tr>
</tbody>
</table>

Key parameters that the design is based upon comprise tunable varactor diode capacitance $C_{var}$, characteristic impedance of an unloaded transmission
line $Z_t$ and electrical length of an unloaded transmission-line section between
varactor diodes $\varphi_{sp}$, which can be converted to physical length of varactor
diodes spacing $l_{sp}$ using relation (2.17). Criteria for an optimal choice of their
values are given by the design objectives as well as by the Bragg frequency
condition that is explained in section 2.4.

Since Bragg frequency $f_B$ represents a cutoff frequency of the distributed
phase-shifter periodic structure, the circuit has to be designed and optimized
carefully to ensure that the Bragg frequency is much higher than the max-
imum frequency of operation $f_{max}$. Generally, the operational frequency is
approximately half of the Bragg frequency [26]. This condition can than be
written as follows:

$$f_B \geq 2 \cdot f_{max}.$$  (3.1)

Contribution of the key design parameters to the phase-shifter circuit
characteristics is discussed below. Also, technological constraints, which limit
the choice of their values, are presented.

**Varactor diode capacitance $C_{var}$.** As is evident from expression (2.25),
the wider the range of achievable varactor diode capacitances $C_{var}$—
i. e. the greater the difference between $C_{var}^{max}$ and $C_{var}^{min}$, the higher
the maximum differential phase shift of a varactor-loaded transmission
line $\Phi_L$.

On the other hand, analytical computations based on formulas (2.42)–
(2.52) show that high $C_{var}$ values lead to high reflections of an incident
signal and thus to unacceptably low return loss $RL$. The cause of such
behavior is partially visible from relation (2.26). Adjusting $C_{var}$ to
high values alters greatly the characteristic impedance of a synthetic
line $Z_L$ and thus leads to significant impedance mismatch with the
phase-shifter input and output port impedance.

Another fact which imposes an upper limit of the varactor diode capac-
itances range is the dependence of the Bragg frequency $f_B$ on $C_{var}$ as
expressed in formula (2.33). The higher the varactor diode capacitance
$C_{var}$ used, the lower the resulting Bragg frequency $f_B$, as can be seen
in Fig. 2.4 and Tab. 2.1.

As far as technological constraints are concerned, the BB857 varactor
diodes, manufactured by Infineon Technologies AG, which are to be
used for the circuit fabrication, provide the $C_{var}$ values that range from
0.52 pF to 6.6 pF [34]. Detailed characteristics and parameters of the
BB857 varactor diode are recorded in Appendix B [34].
**Characteristic impedance** $Z_i$. If a transmission line is periodically loaded with varactor diodes, characteristic impedance of the resulting synthetic transmission line $Z_L$ is proportionally lower in comparison to characteristic impedance of the unloaded transmission line $Z_i$ as can be seen from equations (2.11) and (2.26). In order to keep the value of $Z_L$ in the vicinity of the phase-shifter input and output port impedance and thus to avoid unacceptable impedance mismatch, proportionally higher characteristic impedance of the unloaded line $Z_i$ has to be used [3].

Analytical computations based on formulas (2.42)–(2.52) demonstrate that higher values of $Z_i$ allow for wider range of capacitances $C_{\text{var}}$ to be employed—and thus providing higher maximum differential phase shift $\Phi_L$ as mentioned above—while still keeping the return loss of the circuit acceptable. It has also been proved by these computations that higher $Z_i$ values provide steeper slope of the $s_{21}$ phase dependence on tunable capacitance $C_{\text{var}}$ and thus guarantee higher differential phase shift $\Phi_L$ for the same $C_{\text{var}}$ range.

On the other hand, higher values of $Z_i$ reduce the resulting Bragg frequency $f_B$ in accordance with equation (2.33), as is demonstrated in Fig. 2.6 and Tab. 2.1. In addition, implementation of higher values of $Z_i$ by planar technology may be more challenging in terms of fabrication precision.

The phase shifter is to be implemented in the microstrip transmission-line structure. The microstrip structure will be fabricated on the RO4350B substrate, manufactured by Rogers Corporation. The available technology provides a range of potential $Z_i$ values that spans from 10 Ω to 100 Ω. Detailed characteristics and parameters of the RO4350B substrate are recorded in Appendix C [35].

$$w < \frac{\lambda_g}{2}$$  \hspace{1cm} (3.2)

Width of a microstrip transmission line $w$ has to be designed in accordance with formula (3.2) to prevent higher-order modes from being generated in the microstrip structure [36]. Taking into account that $Z_i$ of a microstrip transmission line is inversely proportional to $w$ [36], condition (3.2) imposes the lower limit of $Z_i$.

Since a guided wavelength on a transmission line $\lambda_g$ is inversely proportional to frequency $f$, as can be proven by combining relations (2.18) and (2.13), compliance of $w$ with condition (3.2) should be calculated
at $f = f_{\text{max}}$. In addition, computational software of AWR Corporation recommends that
\[
\frac{w}{h} \leq 20. \quad (3.3)
\]
Smaller values of the $w/h$ fraction decrease an approximation error when characteristics of a microstrip structure are computed [36]. For a microstrip structure whose physical parameters are listed in Tab. 3.10, the lower limit of $Z_i$ is then, with respect to the above mentioned facts, established as 10 $\Omega$.

Analogically, the maximum value of $Z_i$ is limited by the minimum $w$ that can be fabricated. The available technology provides the minimum value of $w$ equal to 0.2 mm. However, a transmission line fabricated at this limit would likely be inferior in terms of precision of its $w$ parameter and hence the designed characteristic impedance $Z_i$ would not be guaranteed.

For this reason, it seems convenient to increase the minimum permitted value of $w$. Width $w$ of 0.376 mm is sufficiently higher, and for the microstrip structure described in Tab. 3.10, it produces the transmission-line characteristic impedance $Z_i$ of 100 $\Omega$, which in turn represents the upper limit of $Z_i$.

**Electrical length $\varphi_{sp}$.** Electrical length of an unloaded transmission-line section between varactor diodes $\varphi_{sp}$—convertable to physical length of varactor diodes spacing $l_{sp}$ using relation (2.17)—determines density of the transmission-line loading by varactor diodes. According to formula (2.25), smaller values of the varactor diodes spacing $l_{sp}$ lead to greater tunability of the electrical length of a varactor-loaded transmission line $\varphi_L$ by means of $C_{\text{var}}$ change and thus to greater achievable differential phase shift $\Phi_L$. As a result, circuit dimensions can be minimized.

Moreover, smaller values of $l_{sp}$ or corresponding $\varphi_{sp}$ provide better compliance with the Bragg frequency constraint written in relation (3.1) as can be deduced from equation (2.33) and seen in Fig. 2.5 and Tab. 2.1.

On the other hand, when bigger spacing between varactor diodes $l_{sp}$ is chosen, a smaller total number of the diodes has to be employed. Since each varactor diode comprises a certain amount of loss, the configuration with fewer diodes results in lower insertion loss $IL$. 

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As can be seen, the choice of the values that are to be employed in the phase-shifter design is a trade-off between several contradictory constraints. Constraints which are imposed by technological and functional limitations are summarized in Tab. 3.2.

Tab. 3.2: Technological and functional constraints of the phase-shifter design

<table>
<thead>
<tr>
<th>Constraint</th>
<th>Constraint Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Varactor diode capacitance</td>
<td>$0.52 \text{ pF} \leq C_{\text{var}} \leq 6.6 \text{ pF}$</td>
</tr>
<tr>
<td>Characteristic impedance of the unloaded trans. line</td>
<td>$10 \Omega \leq Z_i \leq 100 \Omega$</td>
</tr>
<tr>
<td>Bragg frequency</td>
<td>$f_B \geq 2 \cdot f_{\text{max}}$</td>
</tr>
</tbody>
</table>

To sum it up, from the functional point of view and with respect to the design objectives, it appears convenient to employ as high value of the unloaded-line characteristic impedance $Z_i$ as possible to allow wider range of variable capacitance $C_{\text{var}}$ and thus higher achievable differential phase shift $\Phi_L$ while still maintaining acceptable circuit return loss $RL$.

Physical spacing of the varactor diodes $l_{\text{sp}}$—or electrical length of the unloaded transmission-line section between varactor diodes $\phi_{\text{sp}}$—needs to be chosen carefully to balance between insertion loss $IL$ of the circuit and the Bragg frequency constraint. This constraint can also be better met by limiting the maximum value of $C_{\text{var}}$ employed.

3.2 Design of the Phase Shifter with Ideal Elements

Section 3.1 discussed contribution of the key design parameters to the phase-shifter circuit characteristics trends and presented related technological constraints. This section aims to determine particular values of the design parameters in order to make up the whole 90° phase-shifter circuit that meets both the objectives depicted in Tab. 3.1 and the constraints summarized in Tab. 3.2. Ideal elements, namely ideal variable capacitors, ideal transmission lines and ideal grounds, will be used for the purpose of the design in this section.

**Characteristic impedance** $Z_i$. With respect to conclusions of section 3.1, as high value of the unloaded-line characteristic impedance $Z_i$ as possible will be employed in the design. This will provide wider range of variable capacitance $C_{\text{var}}$ and steeper slope of the $s_{21}$ phase dependence on tunable capacitance $C_{\text{var}}$. 

28
As a result, higher achievable differential phase shift $\Phi_L$ will be achieved while still maintaining acceptable circuit return loss $RL$. As is stated in section 3.1, the upper limit of $Z_i$ imposed by the available technology is 100 $\Omega$. This very value of $Z_i$ will therefore be used in the phase-shifter design.

**Varactor diode capacitance $C_{\text{var}}$.** In order to avoid unacceptable, i.e. lower than 10 dB, circuit return loss $RL$, the maximum capacitance of the varactor diodes $C_{\text{var}}^{\text{max}}$ needs to be limited. Another beneficial effect of the lower $C_{\text{var}}^{\text{max}}$ value is better compliance with the Bragg frequency condition. These mechanisms are further explained in section 3.1.

Based on circuit simulations, previous theses and the BB857 varactor diode C-V dependence, it seems convenient to choose an empirical value of 2.3 pF as the maximum capacitance of the varactor diodes $C_{\text{var}}^{\text{max}}$. The lower limit of the varactor diodes capacitance $C_{\text{var}}$, i.e. $C_{\text{var}}^{\text{min}}$, remains unchanged and equals to 0.52 pF.

**Electrical length $\varphi_{\text{sp}}$.** When the optimal values of the characteristic impedance $Z_i$ and the maximum capacitance $C_{\text{var}}^{\text{max}}$ are chosen, the third key design parameter—maximum electrical length of the unloaded transmission-line section between varactor diodes $\varphi_{\text{sp}}^{\text{max}}$—can be computed by formulas (2.33) and (3.1) provided that the Bragg frequency condition, expressed in (3.1), has to be met.

For $Z_i = 100 \, \Omega$, $C_{\text{var}}^{\text{max}} = 2.3$ pF and $f_{\text{max}} = 1.4$ GHz, the resulting maximum electrical length $\varphi_{\text{sp}}^{\text{max}}$ of one phase-shifter section equals to 16.8° at the center frequency of operation $f = 1$ GHz. This upper limit of $\varphi_{\text{sp}}$ will be employed in the phase-shifter design. The derived optimal values of the phase-shifter design parameters are recapitulated in Tab. 3.3.

<table>
<thead>
<tr>
<th>Tab. 3.3: Optimal values of the phase-shifter design parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Varactor diode capacitance</td>
</tr>
<tr>
<td>Characteristic impedance of the unloaded trans. line</td>
</tr>
<tr>
<td>Electrical length of one phase-shifter section</td>
</tr>
<tr>
<td>Number of sections</td>
</tr>
</tbody>
</table>

The last step needed to complete a set of the phase-shifter design parameters inhere in a determination of the number $N$ of the phase-shifter sections or unit cells that are to be put in a cascade. The criterion for such a decision
is given by the maximum differential phase shift \( \Phi_L \) that can be achieved by the whole circuit and that is proportional to \( N \). This mechanism is expressed in equation (2.54).

One of the phase-shifter design objectives, presented in section 3.1, is to guarantee that \( \Phi_L \) will be no less than 90° within the whole frequency band of operation. Therefore, the number \( N \) should be chosen properly to achieve this goal. At the same time, \( N \) should be as low as possible to keep the resulting circuit simple, compact and inexpensive.

As is evident from formula (2.53), differential phase shift of the distributed phase-shifter unit cell \( \Phi_u \) for given angular frequency \( \omega \) is accomplished by tuning varactor diode capacitance \( C_{var} \). The particular value of \( \Phi_u \) can then be either analytically computed employing equations (2.42)–(2.52) or determined by simulation in a CAD circuit simulation program.

The latter is used in this thesis. Several phase-shifter circuits with ideal elements and incrementing number of identical sections have been examined in AWR Microwave Office. The goal was to determine the minimum number of sections \( N \) which are necessary to achieve the 90° phase shift in accordance with relation (2.54).

![Fig. 3.1: Minimum number of phase-shifter sections determination](image)

For the purpose of these simulations, each identical phase-shifter section has been assembled from ideal elements whose parameters—\( C_{var} \), \( Z_i \) and \( \varphi_{sp} \)—are summarized in Tab. 3.3. Subsequently, characteristics of a total of twelve circuits have been examined provided that the \( N \)-th circuit consists of \( N \) identical phase-shifter sections. This approach is depicted in Fig. 3.1 and results of the simulations are recorded in Tab. 3.4.

As can be deduced from equations (2.53) and (2.54), maximum differential phase shift \( \Phi_L \) corresponds to the difference between arguments of the transmission scattering coefficient \( s_{21} \) at \( C_{var}^{min} \) and \( C_{var}^{max} \). The varactor diodes capacitance was therefore tuned within the designed range of the \( C_{var}^{min} \) and \( C_{var}^{max} \) values and the respective value of \( \Phi_L \) has been recorded.

30
It is essential that the resulting $\Phi_L$ of the circuit be examined at the minimum frequency of operation $f_{\text{min}}$, which represents the worst case, since maximum differential phase shift $\Phi_L$ is proportional to frequency $f$, as can be seen in formula (2.25). Data in Tab. 3.4 then show that the minimum number of sections $N$ needed to achieve the $90^\circ$ phase shift is 8 and that for $N$ equal to 1 through 8, the whole range of $C_{\text{var}}$ presented in Tab. 3.3 has to be exploited.

Tab. 3.4: Simulated characteristics of distributed phase shifters composed of $N$ sections where $\Phi_L$ is the maximum differential phase shift at $f_{\text{min}} = 0.7$ GHz. $C_{\text{var}}^{\text{max}}$ and $C_{\text{var}}^{\text{min}}$ indicate the maximum and the minimum employed capacitance of varactor diodes, respectively, and $RL_{\text{min}}$ stands for minimum return loss of the circuit.

<table>
<thead>
<tr>
<th>$N$</th>
<th>$\Phi_L$ [°]</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>11.53</td>
<td>22.78</td>
<td>33.92</td>
<td>45.40</td>
<td>57.80</td>
<td>70.80</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>2.3</td>
<td>2.3</td>
<td>2.3</td>
<td>2.3</td>
<td>2.3</td>
<td>2.3</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td>&gt;10</td>
<td></td>
</tr>
</tbody>
</table>

Besides maximum differential phase shift $\Phi_L$, minimum return loss $RL_{\text{min}}$ of each of the circuits with different number of sections $N$ can be obtained from the simulations. The return loss was examined within the designed range of the $C_{\text{var}}^{\text{max}}$ and $C_{\text{var}}^{\text{min}}$ values at three significant frequencies, $f = f_{\text{min}} = 0.7$ GHz, $f = f_0 = 1$ GHz and $f = f_{\text{max}} = 1.4$ GHz.

When the number of sections $N$ increases, the range of $C_{\text{var}}$ needed to provide the $90^\circ$ phase shift can be narrower. As a result, characteristic impedance of the synthetic transmission line $Z_L$ is less altered from the state of impedance match and the return loss $RL$ is thus higher. This effect is visible in Tab. 3.4 for $N$ equal to 9 through 12 and the corresponding $RL$ versus $N$ dependence is plotted in Fig. 3.2.

Unlike an ideal variable capacitor, employed so far in this section, a real varactor diode contains, among others, a nonzero series resistance $R_s$ [2], which contributes to the diode losses and consequently to the overall circuit insertion loss $IL$.

Therefore, if every ideal variable capacitor in the simulation schematics is replaced by a varactor-diode equivalent circuit, which is covered in
sections 2.6 and 3.3, maximum insertion loss $IL_{\text{max}}$ of a distributed phase-shifter circuit can be obtained for each number of sections $N$. For example, the simulations show that $IL_{\text{max}}$ equals to 0.83 dB, 1.87 dB and 2.39 dB for two, four and six sections, or unit cells, respectively.

![Simulated return loss $RL$ of distributed phase-shifter circuits versus number of their sections $N$](image)

In conclusion, the number $N$ of eight phase-shifter sections has been chosen as an optimal value for the phase-shifter design. This value seems just right in order to meet the objective of the $90^\circ$ phase-shift and to balance between the circuit return loss $RL$ on one side and the circuit insertion loss $IL$, its complexity and dimensions on the other side, all of which being proportional to $N$. This value of $N$ has thus completed the set of the optimal phase-shifter design parameters, written in Tab. 3.3, which lays the foundations for further phase-shifter circuit characteristics examination.

This section explored the phase-shifter design employing mainly ideal circuit elements. In order to simulate behavior of the real circuit more precisely, real elements should be taken into account. Section 3.3 presents an equivalent circuit of a BB857 varactor diode and via holes positioned between the diode and a ground plane. Section 3.5 covers recomputation of the ideal transmission lines to real microstrip transmission lines including discontinuities such as bends and steps in the lines width. Section 4.2 then presents final simulation results based on modeling of the entire phase-shifter circuit schematic with real elements in comparison with the measured data.
3.3 BB857 Varactor-Diode Modeling

Inaccurate elements description or missing elements parasitics may lead to imprecise results of microwave systems design because of significant differences between the designed circuit simulations and measurements of the manufactured circuits. Care must be taken to rely on as precise element models as possible in order to obtain acceptable agreement between simulated and measured data.

This section focuses on developing a nonlinear model of a BB857 varactor diode, manufactured by Infineon Technologies AG, that would precisely simulate the real element within the frequency range of 0.5 GHz through 3 GHz for nine significant diode bias voltages. The diode package parasitics and the chip bias-dependent character are taken into account as well. Moreover, the model also involves inductive character of via holes positioned between the diode and a ground plane.

The resulting model is then used to enhance the precision of the varactor-loaded transmission-line phase-shifter design, particularly in section 3.5. And it is partly employed in section 3.2 as well. Since the phase-shifting feature in such a circuit is implemented by a varactor-diode adjustable capacitance, the precise varactor-diode modeling is crucial to the phase shifter design and attention will especially be paid to achieve sufficient agreement between measured and modeled data in terms of the reflection scattering coefficient $s_{11}$ phase.

![Varactor-Diode Model](image)

The manufacturer of the BB857 varactor diodes suggests that the chip itself be simulated by a SPICE diode model and the parasitics, including the package parasitics, by reactance elements put around the chip. The entire
model schematic is shown in Fig. 3.3. The SPICE parameters of the chip and the parasitics data provided by the manufacturer are then listed in Tab. 3.5 and Tab. 3.6, respectively, accompanied by necessary explanation. According to the manufacturer, the model should be valid up to 6 GHz [34].

In the model displayed in Fig. 3.3, inductances $L_{s1}$, $L_{s2}$ and $L_{s3}$ form overall series inductance of the diode $L_s$ and capacitance $C_p$ stands for the diode package capacitance. Parallel capacitance $C_{cv}$ and a very high value for the bottom built-in voltage are applied for better C-V curve approximation between reverse bias voltages of 0.5 V through 28 V [34]. The SPICE diode model has been implemented by the SDIODE element in the CAD circuit simulation program, AWR Microwave Office.

Tab. 3.5: The BB857 chip model data provided by the manufacturer [34]

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_s$</td>
<td>1.35 fA</td>
<td>Reverse saturation current</td>
</tr>
<tr>
<td>$R_S$</td>
<td>1.5 Ω</td>
<td>Series resistance</td>
</tr>
<tr>
<td>$N_N$</td>
<td>1.074</td>
<td>Bottom ideality factor</td>
</tr>
<tr>
<td>$T_T$</td>
<td>70.0 ns</td>
<td>Storage time</td>
</tr>
<tr>
<td>$C_{J0}$</td>
<td>9.122 pF</td>
<td>Zero-voltage bottom junction capacitance</td>
</tr>
<tr>
<td>$V_J$</td>
<td>6.223 V</td>
<td>Bottom built-in voltage</td>
</tr>
<tr>
<td>$M$</td>
<td>2.42</td>
<td>Bottom junction grading coefficient</td>
</tr>
<tr>
<td>$F_C$</td>
<td>0.5</td>
<td>Bottom depletion capacitance linearization parameter</td>
</tr>
<tr>
<td>$B_V$</td>
<td>32.0 V</td>
<td>Breakdown voltage</td>
</tr>
<tr>
<td>$I_{BV}$</td>
<td>5.0 µA</td>
<td>Current at breakdown voltage</td>
</tr>
<tr>
<td>$E_G$</td>
<td>1.16 eV</td>
<td>Energy gap at nominal temperature</td>
</tr>
<tr>
<td>$X_{TI}$</td>
<td>3.5</td>
<td>Temperature scaling coefficient</td>
</tr>
<tr>
<td>$C_{cv}$</td>
<td>0.27 pF</td>
<td>Capacitance for a better C-V curve approximation</td>
</tr>
</tbody>
</table>

Tab. 3.6: The BB857 parasitics model data provided by the manufacturer [34]

<table>
<thead>
<tr>
<th>Inductance</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{s1}$</td>
<td>0.45 nH</td>
<td>Series inductance no. 1</td>
</tr>
<tr>
<td>$L_{s2}$</td>
<td>0.15 nH</td>
<td>Series inductance no. 2</td>
</tr>
<tr>
<td>$L_{s3}$</td>
<td>0.10 nH</td>
<td>Series inductance no. 3</td>
</tr>
<tr>
<td>$C_p$</td>
<td>90 fF</td>
<td>Package capacitance</td>
</tr>
</tbody>
</table>

In order to employ as precise model of a BB857 varactor diode as possible, this thesis does not rely on the model data provided by the manufacturer. Instead, own measurements of a BB857 varactor diode have been carried out and the manufacturer’s model data have been used as a starting point for the BB857 model optimization. The BB857 varactor-diode measurement and the respective calibration is covered in section 3.4.
Several parameters of the BB857 equivalent circuit were then optimized in AWR Microwave Office so that the $s_{11}$ phase characteristics of the model fit the characteristics of the measured data. The BB857 varactor-diode model has thus been improved. The resulting values of the optimized model variables as well as their comparison with the former values, suggested by the manufacturer, are listed in Tab. 3.7 and 3.8.

Tab. 3.7: Comparison of the manufacturer’s and the improved BB857 chip model data

<table>
<thead>
<tr>
<th>Element</th>
<th>Manufacturer’s model</th>
<th>Improved model</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_S$</td>
<td>1.50 Ω</td>
<td>1.58 Ω</td>
</tr>
<tr>
<td>$C_{J0}$</td>
<td>9.122 pF</td>
<td>8.559 pF</td>
</tr>
<tr>
<td>$V_J$</td>
<td>6.223 V</td>
<td>6.291 V</td>
</tr>
<tr>
<td>$M$</td>
<td>2.42</td>
<td>2.40</td>
</tr>
<tr>
<td>$C_{cv}$</td>
<td>0.27 pF</td>
<td>0.27 pF</td>
</tr>
</tbody>
</table>

Tab. 3.8: Comparison of the manufacturer’s and the improved BB857 parasitics model data

<table>
<thead>
<tr>
<th>Element</th>
<th>Manufacturer’s model</th>
<th>Improved model</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{s1}$</td>
<td>0.45 nH</td>
<td>1.28 nH</td>
</tr>
<tr>
<td>$L_{s2}$</td>
<td>0.15 nH</td>
<td>0.13 nH</td>
</tr>
<tr>
<td>$L_{s3}$</td>
<td>0.10 nH</td>
<td>0.14 nH</td>
</tr>
<tr>
<td>$C_p$</td>
<td>0.09 pF</td>
<td>0.22 pF</td>
</tr>
</tbody>
</table>

The agreement between the simulated and measured characteristics of the BB857 varactor diode can be verified by plotting frequency dependences of the $s_{11}$ phase and magnitude while applying different bias voltages to the diode. These plots are displayed in Fig. 3.4 through 3.9. Accuracy of the model provided by the manufacturer in comparison with the improved model can thus be observed in these figures.

The maximum absolute deviations in the $s_{11}$ phase of the improved BB857 model from the measured data for significant diode bias voltages are presented in Tab. 3.9.

As can be seen in Fig. 3.4 through 3.6, the improved varactor-diode model provides excellent agreement between the measured and the simulated phase of $s_{11}$ over the frequency range and its phase characteristics are superior to the ones of the manufacturer’s model. The values of the maximum absolute deviations in the $s_{11}$ phase of the improved BB857 model from the measured data, depicted in Tab. 3.9, are very low.
Tab. 3.9: Maximum absolute deviations in the $s_{11}$ phase of the improved BB857 model from the measured data versus applied reverse bias voltage

<table>
<thead>
<tr>
<th>Reverse bias voltage [V]</th>
<th>Max. negative deviation in the $s_{11}$ phase [°]</th>
<th>Max. positive deviation in the $s_{11}$ phase [°]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-1.23</td>
<td>0.35</td>
</tr>
<tr>
<td>2</td>
<td>-0.81</td>
<td>1.12</td>
</tr>
<tr>
<td>4</td>
<td>-0.54</td>
<td>1.46</td>
</tr>
<tr>
<td>6</td>
<td>-1.58</td>
<td>0.81</td>
</tr>
<tr>
<td>8</td>
<td>-2.38</td>
<td>2.11</td>
</tr>
<tr>
<td>10</td>
<td>-1.06</td>
<td>1.13</td>
</tr>
<tr>
<td>15</td>
<td>-0.68</td>
<td>1.56</td>
</tr>
<tr>
<td>20</td>
<td>-0.71</td>
<td>1.54</td>
</tr>
<tr>
<td>25</td>
<td>-1.41</td>
<td>1.11</td>
</tr>
</tbody>
</table>

Fig. 3.4: The phase of $s_{11}$ versus frequency $f$ dependence of the proposed BB857 varactor-diode model in comparison with the measured characteristics and the model data provided by the manufacturer at reverse bias voltage of 1 V, 2 V and 4 V

On the other hand, the agreement between the measured and the simulated magnitude of $s_{11}$ is worse, as is visible in Fig. 3.7 through 3.9. This is due to the fact that the parameters of the BB857 equivalent circuit have been optimized to provide the best possible agreement in phase of $s_{11}$ to the detriment of the magnitude. This approach corresponds to the primary application for which the varactor diode is employed in the circuit—to shift the phase of a signal.

Yet, the agreement in trends of the measured and the simulated frequency characteristics of the $s_{11}$ magnitude is acceptable for the frequency band
Fig. 3.5: The phase of $s_{11}$ versus frequency $f$ dependence of the proposed BB857 varactor-diode model in comparison with the measured characteristics and the model data provided by the manufacturer at reverse bias voltage of 6 V, 8 V and 10 V.

Fig. 3.6: The phase of $s_{11}$ versus frequency $f$ dependence of the proposed BB857 varactor-diode model in comparison with the measured characteristics and the model data provided by the manufacturer at reverse bias voltage of 15 V, 20 V and 25 V.

of 0.7 GHz through 1.4 GHz where the phase shifter operates. Also, the characteristics of the improved model are still slightly closer to the measured characteristics of a real BB857 diode than the ones of the manufacturer’s model. In order to reach better agreement in terms of the $s_{11}$ magnitude, a more complex equivalent circuit of the varactor diode should likely be employed.
Fig. 3.7: The magnitude of $s_{11}$ versus frequency $f$ dependence of the proposed BB857 varactor-diode model in comparison with the measured characteristics and the model data provided by the manufacturer at reverse bias voltage of 1 V, 2 V and 4 V.

Fig. 3.8: The magnitude of $s_{11}$ versus frequency $f$ dependence of the proposed BB857 varactor-diode model in comparison with the measured characteristics and the model data provided by the manufacturer at reverse bias voltage of 6 V, 8 V and 10 V.
Fig. 3.9: The magnitude of $s_{11}$ versus frequency $f$ dependence of the proposed BB857 varactor-diode model in comparison with the measured characteristics and the model data provided by the manufacturer at reverse bias voltage of 15 V, 20 V and 25 V.
3.4 Varactor-Diode Measurement and Respective Calibration

As has been mentioned in section 3.3, own measurements of a BB857 varactor diode have been carried out in order to gain s-parameters of the real element. Besides better control over the measured data and the employed calibration technique, this approach allows to measure the diode in the same circuit configuration as in the case of the final phase-shifter circuit.

At the beginning of the vector measurement, an Agilent E8364A vector network analyzer was roughly pre-calibrated by the SOLT calibration technique, i.e. Short, Open, Load and Thru calibration standards were used. Afterwards, raw complex scattering coefficients of TRL calibration standards, i.e. Thru, Reflect and Line, and of the varactor diode itself were measured up to 3 GHz.

The raw measured data were then numerically calibrated by the TRL calibration method in the MATLAB environment. As a result, de-embedded s-parameters of the BB857 varactor diode were obtained of which only an $s_{11}$ vector was of practical value with respect to the diode one-port configuration.

In order to reduce systematic errors of the measurement to the maximum possible extent, own set of on-wafer TRL calibration standards has been developed in a microstrip structure and used. The same structure is employed for the designed phase shifter, i.e. the varactor diode has been measured in the same configuration as in the case of the phase-shifter circuit.

The dimensions of all the designed TRL calibration standards boards needed to be the same since a microwave holder of fixed dimensions had to be used. Therefore, the Line standard could not be implemented as a straight transmission line. In order to determine the amount of reflections when a meander line is used, simulations of the EM field have been carried out in AWR EM Sight.

Layouts of the designed TRL calibration standards are depicted in Fig. 3.10 through 3.15 and the results of the simulations are plotted in Fig. 3.17 through 3.20.

Conclusions published in [38] show that the level of the reflections is acceptable when the meander lines are employed instead of the straight lines and are outweighed by the reduced dimensions of the microstrip structure. The same conclusion has been confirmed by the simulations.

The following from the designed calibration standards have been used for the purpose of the BB857 varactor-diode measurement calibration: Thru,
Reflect, Line no. 1. The layout of the circuit where the varactor diode is to be placed, is depicted in Fig. 3.16. All the designed circuits consist of 50 Ω microstrip transmission lines whose parameters are identical with the ones in Tab. 3.10.

Fig. 3.10: Designed Line standard no. 1

Fig. 3.11: Designed Line standard no. 2

Fig. 3.12: Designed Line standard no. 3

Fig. 3.13: Designed Line standard no. 4
Fig. 3.14: Designed Reflect standard

Fig. 3.15: Designed Thru standard

Fig. 3.16: Designed circuit for the varactor-diode measurement
Fig. 3.17: Simulated magnitude of $s_{11}$ versus frequency $f$ of the designed TRL calibration standards

Fig. 3.18: Simulated phase of $s_{11}$ versus frequency $f$ of the designed TRL calibration standards
3.5 Modeling of the Phase Shifter with Real Elements

This section covers simulations of the entire phase-shifter circuit schematic with real elements in AWR Microwave Office. The equivalent circuit of a BB857 varactor diode developed in section 3.3 is employed for this purpose.
The model also involves parasitic series inductance of via holes positioned between the diode and a ground plane. In addition, the ideal transmission lines are recomputed to real microstrip transmission lines.

The simulation circuit is based on the determined optimal values of the phase-shifter design parameters written in Tab. 3.3. The phase-shifter circuit consists of eight \( (N = 8) \) identical sections of a microstrip transmission line that are loaded in the middle with the BB857 varactor-diode model subcircuit and that are put in a cascade. Port 1 of the BB857 varactor-diode model subcircuit, shown in Fig. 3.3, is thus connected to the transmission line and port 2 is grounded.

| Tab. 3.10: Physical characteristics of microstrip transmission lines fabricated on the RO4350B substrate as reported in [35] or computed in accordance with [36] at frequency \( f = 1 \) GHz |
|---|---|
| Width of the microstrip transmission line of \( Z_i = 10 \) Ω | 13.16 mm |
| Width of the microstrip transmission line of \( Z_i = 50 \) Ω | 1.63 mm |
| Width of the microstrip transmission line of \( Z_i = 100 \) Ω | 0.376 mm |
| Height of the microstrip-line substrate | 0.762 mm |
| Thickness of the microstrip-line metallization | 35 μm |
| Dielectric constant | 3.66 |
| Loss tangent | 0.0031 |

The determined optimal value of characteristic impedance \( Z_i \) equals to 100 Ω. Therefore, the microstrip transmission-line width have been computed to provide this impedance. The values of the microstrip-structure physical parameters as well as the microstrip transmission-line widths for three significant characteristic impedances \( Z_i \) are written in Tab. 3.10. The microstrip structure is implemented on the RO4350B substrate, manufactured by Rogers Corporation, whose detailed characteristics and parameters are recorded in Appendix C [35].

At frequency \( f = 1 \) GHz, the electrical length of the transmission-line sections between varactor diodes \( \varphi_{sp} = 16.8° \) corresponds to the physical length of the varactor-diodes spacing \( l_{sp} = 8.76 \) mm. And this very value of \( l_{sp} \) is used in the phase-shifter circuit schematic.

Based on data in Appendix B, reverse bias voltage \( V \) applied to the varactor diodes should be tuned from 5 V to 28 V in order to exploit the range of varactor-diodes capacitance \( C_{var} \) of 0.52 pF through 2.3 pF, where \( V = 5 \) V corresponds to \( C_{var} = 2.3 \) pF and \( V = 28 \) V results in \( C_{var} = 0.52 \) pF.

The designed structure of the phase-shifter circuit can be seen in Fig. 3.21.
Fig. 3.21: Designed phase-shifter circuit schematic

Besides the eight phase-shifter unit cells, the circuit consists of two 50 Ω microstrip transmission-line sections that connect the circuit to its SMA connectors, an ideal DC voltage source applying the reverse bias voltage to the varactor diodes and grounded by an inductor of sufficiently high inductance ($10^4$ nH), and a 50 Ω section of an ideal transmission line.

This section of an ideal transmission line compensates the phase difference between the simulated and the measured data due to the fact that the position of the calibrated reference plane during the measurement was different from the phase-shifter circuit input port. The electrical length of this compensation section has been determined to be 72° at frequency $f = 1$ GHz.

The simulations have also involved the influence of discontinuities in the microstrip structure, namely bends and steps in the lines width, but for simplicity, these are not depicted in Fig. 3.21.

The simulation results based on modeling of the entire phase-shifter circuit schematic with real elements are presented along with the measured characteristics in section 4.2, in Fig. 4.2 through 4.13.
Chapter 4

Phase-Shifter Circuit Fabrication and Measurement

4.1 Circuit Fabrication

The designed varactor-loaded transmission-line phase shifter is fabricated on the RO4350B substrate in the microstrip structure covered in section 3.5.

The layout of the designed phase shifter has been optimized so that the board with the fabricated circuit fits the horizontal dimensions of the AH101 metal box, 67 mm to 46 mm. The existence of the bends in the microstrip structure has thus been outweighed by its more compact footprint. The resulting layout of the circuit can be seen in Fig 4.1.

After its optimization, the layout was exported from AWR Microwave Office into the Gerber and Excellon data files for the purpose of the copper-layer fabrication using a photoplotter and the control of drilling the via holes through the substrate, respectively.

The available technology provides the minimum value of the microstrip transmission-line width $w$ equal to 0.2 mm. However, as has been already mentioned in section 3.1, a transmission line fabricated at this limit would likely be inferior in terms of precision of its $w$ parameter and hence the designed characteristic impedance $Z_i$ would not be guaranteed. A higher minimum value of $w$ has therefore been chosen.

The BB857 varactor diodes have been soldered onto the phase-shifter circuit board with the cathode connected to the microstrip transmission line and the anode to a pair of via holes. In order to feed the RF signal into the phase shifter, SMA connectors have been mounted on the phase-shifter.
metal box and soldered onto the circuit. An external bias tee has been used to apply the reverse bias control voltage to the varactor diodes.

Fig. 4.1: Designed phase-shifter circuit layout

4.2 Results of the Small-Signal Measurements

RF measurements using a small signal have been made on an Agilent E8364A vector network analyzer calibrated by SOLT calibration standards. The two-port s-parameters of the fabricated phase-shifter circuit were recorded at frequencies of 45 MHz through 3 GHz and reverse bias voltages of 1 V through 28 V.

The measured characteristics along with the corresponding simulated data are plotted in Fig. 4.2 through 4.13. Tab. 4.1, 4.2, and 4.3 summarize the achieved qualities of the designed phase shifter in terms of the maximum differential phase shift $\Phi_L$, the minimum circuit return loss $RL_{min}$ and the maximum insertion loss $IL_{max}$.

Fig. 4.2 through 4.9 show frequency characteristics of the designed phase-shifter s-parameters, namely phase of $s_{21}$, magnitude of $s_{11}$ and magnitude of $s_{21}$, which can be used to evaluate the agreement between the simulated and the measured data.
Simulated and measured magnitudes of $s_{21}$ versus frequency $f$ that goes beyond the maximum frequency of operation $f_{\text{max}}$ are depicted in Fig. 4.8 and 4.9 so that the impact of the Bragg frequency $f_B$ on the transmission scattering coefficient $s_{21}$ of the phase-shifter circuit can be seen.

Analogically, Fig. 4.10 through 4.13 depict the simulated and the measured $s$-parameters as functions of applied reverse bias voltage and thus enable to verify the function and parameters of the designed phase shifter during its operation at a chosen frequency.

![Graph showing simulated and measured phase of $s_{21}$ versus frequency $f$.](image)

**Fig. 4.2:** Simulated and measured phase of $s_{21}$ versus frequency $f$ of the designed phase shifter at reverse bias voltage of 5 V, 6 V, 7 V and 8 V

<table>
<thead>
<tr>
<th>$f$ [GHz]</th>
<th>Simulated $\Phi_L$ [°]</th>
<th>Measured $\Phi_L$ [°]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.7</td>
<td>106.2</td>
<td>110.1</td>
</tr>
<tr>
<td>1.0</td>
<td>166.1</td>
<td>166.8</td>
</tr>
<tr>
<td>1.2</td>
<td>216.6</td>
<td>212.7</td>
</tr>
<tr>
<td>1.4</td>
<td>293.2</td>
<td>276.7</td>
</tr>
</tbody>
</table>

The primary function of the circuit and the main objective of the design—to provide the maximum differential phase shift $\Phi_L$ of no less than $90$° within the whole frequency band—has been successfully met, as is evident from Tab. 4.1 and Fig. 4.11. The fabricated phase shifter provides the maximum differential phase shift $\Phi_L$ of $110.1$° at the minimum frequency of operation $f = f_{\text{min}} = 0.7$ GHz.
Different values of $\Phi_L$ recorded in Tab. 3.4 for number of sections $N = 8$ and in Tab. 4.1 can be explained by the fact that varactor-diode capacitance $C_{\text{var}}$ was supposed to be a linear function of applied reverse bias voltage in section 3.2. However, a real varactor-diode capacitance $C_{\text{var}}$ is a nonlinear function of applied reverse bias voltage, based on expression (2.57), and thus the $s_{21}$ phase dependence of the phase-shifter circuit versus applied reverse bias voltage is nonlinear as well. Therefore, both mentioned simulated values of $\Phi_L$ are slightly different.
Fig. 4.5: Simulated and measured magnitude of $s_{11}$ versus frequency $f$ of the designed phase shifter at reverse bias voltage of 10 V, 15 V, 20 V and 28 V.

Fig. 4.6: Simulated and measured magnitude of $s_{21}$ versus frequency $f$ of the designed phase shifter at reverse bias voltage of 5 V, 6 V, 7 V and 8 V.

Fig. 4.2, 4.3 and 4.10 show excellent agreement between the measured and the simulated characteristics of the designed phase-shifter circuit in phase of $s_{21}$ over the whole range of frequencies and reverse bias voltages, which is crucial for the successful function of a phase shifter.

Another design objective requires that the minimum circuit return loss $RL_{\text{min}}$ of 10 dB is achieved. As is depicted in Tab. 4.2 and Fig. 4.12, the fabricated phase shifter successfully meets this requirement at the four significant frequencies $f$. Even in the worst case, which is visible in Fig. 4.4 at
Fig. 4.7: Simulated and measured magnitude of $s_{21}$ versus frequency $f$ of the designed phase shifter at reverse bias voltage of 10 V, 15 V, 20 V and 28 V.

Fig. 4.8: Simulated and measured magnitude of $s_{21}$ of the designed phase shifter at reverse bias voltage of 5 V, 6 V, 7 V and 8 V versus frequency $f$ that goes beyond the maximum frequency of operation $f_{\text{max}}$ in order to depict the impact of the Bragg frequency $f_B$ on the transmission scattering coefficient of the phase-shifter circuit.

Reverse bias voltage $V$ of 5 V, the limit of 10 dB is exceeded by the fabricated circuit by a mere 0.05 dB.

Fig. 4.4 shows very good agreement between the measured and the simulated $s_{11}$ magnitude of the designed phase shifter over the frequency range and at reverse bias voltage of 5 V, 6 V, 7 V and 8 V. Fig. 4.5 depicts the same dependences at reverse bias voltage of 10 V, 15 V, 20 V and 28 V.
Fig. 4.9: Simulated and measured magnitude of $s_{21}$ of the designed phase shifter at reverse bias voltage of 10 V, 15 V, 20 V and 28 V versus frequency $f$ that goes beyond the maximum frequency of operation $f_{\text{max}}$ in order to depict the impact of the Bragg frequency $f_B$ on the transmission scattering coefficient of the phase-shifter circuit.

Fig. 4.10: Simulated and measured phase of $s_{21}$ versus reverse bias voltage $V$ of the designed phase shifter at frequency of 0.7 GHz, 1.0 GHz, 1.2 GHz and 1.4 GHz where the agreement is worse but still satisfactory. The agreement of the $s_{11}$ magnitude characteristics plotted in Fig. 4.12 is also satisfactory but with noticeable difference between the measured and the simulated data.

The difference between the measured and the simulated characteristics at lower values of the $s_{11}$ magnitude may be explained by reflections of the used SMA connectors, which may thus mask reflections of the measured circuit.
Fig. 4.11: Simulated and measured differential phase shift $\Phi_L$ of the designed phase shifter, which equals to the difference in phase of $s_{21}$ and is normalized to the $s_{21}$ phase value at the maximum varactor-diodes capacitance $C_{\text{var}}^{\text{max}}$, versus reverse bias voltage $V$ at frequency of 0.7 GHz, 1.0 GHz, 1.2 GHz and 1.4 GHz.

Fig. 4.12: Simulated and measured magnitude of $s_{11}$ versus reverse bias voltage $V$ of the designed phase shifter at frequency of 0.7 GHz, 1.0 GHz, 1.2 GHz and 1.4 GHz.

that are below a certain level.

Values of simulated and measured maximum insertion loss $IL_{\text{max}}$ of the designed circuit at four significant frequencies $f$ are recorded in Tab. 4.3 and the corresponding characteristics are shown in Fig. 4.13. As can be seen, the circuit insertion loss $IL$ increases with increasing frequency $f$ and decreasing reverse bias voltage $V$. 

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Fig. 4.13: Simulated and measured magnitude of $s_{21}$ versus reverse bias voltage $V$ of the designed phase shifter at frequency of 0.7 GHz, 1.0 GHz, 1.2 GHz and 1.4 GHz

Tab. 4.2: Simulated and measured minimum return loss $RL_{min}$ of the designed phase shifter at four significant frequencies $f$

<table>
<thead>
<tr>
<th>$f$ [GHz]</th>
<th>Simulated $RL_{min}$ [dB]</th>
<th>Measured $RL_{min}$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.7</td>
<td>11.0</td>
<td>11.8</td>
</tr>
<tr>
<td>1.0</td>
<td>17.8</td>
<td>18.8</td>
</tr>
<tr>
<td>1.2</td>
<td>8.7</td>
<td>10.1</td>
</tr>
<tr>
<td>1.4</td>
<td>8.9</td>
<td>11.4</td>
</tr>
</tbody>
</table>

Tab. 4.3: Simulated and measured maximum insertion loss $IL_{max}$ of the designed phase shifter at four significant frequencies $f$

<table>
<thead>
<tr>
<th>$f$ [GHz]</th>
<th>Simulated $IL_{max}$ [dB]</th>
<th>Measured $IL_{max}$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.7</td>
<td>0.73</td>
<td>1.18</td>
</tr>
<tr>
<td>1.0</td>
<td>0.81</td>
<td>1.73</td>
</tr>
<tr>
<td>1.2</td>
<td>1.81</td>
<td>3.16</td>
</tr>
<tr>
<td>1.4</td>
<td>2.66</td>
<td>4.47</td>
</tr>
</tbody>
</table>

Based on formula (2.57), varactor-diodes capacitance $C_{var}$ increases with decreasing reverse bias voltage $V$. And both higher values of varactor-diodes capacitance $C_{var}$ and frequency $f$ result in increased varactor-diode admittance $Y_{var}$, as is evident from equation (2.36). When varactor-diode admittance $Y_{var}$ increases, an amount of an RF current that flows through the diode—and through its series resistance $R_s$—increases as well. Insertion loss $IL$ of the circuit therefore increases with increasing frequency $f$ and decreas-
Fig. 4.6, 4.7 and 4.13 present acceptable agreement in trends between the measured and the simulated characteristics of the designed phase-shifter circuit in magnitude of $s_{21}$ over the range of frequencies and reverse bias voltages. However, a more complex model of the varactor diode would probably need to be employed in order to simulate the circuit insertion loss $IL$ more precisely.

Bragg frequency $f_B$ of the designed phase-shifter periodic structure is well beyond maximum frequency of operation $f = f_{max} = 1.4$ GHz for the whole range of used reverse bias voltages, as can be seen in Fig. 4.8 and 4.9.

### 4.3 Large-Signal Behavior of the Circuit

The BB857 varactor-diode nonlinear model has also been exploited to analyze the large-signal behavior of the resulting phase-shifter circuit since the nonlinear dependence of the PN junction capacitance on the applied bias voltage represents the main source of the circuit nonlinearity.

The two-tone third-order output intercept point $OIP_3$ of the designed phase shifter has been simulated and measured, whereas the 1 dB compression point $P_{-1dB}$ has only been simulated and not measured due to technical limitations of the available input power level.

Analysis of the large-signal behavior of the designed circuit has been carried out for tones at $f_1 = 1.00$ GHz and $f_2 = 1.01$ GHz and for reverse bias voltage $V$ of 6 V applied to the varactor diodes. AWR Microwave Office has been used for the simulations.

The two-tone third-order output intercept point $OIP_3$ can be determined as the y-axis coordinate of an intersection point of tangents to the output-power characteristics of the first-harmonic and the third-order intermodulation (IM) products [39]. The 1 dB compression point $P_{-1dB}$ is the y-axis coordinate of a point on the first-harmonic output-power characteristic where the difference between the compressed output-power characteristic and its tangent equals to 1 dB [39].

The characteristics obtained from the large-signal behavior analysis of the circuit are depicted in Fig. 4.14 and the particular resulting values of $OIP_3$ and $P_{-1dB}$ are recorded in Tab. 4.4. These characteristics provide good agreement between the measured and simulated behavior of the phase shifter.
Fig. 4.14: Output power $P_{\text{out}}$ of simulated and measured first harmonics and third-order IM products of the designed phase-shifter circuit versus input power $P_{\text{in}}$ in order to determine the two-tone third-order output intercept point $OIP_3$ and the 1 dB compression point $P_{\text{1dB}}$

Tab. 4.4: Resulting values of the two-tone third-order output intercept point $OIP_3$ and the 1 dB compression point $P_{\text{1dB}}$ of the designed phase shifter

<p>| | | |</p>
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<tr>
<td>Simulated $OIP_3$ [dBm]</td>
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<td>Measured $OIP_3$ [dBm]</td>
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<td>Simulated $P_{\text{1dB}}$ [dBm]</td>
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Chapter 5

Conclusion

The thesis has presented theory behind the function principles of a phase shifter based on a varactor-loaded transmission line. The function of the phase shifter has been examined in relation to its physical parameters, key variables of the structure have been specified and an appropriate choice of their values has been discussed in order to design and optimize the phase-shifter circuit.

The main objective of the thesis was to design and fabricate a phase shifter that operates within a one-octave frequency band for the center frequency of 1 GHz or higher. The maximum differential phase shift of the circuit was required to be at least 90° within the whole frequency band and its return loss to be no less than 10 dB.

This objective has been successfully met. The developed phase shifter operates within the frequency range of 0.7 GHz through 1.4 GHz, provides the maximum differential phase shift of 110.1° at the minimum frequency of operation, the return-loss limit of 10 dB has been exceeded by the fabricated circuit by a mere 0.05 dB in the worst case and good agreement between the simulated and the measured characteristics of the phase-shifter circuit has been obtained. The third-order output intercept point of the phase shifter has been measured to be 25.4 dBm.

The model of a BB857 varactor diode provided by the manufacturer has been improved based on own measurements of the element for the purpose of the phase-shifter simulations. However, a more complex model of the varactor diode would probably need to be employed in order to simulate the circuit insertion loss more precisely.
Acknowledgements

First of all, I wish to acknowledge kind support and guidance of my supervisor, Ing. Jan Šístek, Ph.D., that he provided to me during my work on this thesis as well as highly qualified instruction and suggestions I received from him.

I would also like to express my appreciation to prof. Ing. Karel Hoffmann, CSc. for helpful discussions and assistance in the field of microwave measurements.

And last but not least, I wish to thank my parents for their encouragement and loving support not only for the work on this thesis but also throughout my whole university studies.

Matěj Vokáč
May 22, 2009
Prague
Bibliography


[34] Infineon Technologies AG: *BB857 Data Sheet* and *BB857 Simulation Data*, http://www.infineon.com, January 2009


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Appendix A

Chain to Scattering Matrix Conversion

Theory behind conversion of a chain matrix $[A]$ into a scattering matrix $[S]$ is presented in [40] and leads to formulas (A.1)–(A.5), where $Z_{\text{in}}$ and $Z_{\text{out}}$ stand for reference impedance at the input and the output port, respectively.

$$[A] = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \quad \rightarrow \quad [S] = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix}$$ (A.1)

$$s_{11} = \frac{a_{11}Z_{\text{out}} + a_{12} - a_{21}Z_{\text{in}}^*Z_{\text{out}} - a_{22}Z_{\text{in}}^*}{a_{11}Z_{\text{out}} + a_{12} + a_{21}Z_{\text{in}}Z_{\text{out}} - a_{22}Z_{\text{in}}}$$ (A.2)

$$s_{12} = \frac{\det[A] \cdot 2 \sqrt{\Re\{Z_{\text{in}}\} \Re\{Z_{\text{out}}\}}}{a_{11}Z_{\text{out}} + a_{12} + a_{21}Z_{\text{in}}Z_{\text{out}} - a_{22}Z_{\text{in}}}$$ (A.3)

$$s_{21} = \frac{2 \sqrt{\Re\{Z_{\text{in}}\} \Re\{Z_{\text{out}}\}}}{a_{11}Z_{\text{out}} + a_{12} + a_{21}Z_{\text{in}}Z_{\text{out}} - a_{22}Z_{\text{in}}}$$ (A.4)

$$s_{22} = \frac{-a_{11}Z_{\text{out}}^* + a_{12} - a_{21}Z_{\text{in}}Z_{\text{out}}^* - a_{22}Z_{\text{in}}}{a_{11}Z_{\text{out}} + a_{12} + a_{21}Z_{\text{in}}Z_{\text{out}} - a_{22}Z_{\text{in}}}$$ (A.5)
BB837/BB857... Silicon Tuning Diode
• For SAT tuners
• High capacitance ratio
• Low series resistance
• Excellent uniformity and matching due to "in-line" matching assembly procedure
• Pb-free (RoHS compliant) package 1)
• Qualified according AEC Q101

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Maximum Ratings at $T_A = 25^\circ C$, unless otherwise specified

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1) Pb-containing package may be available upon special request.
Electrical Characteristics at $T_A = 25^\circ C$, unless otherwise specified

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AC Characteristics

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Capacitance matching

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For details please refer to Application Note 047.
Date Code marking for discrete packages with one digit (SCD80, SC79, SC75) CES-Code

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1. New Marking layout for SC75, implemented at October 2005.

Standard Reel with 2 mm Pitch

Reel Ø180 mm = 3.000 Pieces/Reel
Reel Ø180 mm = 8.000 Pieces/Reel (2 mm Pitch)
Reel Ø330 mm = 10.000 Pieces/Reel
Attention please!
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RO4000® Series High Frequency Circuit Materials are glass reinforced hydrocarbon/ceramic laminates (Not PTFE) designed for performance sensitive, high volume commercial applications. RO4000 laminates are designed to offer superior high frequency performance and low cost circuit fabrication. The result is a low loss material which can be fabricated using standard epoxy/glass (FR4) processes offered at competitive prices.

The selection of laminates typically available to designers is significantly reduced once operational frequencies increase to 500 MHz and above. RO4000 material possesses the properties needed by designers of RF microwave circuits and allows for repeatable design of filters, matching networks and controlled impedance transmission lines. Low dielectric loss allows RO4000 series material to be used in many applications where higher operating frequencies limit the use of conventional circuit board laminates. The temperature coefficient of dielectric constant is among the lowest of any circuit board material (Chart 1), and the dielectric constant is stable over a broad frequency range (Chart 2). This makes it an ideal substrate for broadband applications.

RO4000 material's thermal coefficient of expansion (CTE) provides several key benefits to the circuit designer. The expansion coefficient of RO4000 material is similar to that of copper which allows the material to exhibit excellent dimensional stability, a property needed for mixed dielectric multilayer boards constructions. The low Z-axis CTE of RO4000 laminates provides reliable plated through-hole quality, even in severe thermal shock applications. RO4000 series material has a Tg of >280°C (536°F) so its expansion characteristics remain stable over the entire range of circuit processing temperatures.

RO4000 series laminates can easily be fabricated into printed circuit boards using standard FR4 circuit board processing techniques. Unlike PTFE based high performance materials, RO4000 series laminates do not require specialized via preparation processes such as sodium etch. This material is a rigid, thermoset laminate that is capable of being processed by automated handling systems and scrubbing equipment used for copper surface preparation.

Features:
- Excellent high frequency performance due to low dielectric tolerance and loss
- Stable electrical properties versus frequency
- Low thermal coefficient of dielectric constant
- Low Z-Axis expansion
- Low in-plane expansion coefficient
- Excellent dimensional stability
- Volume manufacturing process

Some Typical Applications:
- LNB's for Direct Broadcast Satellites
- Microstrip and Cellular Base Station Antennas and Power Amplifiers
- Spread Spectrum Communications Systems
- RF Identities Tags

The world runs better with Rogers.
The information contained in this fabrication guide is intended to assist you in designing with Rogers' circuit materials and prepreg. It is not intended to and does not create any warranties, express or implied, including any warranty of merchantability or fitness for particular purpose. The user is responsible for determining the suitability of Rogers' circuit materials and prepreg for each application.

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Rogers recommends that the customer evaluate each material and design combination to determine if the values listed.

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Typical values are a representation of an average value for the population of the property. For specification values contact Rogers Corporation.

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Fabrication Guidelines for RO4000® Series High Frequency Circuit Materials

RO4000® High Frequency Circuit Materials were developed to provide high frequency performance comparable to woven glass/PTFE substrates, with the ease of fabrication associated with epoxy/glass laminates. RO4000 materials are a silica-reinforced hydrocarbon/ceramic filled thermoset material with very high glass transition temperature (Tg >280°C). Unlike PTFE-based microwave materials, no special through hole treatment or handling procedures are required. Therefore, RO4000 material circuit processing and assembly costs are comparable to epoxy/glass laminates.

Some basic guidelines for processing double sided RO4000 panels are provided below. In general, process parameters and procedures used for epoxy/glass boards can be used to process RO4000 boards.

**DRILLING:**

**ENTRY/EXIT MATERIAL:**

Entry and exit materials should be flat and rigid to minimize copper burrs. Recommended entry materials include aluminum and rigid composite board (0.010" to 0.025" (0.254 - 0.635mm)). Most conventional wall materials with or without aluminum cladding are suitable.

**MAXIMUM STACK HEIGHT:**

The thickness of material being drilled should not be greater than 70% of the flute length. This includes the thickness of entry material and penetration into the backer material.

For example:
- **Flute Length:** 0.300" (7.62mm)
- **Entry Material:** 0.015" (0.381mm)
- **Backer Penetration:** 0.030" (0.762mm)
- **Material Thickness:** 0.020" (0.508mm)

Maximum Stack = 0.70 x 0.300" (7.62mm) = 0.210" (5.33mm) (available flute length)

Minimum Stack per stack = 0.165" (4.19mm) (available for PCBs)

Boards per stack = [0.165" (4.19mm)] / [0.023" (0.58mm)] = 7.2 boards per stack (round down)

DRIING CONDITIONS:

Surface speeds greater than 500 SFM and chip loads less than 0.002" (0.05 mm) should be avoided, whenever possible.

Recommended Ranges:
- **Surface Speed:** 300 - 500 SFM (90 to 150 m/mm)
- **Chip Load:** 0.002" - 0.004"/rev. (0.05 - 0.10 mm/rev)
- **Retract Rate:** 500 - 1000 IPM (12.7 m/min) for tool less than 0.0135" (0.343 mm), 1000 IPM (25.4 m/min) for all others.
- **Tool Type:** Standard Carbide
- **Tool Life:** 2000-3000 hits

Hole quality should be used to determine the effective tool life rather than tool wear. The RO4003™ material will yield good hole quality when drilled with bits which are considered worn by epoxy/glass standards. Unlike epoxy/glass, RO4003 material hole roughness does not increase significantly with tool wear. Typical values range from 6-25 um regardless of hit count (evaluated up to 8000 hits). The roughness is directly related to the ceramic filler size and provides topography that is beneficial for hole-wall adhesion. Drilling conditions used for epoxy/glass boards have been found to yield good hole quality with hit counts in excess of 2000.

**CALCULATING SPINDLE SPEED AND INFEED:**

\[
\text{Spindle Speed (RPM)} = \frac{12 \times \text{Surface Speed (SFM)}}{3.14 \times \text{Tool Diameter (in.)}}
\]

\[
\text{Feed Rate (IPM)} = \frac{\text{Spindle Speed (RPM)} \times \text{Chip Load (in/rev.)}}{12}
\]

**Example:**
- Desired Surface Speed: 400 SFM
- Desired Chip Load: 0.003" (0.076 mm/rev)

Spindle Speed = \[\frac{12 \times 400}{3.14 \times (0.023\text{in.})} = 51,800 \text{ RPM}\]

Feed Rate = \[\frac{51,800 \text{ RPM}}{12} = 4,317 \text{ IPM}\]

The information contained in this fabrication guide is intended to assist you in designing with Rogers’ circuit materials and processing. It is not intended to create any legal or other obligation on the part of Rogers. The user is responsible for determining the suitability of Rogers’ circuit materials and processing for each application.
The information contained in this fabrication guide is intended to assist you in designing with Rogers' circuit materials and prepreg. It is not intended to or does not create any warranties, express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this fabrication guide will be achieved by a user for a particular purpose. The user is responsible for determining the suitability of Rogers' circuit materials and prepreg for each application.

**COPPER PLATING**

No special treatments are required prior to electrolite copper plating. A board should be processed using conventional epoxy/glass procedures. Interior of all vias is not typically required, as the high Tg (280°C + [536°F]) resin system is not prone to smearing during drill. Resin can be removed using a standard 45°C plasma cycle or a double pass through an alkaline permanganate process should smear result from aggressive drilling practices.

**IMAGING/ETCHING**

Panel surfaces may be mechanically and/or chemically prepared for photoresist. Standard aqueous or semi-aqueous photoresists are recommended. Any of the commercially available copper etchants can be used.

**SOLDERMASK**

Any screenable or photoimageable solder masks typically used on epoxy/glass laminates bond very well to the surface of RO4000. Mechanical scrubbing of the exposed dielectric surface prior to solder mask application should be avoided as an "as etched" surface provides for optimum bonding.

**ROUTING CONDITIONS**

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### QUICK REFERENCE TABLE

<table>
<thead>
<tr>
<th>Tool Diameter</th>
<th>Spindle Speed (SFPM)</th>
<th>Infeed Rate (IPM)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.0100&quot; (0.254mm)*</td>
<td>95.3</td>
<td>190</td>
</tr>
<tr>
<td>0.0130&quot; (0.330mm)*</td>
<td>70.7</td>
<td>141</td>
</tr>
<tr>
<td>0.0160&quot; (0.406mm)*</td>
<td>55.5</td>
<td>190</td>
</tr>
<tr>
<td>0.0195&quot; (0.500mm)*</td>
<td>77.6</td>
<td>190</td>
</tr>
<tr>
<td>0.0255&quot; (0.640mm)</td>
<td>40.0</td>
<td>180</td>
</tr>
<tr>
<td>0.0298&quot; (0.759mm)</td>
<td>60.0</td>
<td>180</td>
</tr>
<tr>
<td>0.0295&quot; (0.759mm)</td>
<td>51.8</td>
<td>155</td>
</tr>
<tr>
<td>0.0295&quot; (0.759mm)</td>
<td>45.2</td>
<td>130</td>
</tr>
<tr>
<td>0.0295&quot; (0.759mm)</td>
<td>38.8</td>
<td>114</td>
</tr>
<tr>
<td>0.0305&quot; (0.775mm)</td>
<td>33.7</td>
<td>101</td>
</tr>
<tr>
<td>0.0495&quot; (1.257mm)</td>
<td>31.1</td>
<td>93</td>
</tr>
<tr>
<td>0.0531&quot; (1.349mm)</td>
<td>28.8</td>
<td>86</td>
</tr>
<tr>
<td>0.0625&quot; (1.588mm)</td>
<td>24.5</td>
<td>74</td>
</tr>
<tr>
<td>0.0750&quot; (1.905mm)</td>
<td>16.5</td>
<td>50</td>
</tr>
<tr>
<td>0.1000&quot; (2.540mm)</td>
<td>15.0</td>
<td>45</td>
</tr>
</tbody>
</table>

* C tol. as lead-free per EIA J-STD-003, up to 2 oz Au and 2 oz Pd.
The information in this data sheet is intended to assist you in designing with Rogers' circuit material laminates. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this data sheet will be achieved by a user for a particular purpose. The user should determine the suitability of Rogers' circuit material laminates for each application.

Prolonged exposure in an oxidative environment may cause changes to the dielectric properties of hydrocarbon based materials. The rate of change increases at higher temperatures and is highly dependent on the circuit design. Although Rogers' high frequency materials have been used successfully in innumerable applications and reports of oxidation resulting in performance problems are extremely rare, Rogers recommends that the customer evaluate each material and design combination to determine fitness for use over the entire life of the end product.

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The information in this data sheet is intended to assist you in designing with Rogers' circuit material laminates. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose.

Tool Diameter | Spindle Speed | Lateral Feed Rate
--- | --- | ---
1/32 | 40 x 97 RPM | 30 IPM
1/16 | 25 x 97 RPM | 31 IPM
3/32 | 20 x 97 RPM | 25 IPM
1/8 | 15 x 97 RPM | 19 IPM

World Class Performance

Rogers Corporation (NYSE:ROG), headquartered in Rogers, Conn., is a global technology leader in the development and manufacture of high performance, specialized materials and products for a diverse, applications-driven market including mobile communications, wireless infrastructure, specialty chemicals, advanced electronics, medical, energy, aerospace, defense and transportation. Rogers Corporation is active in a global marketplace where customer satisfaction is the key to success. The company operates facilities in the United States, Europe and Asia.

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