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Coplanar Waveguide Phase Shifter on Silicon at LMCS Frequencies

by

Stéphan D. Picard

**A thesis submitted to the Faculty of Graduate Studies and Research
in partial fulfillment of the requirements for the degree of**

Master of Engineering

Ottawa-Carleton Institute for Electrical Engineering

Department of Electronics

Carleton University

Ottawa, Ontario, Canada

October 2001

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October 2001

Abstract

This thesis presents the design, simulation, construction and evaluation of novel coplanar-waveguide (CPW) phase shifters for Local Multipoint Communication Service (LMCS) applications. To meet the LMCS system requirements, a 2 gigahertz (GHz) bandwidth, centered at 28 GHz, was required. Two design strategies were implemented: reflection and transmission phase shifters. The reflection phase shifters used branch-line coupler in conjunction with delay lines. The transmission phase shifter was based on two distributed-element filters. Both approaches took advantage of the uniplanar nature of CPW to simplify the fabrication process, which is described in some details.

Short-circuiting of the delay lines and switching between the filters is to be realized with micromachined (MEMS) switches, which are currently under development.

The measurement results of the reflection phase shifter showed behaviour approaching that expected, but several fabrication problems caused significant deterioration of the results. In the case of the transmission phase shifter, the measured phase difference of 161° matched the simulated one of 176° . The results are explained in detail and suggestions for improving upon them are given.

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Acronyms and Abbreviations

AC	Alternating current
ADS	Advanced Design System
Au	Gold
ATM	Asynchronous transfer mode
BLC	Branch-line coupler
BLC-PS	Branch-line coupler phase shifter
BPF	Band-pass filter
BSF	Band-stop filter
BW	Bandwidth
CAD	Computer-aided design
CATV	Cable television
CITR	Canadian Institute for Telecommunication Research
CPW	Coplanar waveguide

CPWG	Coplanar waveguide with lower ground plane
CRC	Communication Research Centre
CRC-TWS	Communication Research Centre - Terrestrial Wireless Systems
DC	Direct current
E-field	Electric field
EM	Electromagnetic
EMFS	Electromagnetic fields solver
FDTD	Finite difference time domain
GaAs	Gallium arsenide
H-field	Magnetic field
HP	Hewlett Packard
HPF	High-pass filter
IEEE	Institute for Electrical and Electronics Engineers
IF	Intermediate frequency
IP	Internet protocol
LAN	Local area network
LIGA	Lithographie, galvanoförmung, abförmung (german for Lithography, electroplating and molding)

LMCS	Local multipoint communication service
LMDS	Local multipoint distribution system
LNA	Low-noise amplifier
LPF	Low-pass filter
MEMS	Micro electro-mechanical systems
MIC	Monolithic integrated circuit
MMIC	Monolithic microwave integrated circuit
MOEMS	Micro (opto)electro-mechanical systems
No.	Number
OSLT	Open, short, load, through
p/pp	page/pages
PIN	P-doped/intrinsic/n-doped
PMP	Point to multipoint
PR	Photoresist
PTP	Point to point
POTS	Plain old telephony system
RF	Radio frequencies
RF-MEMS	Radio frequency micro-electromechanical system

SA	Smart antenna
Si	Silicon
Si₃N₄	Silicon nitride
SiGe	Silicon germanium
SEM	Scanning electron microscopy
SDMA	Spatial division multiple access
S-Parameters	Scattering parameters
SPICE	Simulation Program with Integrated Circuit Emphasis
TEM	Transverse electromagnetic
TiW	Titanium tungsten
TRL	Through, reflect, line
UV	Ultraviolet
Vol.	Volume
WAN	Wide area network

Symbols

Symbol	Description	Units
B_i	Voltage wave amplitude at the i^{th} port	V
C	Capacitor/capacitance	F
G	Gap spacing (of a CPW)	m
k_1, k_2, k_3	Modulus of the complete elliptic integral	—
k'_1, k'_2, k'_3	Complementary modulus of the complete elliptic integral	—
L	Inductor/inductance	H
N	Ground plane width (of a CPW)	m
P_i	Incident power	W or dBm
P_r	Reflected power	W or dBm
P_t	Transmitted power	W or dBm
S_{11}	Forward reflection scattering parameter	—
S_{12}	Forward transmission scattering parameter	—
S_{21}	Reverse transmission scattering parameter	—
S_{22}	Reverse reflection scattering parameter	—
q	Filling factor	—
R	Probe pitch	m

$\tan \delta$	Dielectric loss tangent	—
$\tan \mu$	Magnetic loss tangent	—
W	Centre conductor width	m
Z	Impedance	Ω
Z_c	Coplanar-waveguide characteristic impedance	Ω
Z_i	Input impedance	Ω
Z_L	Load impedance	Ω
Z_0	Characteristic impedance	Ω
β	Propagation constant	rad/m
Γ	Reflection coefficient	—
Δl	Physical length of a transmission line	m
$\Delta \theta$	Electrical length of a transmission line	° or rad
δ	Skin depth	m
ϵ_{eff}	Effective relative permittivity (effective dielectric constant)	—
ϵ_r	Relative permittivity in material	—
ϵ_0	Permittivity in vacuum	F/cm
λ	Wavelength	m
μ_r	Relative permeability in material	—
μ_0	Permeability in vacuum	H/cm
ρ	Metal resistivity	$\Omega \cdot \text{m}$
σ	Dielectric conductivity	
ω	Angular frequency	rad/s

Chapter 1

Introducing the Project

1.1 Getting a Global Perspective on the Project

Telecommunications is evolving at a fast pace and public demand for these services is increasing¹. To satisfy this increase, it is important to have an accessible, enabling infrastructure to reach the end-users. It is desirable to have a flexible, low-cost scalable² infrastructure which provides adaptability and expandability to suit a rapidly changing market.

Installing wireline links from a service provider to every potential client is not flexible and not scalable. These links also rapidly become obsolete as technology improves in terms of performance.

Wireless links are easier to tailor to a market in motion as they facilitate resource reassignment to the clients. Greater flexibility and scalability make this approach

¹According to Statistics Canada, the number of cell phone users increased by 28% from 1999 to 2000. The number of households connected to the Internet increased by four fold between 1996 and 1999. [2]

²Scalable means it “is designed to support growth to an arbitrarily large size [...]” [3].

potentially more viable and profitable. Also, the overall cost — as opposed to an individual user cost — is reduced since the resources are solely allocated to subscribing clients, rather than a whole neighborhood. This flexibility also improves the quality of service as the new technological advances are readily implemented within the service distribution chain as they become available.

1.1.1 Introducing Local Multipoint Communication Services

The world of telecommunications is facing new challenges that are described by the Canadian Institute for Telecommunication Research³ (CITR) in the following terms:

“The market for providing broadband services to homes and offices, especially high-speed Internet access, is heating up fast. Fixed wireless communications make the market more competitive.”[4]

The public is demanding faster and cheaper facilities for multimedia applications. To meet these requirements, industry, supported by university research, has to comply by investigating new areas.

Defining LMCS

Local Multipoint Communication Services⁴ (LMCS) is a fast growing⁵ and promising venue to provide this kind of service to the public. There is a good tutorial on LMDS [5] that explains the characteristics of such system:

³CITR, McGill University, 3480 University, Suite 753, Montreal, Quebec, Canada, H3A 2A7.

⁴Otherwise known in the United States as LMDS (Local Multipoint Distribution Systems) and in Europe as MVDS (Multipoint Video Distribution Services).

⁵Market prediction of \$0.192B, \$1B and \$1.5B respectively in 1999, 2000 and 2001 according to Telecompetition Inc.

- Short range due to atmospheric attenuation of the 20-50 GHz frequency range used,
- Point-to-multipoint service from one provider to multiple clients.

Comparing the Cellular Nature of LMCS and of Cellular Telephony

LMCS, like mobile telephony, is cellular in nature even though it has one major difference: the client is fixed. An example of a client is a university campus providing Internet and telephone access to its students and staff. The receiving antenna would be located on the highest building of the campus and directed toward the service provider.

The fixed receiving antenna of LMCS greatly simplifies the link budgeting as compared to cellular telephony for the following reasons:

The antenna direction. A cellular phone is subject to movement in all three dimensions. The difficulty in predicting the behaviour of the users prevents the use of a directional antenna. With LMCS, the antenna is fixed and allow greater directivity. This aspect may become less significant with the development of smart antennas, whose beam can be dynamically steered toward the source of the signal.

The environment. This factor affects the link budget. Cellular phones are used in environments that vary widely. Particularly difficult environments include urban topographies where tunnels and buildings may significantly attenuate the signal level. This constraint is alleviated for LMCS, since once the antenna is installed the received signal level is predictable and less prone to deterioration from unanticipated events.

Figure 1.1: LMCS cellular nature.

Some factors are still significant and need to be studied further, for example foliage and atmospheric precipitation but these are beyond the scope of this project.

Although it does not influence the link budget, another aspect of the fixed antenna is of importance in LMCS: spatial division multiple access (SDMA). SDMA is possible because the location of the receiver is known. In directing a signal solely toward the intended receiver, it is possible to use the same frequency, at the same time, with two receivers located in non-contiguous regions. Figure 1.1 illustrates frequency re-use (SDMA), the cellular nature of LMCS, and the use of an additional base station to cover a region shadowed by an obstacle.

1.1.2 Glancing at the Supporting Networks

LMCS provides an infrastructure for telecommunication but does not provide control over traffic circulation. To allow traffic to reach a target, a network must be built on this infrastructure.

The hubs of existing networks are interconnected over long distance via cable, optical fibre, or satellite link. The connection between a hub and a client is called

Figure 1.2: The “last mile” network connection.

“the last mile”. In the case of LMCS, the service provider collects the signals from a hub and relays them to the consumer by microwave link. Similarly, it receives signals from a consumer. It delivers this signal to the hub for long distance transmission. Alternatively, it can transmit the information directly to another user, in the case of a local receiver. Figure 1.2 illustrates the relation between the service provider, the customer and the network to which they are connected.

Network Architecture

LMCS offers an alternative for the local loop (or “last mile”) to carry voice, data and video. Originally, different telecommunication networks (for example telephony, cable television, local and wide area networks, and even the Internet) were distinct because of the specific requirements of each type of transmitted signals. For example, the delay between pieces of information sent over telephone lines is much more critical

than it is for data transmissions. However, large bandwidth is required for data, often in unpredictable bursts, while it is not the case for voice applications.

These different networks seem to be converging and the boundaries between them are being blurred. Some new architectures, especially if integrated with proper protocols, are capable of handling the parameterization of the channel characteristics. The different signals, each with proper characteristics, could be managed in an LMCS by protocols such as asynchronous transfer mode (ATM) and Internet protocol (IP). These two protocols can easily be laid on top of the LMCS architecture since they are independent of the physical layer.

Integrating LMCS structure with a scalable network architecture, overlaid with a protocol such as ATM or IP, allows networks to achieve three goals: blurring the boundaries among existing networks, allowing the network to grow with the market demand, and facilitating the integration of upgrades as technological improvements become available.

Asynchronous Transfer Mode

The ATM is independent of the physical layer. The data stream is passed to the ATM Adaptation Layer (AAL) which interfaces existing applications for compliance with the ATM standards. The AAL separates the stream into packets of 48 bytes (the “payload”) and passes it down to the following layer. The ATM layer adds a few bytes of header consisting of the next destination address and some packet management parameters. The aggregation of the payload and the header constitutes a cell⁶. The cell is then sent to the physical layer for transport toward the next destination, either

⁶This term is used specifically for ATM to define a fixed-length packet

Figure 1.3: ATM System Architecture.

the next network switch or the final destination. Figure 1.3 (redrawn with permission from [6]) illustrates this layering.

The physical layer may be a fibre optic network, a satellite link or an LMCS. This shows the ease of interfacing a fibre optic link (possibly between two major cities) with an LMCS for the local loop (at the network hub close to the client).

1.1.3 Presenting the System Components

The components that are part of LMCS are distributed in two physical locations: the base station and the consumer sites. The base station includes a transmitter, a receiver (or a transceiver), and an antenna. The consumer location combines an antenna and a transceiver which provides signals to a network interface unit. These components are part of the physical layer described earlier. A block diagram of the physical layer is presented in Figure 1.4.

The network interface unit. The interface unit provides the client with an interface between the transmitting/receiving module and the local area network (LAN) or targeted equipment. It comprises a radio-frequency (RF) section and a modem section.

(a) Physical layer

(b) RF blocks

Figure 1.4: Physical layer and RF block of LMCS.

Only a few blocks constitute the RF section as can be seen in Figure 1.4. In the receive path, the preselect filters preserve the desired channel while attenuating the adjacent channels and the noise outside the desired channel frequency band. The low-noise amplifier (LNA) amplifies the received signal to a usable level while adding little noise. The multiplexer down-converts the signal from RF near 28 GHz to an IF in the 1–2 GHz range. In the transmit path of the RF block, the desired signal is up-converted from the IF to the RF. The power amplifier brings the signal to the appropriate power level before passing it to the antenna.

This unit converts the output signal of the receiver to a signal usable by the client's equipment. The modem extracts the useful information from the IF signal, making this information readily available to the multiple applications (such as a LAN, a telephony system or a closed circuit television).

On a final note, it is worth mentioning that Hakegard wrote an excellent background document on the topic [7].

1.2 Focusing on the Project

As mentioned earlier, the LMCS antenna directs most of the signal toward the recipient of the information (the consumer for downlink, the service provider for uplink). An antenna designed with high gain and physically pointed in the right direction achieves this.

Another approach enabling strong directivity is an old technology that has recently been revisited: smart antennas. They consist of an array of antennas whose elements are fed with the same signal but at different phases. The elements' proximity allows constructive and destructive interferences to occur among the EM waves emanating

Figure 1.5: Example of antenna patterns.

from the elements of the array. Proper phases fed to a well-designed array allows for constructive interference to occur in the receiving antenna area, and destructive interference elsewhere. The appropriate phase can be selected at the transmitter by means of electronic devices, such as a transistor.

Figure 1.5 illustrates this concept with three types of antenna patterns. The omnidirectional antenna transmits equally in all directions. The dipole antenna transmits in a “doughnut” or “figure-eight” pattern. The pattern of the array of transmission antennas depends on the number of antennas, the configuration of the array, and the phase on each of these antennas.

An example of an array of antenna is shown in Figure 1.6. A signal source feeds the antenna array elements’ phase shifter. The phase shifter (PS) are controlled to properly modify the the phase of the signal, taking into account path length differences between the signal source and the phase shifter and between the phase shifter and the antenna element.

Figure 1.6: Array of antenna with phase shifters.

1.2.1 Shifting the Phase of a Signal

The appropriate phases can be established at the transmitter by means of an electronic device: the phase shifter.

This thesis investigates the realization of a practical low-loss phase shifters. In specific, this research focuses on the realization of a phase shifter operating at 28 GHz on high resistivity silicon. Specifically, the phase shifter is realized using coplanar waveguide (CPW) transmission lines and including micromachined switches.

Two different topologies for phase shifters were explored and compared. The first category of phase shifter topology is the reflection phase shifter. As the name implies, these phase shifters are based on a reflection of the signal providing an appropriate phase shift in a compact form. The second category is the transmission phase shifter. In this case, the phase shift occurs as the signal is transmitted through a device.

1.2.2 Projecting the Incidence of the Phase Shifter on LMCS

In LMCS applications, the losses are critical. From a marketing point of view, system manageability is crucial to allow swift modifications with a change in consumers' preferences. Low-loss phase shifters meet this requirement by helping to improve the usability of smart antennas that facilitate LMCS development and application.

Results of this research should facilitate the expansion of broadband distribution networks by providing the developers of distribution systems with a practical solution, reducing costs and improving services to clients.

From a broader point of view, it is sensible to envision long-term use of smart antennas and phase shifters. If each transceiving device was to be enabled with highly directive energy transmitting capabilities, issues such as human-made EM noise and effects on human tissues would be significantly reduced.

1.2.3 Listing the Objectives of the Project

The primary goal was to study CPW phase shifter topologies that may be integrated with micromachined switches. As mentioned earlier, these switches have been studied for a few years. The performance still needs to be improved, and it is important to explore their potential uses.

Coplanar waveguides are well understood. They are uniplanar (which reduces costs, fabrication time, and complexity), are reasonably low loss, and more flexible than microstrip or stripline. In this thesis, a simple approach incorporating directional couplers and shorted transmission lines was considered for the design of the phase shifter. Also, recent understanding of series CPW stubs [8, 9, 10, 11] allows for innovative applications of filters into the design of phase shifter.

Another advantage of this design is that it can easily be ported to other frequency bands through simple adjustments to the dimension of the device.

1.3 Organizing the Document

This section outlines the present document. Aside from the introduction, five chapters constitute the thesis. Chapter 2 summarizes the technology and presents the preliminary choices involved in the design process, the microtechnologies and the micromachined switches, and the software tools used for this project.

Chapter 3 details the design of a reflection and a transmission phase shifter. The characteristics of some basic elements and discontinuities are presented, and are followed by the theoretical design procedure. Chapter 3 also imparts the simulation results obtained with ideal elements (i.e., mathematically modeled) to refine the theoretical design. Another set of simulations (involving numerical solutions of Maxwell's equations) was then performed and some final refinements were brought into the design, and these are also described.

Chapter 4 specifies the fabrication material and techniques used to realize the two previously mentioned phase shifters. It also introduces all the components included in the layout of the wafer.

Chapter 5 describes the measurement procedure and reports the results. It compares the simulation results with the measured results and comment on the validity of the results.

Finally, Chapter 6 briefly concludes the thesis, highlighting the important results and providing some directions for future researches.

Chapter 2

Outlining the Technology

Before discussing the design of the phase shifter *per se*, some background information should be introduced. The first part of the chapter explains some early choices related to the material used in the fabrication of the phase shifter. The second part introduces microtechnologies and, more specifically, micromachined switches. It compares the use of mechanical switches to electronic switches. Finally, the third part presents the software packages involved in the design process.

2.1 Making Preliminary Technological Choices

Some choices were made at an early stage of the design and they frame the design decision process. These decisions were constrained by availability of the material, access to the fabrication technology and general directions adopted by the Communications Research Centre Terrestrial Wireless Systems¹ (CRC-TWS). These choices

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were related to the base material (the substrate) and the type of transmission line chosen. They are presented in this section.

2.1.1 High-Resistivity Silicon

Several alternatives exist to constitute the base of the design. Silicon (either standard or high resistivity), gallium arsenide (GaAs), quartz and sapphire are a few of the possible materials available to build electronic components. Silicon has been traditionally used for the fabrication of chips because of its properties, as pointed out by Cressler [12]:

- Easily grown high-quality dielectric,
- can be grown in large single crystal (300 mm at the time of the paper),
- thermal properties allowing fast dissipation of heat,
- large dynamic range of doping with n- and p-type impurities,
- mechanically strong, facilitating handling, and
- easy to make low-resistance ohmic contact to Si.

Also, being the second most abundant element in Earth's crust (after oxygen), Si is widely available. "The ability to control the electrical properties of silicon, and its abundance in nature, have made possible the development and widespread application of transistors and integrated circuits used in the electronics industry." [13]

Silicon has proven useful for electronics applications until recently, when problems started to arise with the increase of the operating frequency. An intrinsic limitation of pure silicon prevents its use at high frequency: electron and hole mobilities.

To circumvent this problem, researchers started using another material: gallium arsenide. Its electrical properties suit microwave frequency applications. However, GaAs is not as readily available and inexpensive as Si. Even though it shows better performance at high frequencies, it is not the material of choice for the mass market.

An approach at resolving the issue of low electron mobility is the use of silicon germanium (SiGe) has been considered since the 1960's but has been successful 13 years ago [12]. It consists of an additional step in the fabrication process that dopes the silicon wafer, with germanium, in the base region of transistors. This doping provides a much higher electron mobility than pure silicon at a more reasonable price than with GaAs. However, being an additional step, it increases the fabrication costs.

High resistivity silicon² is a different substrate and is costly. It would be preferable to work in the current market mainstream (favouring the affordable standard silicon). However, considering the early stage of development of micromachining and local multipoint communication system (LMCS) devices, a proof of concept on high resistivity silicon is a valid approach. It should be kept in mind that it would be desirable to move this work to a cheaper technology to pierce the mass market with arguments such as good performance at low costs.

2.1.2 Coplanar Waveguide

The CRC-TWS (collaborating with Carleton University³ for this project) is migrating towards a uniplanar electromagnetic (EM) structure for simplicity, cost reduction, and fabrication complexity reduction. This is because only a single metal layer is

²High resistivity silicon has a resistivity of 10–100 k Ω -cm, compared to the 0.1 Ω -cm of regular Si.

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Figure 2.1: Definition of CPW dimensions.

required. In some case, air bridges are also required and they are built with the monolithic microwave integrated circuits (MMIC) capacitors. This is done by laying the main CPW conductors on the same layer as the bottom capacitor plates and the air bridges along with the top capacitor plates. Coplanar waveguide (CPW) provides flexibility which is harder to achieve with other types of transmission line.

CPW offers a wide range of geometries for a given application: from the miniature structure, when real estate is critical, to the larger one allowing complex and intricate patterns. Control over the centre conductor and gap widths permits easy variation of the characteristic impedance. Conversely, it gives access to a wide range of total dimension for a given characteristic impedance. More details about the working principles of the CPW transmission lines are presented in Section 3.3.1.

A second advantage of the CPW is the availability of the ground plane. Lying beside the centre conductor, it becomes simple to connect shunt elements, short-circuit the centre conductor, or use the ground plane to build stubs within the centre conductor. No additional fabrication steps, such as the use of vias, are required. An example of CPW line is illustrated in Figure 2.1, where W represents the centre

conductor width, G the gap widths, L the length of the transmission line, and H , the thickness of the substrate.

2.2 Introducing the Microtechnologies

Microelectromechanical systems (MEMS) and on-chip transmission lines are evolving in parallel but are expected to merge at some point in the future, explaining the interest in exploring transmission line and components (such as switches) on Si. This material proves to be unacceptably lossy but techniques like adding an electrical insulator before depositing the transmission line or etching the substrate just below the transmission line improve the performance notably.

Microtechnology is a very recent area of study and a significant amount of work is required to understand it well. At this physical level, concepts such as surface tension, the flow of electrons and mechanical inertia take on much greater significance than in our common understanding of the world. When stiction⁴ prevents two solid objects from being separated, when one electron at a time flows in a transistor, and when 100 G acceleration⁵ does not cause unattached objects to move noticeably with respect to their support, it becomes mandatory to consider factors that are neglected in traditional mechanics and electronics.

This section briefly introduces microtechnologies, the fabrication steps involved in their creation, and micromachined switches.

⁴Stiction: The force required to cause one body in contact with another to begin to move. [14]

⁵Acceleration unit. 1 G correspond to the earth's gravitational acceleration (9.81 m/s^2).

Figure 2.2: Microtechnologies.

2.2.1 Presenting the Microtechnology Categories

The recent development achieved in microelectronics are impacting upon other fields. Optical and mechanical technologies are benefiting from these advances. Figure 2.2 illustrates the relations between the different fields while the following paragraphs describe the vocabulary associated with these new concepts.

Optical technology involves materials and structures capable of altering the propagation and behaviour of light, both visible and non-visible. It also includes devices capable of generating or absorbing light. In microtechnology, optics is involved, for example, in the synthesis, capture and processing of images on a chip. Another example would be the transmission and reception of information over optical fibre. Both geometrical and physical optics devices can be produced in microtechnology.

Electrical technology encompasses the production and utilization of electrical energy as well as its transformation from and to other forms of energy. An example would be the chemical energy of a battery used to provide power to a calculator. The direct current (DC) voltage powers the electronics that perform

Figure 2.3: Photograph of a micromachined motor [1].

mathematical operations. The electricity also causes physical changes in the liquid crystals of a display to show the results. In microtechnology, electrical technology consists of processing the information under the form of electrical signals or flows of current.

Mechanical technology involves moving objects and their reaction to forces. It encompasses the concepts of force, acceleration, torque, rotational velocity, friction, strain and stress, among others. In microtechnology, a typical application of this might be a rotating gear system which moves a physical beam on a silicon chip, as can be seen in Figure 2.3.

Opto-mechanics relates to an overlap between optical and mechanical technologies. An example of an opto-mechanical device is the Bragg cell. This device permits the modulation of light by a mechanical wave applied to the light transmitting medium.

Optoelectronics is a combination of optical and electrical technologies. Typical applications of opto-electronics are found in fibre optics. Lasers and detectors are devices that respectively convert electrical energy into coherent light and light into electrical energy.

Microelectromechanical Systems form the intersection of electrical and mechanical technologies. They enable mechanical-to-electrical or electrical-to-mechanical energy conversion. A popular example of a MEMS is the electrical motor that can be fabricated on the surface of a chip. Constituted of the same basic elements as a “normal size” motor, this device has a rotor revolving under the application of the required current. Another example is an accelerometer whose mass pulls down a spring loaded arm when submitted to acceleration. The electrical characteristics of the arm change proportionally with the strain applied on the material.

Micro-optoelectromechanical Systems (MOEMS) are the overlap of the three preceding technologies. A typical example is an optical fibre duplexer. In this device, a laser source converts electrical energy to optical energy. A length of fibre connects this laser source to the duplexer which is switched by a gear system. Finally, the collecting fibre ends on a detector that converts the optical energy back into electrical energy.

Additional Comments

Other branches of research are also involved in microtechnology for some specific applications: biology, biochemistry and thermodynamics, to name just a few, also benefit from the techniques devised for optics, electronics and mechanics.

The microtechnology categories can have deeper subdivisions.. An important one related to this project is radio-frequency microelectromechanical systems. RF-MEMS relate to the use of devices at millimeter-wave and microwave applications. RF-MEMS

meet the conditions to be considered as MEMS but also meet additional requirements to suit high frequency applications, such as low loss.

The principal requirement for RF-MEMS is isolation of the structure from the lossy substrate material. As mentioned earlier (in Section 2.1.1), substrate losses with “ordinary” silicon are too large at high frequencies, causing deterioration of components performances. Two solutions are possible to reduce these losses: using a substrate showing low dielectric loss at the desired frequencies, or micromachining the substrate so as to leave only a thin layer of substrate for structural support. From an EM point of view, the second solution is nearly equivalent to having the structure built with air dielectric.

2.2.2 Introducing the Techniques

Micromachining consists of the fabrication of miniature devices falling into at least one of the following categories: electrical, optical and mechanical. They are fabricated using standard batch-process chip technology. These devices range in size from a few microns to a few millimeters. A wide variety of devices can be fabricated and the possibilities are limited only by the fabrication tolerances and the designer’s imagination and skill.

This section outlines the principal micromachining techniques. This section is not exhaustive as the field of micromachining is constantly expanding and evolving. Sze [15] and a special issue from the Proceedings of the Institute of Electrical and Electronics Engineers (IEEE) [16] are excellent references for further reading on the topic of micromachining.

Several micromachining techniques exist: bulk micromachining, surface micromachining, single point micromachining and *Lithographie, Galvanoformung, Abformung* (LIGA - German for lithography, electroplating and molding), to name just a few. These are presented in the following paragraphs.

Bulk micromachining

It is a common technique which consists of etching the substrate (usually silicon) in such a way that the desired device “emerges” from the raw material. It is similar to a sculptor carving marble away to leave a finished work.

Limitations are imposed on the complexity of the devices. The principal limitation is due to the cleavage planes of the silicon crystal. A (100) wafer [17], for example, allows for different structures than a (110) wafer [15, pp. 43-46]. Another important limitation is the precision on the penetration of etch-stop in the wafer, bounding the complexity of the design emerging from the silicon.

Surface Micromachining

It consists in depositing onto the substrate materials forming the object. This technique, invented in 1967 by Nathanson [1], often involves sacrificial layers holding the parts of the object in place during fabrication. For example, the first layer of gold atoms of an air bridge would not stand in air, it needs to lay on top of some material. Dielectric is deposited to favor gold deposition. After fabrication, the material thickness provides enough mechanical strength to support itself. The sacrificial layer is washed away, leaving the object as intended.

The limitations were of two kinds: economical and technological. Economical limitations arise because the more vertically complex and intricate the structure is,

the larger the number of masks required. Technological ones arise because of the characteristics of the available etchants. The choices are limited and interactions with all exposed materials at time of use have to be considered. It is necessary to insure that the steps of a process will not adversely influence or destroy the elements made in a previous step.

Single point micromachining

This technique is closer to the common understanding of machining in its mechanical sense. It uses a tiny sharp diamond edge to literally machine the material. It usually is used to obtain grooves for optical fiber aligners.

Limitations on this type of micromachining depend on the precision of the equipment and the structure that is desired. While chemical etchants are able to reach in fine interstices or even under existing structure, a diamond at the tip of a shaft does not have the flexibility and the space necessary to machine the undersides of elements of the structure.

LIGA

This acronym originates from the German term for Lithography, Electroplating and Molding. Lithography is used to pattern the desired structure (e.g., a gear), in a thick layer of photosensitive resist (photoresist). Electroplating is then used to build a metal form of the desired structure by growing metal in the space made in the photoresist. Plastic is then injected around the metallic mold to obtain multiple replicas of the desired structure.

2.2.3 Switching to MEMS

Electronic switches, such as p-doped/intrinsic/n-doped (PIN) diodes and transistors, are possible alternatives to perform switching at microwave frequencies. However, their relatively high losses and parasitic effects, combined with the research in microtechnology, encourage a new alternative: micromachined switches.

Among the elements that can be micromachined, MEMS switches are of particular interest in a phase shifter. This section explains their working principles, and points out their limitations at the current stage of development.

Introduction to MEMS Switches

MEMS switches typically consist of a bottom plate and a top plate of metal separated by a few microns of distance. The working principle is simple. The bottom plate lies on the wafer and does not move. In the open position, the rigidity of the material maintains the top plate straight and distant from the bottom plate. If a difference in potential is applied between the plates, an electrostatic force is induced and the plates attract each other. If the difference in potential is sufficient, the electrostatic force overcomes the rigidity of the material, and the top plate bends and contacts the bottom plate. When the difference in potential between the plates falls below the operating voltage of the switch, the elasticity of the material brings the switch back to its open position.

There are two types of MEMS switches: capacitive and resistive. The main noticeable difference resides in the presence or the absence of a dielectric layer on the bottom plate. Their working principles differ also. If a DC block is required and the switch is used at high frequencies, the capacitive version should be used. If DC blocking is

Figure 2.4: MEMS Switches.

not important or the switch is intended for use at low frequencies, the resistive switch would be more appropriate. The control voltage is brought either through the centre conductor for the capacitive switch or through a dedicated conductor for the resistive switch. Diagrams of the two types of switches are presented in Figure 2.4.

Design Trade-off

The above explanations of the working principles of the two types of MEMS switches allows to now introduce the factors that are traded-off to achieve a good switch design. These factors are presented in the following paragraphs.

The major trade-off factor in a MEMS switch design is between parasitic capacitance and operating voltage. Negligible parasitic capacitance – both in the up state (open) and in transition between the up and down states – and low actuation voltage depend on common design parameters. These parameters typically affect the capacitance and the voltage in opposite ways. On one hand, a small steady-state

open parasitic capacitance requires small and widely spaced plates. Also, for a small transition parasitic capacitance, a large coefficient of elasticity in the plate material is required: the rigidity insures short transition times and small mechanical bouncing⁶ of the top plate. Bouncing varies the parasitic capacitance to ground and thus, introduces an undesired modulation into the signal going through the line.

On the other hand, a small actuation voltage requires large area (for greater total electrostatic force) and small distance between the plates (as the electrostatic force is inversely proportional to the square of the distance [1]). A small coefficient of elasticity is preferable to reduce the required actuation voltage since a smaller mechanical force has to be overcome by the electrostatic force.

State of the Art

Multiple attempts have been made to fabricate MEMS switches [18, 19, 20, 21]. The main concern of the researchers is still reducing the actuation voltage while maintaining low parasitic capacitances [20]. Recently published research presents switches with reasonably low losses (often below 0.5 dB) [20, 22] requiring an actuation voltage that varies significantly, depending on the processing, from 14 V to 50 V [23, 20]. This area needs to be further researched as newly available chipsets have operating voltages below 2 V.

⁶The elasticity of the top plate brings it back to its upper position. Because of this elasticity, the plate actually reaches a slightly higher point than its original position and then lowers back to the upper position, oscillating for a short period of time around its upper position. This is called mechanical bouncing.

The CRC-TWS group lead by Dr. Malcolm Stubbs⁷, is collaborating with Dr. Niall Tait⁸ of the Department of Electronics at Carleton University on micromachined switches. Unfortunately, their research was not at a stage allowing inclusion into the design when this program was initiated.

2.3 Presenting the Software Tools

Two types of tools are used in the design process: Simulation Program with Integrated Circuit Emphasis (SPICE) simulator and Electromagnetic Field Solver (EMFS). The former uses component models (typically in terms of mathematical equations) and solve matrices to satisfy Kirchoff's laws. They are fast but their precision is limited by the complexity of the models. EMFS, as the name implies, solves the EM field in and near a structure. It is based on numerical solutions of Maxwell's equations through finite difference time domain (FDTD), integral equation techniques or finite element techniques. EMFS is more versatile and more precise than the SPICE simulators but require greater computer resources, and are more time consuming.

Practical designs make use of both types of tools. After generating an electrical model (based on electrical characteristic of transmission line) or a physical model (based on the physical characteristics of the transmission lines), the tools using lumped element models can promptly confirm the validity of this first approximation. Refining the design at this stage generates a device ready for more advanced

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simulation which can be ported to an EMFS. The design should be as close as possible to its final stage since these simulation are time consuming. The EMFS takes into account finer details, considers the discontinuities between the elements and includes the details that would be overwhelming for theoretical consideration or too specific for a mathematical model.

The software tools that were used for this project are:

- Hewlett Packard⁹ Linecalc®
- Hewlett Packard Advanced Design System®
- IC Edit®

More informations can be found at their respective manufacturer's web site: <http://www.agilent.com> and <http://www.iceditors.com>.

⁹Now Agilent Technologies.

Chapter 3

Designing the Phase Shifters

This chapter introduces the working principles of a phase shifter and describes a few possible topologies. It then describes the topology selection and the design processes for two 28 GHz LMCS-based phase shifters.

The first approximations of the phase shifters were based on ideal transmission lines. Thus neglecting undesired losses, impedance mismatches and discontinuity effects simplified the validation of the basic concept. Refining the design using realizable coplanar waveguide (CPW) transmission lines, corrections were made to account for some discontinuities and peculiarities of CPW. Electromagnetic (EM) simulations of the complete designs then helped to confirm their accuracy. Finally, the simulations and their results are presented along with a comparisons of the performances.

3.1 Laying the Fundamentals of Phase Shifters

A component is said to introduce a phase shift into a signal if it alters the phase and preserves the other characteristics of this signal. The simplest phase shifter is a delay line. It introduces a phase shift corresponding to its electrical length.

Reflection and transmission phase shifters were considered for this project. In the former type, a signal enters the input port, travels across the device, is reflected from a high reflection coefficient element, travels backward across the device and leaves through the output port. In the latter, a signal enters the device on the input port, travel once across the device and exits the device at the output port.

The reflection phase shifter is slightly less precise, since the phase shift error is doubled because the signal travels twice through the element that introduces the phase shift. For the same reason, however, it occupies less real estate than the transmission phase shifter. The area is reduced even though an additional element is required to separate the input and the output signals. The element separating the signals is typically a directional coupler (e.g., quadrature hybrid, ring hybrid, tapered coupled line hybrid or coupled-line couplers).

As mentioned, the transmission phase shifter is slightly more precise but occupies more area. Different topologies can form a transmission phase shifter, such as a delay line, a loaded line, or filters. They consist of two (or more) lines with different phase characteristics between which the signal is switched. For the delay line topology, the difference in physical (and thus electrical) line lengths affects the phase of the signal. The loaded line and filter phase shifters work similarly: the phase characteristics of the reactive elements contribute to the phase of the signal and switching between different branches selects which elements are active.

3.2 Choosing the Topologies

This section presents the potential topologies for the reflection and the transmission phase shifters. It also explains the choice of the final topology for each type of phase shifter.

Reflection Phase Shifter

The couplers available for the reflection phase shifter have different characteristics. The final choice of which to use is based on:

- bandwidth,
- losses,
- ease of fabrication,
- compatibility with CPW, and
- area occupied.

Some of the possibilities have been relinquished for various reasons. The coupled line coupler is difficult to realize with -3 dB on each of the outputs because the design would be too sensitive to fabrication tolerances. The ring hybrid is difficult to realize with some fabrication processes since some of the processes limit the geometry angles to multiples of 45° . Also, CPW implementation interferes with the proper behaviour of ring hybrid due to unequal lengths of the inner and outer ground planes (the outer ground plane is 12% longer than the inner ground plane). Also, the arm directions makes this coupler difficult to use with delay lines used for the phase shifting. The tapered coupled line presents the same fabrication difficulties as the ring hybrid.

Type of Coupler	Dimension (μm)	Area(mm^2)	0.5 dB BW (%)
Ring hybrid	ϕ 2174	3.7/4.7†	28
2-branch branch-line	1339×1279	1.7	18
3-branch branch-line	2165×1396	3.0	29
Coupled line	766×1099	0.85	51

† 3.7 mm^2 is the area of the circle occupied; 4.7 mm^2 is the area of the square circumscribing this circle.

Table 3.1: Comparison of four types of coupler.

Figure 3.1: Prototype of a reflection phase shifter.

The branch-line coupler (also called the quadrature hybrid) offers low losses, enough bandwidth for the desired application and flexibility in the choice of the two- or the three-branch version. Table 3.1 summarizes the characteristics of four types of coupler.

Micromachined switches provide the short circuit required to introduce the desired phase shift. They electrically connect the CPW centre conductor to ground. In the open state (up), the switches are assimilated to air bridges, keeping the ground planes at the same potential and introducing a hint of parasitic capacitance (less than one femtofarad, yielding a parallel impedance 100 times larger than the characteristic impedance of the line). In the closed state (down), the switches introduce only a few ohms of contact resistance. Figure 3.1 shows a prototype of the chosen topology for the reflection phase shifter.

Type of coupler	Dimension (μm)	Area(mm^2)
Delay line	1575×2142.5	3.4
Filter	1474×1000	1.5

Table 3.2: Comparison of 1-bit transmission phase shifters.

Transmission Phase Shifter

The filter phase shifter is the topology offering the most flexibility with low losses and compact dimensions. It occupies less area than the delay line (typically 65% of the area occupied by a 1-bit delay line phase shifter), and is more flexible than the loaded line since it has more degrees of freedom. Table 3.2 outlines the area occupied by the different types of transmission phase shifters.

Even though more elements are present in the filter phase shifter than in the loaded line, the area is kept small. This is possible because of the recent advances that have been realized in shunt and series stubs for CPW [8, 9, 10, 11].

It follows that switching between two (or more) filters to obtain the desired phase shift is the most desirable alternative. Figure 3.2 illustrate the working principle of the transmission phase shifter.

3.3 Characterizing the Basic Elements

The following section presents the primitive elements used to construct the branch-line coupler and filters phase shifters. The first part considers the elements common to both topologies; this is followed by a presentation of the elements specific to the reflection or the transmission phase shifter.

Figure 3.2: Prototype of the transmission phase shifter.

3.3.1 Common Elements

This section introduces some of the building blocks used in the final design. They are the transmission line and the two loads (short circuit and open circuit).

Transmission Line

Aside from specific details, coplanar waveguide transmission lines obey the rules applying to generic transmission lines. Specifically, they operate in the quasi-transverse electromagnetic (TEM) regime because the dielectric is non-uniformly distributed around the line – it has dielectric material below the plane of the line and air above it. Its symmetric geometry supports odd and even modes (Figure 3.3). The even mode, also called the slot-line mode, was not desirable for this realization. At discontinuities, air bridges were installed to maintain the two ground planes at the same potential, which suppress the even mode.

Figure 3.3: Coplanar waveguide propagation modes.

Figure 3.4: Definition of CPW dimensions.

The characteristic impedance of a CPW has been determined by conformal mapping. This technique reduces the complex structure of the CPW into a parallel plate problem. Collin [24] developed the conformal mapping for a coplanar waveguide. The general case is for a finite substrate thickness, H , and a cover above the CPW at a given height, H_1 (Figure 3.4).

A simple series of equations can be solved, yielding a numerical value of the characteristic impedance. First, the modulus for the complete elliptic integral is determined using¹:

¹There is a typographical error in the equation for k_1 in [24]. There should be a factor of 4 in the denominator of the argument of the hyperbolic tangent present in the denominator of the equation for k_1 . This factor is included here.

$$\begin{aligned}
k_1 &= \frac{\tanh(\pi G/4H)}{\tanh(\pi(G+2W)/4H)} \\
k_2 &= \frac{\tanh(\pi G/4H_1)}{\tanh(\pi(G+2W)/4H_1)} \\
k_3 &= \frac{\sinh(\pi G/4H)}{\sinh(\pi(G+2W)/4H)}
\end{aligned} \tag{3.1}$$

as well as the complementary modulus, given by:

$$\begin{aligned}
k'_1 &= \sqrt{1 - k_1^2} \\
k'_2 &= \sqrt{1 - k_2^2} \\
k'_3 &= \sqrt{1 - k_3^2}
\end{aligned} \tag{3.2}$$

It can be shown that if the cover is sufficiently far from the transmission line, $k_2 \simeq k$ (i.e., $\lim_{H_1 \rightarrow \infty} k_2 = k$). For this investigation, the absence of a cover made it unnecessary to evaluate k_2 using equation 3.1.

It is then possible to evaluate the ratios of the complete elliptic integral using:

$$\frac{K(k)}{K(k')} = \frac{K(k)}{K'(k)} = \frac{K}{K'} = \begin{cases} \frac{1}{\pi} \ln \left(2 \frac{1+\sqrt{k}}{1-\sqrt{k}} \right) & 0.7 \leq k \leq 1 \\ \left[\frac{1}{\pi} \ln \left(2 \frac{1+\sqrt{k'}}{1-\sqrt{k'}} \right) \right]^{-1} & 0 \leq k \leq 0.7 \end{cases} \tag{3.3}$$

$$K' = K(k') = \frac{2}{1+k} K \left(\frac{1-k}{1+k} \right)$$

The filling ratio (an expression of the proportion of the total field in the dielectric to that in the air) is computed following:

$$q = \frac{\frac{K(k_3)}{K(k'_3)}}{\frac{K(k_2)}{K(k'_2)} + \frac{K(k)}{K(k')}}.$$

which allows the computation of the effective dielectric constant, ϵ_{eff} , with

$$\epsilon_{eff} = 1 + q (\epsilon_r - 1)$$

Finally, the characteristic impedance of the coplanar waveguide can be calculated using

$$Z_c = \frac{Z_0}{2\sqrt{\epsilon_{eff}} \left[\frac{K(k_2)}{K(k'_2)} + \frac{K(k)}{K(k')} \right]} \quad (3.4)$$

where Z_0 = characteristic impedance of vacuum, 120π .

The final comment regarding the CPW concerns the z-axis, the length of the transmission line. Using

$$\beta \cdot \Delta l = \Delta\theta \quad (3.5)$$

where β is the propagation constant, Δl is the physical length, and $\Delta\theta$ is the electrical length, it becomes trivial to determine the required physical length for a desired electrical length since the propagation constant is determined by

$$\beta = \omega \sqrt{\mu_0 \epsilon_0 \epsilon_{eff}}$$

where ω is the signal frequency in $\frac{rad}{s}$ and ϵ_{eff} is the effective dielectric constant (calculated above).

Open Circuit and Short Circuit Terminations

This section presents the open and short circuits as design components as loads applied to the end of a transmission line. These discontinuities are presented in Section 3.4.

The cases of open- and short-ended transmission lines are two special cases of a terminated transmission line. For simplification, let us consider the case of a lossless line. The reflection coefficient at the load is given by

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0}$$

where Z_L is the impedance of the load and Z_0 is the characteristic impedance of the line. Considering a 50Ω transmission line, the reflection coefficient becomes:

Open circuit	Short circuit
$\Gamma = \frac{\infty - 50}{\infty + 50}$	$\Gamma = \frac{0 - 50}{0 + 50}$
$\Gamma = 1$	$\Gamma = -1$

showing that reflection occurs in both cases. The only difference is the negative sign in the case of a short circuit. It represents a total reflection with inversion of phase. This inversion of phase is necessary to ensure a zero sum of the incident and the reflected voltage waves, to respect the condition imposed by the short circuit. Conversely, for the open circuit, the positive sign shows that the total reflection is in phase with

the incident wave and that is to ensure the correct potential difference between the transmission line and the ground as specified by the condition on the open circuit.

An open circuit presents a large load impedance to the line (infinite impedance in the ideal case). Considering the definition of impedance in terms of fields ($Z=E/H$), the large impedance calls for a large E-field and a small H-field, suggesting a reactive behaviour of capacitive nature.

A short circuit fields definition (again, $Z=E/H$) requires a small E-field and large H-field to meet the imposed condition of $\sim 0\Omega$ (a short circuit). The small electric field and the large magnetic field is characteristic of an inductive behaviour.

3.3.2 Reflection Phase Shifter

The devices presented in Section 3.2 are directional couplers. Such a four port device presents interesting characteristics for phase shifter applications. When properly designed, a directional coupler can split a signal or combine two signals. When presented with a signal on its input (refer to Figure 3.5), it splits the power of this signal between the thru and coupler ports. These output signals have phase differences of 90° and 180° with the input signal, respectively. The passive and reciprocal nature of this kind of coupler allows the combination of two signals applied with appropriate phase at the thru and coupled ports. The sum is collected on the fourth (isolated) port of the device. Figure 3.5 illustrates a generic four port coupler, which can be used in dual mode.

The reciprocal nature of the coupler forms the foundation of a phase shifter. A signal applied to the input port of the coupler is split equally between the thru and

Figure 3.5: Generic four port coupler.

Figure 3.6: Branch line couplers.

coupled ports. It travels along the connected transmission lines which introduce half of the desired phase shift. The signals reflect at short circuits (ends of transmission line or closed switches) and travel back on the transmission lines, introducing the second half of the phase shift. The signals are recombined in the coupler, which now acts as a power combiner. The recombination is possible because the relative phase of the two halves of the signal is ensured by the use of equal electrical length of transmission lines. The signal finally leaves the output port of the phase shifters. Note the distinction between output port of the phase shifter and the “output” ports of the coupler as they are usually considered.

Two-branch Branch-line Coupler

Pozar presents, in *Microwave Engineering*, a detailed analysis of the two-branch quadrature hybrid [25, pp. 379-383]. The important results for this application are the voltage waves amplitudes (shown on Figure 3.6 at each port, given by:

$$\begin{aligned} B_1 &= 0 \\ B_2 &= -\frac{j}{\sqrt{2}} \\ B_3 &= -\frac{1}{\sqrt{2}} \\ B_4 &= 0 \end{aligned} \tag{3.6}$$

with an input signal on port 1. Equation 3.6 indicates that port 1 (input port) is matched at the input, port 2 (thru port) receives half of the power with a -90° phase difference with port 1, port 3 (coupler port) also receives half of the power with a -180° phase difference with port 1, and port 4 (isolated port) is isolated from port 1.

The bandwidth of a two port quadrature hybrid is limited by the requirement for quarter-wavelength elements that compose the coupler. With two branches, the 0.5 dB bandwidth is approximately 18% (for an ideal transmission line model).

Three-branch Branch-line Coupler

The same analysis may be carried out for the three-branch branch-line coupler. It shows similar wave amplitudes as with the two branch coupler. The phases are different, however, with -180° and -270° for the thru and coupled port, respectively. The use of an additional branch provides a larger bandwidth. From 18% (two-branch

version), the bandwidth of a three-branch quadrature hybrid increases to 29% (with ideal transmission lines).

Two difficulties arise from this approach: a larger occupied area (almost double) and increased losses. The dielectric and conductor losses both increase because more elements of transmission lines are involved between the input and the thru/coupled ports.

3.3.3 Transmission Phase Shifter

The elements used for the transmission phase shifter are series and shunt stubs. Instead of their well-known use for matching, they are used as distributed reactive elements to build filters. New advances achieved in the research of series stubs [9] allow keeping the dimensions reasonably small while maintaining appropriate filter characteristics.

Series and parallel stubs are well-known in the theory of transmission lines and are explained in the literature [25, pp. 259-266]. The input impedance of a stub is calculated as

$$Z_i = Z_0 \frac{Z_L + j \cdot Z_0 \tan(\beta l)}{Z_0 + j \cdot Z_L \tan(\beta l)}$$

In the case of a short-circuited or open-circuited line, the load impedances are respectively zero or infinite impedance. These translate into input impedances given by, respectively,

$$Z_i = j \cdot Z_0 \tan(\beta l) \quad (3.7)$$

$$Z_i = -j \cdot Z_0 \cot(\beta l) \quad (3.8)$$

Figure 3.7: Series stub.

For practical transmission lines, technical considerations are often an obstacle to their use. Several factors favor their implementation in CPW's: the better understanding of their effects on CPW transmission line, the flexibility offered by CPW and the improvements in the precision of fabrication processes all lead to a better use of short- and open-circuited stubs.

Series Stubs

The two types of series stubs are the short-circuited and open-circuited stubs. By inspection of equations 3.7 and 3.8, we can minimize stub length by using a short-circuited stub when an inductive element is required in the design and an open-circuited stub for a capacitive one. A simple design shows a significant difference in the effect on the transmission line of the short-circuited stubs compared with the open-ended stub. This fact is discussed in the following paragraphs and was appropriately be accounted for in the design. Figure 3.7 illustrates a series capacitive stub and its CPW realization.

To better understand the difference in quality between the short-circuited and open-ended stubs, consider the particular case of a quarter-wavelength long series transmission line (refer to Figure 3.8). If this line is short-circuited at the end, the

Figure 3.8: Effects difference of the CPW series stubs.

input impedance would be very large (ideally infinite). Used as a series stub, this open circuit reflects the incoming signal toward its source. Ideally, no portion of the signal flows through and propagates on the section of the transmission line following the short-circuited stub (or, in other words, the open circuit that is seen at the input of the stub, at the frequency of interest).

The same quarter-wavelength line has an input impedance of nearly zero ohm when open-circuited. Used as a series stub, it translates ideally to a zero reflection coefficient, meaning that no signal is reflected back and the signal (nearly) all flows on the transmission line after the stub. In practice, however, the two transmission lines connected to the stub are not matched. The stub engraved in the centre conductor perturbs the geometry and the normal behaviour of the transmission line. This

modifies the characteristic impedance of this portion of line, resulting in reflection toward the input.

Two methods exist to reduce reflections. One is to reduce the total width — sum of the centre conductor and gap widths — of the stub. This can be achieved with a CPW which allows the reduction of the dimensions while maintaining a constant characteristic impedance. Another possible way to reduce this effect would be to match the stub to the transmission line. It is difficult to properly model the coupling between the centre conductor and the stub (the ground plane of the stub is also the centre conductor of the main transmission line). Engraving the stub in the centre conductor alters the intrinsic behaviour of the transmission line since its geometry is modified.

Shunt Stubs

Similarly, there are also the short- and open-ended shunt stubs. They are used respectively when an inductive or capacitive element is required. There are two methods of implementing shunt stubs in CPW (refer to Figure 3.9). The first approach consists of extending a part of the centre conductor [A] into the ground plane [B], this extension being the centre conductor of the stub [C]; the main transmission line ground plane also is the stub ground plane. In the second approach the main transmission line ground plane [D] extends into the main line centre conductor [E]. The main transmission line ground plane becomes the centre conductor of the stub [F]; the centre conductor of the main transmission line acts as its ground plane.

The first approach offers fewer perturbations of the transmission line characteristic impedance. The stub is simply added to the line, opening the ground plane perpendicularly to the line but not affecting the characteristics of the main line sig-

Figure 3.9: Asymmetric shunt stubs.

nificantly (except for the desired reactive effect). The second approach reduces the space occupied by the stub significantly since it is within the centre conductor instead of occupying the ground plane. The modified geometry affects the characteristics of the main line, however, and so requires some additional design work to compensate for this effect. The results in Chapter 5 show that it is possible to obtain a satisfactory outcome.

Survey of CPW Realization of Basic Element

It is possible to achieve an improvement of the realization of basic elements (such as a distributed capacitor and distributed inductor) and filters with the technique presented above. The amplitude and phase characteristics are acceptable and the area reduction which can be accomplished is appealing.

The transmission line models and CPW realizations of several basic elements and simple circuits can be found in Appendix A.

Figure 3.10: 3-dimensional E-field of a CPW.

3.4 Characterizing the Discontinuities

Before introducing the discontinuities *per se*, a brief explanation of the coplanar waveguide transmission line is presented. For simplicity, the discussion is limited to considering fields in two dimensions, in the plane of the transmission line. It can be extended to three dimensions by allowing the E-field to arch over and under the gaps between the centre conductor and the ground planes as shown in Figure 3.10.

An ideal coplanar waveguide has its electric field going from the centre conductor to the ground plane, perpendicular to the length of the transmission line. Considering a single element of this line (with semi-infinite ground planes), the E-field goes from this element to the ground plane, with an amplitude that is reduced with an increase of distance (between the line and the element of ground plane) as can be seen in Figure 3.11. In an ideal CPW, the E-field components that are not orthogonal to the direction of propagation are cancelled by adjacent elements thus leaving components of the E-field only in the X- and Y-directions, where X and Y are in the plane of the cross-section of the CPW and Z is the direction of propagation (see Figure 3.12).

Figure 3.11: E-field in CPW lines.

Figure 3.12: Definition of the X-, Y-, and Z-directions.

Discontinuities modify the behaviour of the CPW from the theoretical ideal. The following paragraphs present these discontinuities, while compensation techniques are presented in the following section.

The presence of discontinuities in the realization is due to the fact that the ideal transmission line model assumes perfect matching of the components, i.e. no coupling occurs among them and no abrupt change of centre conductor direction occurs. The ground plane is considered semi-infinite (on each side) perpendicular to the transmission line and infinite along the transmission line. In practice, discontinuities such as coupling and effective electrical length modifications affect the behaviour of the device. Coupling arises between nearby elements, and changes in the electrical length of stubs or transmission lines can be caused by open or short circuits. These effects must be taken into account for proper results. Table 3.3 presents these effect and their associated model.

3.5 Applying the Theory to Design

The process which was used to achieve optimal reflection and transmission phase shifters is explained in this section.

In both cases, the dimensions of the CPW structure were chosen to fit the Pico-probe probe pitch used at the Communications Research Centre (CRC) for measurements. A total width ($W+2\times G$) of $178\text{ }\mu\text{m}$ (7 mils) provided $89\text{ }\mu\text{m}$ from mid-centre conductor to ground plane, which safely accommodated the $150\text{ }\mu\text{m}$ probe pitch (R). For a CPW characteristic impedance of $50\text{ }\Omega$, a centre conductor width (W) and a gap width (G) of respectively $83\text{ }\mu\text{m}$ and $47.5\text{ }\mu\text{m}$ provided the chosen $W+2G$ of $178\text{ }\mu\text{m}$. A design rule of thumb calls for a ground plane equal to $W+2G$ to suit-

Discontinuity	CPW realization	Model
Tee junction
Impedance step
Bend
Air bridge
Open circuit
Short circuit

Table 3.3: Summary of CPW discontinuities.

Figure 3.13: Probe and CPW dimensions.

ably approximate a semi-infinite ground plane; this value ($178 \mu\text{m}$) was selected for a minimum ground plane width (N). Figure 3.13 illustrates the probe-CPW dimension relationship.

3.5.1 Reflection Phase Shifter

A quadrature hybrid coupler and transmission lines were assembled to form a reflection phase shifter. As mentioned earlier (in Section 3.3.2), the quadrature hybrid coupler separates the input from the output of the phase shifter. The transmission lines allow the selection of the proper signal phase shifting.

The ideal quadrature hybrid model is well understood [25, pp. 379-383]. It is composed of quarter-wavelength long elements of characteristic impedance Z_0 and $\frac{1}{\sqrt{2}} \cdot Z_0$ ohms for the 2-branch version. The three-branch includes transmission line of $\frac{1}{\sqrt{2}} \cdot Z_0$ and $(\sqrt{2} + 1) \cdot Z_0$ ohms. Figure 3.14 illustrates the two topologies.

The length of the connected transmission lines depends on the desired phase shift. The final design has 360° short-circuited CPW transmission lines. To show the proper modes of operation in the absence of switches, distinct couplers associated with different length of transmission line were fabricated (refer to Section 4.2.2 for more details on the devices included in the fabrication).

Figure 3.14: Characteristic impedance and length of the component of an ideal quadrature hybrid.

Simulation of the phase shifter with ideal elements showed the expected behaviour of the design. The CPW version (with losses but without discontinuity effects) had a similar behaviour. Figure 3.15 presents the ADS simulation file for both designs and Figure 3.16 compares the results of both approaches, showing excellent agreement between the ideal transmission line and the coplanar waveguide (without discontinuity effects) simulations.

3.5.2 Transmission Phase Shifter

For the transmission phase shifter, the main components are filters. As a proof of concept, a 1-bit transmission phase shifter is presented. To obtain this phase shifter, two filters were required. The chosen approach consisted in the design of a low-pass and a high-pass filter. The filters' cut-off frequencies are carefully chosen considering the desired bandwidth (2 GHz) centered on 28 GHz. For these filters, the amplitude characteristic is important but the phase characteristic is more crucial.

An easily realizable CPW filter topology is the tee configuration: two reactive elements in series (either two inductances or two capacitances) and the dual reactive

(a) Ideal elements

(b) CPW realization

Figure 3.15: ADS simulation file of reflection phase shifter.

Figure 3.16: Simulation results of reflection phase shifter.

Figure 3.17: Filters tee configuration.

element (respectively, a capacitor or an inductor) connected between the ground plane and the common point of the first two reactive elements. Series stubs formed the top branches of the tee while the element connected to ground consists of a parallel stub. Figure 3.17 presents these two lumped element configurations.

For the chosen filter configurations, it was necessary to determine the values of the reactive elements providing a 180° phase difference at 28 GHz between two lumped element filters for the 1-bit phase shifter. The amplitude characteristics at this frequency were also kept in mind for the design.

At this stage, ADS was used to optimize the values of the reactive elements. Referring to Figure 3.17, the values for C1, C2 and C3 were chosen to be the same, as

were the values for L1, L2, and L3. The optimization goal was to obtain an absolute value of the difference of the phase characteristics, $|\angle S_{21}(HPF) - \angle S_{21}(LPF)|$, at 28 GHz, to be between 179° and 180° . The optimizer returned the values of 0.17 pF for the capacitors and 0.19 nH for the inductors. They correspond to an impedance magnitude of $33.4\ \Omega$ in both cases ($\mp j \cdot 0.67\ \Omega$ respectively for the capacitor and the inductor respectively, normalized to $Z_0 = 50\ \Omega$).

Knowing the impedance values of the reactive elements, the physical length of the stubs is determined either graphically (using a Smith chart) or numerically (using Equations 3.7 and 3.8). The numerical approach gives a more accurate length but at this stage of the design, such an additional accuracy is not required: the discontinuities effect will have a greater effect than the precision loss of the graphical approach.

Transferring impedances onto a Smith chart ($0+j0.67$ for the short-circuited inductive stub, and $0-j0.67$ for the open circuit capacitive one), the stub lengths are determined to be 0.094λ for the inductive stub (starting at $Z=0\ \Omega$) and 0.156λ for the capacitive stub (starting at $Z=\infty\ \Omega$). These electrical lengths correspond to physical dimensions of respectively $410\ \mu\text{m}$ and $681\ \mu\text{m}$, using HP Linecalc.

3.6 Simulating and Optimizing the Design

The CPW designs were verified and adjusted to match the ideal element versions as closely as possible. After an introduction to the techniques of optimization iterations, simulation results are presented and compared to the ideal-element simulation results.

3.6.1 Reflection Phase Shifter

After simulation of the CPW version of the phase shifter (without the discontinuities), the design was captured into (i.e., converted for analysis by) HP Momentum. This simulator showed the deficiencies of the design at the then-current stage, since the discontinuities affected the proper behaviour of the phase shifter. When the discontinuities were taken into account, the required length and characteristic impedances of the components were altered to obtain better performances of the passive structure.

The optimization capabilities of the simulator (HP Momentum) could be used to adjust the design. However, for a relatively complex structure such as a complete phase shifter consisting of a three-branch branch-line coupler and transmission lines, varying all the possible parameters (centre conductor and gap widths and lengths) of each of the components in the design would be prohibitively time-consuming². A few simplifications were employed to reduce the computational intensity.

Simplifications Based on Physical Considerations

Some simplifications are easy to implement and reduced the problem to a much more reasonable one.

- The distance between the transmission lines was assumed to be large enough to neglect coupling.
- The lines were assumed to be wide enough to neglect the effect of metal thickness on the characteristic impedance of the line. Even though some stubs were in

²Considering the number of parameters that is possible to vary and the fact that a single simulation requires approximately 12 hours to complete.

fact quite narrow, the precision on their characteristic impedances was not of great importance.

- Good models of the short and the open circuits were available and reduced simulation time significantly.

Based on these assumptions, transmission lines, short circuits and open circuits could reasonably be simulated using ADS. Simulations were significantly faster since they are based on the solution of an equation set. The main problem resided in the absence of models for the tee junction and its air bridges. A possible solution is provided by a feature of ADS: it is possible to simulate the tee junction in HP Momentum and to use the results as a 3-port S-parameter block in ADS. The simulation time was reduced because resources were not used to solve for the fields in the well-known elements such as transmission lines or short circuits. Precision was maintained by performing the full-wave analysis on elements with no established solution.

With the transmission line and short circuit mathematical models, and with the 3-port S-parameter block, a complete phase shifter was assembled. The parameters obtained in the theoretical design (Section 3.5) were used for an optimization in ADS. The characteristic impedances of the thru, coupled and shunt³ arms were converted into centre conductor and gap widths while the electrical lengths of the same arms were translated into physical lengths (originally, quarter wavelengths). An optimization process was then run on this simplified design. The optimization variables were the centre conductor and gap widths of every transmission line and the length of each arm. This leaves the two-branch branch-line coupler with 12 variables and the

³The thru arm connects the input port to the thru port. The coupled arm connects the isolated port to the coupled port. The shunt arms connect the input port to the isolated port, and the through port to the coupled port.

three-branch version with 21 variables. These steps represented a major reduction in computation effort compared with the optimization in Momentum; but further improvements were achieved by using some simple logic.

Simplification Based on Symmetry

A directional coupler analysis (e.g., [25, p. 354-359]) is based on the assumption that the lengths of each arm are the same and equal to a quarter wavelength. Based on the physical symmetry of the directional coupler, it is reasonable to assume similar discontinuity effects for both shunt arms, and likewise for the coupled and thru arms. Also, due to realization considerations, it was convenient to keep the length of opposite arms consistent.

Based on the analysis mentioned in the previous paragraph, the impedances of opposite arms (the two shunt arms and the thru and coupled arms) are the same. Considering the symmetrical nature of the directional coupler, the discontinuity effects were considered to be similar on each pair of opposite arms.

Considering the above assumptions, the numbers of variables for the two-branch and three-branch branch line couplers were reduced from twelve and twenty one to six and eight, respectively: two and three centre conductor widths, two and three gap widths, and two arm lengths, still respectively.

This approach is good but has one inherent problem: the S-parameter block of the tee-junction (based on the EM simulation of a particular structure) is valid for only one set of CPW gap and centre conductor widths. The ADS optimization modifies these widths, but these modifications are not reflected in the S-parameter block of the tee-junction.

This difficulty was addressed by running a few iterations of a corrective procedure. First, the design with ideal transmission lines was used, the tee-junctions EM-simulated and the S-parameters used to optimize the original design. After this first optimization, the new widths were used to adjust the tee-junction drawing before re-simulation. These new S-parameters were used for a second iteration of the ADS optimization. This procedure converged rapidly and only a few iterations were required before the adjustments of the widths became insignificant.

Once this procedure was completed, an EM-simulation of the whole structure validated the soundness of the procedure.

3.6.2 Transmission Phase Shifter

For the transmission phase shifter, there were fewer variables to consider. The first consideration concerned the impedance of the stubs, which was not critical. This can be easily explained as follows. Referring to Equations 3.7 and 3.8, variation of the characteristic impedance impacts on the length of the stub for a given (purely reactive) input impedance. Because most series stub have a finite ground plane which affects the characteristic impedance of the line, it is convenient to choose a stub length based on a set characteristic impedance. While doing this, shorter stubs were preferred due to their broader bandwidth.

The second consideration relates to the end effect of the stub. The discontinuities of an open circuit and a short circuit translate into different electrical lengths than those suggested by the physical lengths. Thus, an appropriate modification is required, as was previously discussed.

Simulation Methodology

By applying the techniques described above, the simulations of the transmission phase shifter and the modifications due to the discontinuities were greatly simplified. The first and simplest procedure consisted of simulating each reactive element independently and adjusting it to match the theoretical design characteristics.

Few additional steps were required to assemble the elements together to form the complete filters. Very little coupling occurs between the different stubs since the ground plane isolates them from each other. The coupling that occurs between the stub and the main transmission line has already been taken into account in the simulation and tuning of the separate items. By minimizing the steps in width while assembling the filter elements, few new discontinuities were introduced.

Duplexing

In the case of a one-bit phase shifter, two different filters are used and switching between them allows the desired phase shift to be applied to the signal. The concept is simple but in practice, some additional components are required to implement it. It is possible to have the two filters adjacent on the wafer and have Y connections between their inputs and between their outputs as illustrated in Figure 3.18 a. Simultaneously switching the inputs and outputs allows the signal to travel through the first or the second filter and the phase to be shifted accordingly.

Micromachined switches were used at the connecting point of the Y. They close the path in one of the branch and allow the signal to enter (or leave) the appropriate filter. Figure 3.18 b shows two ways of using micromachined switches for this purpose.

(a) Duplexer.

(b) Duplexer with micromachined switches

Figure 3.18: Duplexer used to shift the phase of the signal.

Chapter 4

Manufacturing the Phase Shifters

This chapter covers the fundamental principles and practical details of the design fabrication process. The first section introduces the processing steps used and discusses the solutions to the problems encountered during fabrication. The second section itemizes each component included in the final design, for both the reflection and the transmission phase shifters. It also outlines the purpose of these components.

4.1 Introducing the Fabrication Process

The fabrication procedure used for this project was derived from the monolithic integrated circuit (MIC) process used at the Communications Research Centre¹ (CRC) fabrication lab. This section presents the fabrication steps, discussing the layers in order of deposition on the silicon wafer. It also describes the problems encountered and the preferred solutions.

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4.1.1 Layering the Material

The current section discusses, step by step, the procedure involved in the fabrication of the circuit, and comments on the different layers. The letters in square brackets used in this section refer to Figure 4.1.

Before introducing the layers, a brief discussion of the masks is necessary. A positive mask is a glass plate on which deposited chromium shows the desired structure. For example, a coplanar waveguide (CPW) positive mask would show the centre conductor and the ground planes. A negative mask shows everything else. For the same example, a negative mask would show the gaps of the CPW line and the space around this line. Four masks are used for different layers: first layer metal, resistors, posts and air bridges.

High Resistivity Silicon As mentioned earlier (in Section 2.1.1), the fabrication process starts with a high resistivity silicon wafer. This type of material reduces the dielectric losses of the coplanar waveguide (CPW) transmission line. The three-inch diameter wafer is approximately 500 microns thick. Its dielectric constant is approximately 11.6 and the resistivity varies between 10 k Ω -cm and 100 k Ω -cm. The loss tangent used in simulations was 0.0014, but this information was not available for the particular wafer used.

Silicon Nitride The first deposited layer is silicon nitride (Si₃N₄). 4000 Å of this dielectric² was laid [a] to prevent metal diffusion (gold or titanium-tungsten) in the substrate during the following steps.

²1 Å = 0.1 nm = 1×10^{-10} m

Titanium-tungsten This seed layer is required for the gold to adhere properly to the wafer. The 300 Å of titanium-tungsten (TiW) was sputtered [b] onto the silicon nitride.

Gold This material was chosen to reduce the losses of and microwave passive structures. The patterning of the deposited gold layer [c] was done by etching [d, e, and f].

Resistor Resistors were created by patterning the underlying titanium-tungsten (TiW) layer [h, i, and j]. The thickness of this layer is constant across the wafer and has a specific sheet resistivity. The correct resistance values were obtained by forming rectangles with the appropriate length to width ratio.

The base sheet resistivity³ of TiW is approximately 40 $\Omega/\square$. Baking the wafer at 375 °C increased the resistivity to the desired level of 100 $\Omega/\square$. The proper 2:1 aspect ratio provided a resistor of 50 Ω .

Posts and Air Bridges Posts and air bridges are made of gold and they connect the first layer of metal (typically, on the ground planes) to the air bridges. Again, gold was used to reduce the conductor losses and maintain the lowest voltage difference between the two ground planes that are connected. The spacing between the air bridge and the centre conductor is two microns. The posts were first fabricated [k, l, and m] and the air bridges were grown on top [n and o]. The structure was then readied for measurement [p].

³Ohm/square is the unit of sheet resistivity. Resistivity of a block of material is defined as $R \propto \frac{L}{W \cdot H}$ (R=resistivity, L=length, W=width, and H=height). For a given thickness of material (i.e., for a given H), the resistance of a square is independent of the dimensions since L=W and, thus, their effect disappear in the relationship for the resistance.

Figure 4.1: Fabrication process steps.

4.1.2 Encountering Problems

This section presents the two major problems encountered during fabrication, as well as the solutions adopted.

Diffusion of the Titanium-tungsten in the Wafer

During the first run of fabrication, a serious problem occurred. At the baking stage, the absence of silicon nitride caused metal (TiW or Au) to diffuse into the substrate. This was problematic since a high resistivity silicon wafer was used to reduce the dielectric losses of the circuit. Diffused metal within the substrate significantly decreased the resistivity of the wafer and thus increased the substrate losses.

To solve this problem, a dielectric layer (silicon nitride), was deposited. This technique successfully stopped the diffusion of the metal into the substrate.

The silicon nitride should not influence the devices for two reasons: it is an insulator, and it is very thin. It makes up under one percent of the gap width of the coplanar waveguide, and thus has little effect on the electromagnetic field distribution of the structure. The fact that it is an insulator ensures that no unwanted electric short circuits occur.

Lateral Over-etching of the Gold

A note should be made about etching the gold layer. It was expected that as the gold was etched in the vertical direction, some lateral etching (undercutting) would occur underneath the photoresist layer. Lateral etching equal to the etching thickness is usual. Common practice at the CRC fabrication lab is to work with a transmission line thickness of one micron. Thus, the original design included compensation for

one micron undercut. However, some unexpected lateral over-etch occurred in the process. This might have been due to the different wafers used or to the presence of a silicon nitride layer. This lateral over-etch was a problem because precision on the structural dimensions are essential for the proper behaviour of the device. Such a decrease of the centre conductor width and increase of the gap width increases the characteristic impedance of every transmission line compared to that specified in the design. Another problem was the disappearance of fine structures. For example, the smallest stubs were three microns wide. An over-etch of one micron on each side added to the half micron tolerance on the process was likely to make the stub completely vanish.

To compensate, the thickness of the gold layer was reduced to a half a micron. The excess undercut at this thickness corresponded to the desired lateral etch of the original one micron thickness. A rapid Linecalc simulation showed this reduction of thickness increased the impedance by approximately 0.6%, which was well within tolerance.

4.2 Finalizing the Layout

This section presents the components included in the masks used for fabrication. The three-inch wafer accommodated eight reproductions of a reticle, but all eight are truncated in different ways due to the round shape of the wafer and because of the presence of the alignment marks. Figures 4.2 and 4.3 respectively represent views of one reticle and of the complete wafer. The numbers in parentheses appearing in this section refer to the corresponding element of the reticle view. The first subsection

describes the basic elements while the following two subsections present the elements that are specific to the reflection phase shifter and the transmission phase shifter.

4.2.1 Basic Elements

This subsection deals with the elements included in the final work that are not part of any phase shifter. They are transmission lines and calibration elements.

Variation of the Width The upper left corner (in Figure 4.2) consists of transmission lines. The objective of including these elements was to study the effect of varying the dimensions. Blocks 1 to 11 are transmission lines terminated with 50 ohms, open circuits and short circuits. The centre conductor and gap widths were varied to evaluate the effects of the etching tolerances on the final characteristic impedance of the line. Blocks 7 and 10 are identical except for air bridges, which are present in block 7 but not block 10. Comparing these results allowed the evaluation of the air bridge effects on the characteristics of the transmission line.

Variation of the Length Blocks 12 through 15 are transmission lines ending on short circuits (12-13) or open circuits (14-15). Their lengths were set to fractions ($1/4$, $1/2$, $5/8$, $3/4$, $7/8$, 1) of an arbitrary reference ($2762 \mu\text{m}$). They enabled comparison of the measured and simulated phase characteristics.

Calibration of the Measure Blocks 28 and 29 are Through, Reflect, Line (TRL) calibration sets. This type of calibration is commonly used at radio-frequency (RF) and microwave frequencies. They allow measurement of the parasitic effects of the measuring equipment: the network analyzer, cables, probes and connectors. This

(a) Complete reticle.

(b) Outline of the elements.

Figure 4.2: View of a reticle.

(a) Complete wafer.

(b) Outline of reticles.

Figure 4.3: View of the complete wafer.

measurement enables the determination of the parasitic effects of the launch pads and de-embedding by a fixed length of transmission line to calibrate out proximity effects of the probe on the measured circuit. Block 28 was for a straight transmission line; block 29, for a transmission line with a 90-degree bend and air bridges. This bend will be discussed in further detail in the following subsection.

4.2.2 Reflection Phase Shifter

This subsection presents the circuits and elements that were included in the final layout that are related to the reflection phase shifter. Some brief comment are made, followed by a description of these elements and circuits.

Comments on the Practical Realization

The first comment about the realization of the reflection and transmission phase shifters concerns the micromachined switches. They were originally intended to control switching between the states of the phase shifter. Unfortunately at the time of fabrication of the circuit, their design was too immature to include them in this project.

To validate the approach in the absence of the micromachined switches, an alternative route had to be taken. The same complete circuit is duplicated in different modes of operation. Closed switches are replaced with short circuits, and open switches with air bridges. The measurements of these altered circuits can be assumed to approximate the switch effects, allowing easy integration of the switches when they become available.

The second comment concerning the reflection phase shifter regards the input and output. A glance at the schematic shows that the input and the output transmission lines of the phase shifter are bent at a 90 degree angle. This is imposed by measurement constraints. Not bending the input and output would mean that both ports sit on the same side. Measurements would be nearly impossible since the dimensions of the probes requires ports to be 90 degrees or 180 degrees apart from each other.

These bends are not a problem for two reasons. Firstly, the relative phase difference is the primary issue for the phase shifter, not the absolute phase difference. Adding the bends to all the copies of the circuit introduces the same phase shift to all measurements and thus does not affect the relative difference between them. The second reason is the possibility of calibrating out the parasitic effects associated to the bends by using an appropriate TRL calibration set (block 29).

Elements Included in the Design

Considering the blocks in detail, two types of reflection phase shifters should be discussed: the two-branch (block 30-34) and the three-branch (block 16-26) branch line coupler phase shifters.

The three-branch branch-line coupler phase shifter. Block 16 is the “zero phase shift” reference; blocks 17, 18, 19, 23 and 24 respectively introduce phase shifts of 22.5° , 45° , 90° , 180° and 360° . The 360° version has been included for comparison with the reference and to study the losses introduced by the phase shifter.

Blocks 19 and 20 are similar (meaning that they should both introduce 90° phase shifts) but the input and output are at different angles. The bent transmission lines

of the input and output have been included in the tee junctions section of the three-branch branch line coupler.

Blocks 21 and 22 were meant to measure the isolation of the branch-line coupler but a design mistake renders them useless. A resistor only at the coupled port was included in the design while one was also required at the thru port to properly measure the isolated port.

Block 25 is a simple branch line coupler with the input, through and isolated port connected to measuring pads, and the coupled port terminated with a 50 ohm load. This block was used for the study of the branch line coupler with three probes. Two of the probes were measuring probes while the third one is a matched load.

Block 26 is the same as block 24 except for the presence of air bridges to simulate the parasitic effects introduced by the switches of a 4-bit phase shifter.

Block 27 shows a pair of branch-line couplers connected back to back. It was included to study the separation of an input signal into two parts (the ones going to the thru and coupled ports) and the recombination of these parts into one output signal. Because of the mirror effect of the couplers, the input (which would have acted as output port of the two-coupler combination) and isolated ports were inverted on the second coupler, rendering the combination virtually useless.

The two-branch branch-line coupler phase shifter. Blocks 30 and 31 are two-branch equivalents of block 25. They allowed study of the behaviour of the branch line coupler, and have three ports (input, output and through) available for measurements and the fourth port (coupled) connected to a 50 ohm load.

Block 32 is the phase shifting reference. Blocks 33 and 34 are designed to introduce 45° and 90° phase shifts, respectively.

The resistor

The resistors integrated into the design absorb the energy at one or two ports of the three- or four-port devices. Two-port measurements are more practical than three- or four-port ones. It is in fact possible to obtain the full S-parameters from a set of two-port measurements with the additional port(s) connected to matched loads ($50\ \Omega$ resistors in this case). For example, $50\ \Omega$ resistors at the isolated and thru ports of a directional coupler allow measurement of the coupled port energy with respect to the input port energy without having to account for reflections from the isolated and thru ports.

4.2.3 Transmission phase shifter

The realization of the transmission phase shifter was fairly simple after the design stage. Its intrinsic nature of having its input and output on opposite sides simplified the measurement: the problem of probe positioning encountered with the reflection phase shifter was not an issue here.

As mentioned in the previous chapter, it is necessary to have wye connections to duplex the signal to one of the two filters. Switching between the filters selects the phase shift. The simple wye connection has been included in the design (blocks 38 through 49). Blocks 41, 34 and 46 allowed the measurement of losses incurred by the device. Blocks 38, 42, 45, and 48 permitted to measure the isolation in the unused arm.

Also, to allow for the possibility of studying the behaviour of each of these components separately, series stub inductors and capacitors were laid down (blocks 35 and 36). The filter, along with a length of transmission line, was included to allow

study of the amplitude and phase characteristics of the filters without the influence of the duplexer (in block 37). Blocks 50 and 51 are the same filters but with one of the duplexers⁴. The final designs (again excluding the switches) with the two duplexers were also constructed to permit measurement of the effects of the proximity of the second arm (blocks 52 and 53). Block 54 allowed measurement of the amount of energy coupled from one filter to the other.

⁴Two duplexers are included: one each at the input and the output.

Chapter 5

Validating the Experiment

Experimental validation is an important step of the research process. It confronts the theory and the simulation results with actual measurements and allows correction of the mistakes made in the earlier stages. The first part of this chapter consists of a few scanning electron microscopy (SEM) photographs followed by an introduction to the measurement methodology and the equipment used. The second part of the chapter presents the measurement results, for both the reflection and transmission phase shifters. Finally, several potential sources of discrepancies between the simulations and the measurements are explained.

5.1 Viewing the Devices

The design and fabrication of the circuits was mostly “unseen” due to the minute dimensions involved. The use of a microscope gives a good idea of the final results but is limited as far as depth perception is involved. The SEM allows greater magnification of parts of the circuit and permits side or angled views, providing a better

Figure 5.1: TRL Calibration Elements.

representation of the depths and layering of the circuits. The figures in this section present close-ups of the circuits, the first few views are from the optical microscope and the rest, from the SEM.

5.1.1 Viewing Through the Optical Microscope

This subsection presents views of some of the circuits and circuit elements that were fabricated.

TRL Calibration Elements

In Figure 5.1 can is seen the thru, reflect and line (TRL) elements that were used for calibration of the network analyzer.

CPW Transmission Line

Figure 5.2 illustrates simple coplanar waveguide (CPW) transmission lines. These three lines of different lengths are terminated by open circuits.

Figure 5.2: Photograph of a CPW Transmission Line.

Figure 5.3: Photograph of BLC-PS'.

Figure 5.4: Photograph of the Transmission phase shifter.

Branch-Line Coupler Phase Shifter

A 2-branch branch-line coupler phase shifter (BLC-PS) is represented in the photograph of Figure 5.3. This particular circuit was meant to provide 90° of phase shift at 28 GHz.

Transmission Phase Shifter

The last of the optical microscope images (Figure 5.6) shows a transmission phase shifter, complete with wye connections and feed lines. The upper arm contains the low-pass filter (LPF) while the lower arm has the high-pass filter (HPF).

Figure 5.5: Photograph of the High-Pass Filter.

Figure 5.6: Photograph of the Low-Pass Filter.

5.1.2 Viewing Through the Scanning Electron Microscope

This subsection presents some SEM views of some parts of circuits and also of some peculiarities of the wafer. It also briefly comments on a few issues that arose during fabrication.

Short Circuit

Figure 5.7 shows an angled view of half a short-circuited CPW transmission line. The centre conductor, ground plane, short circuit and gap can be seen in this image.

The darker protuberances that can be seen at the surface of the metal are of unknown nature. There were two hypotheses advanced as to their nature, but neither could be retained after further experimentation. The first one was that they are residue from a photoresist removal process step. This hypothesis is not considered anymore since recent exposure to an oxygen plasma should have ashed these residue and it did not. The second hypothesis supposed a metallurgical phenomenon that combines the upper gold layer to the lower titanium-tungsten layer into an alloy that “pierces” the surface and appears on top of the gold. That latter hypothesis also has been discarded since exposure to heat in the drying oven at 375°C for 30 mins (parameters for the resistors growth) did not increase noticeably the presence of the bumps.

Phantom Lines

Another concern about the fabrication that was raised concerned lines that are present in some locations on the wafer. Such a line can be seen crossing the gap and going along the metal of the ground plane, as can be seen in Figure 5.7. It simply is the

Figure 5.7: One gap of a short-circuited CPW line.

Figure 5.8: Over-etch from a former process step.

over-etch (see Figure 5.10) from a previous fabrication run on the same wafer. As can be seen in Figure 5.8, this defect is rather small and should not have significant effects on the circuits' performance.

Air Bridges

Figure 5.9 illustrates an air bridge suspended above the centre conductor of a CPW transmission line, connecting the two ground planes. The crystalline structure gold acquires when it grows can be seen in the thickness of the air bridge. The shape acquired by the air bridge as it followed the profile of the photoresist that was present at the time of fabrication can also be observed. Referring to Section 4.1.1, photoresist

Figure 5.9: Air bridge above the centre conductor of a CPW line.

was deposited on top of the first layer of metal, leveling the surface to a large extent, but some of the larger level variations remained visible. This translated into an arched air bridge over the centre conductor.

Metal-Substrate Interface

Figure 5.10 offers a view of the cross section of the metal-substrate interface. In this view, the dielectric layer, Si_3N_4 , is clearly seen to have been washed away from the substrate in the cleaning process. That same layer is also undercut, being removed from under the metal while being removed from the top of the substrate. It can be noted that the undercut is approximately the same length as the thickness of this layer (approximately $0.5\ \mu\text{m}$). The over-etch that can be observed in the silicon near the dielectric-substrate interface is due to stagnant chemicals that accumulated in the edge formed by the dielectric and the substrate, allowing more etching than desired.

These figures illustrate some of the elements that are part of the circuit designs for this project. The photographs reduce some concerns such as the over-etch that occurred in some circuits (Figure 5.8). However, they also raised some new questions,

Figure 5.10: Cross sectional view of gold on silicon.

such as the “bumps” that appears on the surface of the metal (Figure 5.7). These issues are discussed in further detail in Section 5.5.

5.2 Introducing the Measurement Methodology

To facilitate experimental reproducibility, it is essential to include details of the equipment and the techniques that were employed to obtain the measurement results. These data are provided in the following paragraphs.

Equipment

The Pico-probe used had a $150\ \mu\text{m}$ probe pitch. They were mounted on the probe station, and connected to the network analyzer via semi-rigid coaxial cables. The data were collected by a Wiltron 306B network analyzer and transferred to a micro-computer. Custom-made software initiated the transfer and stored the data for later analysis. The data were ultimately imported into ADS using a dataset format for comparison with the simulation results. Figure 5.11 illustrates the configuration of the equipment.

Figure 5.11: Equipment configuration.

Measurement Technique

After the proper calibration (which will be discussed further in Sections 5.3 and 5.4) has been performed, the wafer was brought under the probes for contact. The probes were lowered until they just made contact with the desired circuit. A few microns of over-travel¹ insured a sufficiently conductive contact between the probes and the circuit. The microcomputer was then set to initiate measurements by the network analyzer and transfer the data to a custom file format. The wafer was then ready to be moved for measurement of the next circuit. After the measurements were completed, the data was transferred to a more standard format for analysis.

5.3 Measuring the Reflection Phase Shifter

Measured by the procedure just described, this section introduces the results for the reflection phase shifter and some of its building blocks. After brief comments on the calibration, results for the BLC are followed by the results for the 2-branch and the 3-branch versions of the BLC-PS.

¹There is contact as soon as the tips touch the circuit. To insure a good contact, however, the probe tips need to flex and press down on the circuit, and thus insure a low-resistance contact.

Figure 5.12: Representation of a CPW two-branch BLC.

For each element, the measurement results are presented alongside with the simulation results for comparison purposes. A later section (Section 5.5) explains and comments on these results.

5.3.1 Comments on Calibration

The calibration employed for the reflection phase shifter is TRL. This calibration compensates for the parasitic effects of measuring equipment, cabling, probes, probe-to-pad transitions and feed lines (c.f. Section 4.2.1).

5.3.2 Two-Branch Branch-Line Coupler

The first element measured was the BLC, one of the main components of the reflection phase shifter considered in this project. It was important to verify that it behaved properly before considering the complete phase shifter. Figure 5.12 presents the final layout of the BLC. The coupled port was terminated with a $50\ \Omega$ resistor. Figures 5.13 and 5.14 respectively present the thru- and isolated-port measurements.

Figure 5.13: Thru measurements of the two-branch BLC.

Figure 5.14: Isolation port of the two-branch BLC.

Figure 5.15: Representation of the 2-branch BLC-PS.

2-branch BLC-PS	Phase (degree)		Phase Shift† (degree)	
	Simulated	Measured	Simulated	Measured
Zeroth state (phase reference)	36.543	73.050	0	0
First state	-1.358	14.932	-37.901	-59.653
Second state	-47.572	-92.533	-84.115	-165.580

†Phase shift = Phase - (Phase reference)

Table 5.1: Phase of 2-branch BLC-PS at 28 GHz.

5.3.3 Two-Branch Branch-Line Coupler Phase Shifter

The 2-branch BLC-PS was then measured. The results for the phase reference and two phase increments (45° and 90°) are presented below. Tables 5.1 and 5.2 highlight respectively the phase and amplitude of these phase shifters at 28 GHz, and Figure 5.15 illustrates the 2-branch BLC-PS.

2-branch BLC-PS	Amplitude (dB)	
	Simulated	Measured
Zeroth state (phase reference)	-0.041	-3.504
First state	-0.216	-3.086
Second state	-0.386	-2.592

Table 5.2: Insertion loss of 2-branch BLC-PS at 28 GHz.

Before introducing the results for the BLC-PS, it is necessary to provide some additional information. The form used to refer to the different states of the phase shifters originates from the digital control used to operate them. Each of the switches having two states (open and closed), it is natural to represent their status using binary notation. A one-bit phase shifter would have one switch that could be either closed or open. To refer to the state of the three-bit phase shifter presented in this section, binary notation is used where the n^{th} state is that in which the phase shift is $n \cdot 45^\circ$. In this instance, eight different states are possible (i.e., $n \in [0, 1, \dots, 7]$).

As a short example, the zeroth state of a 3-bit phase shifter would be represented in binary, by “000”, which corresponds to a 0° phase shift. The fifth state would be represented by “101” and corresponds to a $(5 \cdot 45^\circ) = 225^\circ$ phase shift.

Zeroth State (Phase Reference)

The first measurement of the two-branch BLC-PS performance was the phase reference. A length of transmission line was added after the coupler to reduce the undesired coupling effects among the switches, coupler, and transmission line. The important parameter of the BLC-PS is the difference of phase between its different states. The measurement results of the phase reference are presented in Figure 5.16.

First State

The first step of phase shift (with respect to the phase reference) brings the phase shifter to its second state, 45° phase shift. The measurement results for this state are compared to the simulation results and presented in Figure 5.17.

Figure 5.16: 2-branch BLC-PS (zeroth state — phase reference).

Figure 5.17: 2-branch BLC-PS (first state).

Second State

The third and final phase shifter state included in the layout was for a shift of 90° . The measurement and simulation results are displayed in Figure 5.18.

5.3.4 Three-branch Branch-line Coupler

This subsection presents the measurement and simulation results for the three-branch BLC, similarly to Subsection 5.4.3.

5.3.5 Three-Branch Branch-Line Coupler Phase Shifter

The three-branch BLC-PS was likewise designed; this subsection presents the measurement and simulation results for the phase reference and four phase increments (22.5° , 45° , 90° , 180°). A 360° version of the phase shifter was also included for verification. Figure 5.22 represents a typical three-branch BLC-PS. Tables 5.3 and 5.4 summarize the characteristics of the phase-shifter states for phase and amplitude, respectively.

It should be noted that in Table 5.3, the phase shift (last column) shows the phase difference with respect to the first state. This is due to a mistake made in the final layout of the phase-reference state. The bend at the output port is shorter compared to all the other states of the phase shifter. This would introduce apparent error on all of the phase shifts, even though only the phase-reference circuit is incorrect.

The comment made regarding the states of the two-branch BLC-PS also apply for the three-branch version. The only difference, for this particular project, is that the three-branch BLC-PS is a 4-bit phase shifter (meaning there is 16 possible states).

Figure 5.18: 2-branch BLC-PS (second state).

Figure 5.19: Representation of a CPW three-branch BLC.

3-branch BLC-PS	Phase (degree)		Phase Shift† (degree)	
	Simulated	Measured	Simulated (Expected)	Measured
Zeroth state (phase reference)	-13.248	-159.49	-92.47 (-22.5)	-2.85
First state	-36.740	108.040	0 (0)	0
Second state	-56.728	107.540	0.5 (22.5)	45.64
Fourth state	-103.398	69.920	38.12 (67.5)	83.26
Eighth state	156.709	51.729	56.311 (157.5)	120.283
Sixteenth state	-14.625	-116.74	224.78 (337.5)	238.16

†Phase shift = $\text{Phase}_{\text{element}} - \text{Phase}_{\text{phase reference}+22.5}$

Table 5.3: Phase of 3-branch BLC-PS at 28 GHz.

3-branch BLC-PS	Amplitude (dB)	
	Simulated	Measured
Zeroth state (phase reference)	-0.069	-4.524
First state	-0.063	-7.131
Second state	-0.077	-7.331
Fourth state	-0.077	-4.959
Eighth state	-0.101	-4.451
Sixteenth state	-0.085	-5.099

Table 5.4: Insertion loss of 3-branch BLC-PS at 28 GHz.

Figure 5.20: Through port of the three-branch BLC.

Figure 5.21: Isolation port of the three-branch BLC.

Figure 5.22: Representation of the 3-branch BLC-PS.

Zeroth State (Phase Reference)

The phase reference was measured and the results are presented alongside the simulation results in Figure 5.23.

As noted earlier, this “phase reference” circuit was actually not used as the phase reference because of a layout error on the output port of this circuit.

First State

The second state for the three-branch BLC-PS was intended to provide 22.5° of phase shift. This state was used as the phase reference for comparison with the other circuits, as described above. The measurement and simulation results are reported in Figure 5.24.

Second State

The next phase shifter step (with respect to the original phase reference) of the BLC-PS was 45° . The measurement and simulation results are displayed in Figure 5.25.

Figure 5.23: 3-branch BLC-PS (zeroth state — phase reference).

Figure 5.24: 3-branch BLC-PS (first state).

Figure 5.25: 3-branch BLC-PS (second state).

Fourth State

The following phase shifter state was 90° . The measurement and simulation results are shown in Figure 5.26.

Eighth State

The final phase shifter state that has been included in the design is 180° compared to the phase reference. Its measurement and simulation results are displayed in Figure 5.27.

“Sixteenth” State

As mentioned earlier, the 360° phase shifter state has been included for comparison with the phase reference. Both the phase reference and this circuit should have the same measured phase at 28 GHz. However, inspection of the structure would reveal that the latter is delayed by a full cycle compared to the phase reference. The measurement and simulation results are presented in Figure 5.28.

5.4 Measuring the Transmission Phase Shifter

Measurements were also performed on the transmission phase shifter and its basic components. This section presents the results of those of the wye interconnection, and of the open-ended and short-circuited transmission lines. It then shows the results for the simple series stub, and concludes with those for the LPF and HPF.

After brief comments on the calibration, the results of measurements are presented and compared with the simulations' results, as with the reflection phase shifter. For

Figure 5.26: 3-branch BLC-PS (fourth state).

Figure 5.27: 3-branch BLC-PS (eighth state).

Figure 5.28: 3-branch BLC-PS (sixteenth state).

each component, the figures present the amplitude and phase of the significant S-parameter as a function of frequency.

5.4.1 Comments on Calibration

Before introducing the measurement results themselves, it is important to present the peculiarities of the calibration associated with the measurement of the transmission phase shifter components. Even though the elements required to perform a TRL calibration were included on the wafer, they could not be used for measurements of the transmission phase shifter components. The TRL-calibrated reference plane would be within the series stubs of the HPF, which would corrupt the results. To obtain reliable results, an Open-Short-Load-Through (OSLT) calibration was used instead to compensate for the effects of the network analyzer, cables and probes. This type of calibration does not allow the exclusion of parasitic effects of the transition from probe to launch pad, the launch pads, and the feed lines on the measurements. These effects proved negligible and measurements confirmed the expected behaviour of the circuit elements.

The OSLT calibration standard used was model CS-5 by GGB Industries Inc. The parameters used for the calibration process were provided by the manufacturer and entered appropriately in the network analyzer parameter memory. They represent the parasitic effects associated with the physical elements for the calibration:

- Parasitic open-circuit capacitance (C_0): 6.5 fF,
- Parasitic short-circuit inductance (L_0): 8.8 pH,
- Delay compensated physical length: 340 μm .

Finally, the usual procedure at the Communication Research Centre² (CRC) consists of performing the calibration and then counterchecking it by measuring the calibration elements. The following results are expected:

- Open and Short: visual inspection of the Smith chart to show a very high impedance open and a slightly capacitive short.
- Line: amplitude of forward and reverse reflection under -50 dB and phase error of $\pm 0.1^\circ$.

This method was applied throughout the measurements to confirm the proper calibration of the probes.

5.4.2 Wye Transitions

To connect the incoming and outgoing transmission lines to the LPF and HPF, it was necessary to have a wye connection. To study the loss introduced by the wye connections, it was included in the final layout. This subsection presents the measurement and simulation results for the wye connection.

Figure 5.29 is a representation of a wye connection. A gap can be seen on the lower branch of the wye connection. This is to represent the open state of a switch that would normally be present at this point. Figure 5.30 and 5.31 presents the simulation and measurement results for the insertion loss and the isolation, respectively of the wye connection. It should be noted that the calibration used for the wye transitions is TRL. The absolute phase of this element is not critical since only the relative phase

²Carole Glaser, Communication Research Centre, 3701 Carling Avenue, P.O. Box 11490, Stn. H, Ottawa, ON, Canada, K2H 8S2.

Figure 5.29: Wye connection.

between the filters is important and it seemed more important to obtain accurate loss measurements.

5.4.3 Transmission Lines

These elements are the simplest CPW elements. The important parameters are the amplitude and the phase of the forward reflection (S_{11}). The amplitude measurements show the combined losses due to the non-ideal terminations, probe-pad transitions, and the travel of the waves back and forth on the transmission line. For these terminated transmission lines, the phase of the forward reflection parameter is determined by twice the electrical length of the transmission line, the effect of the termination discontinuity, and the parasitic transition effect at the input of the line.

It should be noted that for the amplitude of the forward reflection, the simulation results for the gap and for the metal are both presented. This is because the gap simulation does not include the metal losses and, thus, does not fully represent the losses encountered in the devices. In the case of the phase of the forward reflection, only the simulation results of the gap are presented. For this case, resonances detected

Figure 5.30: Insertion loss of the Wye connection.

Figure 5.31: Isolation of the Wye connection.

Physical Length	Electrical Path Length		Absolute Difference
	Simulated	Measured	
1662 μm	289.069°	289.057°	0.012°
2681 μm	459.878°	455.753°	4.125°

Table 5.5: Open-ended transmission line at 28 GHz.

by the simulator affect the phase while these resonances generally do not occur in the actual devices (refer to Section 5.5 for details).

As an additional comment before presenting the results, it should be noted that numerous transmission lines were laid out to characterize the effect of centre conductor and gaps width variations (refer to Subsection 4.2.1). These characterizations were not done because of the apparently inconsistent³ distribution of the error in the measurement of the simple transmission lines. The phase variation due to the modification of the widths was expected to be of the same order of magnitude as this “random” errors thus invalidating the process.

Open-Ended Transmission Line

The open-ended transmission line is represented graphically in Figure 5.32. The measurement results for two lengths of transmission lines are shown in Figure 5.33. The 1662 μm line offers good phase agreement between simulation and measurements. The other line, of 2681 μm , has a phase difference of four degrees between the measurement and the simulation results. These results are presented in Figure 5.34. Also, Table 5.5 highlights the measurements by presenting the electrical path lengths⁴, both simulated and measured, for the two transmission lines, at 28 GHz.

³Inconsistent in being unpredictable from one line to another, yet repeatable for the same line.

⁴Total delay introduced on a signal between the input and output on a particular transmission line, at a given frequency.

Figure 5.32: Schematic of an open-ended transmission line.

Physical Length	Electrical Path Length		Absolute Difference
	Simulated	Measured	
1406 μm	241.898°	241.875°	0.0023°

Table 5.6: Short-circuited transmission line at 28 GHz.

Short-Circuited Transmission Line

As for the open-circuited transmission line, a representation of the short-circuited element is given in Figure 5.35. The measurement and simulation results are presented in Figure 5.36 for the 1406 μm short-circuited transmission line. Table 5.6 highlights the results by showing the electrical path length at 28 GHz.

5.4.4 Series Stubs

Series stubs are different from a simple transmission line in the sense that the centre conductor of the transmission line contains a stub, as seen earlier in Chapter 3. This section presents the measurement results for the capacitive and inductive series stubs, both shown in Figure 5.37. The results for the series stubs are presented differently in the sense that the insertion loss is related to the forward transmission S-parameter, while the phase characteristics are related to the forward reflection S-parameter. Table 5.7 presents numerical results at 28 GHz for these stubs.

Figure 5.33: Measurement and simulation of an open-ended transmission line, best match.

Figure 5.34: Open-ended transmission line, worse measurement-simulation match.

Figure 5.35: Schematic of a short-circuited transmission line.

Type	Phase of S11		Absolute Difference
	Simulated	Measured	
Single capacitive (thin stub)	142.685°	151.030°	8.345°
Single capacitive (wide stub)	-164.830°	-172.509°	6.679°
Double capacitive	145.980°	160.870°	14.890°
Inductive	-68.154°	-64.314°	3.840°

Table 5.7: Series stubs at 28 GHz.

Single Capacitive

Two different capacitive stubs have been included in the design (refer to Section 3.3.3). The first one was 7 μm wide and 675 μm long; The measurements results for this device are presented in Figure 5.38. The second capacitive stub is 23 μm wide and 540 μm wide, and the measurement results are presented in Figure 5.39.

Double Capacitive

To distinguish between the effect of the stubs and parasitic effects, two capacitive stubs were put back-to-back. The measurement results are compared to the simulation results in Figure 5.40.

Figure 5.36: Results of short-circuited transmission line.

Figure 5.37: Schematic of series stubs.

Type	Phase of S21		Absolute Difference
	Simulated	Measured	
Low-pass filter	131.470°	123.310°	8.160°
High-pass filter	-44.756°	-37.271°	7.485°
Phase difference	176°	161°	15°

Table 5.8: Third-order passive filters, at 28 GHz.

Single Inductive

Finally, the inductive stub was also simulated and measured. The results are presented in Figure 5.41.

5.4.5 Third-Order Passive Filters

The third order passive filters are based on simple transmission lines and were constructed from series and shunt stubs. Two filters, one low-pass and one high-pass, were designed. Numerical data for 28 GHz are presented in Table 5.8.

Figure 5.38: Series capacitive stub, single and thin.

Figure 5.39: Series capacitive stub, single and wide.

Figure 5.40: Series capacitive stubs, double.

Figure 5.41: Series inductive stub.

Figure 5.42: Schematic of the low-pass filter.

Low-Pass Filter

The low-pass filter, as illustrated in Figure 5.42, was measured. The results of both simulation and measurements are presented in Figure 5.43.

High-Pass Filter

Similarly, the high-pass filter is illustrated in Figure 5.44 while its simulation and measurement results are presented in Figure 5.45.

5.5 Commenting on the Discrepancies

The results having been presented, this section contains comments on these results and attempts at explaining the discrepancies between the simulation and the measurement results. To account for these differences, several potential problems have been considered and explored. The dissimilarities that have been considered are sep-

Figure 5.43: Measurement and simulation of the low-pass filter.

Figure 5.44: Schematic of the high-pass filter.

arated into material-related and simulator-related ones, and they are presented in the first two subsections. The conclusion of the section summarizes the most plausible causes of errors and provides an evaluation of the validity of the measurements.

5.5.1 Material-Related Discrepancy Sources

The first group of discrepancies that is presented is related to the material. They address issues that cannot well be represented in the simulator because they are difficult to characterize.

Resonance

As can be seen in several simulation results, simulation of the metal often leads to resonances that are not observed on the circuits (e.g., Figures 5.33, 5.38 and 5.40). This difference in behaviour is explained by simplifications used by the simulator.

Figure 5.45: Measurement and simulation of the high-pass filter.

To better understand the presence of resonances in the simulation and their absence in the measurements, it is important to understand how the simulations can be set up. There are three port types that can be used in ADS: “coplanar” ports, “single” ports, or “internal” ports. These are explained in the following paragraphs.

The preferred approach is the use of “coplanar” ports. As their name suggests, they are specifically designed to be used with coplanar waveguide structures. With this type of port, the gap of the transmission lines are drawn and ADS simulates the magnetic fields within these gaps. It then uses the converse of the results to determine the electric fields and S-parameters. This method assumes perfect and infinitely thin magnetic conductors, thus not allowing for lossy transmission lines. This explain the very small loss that can be observed in the simulation results for the “Simulation (gap)” curves.

To allow proper comparison between measurements of the lossy lines and simulation results, the second port type would be used. It is possible to modify the simulation to accommodate lossy conductors by drawing the actual conductor of the transmission line and using the “single” port, in association with “ground reference” ports to define the two side conductor as ground. However, these ports require the presence of a lower ground plane which transforms the CPW into a coplanar waveguide with lower ground plane (CPWG). This lower ground plane causes resonances to occur with the upper ground planes of the CPW (parallel plate mode) and the centre conductor (microstrip mode), as can be seen on the aforementioned figures. The lower ground plane also causes a shift in phase since it changes the propagation characteristics of the transmission line.

To simulate the conductor while not having the lower ground plane, the third port type would be preferred. The use of “internal” ports — again in association

with “ground reference” ports — allows this. However, these ports do not perform the proper de-embedding to remove transition effects from port to conductor, and introduces a phase error (compared to the de-embedding that occurs in TRL calibration). Thus, this type of simulation will not lead to accurate phase results.

As can be seen, there is no one simulation method that provides accurate results for both the losses and the phase of CPW structures. This problem has been acknowledged by Agilent and should be resolved by the next ADS release [26]. As recommended by technical support staff [27], simulation of the phase with “coplanar” ports and simulation of the losses with “single” ports is the proper way of obtaining accurate results.

It should be noted that the resonances associated with the “single”-port approach are not present in the measurements. This is explained by the fact that during measurements, there was no such ground plane *per se* on the backside of the wafer. The chuck of the prober is metallic, but several factors contributed to reduce significantly its ground plane effect:

- roughness of the wafer underside (as opposed to the polished top surface),
- roughness of the chuck surface,
- grooves of the chuck, and
- air layer between the wafer and the chuck.

All these factors reduced the parasitic behaviour of the circuits from the “ideal” resonating behaviours that were simulated. Also, simulation of the gaps did not show these resonances, which confirmed that they are observed due to the presence of the lower ground plane. This lead to the conclusion that the resonant behaviour predicted

by the simulation of the metal should not be taken into account during comparison with the measurement results.

Metal Thickness and Uniformity

As seen above, the use of single and ground-reference ports should have given proper losses due to both dielectric conductivity and metal resistivity. However, the measured losses were higher than expected; some questions were thus raised regarding the conductor: skin depth, thickness, thickness uniformity, and contamination of the metal. They are discussed in the following paragraphs.

Part of the discrepancy in loss amplitudes might result from deviations from the desired fabrication procedure. The original intent was to deposit one micron of gold for the first layer metal (i.e., the centre conductor and ground planes). One micron is the thickness generally used at the CRC-TWS for applications in the 30-GHz range because it represents approximately twice the skin depth⁵ at this frequency. Because of unexpected over-etching (c.f. Subsection 4.1.2), the thickness has been reduced to half a micron, which corresponds approximately to the skin depth at 28 GHz, given by,

$$\begin{aligned}\delta &= \frac{1}{\omega \sqrt{\frac{\epsilon\mu}{2} \left(\sqrt{1 + \left(\frac{\sigma}{\epsilon\omega}\right)^2} - 1 \right)}} \\ &= 0.45 \mu m\end{aligned}\tag{5.1}$$

⁵Depth in a lossy conductor at which the amplitude of the electromagnetic (EM) wave is 1/e the amplitude at the surface of the conductor.

where δ is the skin depth, ω is the frequency expressed in rad/s, ϵ is the permittivity of the dielectric, μ is the permeability, and σ is the metal conductivity.

The process tolerances combined with the simulator limitations might partly explain the differences in losses that are present between simulation and measurement results since, at thickness close to the skin-depth, the losses vary rapidly with thickness (compared with loss variation at twice the skin depth, for example).

Finally, in simulations, the conductivity of pure gold ($4.54 \times 10^7 \Omega \cdot \text{m}$ [28]) was used. In reality, contamination and surface roughness certainly decreased the conductivity, which would have increased the losses.

Another issue is related to the metal thickness that has been reduced from the original one micron to half a micron. Simple ADS simulations have shown that for the typical 50Ω lines used in this project, this thickness variation would modify the electrical length of a line⁶ by $0.8^\circ/2540 \mu\text{m}$. This error is doubled in the case of a measurement of S_{11} , as mentioned earlier (Subsection 5.4.3). Also, the rate of variation⁷ of the losses is rather large for the thinner metal, at $0.015 \text{ dB}/\mu\text{m}/\text{mm}$ compared to $0.0015 \text{ dB}/\mu\text{m}/\text{mm}$ at a one micron thickness⁸.

Another concern which might have been problematic was the uniformity of the thickness. Because of the process tolerances and the nature of fabrication, there is variation in metal thickness across the wafer, and thus between the few occurrences of the same circuits. However, spatial repeatability of the measurements⁹ has been assessed and measurements proved to be quite reliable. Appendix B shows evidence

⁶A CPW line of $2540 \mu\text{m}$ is typical in this project. Most simple transmission lines are of that length.

⁷Slope of the variation of the loss (in dB) per micron of width modification, for a line of 1 mm, at the standard width of transmission line.

⁸Evaluation of the slope of losses with respect to metal thickness done with Linecalc.

⁹Repeatability of the measurements across the wafer, on the same circuit at different position on the wafer.

of this statement in the form of the measured standard deviations for several devices. Since the measurements of most devices seem to have presented very small variation in the results, it can be assumed that the metal thickness uniformity plays a secondary role in the discrepancies between simulation and measurements.

The most convincing explanation for the discrepancies is the presence of the bumps on top of the metal (refer to Section 5.1). Being of unknown nature, it is reasonable to think that they might increase surface roughness and affect the metal conductivity, which would increase the losses. If they prove to be an alloy, they would further decrease the conductivity compared to pure gold.

Substrate Resistivity and Uniformity

The resistivity and uniformity of the substrate were considered to estimate their effects on the losses.

Although the high resistivity silicon wafer used for this project has a catalogue resistivity of 10-100 k Ω /cm, some fabrication step might have influence this characteristic. To confirm that there was no deviation from this specification, the resistivity was measured and found to be greater than 10 k Ω /cm. This eliminated the possibility of an issue with this parameter.

Although very unlikely, a second question that was raised about the substrate was the uniformity of the wafer parameters: dielectric constant, resistivity and thickness. For the same reason as with metal uniformity (namely, spatial repeatability), this factor was discarded from the set of potential discrepancy sources.

Process Tolerances and Uniformity

Other important potential sources of error are process uniformity and tolerances. This includes everything that is related, for example, to sheet resistivity of the resistive material, height of the air bridges, quality of the conductor/post and the post/air bridge ohmic contacts, and element dimensions.

Depending on the processing step, two factors could influence the uniformity of the process across the wafer: distribution of photo-resist across the wafer and speed of the chemical reactions, which is difficult to control precisely over the whole wafer because of the wafer surface profile. These factors are included in the process tolerances. They introduce errors on the dimensions of the final passive circuits.

Again, these factors seem negligible since the margins between simulation and measurement results are constant for similar circuits at different positions on the wafer.

Dielectric Layer

There is a dielectric layer (silicon nitride — refer to Section 4.1.1) between the conductor and the substrate which might have an effect on the phase of the signal. To verify this, a simple transmission line has been simulated with the “worst case”: a $0.3\text{ }\mu\text{m}$ layer of air between the substrate and the conductor. The effect proved to be negligible, as the phase difference between cases with the presence and absence of this air layer is of 0.0002° !

Probe Issues

There are two main sources of errors that could be attributed to the probes: poor calibration, and parasitic effects introduced by the probes. The first one is eliminated by measuring the calibration elements before and after a set of measurements to insure that the calibration was properly done and that it wasn't altered in the process of measuring the devices¹⁰. The second one seems to be present when measuring different circuits because the error between simulation and measurements seems to be random (as opposed to constant or proportional). However, the errors are constant when measuring the same circuit at different location and when measuring the circuits at different points in time, which contradict the hypothesis that the error is random.

Presence of the Ground Plane

As discussed above, the chuck of the prober is metallic. It is reasonable to assume that it will shift the phase and introduce an error in the measurements. In simulation, there is a variation of the phase of approximately 4° introduced into the results for simple transmission lines when a ground plane is present. The factors presented in the earlier discussion on resonance might provide an explanation to the apparent randomness of the discrepancies between measurement and simulation results.

Extreme Dimensions

In order to keep the length to width ratio of the stubs reasonably large, the dimensions of the stubs are, in certain cases, rather small and are sometimes pushing the limits of the process tolerances and usual guidelines. Most designs at the CRC avoid

¹⁰Small changes like room-temperature variation and movement of the cables are significant to the precision of phase measurements.

dimensions of elements below $10\text{ }\mu\text{m}$. In this project, dimensions as small as $7\text{ }\mu\text{m}$ (series stubs gap) and $3\text{ }\mu\text{m}$ (shunt stubs gap) were used in many instances. These small dimensions had two important effects. The first one is that process tolerances creates large relative errors in the dimensions. The second one is that these dimensions were of the same order as the metal thickness ($0.5\text{ }\mu\text{m}$). Because of this, the fields along the edge of the metal became comparable to the fields above and below the metal, affecting the effective dielectric constant, and thus, the electrical properties of the stubs and the lines. This latter effect might have been aggravated by the fact that Momentum, being a 2.5D simulator, does not include the effect of metal thickness on field distribution and effective dielectric constant.

The combination of these two effects could have influenced both the losses and the phase. A change of dimensions for the stubs would mean a change of characteristic impedance and of length. These parameters influence the reactance of the stub (refer to Subsection 3.3.3), which modifies the cut-off frequency of filters (and thus, losses) as well as the phase characteristics of these filters. It also could have altered the characteristic impedance of some BLC's, modifying the power distribution among the output arms (thru, coupled and — theoretically — isolated ports), and thus the losses in the reflection phase shifters.

Air Bridges

In some cases, simulation and measurement results diverged significantly for air bridges. For the simple short-circuited transmission line with an air bridge at mid-distance between the input and the short circuit, the phase difference between the simulation and the measurement was 30° at 28 GHz. It seems that more bridge capacitance is measured than simulated, but even a twofold simulated increase of bridge

width decreased the phase difference by a mere 7° . In the case of the open-circuited transmission line with an air bridge, the phase difference was less than 0.5° at 28 GHz but was above 50° in the 40 GHz range. One possibility that has been considered is the presence of remnants of photoresist stuck between the centre conductor and the air bridge. If this was the case, the capacitance of the air bridge might have been significantly increased, depending on the dielectric constant of the photoresist. This possibility was lately proved to be accurate. Figure 5.46 show remains of photoresist at location of removed air bridges.

5.5.2 Simulator-Related Discrepancies Sources

The second category of potential sources of discrepancies are related to the simulator. Because of the complexity of the structure involved, it became necessary to make some simplifications for the software to be reasonably efficient. In most cases, these simplifications would not cause major problems, but in a phase-sensitive application such as a phase shifter, these assumptions might become important.

Presence of Other Nearby Circuits (single vs group simulation)

To represent as closely as possible the measurement conditions, it would have been necessary to simulate the whole wafer with all the circuits included in the layout with one simulation port on the desired circuit. The port position would have been changed for the next circuit and the simulation re-run. Obviously, this procedure would have been prohibitively time-consuming. Even simulating each circuit with its close neighbours would have increased simulation time and complexity to impractical level.

(a) Wye connection.

(b) Low-pass filter.

(c) Low-pass filter (detail)

Figure 5.46: Photoresist remains under air bridges.

For this reason, each circuit was simulated independently. However, during the measurements, all of the other circuits were present, relatively close to the measured circuit. This might influence the measured phase and cause differences from the simulated results. The spatial closeness was especially important in the case of the simple transmission lines. Even though the ground plane width rule of “W+2G” was respected¹¹, it still had an effect that could explain the few degrees of difference that was sometimes present between the simulation and measurement results.

Simulation of H-field vs E-field

To further reduce the simulation complexity and time, Momentum was used to simulate the gaps of a CPW structure since generally, the gap surface is smaller than the metal surface. Also, the “coplanar” port type allowed proper consideration of the field polarization within the gap, ensuring a proper coplanar waveguide mode.

The neglected conductor effects in the magnetic-field simulation, combined with the potential introduction of errors in the conversion from magnetic-field to electric-field results, might have contributed to the discrepancies between simulation and measurement results.

Comparison Between the Model and the Electromagnetic Simulations

Since the design process included both circuit model and electromagnetic (EM) simulations, it was appropriate to perform a comparison of the results obtained with both modules of ADS for the same structure. It seems that there were discrepancies between the ideal model and the EM simulation of a simple transmission line.

¹¹Rule of thumb stating that a CPW ground plane of width equal to the sum of the widths of the center conductor and both gaps can be assumed to be of infinite width with reasonable results.

Using a Different Simulation Package

In order to compare the measurements with another simulation package, simulations were performed using Sonnet. This software detected most of the resonances that Momentum detected (when the lower ground plane was used) and gave similar results in most cases. However, it has been considered unreliable since simple passive structures sometimes have had results showing “losses” (e.g., 2 dB around 38 GHz in a particular case) for no apparent reason.

5.5.3 Discarding and Retaining Discrepancy Sources

The two previous subsections listed the potential sources of discrepancies between the measurement and the simulation results, some being discarded, others remaining in consideration. This subsection summarizes those which have been retained and indicates whether they are most likely to influence the losses or the phase of the circuits.

Loss-Related Sources

The error sources that had a greater influence on the insertion loss of the elements that were considered in this project are:

- metal thickness and uniformity (the “bumps”) and
- extreme dimensions.

Phase-Related Sources

The phase changes which most elements are introducing into the signal are quite sensitive to several factors. The sensitivity is explained by the high frequency at which these devices are used, where minute variations of the conditions — such as the temperature, position of the probes and process tolerances — could modify the phase noticeably. The factors that are most likely to have influenced the phase are:

- lower ground plane,
- extreme dimensions,
- air bridges,
- coupling with other circuits,
- simulator H-field to E-field conversion, and
- simulator non-linearities.

5.5.4 Circuit-Specific Comments

The comments that were made previously regarding the discrepancies apply to all the circuits that were measured and simulated. However, it is necessary to make some additional comments regarding specific circuit and some categories of circuits.

Branch-Line Couplers and Phase Shifters

A glance at the measurement and simulation results that were presented above (Section 5.3) clearly shows some major problem with most of the couplers and most of the phase shifters in regards to the losses and the phase. The possible causes for these

discrepancies are: improper characteristic impedances of the lines, improper length of the coupler arms, and improper behaviour of the air bridges.

Characteristic impedance of the lines. The main effect produced by a characteristic impedance different to the one chosen in the design would have is a wrong distribution of power between the output (i.e., thru and coupled) ports of the coupler. This would have affected the way the waves recombined and also affected the losses of the phase shifter. However, a reasonable variation of the characteristic impedance (within a few tens of ohms) would not have influenced the phase of the output signal at the centre frequency (in this case, 28 GHz).

Length of the coupler arms. This factor, of course, would have influenced the phase of the output signal. Considering that the electrical lengths of the arms are directly related to the physical lengths, however, this factor alone could not explain entirely the large difference in phase between the measured and the simulated phases. A few degrees can reasonably be expected but not as much difference as can be observed in several of the BLC and BLC-PS.

Measurement and simulation planes of reference. To ensure that there were no problems with reference planes in the measurements, these were repeated with the TRL bent-calibration elements, and the simulations were also repeated with the simulator ports at the reference planes of the TRL elements. The measurements and simulations were confirmed as identical to a previous set of measurements and calibrations. Major errors of this nature can thus be discarded from consideration.

For the same reason, the differences between simulation and measurement results was assessed (see Tables 5.1 and 5.3), and there is no constant difference for the

different states of operation. This fact eliminates the possibility of a constant-type error like misplacement of the plane of reference in either the simulations or the measurements.

Air bridges. In many measurements, there were unfortunate instances of short-circuited input or output ports, as can be seen in Figure 5.47. This leads to the possibility that since the air bridges of the BLC are fragile and easy to break, they could have short-circuited some transmission lines. This would certainly explain some of the major phase differences between simulations and measurements, as this would have greatly influenced the electrical length of some of the arms or branches of the coupler. With a number of bridges that varies between 16 and 32 per structure, it certainly would take a long time to determine which bridge or bridges is/are defective for each of the malfunctioning structures. However, it certainly is a quite reasonable assumption that this is the primary source of error, considering that this hypothesis has been verified for some of the bridge near the ports.

Another fact that makes this hypothesis even more plausible is the strong measurement asymmetry that exists between the input and output ports. The BLC and the BLC-PS being very symmetrical in nature (as can be seen both by looking at the layout as well as by inspecting the forward- and reverse-reflection S-parameters), strong asymmetries such as the ones that were observed in the measurements (see Figure 5.48) suggest that some air bridges make contact in locations they were not supposed to.

Figure 5.47: Example of short-circuited input and output port.

Figure 5.48: Dissymmetry of the forward and reverse reflection.

Chapter 6

Discussing the Phase Shifters

This last chapter recapitulates the research carried out in the course of the design and fabrication of coplanar waveguide based phase shifters. It briefly traces the steps that were involved, from the original idea to the final measurements. Finally, it offers to directions for future work, highlighting research paths that would be worth pursuing.

6.1 Summarizing the Work

This section restates the original intent of the project, outlines the design innovations, re-introduces the challenges encountered in fabrication, describes the project achievements, and recapitulates the weaknesses of the final layout.

6.1.1 Stating the Intent

The constant increase in public demand for large bandwidth applications and low-cost equipment requires research at high frequencies to be pursued. To provide reliable and reconfigurable broadband multimedia access to the Internet, the local multipoint

communication service (LMCS) approach is promising, having the required bandwidth and flexibility.

For LMCS resources to be easily re-assigned, the use of smart antennas seems to be a promising solution. The smart antenna — an array of discrete antenna elements whose collective radiation-beam directivity is variable — requires proper alteration of the signals which feed each of the elements. A phase shifter is commonly used to adjust the phase to the desired value.

This thesis investigated new techniques and pushed the limits of fabrication to design and manufacture a low-loss CPW-based phase shifter. Passive elements with proper phase characteristics provide the signal with a modification of phase while micromachined switches will permit selection among the multiple elements, to provide the desired phase modification.

6.1.2 Reviewing the Design

In this subsection, a brief overview of the choices that were made at an early stage of the design is followed by a description of the actual design procedure employed to create the final layouts of both types of phase shifters.

Preliminary Technological Choices

Before starting the design of the components for the phase shifters, some choices were made, constituting the basis of the design procedure. The choice of substrate as high resistivity silicon was dictated by a desire to remain close to a mainstream fabrication process, appreciating the low cost and widespread availability of silicon. The high resistivity variety, although more expensive, significantly reduced the dielectric losses.

Coplanar waveguide (CPW) was adopted in order to keep the fabrication to a minimal complexity level. The planar nature of CPW allowed to readily access the ground plane of the transmission line, whether to short-circuit a transmission line, design stubs or connect elements.

The fabrication performed at the Communications Research Centre used standard techniques such as spin coating, sputtering and electroplating for the deposition of material. The removal of material was performed through etching.

Reflection Phase Shifter

The reflection phase shifter is the first of the two approaches considered to realize a phase shifter. The choice of a branch-line coupler (BLC) — both the two-branch and three-branch versions were studied — for the coupler was based on several factors (e.g., the bandwidth, losses and compatibility with CPW transmission line).

The reflection phase shifter consisted of a BLC to split the signal into two parts. Short-circuited transmission lines then introduced the selected phase delay and reflected the signal toward the BLC, which then recombined the signal parts at the output of the phase shifter, having provided the proper phase shift.

The design of the reflection phase shifter consisted of proper choice of BLC parameters, correct dimensioning of the transmission lines providing the required phase shift, and appropriate positioning of the switches along the transmission line (represented in this project by the position of the short-circuit in the different states of the phase shifter).

Transmission Phase Shifter

The second approach, a transmission phase shifter, exploited the phase characteristics of passive third-order filters to modify the phase of the signal.

The design of the transmission phase shifter was based on a T configuration of low-pass and high-pass filters, because the T configuration was easily realizable with series and shunt stubs. Optimization of the lumped-element filters (an LPF and an HPF) provided the the desired $+90^\circ$ and -90° phase characteristics. A simple conversion of the lumped elements into series and shunt stubs then allowed the realization of the filters used for the transmission phase shifter.

Refining the Designs

Once the simplified designs were satisfactory, they were improved by adding parasitic effects of discontinuities and also by accounting for CPW transmission line effects. Most of these effects were included and simulated with the electromagnetic (EM) field solver, and the results were compared to the simplified design to ensure proper behaviour of the devices. After this performance assessment, the reflection and transmission phase shifters, along with several of their components, were ready for fabrication.

6.1.3 Revisiting the Fabrication

The fabrication of the phase shifters followed the usual fabrication rules that are followed by the lab at the Communications Research Centre¹ (CRC). The essential steps began with depositing a first layer of metal (Au) that formed the main elements

¹Contact: Rob James, Communications Research Centre, 3701 Carling Avenue, P.O. Box 11490, Stn. H, Ottawa, ON, K2H 8S2.

of the circuits. Air bridges were then constructed on top of that first layer of metal to connect the ground planes when required for higher-order mode suppression.

Two major issues arose from the fabrication: metal diffusion and lateral over-etching. In the first case, the baking step of the resistor had all the metal that was deposited (TiW and Au) diffuse into the substrate and “disappear”. To solve this problem, a thin layer of dielectric material (Si_3N_4) was deposited in the following fabrication runs. This layer prevented the metal from diffusing into the substrate.

The second problem was over-etching of the first layer metal. For some reasons that are not yet well-understood, the first layer metal underwent more over-etching than generally expected, based on similar projects. Lateral etching roughly equivalent to vertical etching was expected, but in this project, twice as much as lateral etching occurred. This, combined with fabrication tolerances, meant the disappearance of several fine features of the layout. To resolve this issue, the first layer metal thickness was reduced from the originally intended one micron to half a micron.

6.1.4 Reexamining the Measurements and Results

The measurements of the circuits allowed comparisons between the performances of the actual circuits and the simulations that had previously been done. This subsection summarizes the results for the reflection and transmission phase shifters and states the major problems that were encountered.

Reflection Phase Shifter

To perform measurements on the reflection phase shifter circuits, a thru-reflect-line (TRL) calibration was first performed to compensate for the parasitic effects of the

measuring equipment. Measurements were then performed on the different circuits representing various phase-shift states of the branch-line coupler phase shifter (BLC-PS).

A general comment that can be made about these circuits is that large discrepancies existed between measurement and simulation results. To confirm the absence of procedural error, both the simulations and the measurements were performed again, and results remained constant. Particular attention was given to the reference plane for both the simulations and the measurements, but the absence of consistent error between the simulation and measurement results led to the elimination of that possibility.

As discussed in Chapter 5, the most probable causes of discrepancies were the air bridges. They were responsible for the short-circuit of transmission lines in numerous instances. It also seemed that there was a variable but significant amount of photoresist that remained under the air bridges. This photoresist might have modified the capacitance of the air bridges, and thus spoiled the tee-junction discontinuity compensation that was performed in the design. In other words, the electrical lengths of most arms that constitute the coupler were likely altered. Also, some air bridges could have short-circuited some transmission lines within the coupler, drastically changing its behaviour.

The last potential source of discrepancies that was considered was improper simulation of the air bridges. Since their parasitic effects were compensated for in the design, if the actual air bridges did not behave as predicted by the simulator, the compensation that was performed would not be correct. This would have led to discrepancies between measurement and simulation results.

Transmission Phase Shifter

For the case of the transmission phase shifter, an open-short-load-thru (OSLT) calibration was performed. The OSLT (as opposed to the TRL) was necessary to prevent the calibration reference plane from overlapping the series stub, thus invalidating the results. Measurement of the circuits was then performed.

Wye Connection This small but important component was necessary to connect the two filters together. It will be used to integrate the MEMS switches into the design. At the frequency of interest (i.e., 28 GHz), its insertion loss was approximately 0.6 dB, and its phase characteristic matched the simulated one closely over the whole frequency range. The energy coupling between the two branches of the wye connection was below -40 dB for frequencies up to 35 GHz.

Terminated Transmission Lines Each of the approximately 230°-long transmission lines was terminated with a short circuit, an open circuit or a 50 Ω resistor. The measured losses in the transmission lines were approximately twice as large as the simulated ones. This difference was most likely due to the assumption of having perfectly flat pure gold conductor in the simulations, whereas this was obviously not the case for the actual circuits. The phase characteristics of these transmission lines were in good agreement with the simulations, the difference varying between 0.01° to 4°.

Series Stubs The series stubs were important components of the transmission phase shifter, along with the shunt stubs. To verify their proper behaviour, both the inductive and capacitive stubs were laid out and measured. As explained earlier,

a capacitive stub is usually more difficult to realize than an inductive stub. This fact was confirmed in the measurements results: the losses and phase of the inductive stubs were in better agreement with the simulations than those of the capacitive stubs. Phase differences of 3° and 8° exist for the inductive and capacitive stubs, respectively. It should, however, be noted that in both cases, these phase differences corresponded to measured electrical lengths greater than the simulated electrical lengths. It might therefore be relatively easy to compensate for these discrepancies.

Filters The third-order tee-configuration passive filters designed and manufactured for this project were the core elements of the transmission phase shifter. To demonstrate the potential of this design, a 1-bit phase shifter was designed with low-pass and high-pass filters. The measurements of these filters confirmed the simulations. Phase differences (between measurement and simulation results) of 8.1° and 7.5° existed for the low-pass and the high-pass filters, respectively. Again, the measured losses were slightly larger than simulated.

6.2 General Comments

This section presents some general comments on the measurement and simulation results, based on the explanation given in the previous chapter and above.

6.2.1 On the Measurements

There are a couple of important points that should be considered in regards to the measurements and the discrepancies:

- Repeatability in time, and

- Repeatability in space (across the wafer).

The repeatability in time means that the same circuits have been measured at different point in time (months apart) and the results were quite consistent. The repeatability in space means that the different copies of the same circuit, on the same wafer, were measured and, again, the results were consistent.

The repeatability of the measurements confirms that the results were as accurate as possible for these circuits and measuring tools. It eliminates the possibility of calibration corruption, large temperature variation and similar random errors in the course of measurement. On the other hand, it does not exclude the possibility of a constant mistake that could have been introduced, for example, by erroneous procedural design. If there were some problems with the TRL calibration or in the calibration procedure, there would have been a constant error introduced into the measurements. Such an error does not seem to appear in the measurements, however, since the disagreements with the simulation results were not constant.

6.2.2 On the Results

An additional comment should be made regarding the validity of the results. For the particular application toward which the phase shifters of this project were aimed, the results are considered satisfactory. According to the experience of the antenna group of the CRC², a phase error greater than 20° starts to seriously degrade the directivity of an array of antennas. With lesser phase errors, the directivity of the array of antennas is quite acceptable, although this does not hold for sidelobes. For

²Contact: Michel Cuachi, Communications Research Centre, 3701 Carling Avenue, P.O. Box 11490, Stn. H, Ottawa, ON, Canada, K2H 8S2.

applications in which sidelobes are critical (e.g., desired gain below -40 dB from maximum), the control of phase must be precise — within one or two degrees.

Even though the phase error of the circuits measured was significant, the end application and its tolerances must be considered. If the tolerance of the application can well accommodate the error that was obtained here, then, for that application, further refinement might not be necessary.

No optimization were done based on the measurement results; the comparison between simulation and measurement results that are presented in this report are based on the first design iteration. In a normal design setting, where good results are required, there would normally be additional iterations that would allow improvement of the design.

6.3 Directing Future Work

This final section intends to map out possible future work, presenting ideas of what could be done to strengthen the current work, as well as to lead toward application of the results.

6.3.1 Toward Improving the Current Design

Several ideas arose from the measurements procedure, that could be explored to simplify the measurements and improve the measurement results. These ideas are briefly presented in this subsection. They are meant to investigate hypotheses, improve the understanding of certain phenomena, and facilitate the work of following projects.

Characterizing the photoresist. The photoresist that is used in the fabrication process is meant to be washed away, and is not usually meant to be part of the final circuits. This material has thus not been characterized for its electrical properties. Better knowledge of its dielectric constant and loss tangent would facilitate fabrication. It would then be possible to use this knowledge to characterize the air bridges *avec* and *sans* photoresist and thus to readily determine whether the photoresist has been completely washed away.

Characterizing the coupling between the elements. Since the ground plane separating some of the elements is not very large, there is some chance that coupling occurred between, for example, transmission lines on the same wafer. Surface waves or parallel plate mode might couple power from one transmission line to another, leading to variation of the electrical characteristics.

Varying the ground plane width. It would be interesting to construct few transmission lines with all the same parameters except ground plane width. This would help characterize the effect of ground plane width on the electrical parameters of the transmission lines.

Varying the air bridges' width. The air bridges seem to significantly modify the electrical characteristics of the CPW transmission lines, and this behaviour seems not to be well predicted by the simulator. It would be worthwhile to explore some variations of the air bridge physical parameters in order to characterize this important component.

Additional CPW stubs. Only four series stubs were included in the layout of this project. It would be useful to include additional stubs (both series and shunt) to study their behaviour and distinguish between parasitic and desired effects.

Proper TRL calibration. It should be verified that the TRL calibration elements provide a reference plane that does not overlap the circuit for which it is intended (as happened with the transmission phase shifter).

Positioning of the circuits. To facilitate the measurements, a well-chosen placement of the circuits is important. In the case of this project, the wafer had to be rotated between TRL calibration and measurement of the BLC-PS. This was not very practical and measures should be taken in the future to avoid such situations. Also, simple alignment along one of the two planar wafer axes of the measuring pads of the different circuits would ease the measuring procedure by having to move the probes along only one axis to make measurements all across the wafer.

Simulating with other software packages. It would be useful (albeit rather time consuming) to learn different software packages and simulate the circuits in different environments. It would allow evaluation of different simulators for appropriateness to this application.

Characterizing the conductor. In this project, simplifications were made for the conductor. All the simulations that led to fabrication were done with lossless conductors, which probably led to some small phase errors. Also, the metal conductivity that was used for the post-fabrication simulation of the circuits was that of pure gold. The measurement results showed greater losses than those predicted by the simula-

tor, leading to the idea that the conductivity of pure gold might not be appropriate. Proper characterization at DC and at the frequency of interest might be appropriate in order to predict with better accuracy the losses in future circuits. It might be interesting to characterize all three different metal layers independently — first layer metal, posts and air bridges.

Using similar type of filters. To reduce the discrepancies between measurement and simulation results, it might be advantageous to consider using the same type of filters (either LPF or HPF) with different phase characteristics to obtain the desired phase shifts. By doing so, the phase drift for the various circuits would be in the same direction (instead of being in opposite directions as in this project). Also, it would be interesting to have constructed a larger variety of filters to get a better understanding of them and the effects of discontinuities on the phase.

T-junctions. To confirm that the compensation done in the design of the branch-line coupler was properly done, it would be important to include some T-junctions in order to measure them and compare with the simulation predictions. This would also provide insight on the behaviour of the air bridges and their effects on the characteristics of the T-junction.

Transmission line of different impedances. Finally, the inclusion of transmission lines with various characteristic impedances (to match those constituting the BLC) would be a good idea as well. It would allow to verification of the effects of the characteristic impedance and the fabrication tolerances on the phase of these lines.

6.3.2 Errors that were made

Some errors were made in the layout of this project. They are briefly reviewed here.

Proper load position of the back-to-back couplers. As mentioned earlier, the two-branch BLC's that are back-to-back have the resistor placed on the output port instead of on the isolated port. This rendered the device useless, and it would need to be corrected for proper measurements.

Feed-line length. It is important to ensure that the feed lines that are added to the final design before layout are consistent. In this project, the three-branch BLC-PS phase reference (zeroth state) had its output feed line shorter than all the other phase shifters. This prevented proper comparison of the other states with this phase reference.

6.3.3 Toward Using this Design

The project having been summarized and several ideas having been presented to improve the design and behaviour of the phase shifter, this subsection presents future work that could take advantage of the results presented in this report, exploring applications beyond the simple phase shifter that is described here.

A first step toward the realization of smart antennas would be to integrate MEMS switches into the design for control, and actually measure the phase shift in a signal. This will require a greater maturity of the MEMS development. It also requires some work to ensure that the post-fabrication step of adding the switches will not disrupt the proper behaviour of the surrounding circuits. It would also be interesting to

include the MEMS switches fabrication as additional steps of the same fabrication process instead of doing it after the fabrication of the basic devices.

The following step would be to integrate the phase shifters with an array of antennas to perform measurements of the performance of the smart antenna. Eventually, some processing hardware might be integrated with the phase shifter but that is for a “further future”.

As an additional potential application of the phase shifter, it might be possible, with developing understanding of miniature material structures, to consider using these phase shifters to perform phase modulation of the transmitted signal.

Appendix A

Surveying Simple Stub Structures

This appendix will present twelve structures of passive reactive elements. For each of the structures, the transmission line model are presented along with one, two or three possible coplanar waveguide realization of the structures.

Type	Transmission line model	CPW realization
Series capacitor
Shunt capacitor

Type	Transmission line model	CPW realization
Series inductor
Shunt inductor
Series LC
Shunt LC

Type	Transmission line model	CPW realization
Series resonator
Shunt resonator

Type	Transmission line model	CPW realization
2nd order LPF
2nd order HPF
3rd order LPF (tee)
3rd order HPF (tee)

Appendix B

Discussing Repeatability

This appendix aimed at briefly exposing the repeatability of the measurement, both in space and in time. All the standard deviation and mean difference that are presented in this table relates to the phase of the signal at 28 GHz.

The spatial repeatability is shown as a standard deviation of the measurement of the different copies of several circuits, at two different point in time (measurement 1 and measurement 2 were done one month apart).

The time repeatability is shown as the average of the differences in the measurement results for a given circuit, at two point in time. That is, for each measured copies of a given circuit at two different moments, the difference between the measured values are computed for each copies and all of these differences (for a single circuit) are averaged.

The LnnnX elements are simple terminated transmission lines. The phase is that of the forward reflection. The STnnX circuits are filters and series stubs. The phase is that of the forward transmission.

Element	Spatial repeatability		Time repeatability
	Stand. deviation		Mean difference between
	Measurement 1	Measurement 2	first and second measurements.
L203A	0.10	0.15	0.20
L203B	0.32	0.34	0.43
L203C	0.23	0.32	0.27
L303A	-	0.67	0.20‡
L303B	-	0.32	0.30‡
L303C	-	1.37	0.60‡
L304A	0.41	-	0.35
L304B	0.18	-	0.40
L304C	0.54	-	0.65
L305A	0.86	-	0.60
L305B	0.46†	-	1.45
L305C	0.51	-	0.30
ST08A	1.27†	1.22	1.1
ST08B	3.22	3.11	1.2
ST09A	1.55	1.57	1.6
ST09B	2.04	2.06	1.4
ST10A	0.10	0.23	0.67
ST10B	1.25	0.80	1.5

†Without the odds measurements

‡Only one measurement

Appendix C

Publishing a Paper

NOTE TO USERS

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