ALL-SILICON CMOS PULSE-COMPRESSION NONLINEAR TRANSMISSION LINES

by

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For the Degree of

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Abstract

This thesis studies the possibility of building efficient *all-silicon* pulse-compression nonlinear transmission lines (NLTLs). The theory and history of NLTLs are thoroughly reviewed, and the motivation for this project and possible applications are discussed. Silicon coplanar waveguide (CPW) lines and CMOS varactors, as two key components in NLTLs, are fully researched, and were fabricated using a standard 0.18- μm CMOS process. To reduce the dielectric loss caused by the conductive silicon substrate, a slow-wave transmission line technique was further developed, and silicon slow-wave CPW lines were built. A measured S_{21} loss of only 0.25 dB/mm at 40 GHz is achieved, with an effective relative permittivity as high as 25. Six different types of varactors based on NMOS transistors were investigated. They were divided into two groups: one has monotonic capacitance-voltage curves, and the other has non-monotonic curves. This study demonstrates that the first group is more suited for single-edge, the other for double-edge pulse-sharpening. The NMOS varactors used in the final NLTL designs were fabricated and on-chip measurements up to 55-GHz were made. NLTL transient simulations based on component measurements show a leading-edge rise time reduction of 75% for single-edge, and 60% for double-edge pulse sharpening. Following the investigation of the components, two types of NLTL circuits were designed and built on two CMOS 0.18- μm chips, one was designed for single-edge, the other for double-edge pulse compression. Large-signal measurements, based on the newly developed X -parameter method, show a significant compression with the second type of NLTL. However, because of its large attenuation, only minor compression is obtained with the

first type of NLTL. This research successfully demonstrates that it is possible to build pulse-compression NLTLs on low-cost commercial CMOS chips.

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Table of Contents

Chapter 1	Introduction.....	1
1.1	History of Nonlinear Transmission Lines	1
1.2	Project Motivation.....	4
1.3	All-Silicon Transmission Line Pulse Compression	6
1.4	CMOS Varactors and Silicon CPWs.....	8
1.5	Thesis Contributions and Outline.....	10
1.6	Summary	12
Chapter 2	Pulse Compression Technology Overview.....	14
2.1	Nonlinear Transmission Line Theory	14
2.1.1	Dispersion and Bragg Frequency.....	16
2.1.2	Nonlinearity	20
2.1.3	Dissipation	23
2.2	Physical Realization of NLTLs	26
2.3	Pulse-Compression Technology.....	30
2.4	Summary	33
Chapter 3	Silicon CPW Performance	34
3.1	Characterization of Silicon CPWs.....	34
3.2	Silicon CPW Losses	39
3.3	Slow-wave Technique for Substrate Loss Reduction	43
3.4	Silicon CPW Implementation.....	52
3.5	Conclusion.....	57
Chapter 4	CMOS Varactor Characterization.....	58
4.1	CMOS Varactor structures.....	58
4.2	C(V) Characteristics and Cutoff Frequency.....	60
4.3	NMOS Varactor Implementation.....	66
4.4	High-Frequency Characteristics of CMOS Varactors.....	69
4.5	Conclusion.....	78
Chapter 5	All-Silicon Pulse-Compression NLTL Design	80
5.1	Design Strategies.....	80
5.2	NLTL Chip Layout.....	84
5.3	Simulated NLTL Waveforms.....	86
5.4	Summary	88
Chapter 6	Device Measurements.....	89
6.1	Small-signal Frequency-domain Testing	89
6.2	Large-signal Time-domain Testing.....	95
6.3	NLTL Chip Designs and Measurement Results.....	99
6.4	Conclusion.....	114
Chapter 7	Summary and Future Directions	115
Appendix A:	Propagation Characteristics of CPW Lines.....	119
Appendix B:	Calibration Methods for CMOS small-signal measurements.....	122
Appendix C:	X-parameters	128
References		133

List of Figures

Figure 1-1 Distributed NLTL structures: (a) and (b) are MIS lines, (c) and (d) are Schottky-contact lines. (a) and (c) are microstrip lines, (b) and (d) are CPW lines...	2
Figure 1-2 Periodic NLTL structures: (a) CPW on periodically doped semiconductor, (b) Transmission line periodically loaded with varactor diodes.....	3
Figure 2-1 Circuit diagram of the NLTL configuration.....	15
Figure 2-2 Equivalent circuit of the NLTL configuration.....	15
Figure 2-3 (a) Unit section of an NLTL, (b) its equivalent circuit.....	16
Figure 2-4 Dispersion diagram for an NLTL unit cell.....	18
Figure 2-5 Characteristic impedance as a function of frequency for an NLTL cell.....	20
Figure 2-6 MOSFET RF Equivalent Circuit.....	21
Figure 2-7 Diagram of layout-produced parasitic inductances and capacitances for a NLTL cell.....	29
Figure 3-1 Silicon CPW structure and the field distribution.....	34
Figure 3-2 The cross-section view of a silicon CPW with pertinent dimensions.....	35
Figure 3-3 Quasi-TEM equivalent circuit of a silicon CPW: (a) Shibata's model, (b) modified model, (c) symbolic model.....	37
Figure 3-4 Attenuation of silicon CPWs ($W = 4 \mu\text{m}$).....	41
Figure 3-5 Attenuation of silicon CPWs ($W = 8 \mu\text{m}$).....	42
Figure 3-6 Attenuation of silicon CPWs ($W = 10 \mu\text{m}$).....	43
Figure 3-7 Slow-wave structure and design parameters.....	45
Figure 3-8 EM field distribution of (a) conventional CPW, (b) slow-wave CPW.....	46
Figure 3-9 Effect of overhang projection O and spatial duty cycle on line impedance for a slow-wave CPW.....	48
Figure 3-10 Effect of overhang projection O and spatial duty cycle on line attenuation for a slow-wave CPW.....	49
Figure 3-11 HFSS geometry for a slow-wave CPW simulation. M5 and M6 denote process metal layers.....	50
Figure 3-12 E -field distribution of a conventional CPW.....	51
Figure 3-13 E -field distribution of a slow-wave CPW.....	52
Figure 3-14 Technology stack for the CMOS 0.18- μm process.....	53
Figure 3-15 CMOS $2.0 \times 1.5 \text{ mm}^2$ chip with CPW, S-CPW and NMOS varactor components.....	55
Figure 3-16 Comparison between the measured (—) and simulated (--) $ S_{21} $ for a conventional and a slow-wave CPW.....	55
Figure 3-17 Effective relative-permittivity $\epsilon_{r,eff}$ for a conventional and a slow-wave CPW.....	56
Figure 4-1 Cross-sections of six varactor structures.....	59
Figure 4-2 CMOS intrinsic varactor model.....	61
Figure 4-3 $C(V)$ characteristics of the six CMOS varactor structures.....	63
Figure 4-4 $f_c(V)$ characteristics of the six CMOS varactor structures.....	66
Figure 4-5 Measured and simulated $C(V)$ curves of (a) an SDF varactor and (b) an SDB varactor.....	67

Figure 4-6 The CMOS varactor measurement and de-embedding setup.....	71
Figure 4-7 Simulated and measured data for the AMOS varactor.....	72
Figure 4-8 Simulated and measured data for the IMOS varactor.....	72
Figure 4-9 Simulated and measured data for the NMOS varactor.....	73
Figure 4-10 Simulated and measured data for the SnDp varactor.....	73
Figure 4-11 SnDp varactor versus NMOS varactor at 55 GHz (short-gate = 0.2 μ m, long-gate = 0.5 μ m).....	76
Figure 4-12 Series resistances at 5 GHz for AMOS, SDF and SDB varactors with gate area=0.5 $\times$ 12 μ m ²	78
Figure 5-1 Flowchart of the silicon NLTL design process.....	82
Figure 5-2 Top view of a CMOS NLTL layout (not to scale).....	85
Figure 5-3 Cross-sectional view of the CMOS NLTL shown in Figure 5-2 at plane a-a' (not to scale).....	86
Figure 5-4 Simulated slow-wave-NLTL waveforms using measured S-CPW and varactor data.....	87
Figure 6-1 Principle of operation of a VNA.....	90
Figure 6-2 On-wafer probing station connected to a VNA.....	91
Figure 6-3 (a) An on-wafer GSG probe. (The inset shows a microscopic view of the probe tips) (b) An artist's view of a probe (black) contacting a CPW structure (grey).	92
Figure 6-4 NLTL time-domain measurement setup using an oscilloscope.....	97
Figure 6-5 Simplified configuration of an NVNA and ADS workbench for nonlinear measurements.....	99
Figure 6-6 Early NLTL designs that omit slow-wave CPW and tapered-line techniques.	100
Figure 6-7 The latest NLTL designs, which use both slow-wave CPW and tapered-line technologies.	101
Figure 6-8 Measured SDB NLTL time-domain input (A1, yellow) and output (B2, blue) waveforms for a 13.0 dBm input at 5 GHz.....	102
Figure 6-9 Measured frequency-domain SDB NLTL spectrum for a 13.0 dBm input at 5 GHz.....	103
Figure 6-10 Measured SDB NLTL input (A1, yellow) and output (B2, blue) time-domain waveforms for a 13.5 dBm input at 5 GHz.....	104
Figure 6-11 Measured frequency-domain SDB NLTL spectrum for a 13.5 dBm input at 5 GHz.....	104
Figure 6-12 Measured SDB NLTL input (A1, yellow) and output (B2, blue) time-domain waveforms for a 13.9 dBm input at 5 GHz.....	105
Figure 6-13 Measured frequency-domain SDB NLTL spectrum for a 13.9 dBm input at 5 GHz.....	106
Figure 6-14 Measured SDF NLTL input (A1, yellow), reflected (B1, purple), and output (B2, blue) time-domain waveforms for a 13.0 dBm input at 3 GHz.....	107
Figure 6-15 Measured frequency-domain SDF NLTL spectrum for a 13.0 dBm input at 3 GHz.....	108
Figure 6-16 Measured SDF NLTL input (A1, yellow), reflected (B1, blue), and output (B2, purple) time-domain waveforms for a 14.0 dBm input at 4.5 GHz.....	109

Figure 6-17 Measured frequency-domain SDF NLTL spectrum for a 14.0 dBm input at 4.5 GHz.....	109
Figure 6-18 The SDB NLTL simulated result with 5 GHz sinusoid input, 2.2 V amplitude.....	110
Figure 6-19 The SDB NLTL simulated result with 5 GHz sinusoid input, 2.3 V amplitude.....	111
Figure 6-20 The SDB NLTL simulated result with 5 GHz sinusoid input, 1.9 V amplitude.....	111
Figure 6-21 The SDB NLTL simulated result with 5 GHz sinusoid input, 1.6 V amplitude.....	112
Figure 6-22 Reconstructed waveform by using only 5 lowest output harmonics.....	113
Figure 6-23 Spectrum of an input signal and 5 output harmonics.....	113

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- [2] Ming Li, Robert G. Harrison, Rony E. Amaya, Jean-Marc Duchamp, Philippe Ferrari, and N. Garry Tarr, "CMOS Varactors in NLTL Pulse-Compression Applications", *Proceedings of the 37th European Microwave Conference*, pp. 1405-1408, Munich, Germany, Oct., 2007.
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Chapter 1 Introduction

1.1 History of Nonlinear Transmission Lines

Nonlinear transmission lines (NLTLs) are typically implemented as waveguides which are continuously or periodically loaded with diodes. The waveguides can be microstrip lines or coplanar waveguides (CPWs). The diodes act as voltage-dependent capacitors (varactors). NLTLs loaded continuously with diodes are called distributed or homogeneous NLTLs, while NLTLs loaded periodically with diodes are called periodic or discrete NLTLs.

Research on NLTLs began as early as the 1960's. In 1967, Guckel *et al.* discussed basic metal-insulator-semiconductor (MIS) microstrip lines [1], where an oxide layer insulates the semiconductor wafer from the metallic conductors. Later on, Schottky-contact microstrip lines were developed. In these structures, the insulating region arises from the depletion layer of the Schottky barrier. The depth of the depletion layer depends on the applied voltage. The nonlinearity of the lines is produced by the voltage-dependent depletion-layer capacitance. Well-known researchers in this area include Jäger [2] and Hughes [3, 4]. In 1977, Hasegawa introduced MIS and Schottky coplanar striplines on GaAs substrate [5]. Subsequently, different theoretical approaches were established, including Aubourg's finite-element analysis of lossy

microstrip NLTLs [6], Fukuoka's analysis of slow-wave coplanar waveguide [7], and Sorrentino's study on characteristics of MIS coplanar waveguides [8]. These early developed NLTLs were all distributed NLTLs. Fig. 1-1 shows both MIS and Schottky-contact distributed NLTLs, the upper row being MIS lines and the lower row Schottky-contact lines. Also, in Fig.1, (a) and (c) are microstrip lines, (b) and (d) are coplanar waveguide (CPW) lines. An advantage of the distributed NLTLs is that they can reduce the dimension of integrated devices due to their slow-wave properties. A disadvantage is that their attenuation is high due to longitudinal current loss [9].

Figure 1-1 Distributed NLTL structures: (a) and (b) are MIS lines, (c) and (d) are Schottky-contact lines. (a) and (c) are microstrip lines, (b) and (d) are CPW lines.

The first periodic nonlinear transmission line was mentioned by Jäger in 1987 [10]. Bastida and Donzelli also introduced transmission lines loaded periodically with

lumped nonlinear elements a year later [11]. This is a basic model for nonlinear transmission lines. In 1983, Fukuoka and Itoh proposed a novel periodic coplanar waveguide in order to overcome the large attenuation of the homogeneous strip lines [12]. Some examples of periodic structures are shown in Fig. 1-2. In general, compared with homogeneous NLTLs, discrete structures have much less attenuation and larger characteristic impedances. Discrete NLTLs also have a low-pass filter property, due to the Bragg cutoff frequency generated by the periodic structure [9]. This will be discussed in more detail later.

Figure 1-2 Periodic NLTL structures: (a) CPW on periodically doped semiconductor, (b) Transmission line periodically loaded with varactor diodes.

Since the 1990's NLTLs, especially periodic ones, have become very popular, and have been used in many applications. These include pulse generation, pulse compression, and frequency multiplication. The general principle of periodic NLTLs is that the non-linear elements generate higher harmonics, and the periodic structure of the NLTL leads to a phase coupling of these harmonic signals. The pioneering researchers in this area are Jäger [13] and Rodwell [20].

1.2 Project Motivation

The main applications of NLTLs are (a) in the generation of sharp pulses, which can be used in high-speed sampling and on-chip interconnects, and (b) in frequency multipliers [14]. When NLTLs function in the frequency-multiplier mode, the spectral power is concentrated at one particular higher harmonic, but in the pulse-sharpener mode, the amplitudes and phases of the higher harmonics combine constructively to produce pulses with short rise- and/or fall-times.

Due to rapid advances in the bandwidth and speed of electronic systems, ultrafast electronic and optoelectronic devices are currently in high demand for measuring transient electrical signals with rise times in subpicosecond range. In these measurements, picosecond or subpicosecond pulses need to be generated for signal detection in high-speed sampling circuits and frequency down-conversion in wide-

band instruments. The NLTLs used in sampling circuits and time-domain reflectometry (TDR) act as pulse or stimulus generators. A few particular examples of those applications are generation of high-power microwave bursts in picosecond and subpicosecond measurements [15], generation of a shock wave in time-domain reflectometers [16], and generation of ultra-short electromagnetic pulses in THz spectrometers [17]. Picosecond or subpicosecond electrical pulses can also be produced by optoelectronic switches through optical-electrical conversion of optical pulses from pulsed lasers, but this application is limited by the size and complexity of the lasers.

The other application of NLTLs is in on-chip interconnects. With rapidly increasing operation frequencies and growing chip sizes, global interconnects have become the top challenge in CMOS technology. Clock signals have already been brought into the multigigahertz range, resulting in the inductance and other transmission line effects of long lines, such as dispersion and skin loss, becoming the new concerns. At higher operating frequencies, as input pulses travel down a transmission line they broaden due to dispersion, and amplitudes diminish due to skin loss and substrate attenuation. The already power hungry global clock network is set to consume more power and hence become even more difficult to design. Other global interconnects, such as data bus and control lines, face similar complications. NLTLs can be used as interconnects to perform pulse compression or pulse sharpening and to overcome pulse broadening and attenuation [21].

Homogeneous NLTLs using a sequence of identical sections can generate shock-wave or re-shape pulse signals, but require bandpass filters and phase shifters to adjust harmonic phases. Nonhomogeneous NLTLs using tapered sections can generate pulse signals or solitons from sinusoidal inputs without the need of bandpass filters and phase shifters [18]. In other words, nonhomogeneous NLTLs can transform sinusoidal signals into square waves. This capability can be used in a class-F power amplifier design, where an NLTL can be used at the input of an amplifier or as an interconnect between an amplifier and its load to produce square waves, which in turn creates non-overlapping output voltages and current waveforms, thereby increasing the efficiency and output power [19]. This project focuses on simple and compact nonhomogeneous NLTLs.

This program also focuses on building all-silicon NLTLs in a standard CMOS process, due to its low cost, high yield, short fabrication cycle, and its compatibility with other RF circuitry. More details on this subject will be given in the next section.

1.3 All-Silicon Transmission Line Pulse Compression

NLTLs have received considerable attention for application in high-speed sampling, pulse regeneration [20], integrated-circuit interconnects [21], and measurement circuitry [20]. In such applications, pulse compression and pulse

shaping are of critical importance. As input pulses travel down a transmission line they broaden due to dispersion. NLTLs can be used to overcome pulse broadening and attenuation caused by dispersion and frequency-dependent losses. Previous NLTLs have been mostly designed using conventional microstrip or CPW transmission lines (TLs) loaded either with GaAs Schottky diodes [20], [22] or with heterostructure-barrier varactors [22]. Recently, attention has turned to the possibility of designing all-silicon NLTLs in which both the TLs and the varactors are realized in silicon. The possibility of using CMOS varactors in NLTLs was suggested in [21]. A silicon NLTL of this kind, using a BiCMOS process on a high-resistivity silicon substrate, was investigated in [23]. Another approach is to exploit silicon Schottky diodes expressly designed for NLTL applications [24].

With constantly increasing frequencies in digital equipment, downscaling of semiconductor processes, and increasing demand for low-cost integrated RF components, it is necessary to develop CMOS-compatible NLTL designs that can be realized on lossy or standard low-resistivity silicon substrates. The resulting NLTLs will facilitate integration with other all-silicon digital or analog devices.

In all-silicon NLTL designs, the greatest challenge is the loss in the silicon substrate [24], [25], which contributes to losses in both the CPW and the CMOS varactors. The lower the resistivity of the silicon substrate, the higher the NLTL losses. The aim of this research is to investigate the possibility of realizing all-silicon

CMOS pulse-compression NLTLs, especially on standard low-resistivity silicon substrates.

1.4 CMOS Varactors and Silicon CPWs

In general, NLTLs consist of linear transmission lines periodically loaded with nonlinear devices, such as diodes or varactors. The nonlinearity is provided by the voltage-dependent capacitance of the nonlinear devices. In this project, silicon CPWs are used as distributed linear components and CMOS varactors as lumped-element nonlinear components.

During last 10 to 15 years, silicon CMOS technology has improved dramatically. There has been a continued increase in varactor cut-off frequencies. This enables CMOS varactors to play an important role in the implementation of RF components. In the NLTL designs described in this thesis, CMOS varactors are key components. In these varactors, the gate capacitance of a MOS field-effect transistor (MOSFET) is utilized to implement the voltage-dependent reactance. This gate capacitance varies in a nonlinear fashion as the changing MOSFET dc gate bias voltage sweeps the device through its accumulation, depletion, and inversion modes of operation. By making appropriate connections among the terminals of the MOSFET, the shape of the nonlinear capacitance-voltage (C - V) characteristic can be altered. The two resulting

alternative CMOS varactor topologies can be used for different pulse-compression purposes. This subject will be discussed in more detail in Chapter 4.

The realization of CPW transmission lines on lossy silicon substrates has been studied experimentally and theoretically in last few decades [26, 27, 28, 29, 30, 31]. In the earlier studies, the CPW lines were primarily fabricated on high-resistivity silicon substrates, with resistivities in the 20 Ω -cm to 2500 Ω -cm range. However, such low-loss high-resistivity wafers are more expensive than standard silicon substrates and require substantial modifications to standard CMOS fabrication processes. Recent work has begun to focus on low-resistivity silicon substrates [32], [33], [34], concentrating on how to reduce the silicon substrate losses. In [32], a metal ground plane was fabricated beneath the CPW line and above the substrate so as to block electromagnetic (EM) energy from entering the lossy substrate. In [33], a bow-shaped stacked metal ground was inserted below the CPW line in order to reduce the substrate loss by shaping the electric fields under the signal line. In [34], a micromachined polyimide interface layer was added between the CPW line and the substrate in order to increase the distance between the CPW layer and the substrate, thereby reducing the substrate loss. In such work, CPW insertion losses as low as 0.6 dB/mm have been reported [33]. This thesis presents a new slow-wave silicon CPW structure, using a standard CMOS process with a substrate resistivity of only ~ 10 Ω -cm. A 0.2 - 0.3dB/mm insertion loss at 40 GHz is achieved. A detailed discussion will be given in Chapter 3.

1.5 Thesis Contributions and Outline

This thesis investigates the possibility of realizing NLTL pulse compression circuits in low-resistivity silicon processes. The major challenges in all-silicon CMOS NLTL design are to overcome the high-frequency losses, to structure the CMOS varactors so as to create a strong enough nonlinearity, and to implement lower losses NLTLs in standard, low cost silicon processes. The contributions of this research are expected to be the following:

1. The theoretical and analytical development of more efficient and productive strategies and methods for designing pulse-compression NLTLs by exploiting newly-developed, market-proven software design tools.
2. The investigation and implementation of six different CMOS varactor structures, three of them providing monotonic C - V curves, and the other three providing non-monotonic C - V curves. One type of C - V characteristic is more suited for single-edge-compression, other is better for double-edge-compression NLTL designs.
3. Investigation of the feasibility of building low-loss CPWs on low-resistivity silicon substrates. This involves the development of a new silicon CPW topology, the so called “slow-wave” structure, within a standard CMOS silicon

process. This novel type of silicon CPW has reduced losses and can be used in pulse-compression NLTLs.

4. The development of a nonlinear theory that integrates the linear components (silicon CPWs) and the nonlinear elements (the CMOS varactors) together to form NLTLs that have minimum loss and optimized pulse-sharpening capabilities.
5. New experimental techniques for characterizing NLTLs in on-chip testing. This work includes executing calibration procedures with de-embedding techniques, and developing a new strategy for large-signal, high-frequency nonlinear signal testing.

Summarizing this thesis, progress in new pulse-compression technologies is reviewed in Chapter 2. Theories of nonlinear transmission lines and electromagnetic wave propagation are also reviewed in this chapter.

Chapter 3 mainly concerns silicon CPWs. CPW properties such as losses, impedances, effective permittivity and cutoff frequencies are discussed, and some recently improved technologies are reviewed. A new structure for low-loss CPW transmission lines realized on silicon substrates is presented.

CMOS varactor characteristics are discussed in Chapter 4, where six different CMOS varactor structures are investigated. Four of them were implemented on standard CMOS chips. The high-frequency characteristics of these CMOS varactors are fully studied. Of these four, two particular NMOS structures are selected, one being more suited for single-edge pulse-compression, the other for double-edge pulse-compression.

Chapter 5 presents design strategies for all-silicon pulse-compression NLTLs. The complete NLTL structure is investigated, including the influence of insertion losses on the other design considerations. This chapter includes design simulation results.

Chapter 6 deals with fabrication and laboratory measurements on the NLTL chips, including testing environment and experimental setups. Methodologies, such as on-wafer calibration and de-embedding techniques, are discussed. On-chip small-signal measurements of the NLTLs are presented. A newly developed large-signal testing technique, which is used for measuring the nonlinearity of the NLTLs, is investigated in detail.

Chapter 7 summarizes the project and outlines possible future work.

1.6 Summary

A brief history of the development of nonlinear transmission lines (NLTLs) is presented in this chapter. It includes overviews of early NLTL development, the variety of NLTL structures designed, materials used, and manufacturing processes. The motivations for designing and building all-silicon pulse-compression NLTLs are also discussed. In addition, technologies for NLTL pulse-compression are introduced in this chapter. The contributions of this work, and an outline of the dissertation, are given at the end of the chapter.

Chapter 2 Pulse Compression Technology Overview

Nonlinear transmission lines (NLTLs) have been popular during the last three decades because of their unique functions and small size. An important application of NLTLs is in high-speed sampling and timing systems, where they can be used to sharpen either the rising or falling edge of a pulse, (or both). This shrinking of the rise and/or fall times of the pulse is often referred to as pulse shaping or pulse compression.

2.1 Nonlinear Transmission Line Theory

The NLTLs investigated here consist of CPW lines interleaved with NMOS varactors. Figure 2-1 shows the configuration of an NLTL. Here V_s and R_s characterize the pulse source, Z_0 represents the characteristic impedance of the CPW, d is the section length of the CPW, and $C(V)$ and R_v represent respectively the nonlinear capacitance and the loss resistance of the NMOS varactor. The approximate equivalent circuit for an NLTL is given in Figure 2-2, where L and C_0 are the equivalent linear inductance and capacitance per section, and R represents the series resistance of the CPW.

Figure 2-1 Circuit diagram of the NLTL configuration.

Figure 2-2 Equivalent circuit of the NLTL configuration.

A nonlinear transmission line has three fundamental and significant characteristics: nonlinearity, dispersion, and dissipation. These characteristics define an NLTL's functions and behavior. Nonlinearity is due to the voltage-dependent varactor capacitance, dispersion to the periodicity of the NLTL, and dissipation is due to the

finite conductivity of the CPW conductors (represented by the elements R in Figure 2-2) and the series resistance R_v of each varactor.

2.1.1 Dispersion and Bragg Frequency

Dispersion is variation of phase velocity with frequency. An expression for the dispersion of an NLTL can be derived from a unit cell of the line. Figure 2-3 diagrams a single section of the NLTL and its equivalent circuit.

Figure 2-3 (a) Unit section of an NLTL, (b) its equivalent circuit.

To simplify the dispersion derivation, the NLTL is assumed to be lossless, that is $R = 0$ and $R_v = 0$. For the equivalent circuit of the unit section, the characteristic ABCD matrix is

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & \frac{j\omega L}{2} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ j\omega C & 1 \end{bmatrix} \begin{bmatrix} 1 & \frac{j\omega L}{2} \\ 0 & 1 \end{bmatrix} \quad (2.1)$$

where

$$C = C_0 + C_v = C_0 + \frac{1}{V_H - V_L} \int_{V_L}^{V_H} C(V) dV ,$$

and V is the applied varactor bias voltage.

The propagation constant of the line is $\cosh(\gamma d) = (A + D)/2$, where $\gamma = \alpha + j\beta$ (m^{-1}) is the complex propagation constant, α and β are the attenuation constant (Np/m) and phase constant (rad/m) respectively, while d is the physical length of the NLTL. For a lossless NLTL, $\cosh(\gamma d)$ reduces to $\cos(\beta d)$. From equation (2.1), the lossless dispersion formula is found to be

$$\cos(\beta d) = 1 - \frac{\omega^2 LC}{2}$$

or in terms of phase

$$\beta d = \cos^{-1} \left[1 - \frac{\omega^2 LC}{2} \right] \quad (2.2)$$

The dispersion relationship is shown in Figure 2-4. Here it is assumed that $L = 68 \text{ pH}$, $C = C_0 + C_v = 8.2 + 19 \text{ fF} = 27.2 \text{ fF}$ for the equivalent circuit of a NLTL cell in Figure 2-2 (b). Here we assume small-signal operation, so that C_v can be treated as a linear capacitance.

Figure 2-4 Dispersion diagram for an NLTL unit cell

The small-signal characteristic impedance of the line is $Z = \sqrt{B/C}$. From (2.1), we have

$$Z = \sqrt{\frac{L}{C} - \frac{\omega^2 L^2}{4}}. \quad (2.3)$$

Figure 2-5 shows the characteristic impedance Z_0 as a function of frequency. The L and C element values of the equivalent NLTL cell are the same as those used for Figure 2-4.

Figures 2-4 and 2-5 show that the dispersion relationship has a low-pass characteristic. The Bragg frequency f_B is defined as the low-pass corner frequency at which the propagation factor βd is equal to π and the impedance Z becomes zero. From equation (2.2) or (2.3) we have

$$f_B = \frac{1}{\pi\sqrt{LC}}. \quad (2.4)$$

In Figure 2-5, the Bragg frequency is seen to be 230 GHz.

Figure 2-5 Characteristic impedance as a function of frequency for an NLTL cell.

2.1.2 Nonlinearity

Nonlinearity of the NLTLs is due to the nonlinear components such as diodes or varactors. There are two types of nonlinearity: conductive and reactive. Conductive nonlinearities are defined by $I(V)$ curves while reactive nonlinearities are defined by $C(V)$ curves. In mixer or detector applications nonlinear conductive elements are used. In many other RF applications, due to the dissipative nature of the conductive nonlinearity, reactive nonlinearity is much more preferred.

Figure 2-6 shows a MOSFET RF equivalent circuit [35].

Figure 2-6 MOSFET RF Equivalent Circuit.

The shape of the $C(V)$ characteristic of a diode or varactor depends mainly on the structure of the device. Taking the total gate capacitance of a MOS transistor as example and assuming small-signal conditions, the capacitance C_{gs} between the gate and source is approximately given by [36]

$$C_{gs} = \frac{2}{3} WLC_{ox}, \quad (2.5)$$

where W is the gate width and L is the effective gate length. $C_{ox} = K_{ox}\epsilon_0 / t_{ox}$ is the gate capacitance per unit area, where K_{ox} is the relative permittivity of SiO_2 , ϵ_0 is free space permittivity, and t_{ox} is the thickness of the thin oxide under the gate. This capacitance is primarily due to the change in channel charge as a result of a change in V_{GS} (the voltage between the gate and source).

Also, the capacitance C_{sb} between the source and bulk is given approximately by

$$C_{sb} = (A_s + A_{ch}) \frac{C_{sj0}}{\sqrt{1 + \frac{V_{SB}}{\phi_0}}} \quad (2.6)$$

Here A_s is the area of the source junction, A_{ch} is the area of the channel, V_{SB} is the applied voltage between the source and bulk, C_{sj0} is the depletion capacitance of the source junction at zero bias, and ϕ_0 is the barrier potential at zero bias.

In the large-signal case, the effective gate length and the channel area will be a function of depletion depth but only a very weak function, which in turn is a function of the applied voltage V . The $C(V)$ curves given (2.5) and (2.6) are different for different doping profiles, and numerical methods are often required to obtain the $C(V)$

data. These data can be approximated by various different functional relationships. The most common function used for diode $C(V)$ curves is

$$C_j(V) = \frac{C_{j0}}{(1 - V/V_0)^N} \quad (2.7)$$

where C_j is the junction capacitance, C_{j0} is the fitted zero-bias junction capacitance, V is the junction potential, V_0 is the fitted barrier potential, and N is a fitting coefficient which depends on the junction doping profile. The $C(V)$ of (2.7) fits most diode characteristics because it provides three fitting parameters: C_{j0} , V_0 , and N .

2.1.3 Dissipation

There are two main sources of dissipation in an NLTL. One is loss in the diode series resistance (R_v in Figure 2-3), the other is metal conductor loss (R in Figure 2-3). Diode losses are caused by the non-zero contact and bulk resistances of the semiconductor structure while metallic losses are due to the geometry and finite conductivity of the CPW.

The diode loss can be represented by

$$\alpha_{c-dio} = \frac{G_{shunt}}{2Y_{NLTL}} \quad (\text{dB/m}) \quad (2.8)$$

where G_{shunt} is the shunt conductance of the diode, and Y_{NLTL} is the characteristic admittance of the loaded NLTL. The diode loss increases with the square of the operating frequency.

For high-frequency CPWs built on a silicon substrate, there are more than just resistive metallic losses. The first is the conductor loss which includes metal skin-effect loss. The second is the silicon substrate loss or dielectric loss, and the third is the radiation loss.

The conductor loss can be expressed as [37]

$$\alpha_{c_CPW} = 8.69 \frac{(R_s + R_g)}{2Z_{CPW}} \quad (\text{dB/m}), \quad (2.9)$$

where R_s the unit-length resistance of CPW signal line, and R_g is the unit-length resistance of the CPW ground lines.

The dielectric loss can be written [38]

$$\alpha_{d_CPW} = 8.69 \frac{\pi}{\lambda_0} \frac{\epsilon_r}{\sqrt{\epsilon_{reff}}} \kappa \tan(\delta) \quad (\text{dB/m}), \quad (2.10)$$

where λ_0 is free space CPW guide wave length, ϵ_r and ϵ_{reff} are relative and effective permittivities of the CPW substrate, κ is the structure coefficient of the CPW, and δ is the dielectric loss angle of the substrate.

The radiation loss can be represented by [39]

$$\alpha_{r_CPW} = 8.69 \left(\frac{\pi}{2} \right)^5 \frac{2 \left(1 - \frac{\epsilon_{reff}}{\epsilon_r} \right)^2}{\sqrt{\frac{\epsilon_{reff}}{\epsilon_r}}} \frac{(w+2s)^2 \epsilon_r^{1.5} \left(\frac{f}{c_0} \right)^3}{K(k_0') \cdot K(k_0)} \quad (\text{dB/m}), \quad (2.11)$$

where w is the width of the CPW signal line, s is the gap between the CPW signal and ground lines, $k_0 = w/(w+2s)$, $K(k_0)$ is the complete elliptic integral of the first kind, $K(k_0') = K(\sqrt{1-k_0^2})$, f is the operating frequency, and c_0 is the free-space speed of light.

The conductor loss is proportional to $f^{0.5}$, the dielectric loss to f , and the radiation loss to f^3 , where f is the operating frequency. These formulae only apply for NLTLs at frequencies well below the Bragg frequency. In order to determine the exact effects of diode and skin losses on the NLTL, we need to introduce the losses into the $ABCD$ matrices and extract the real part of complex propagation constant γ . Assume a transmission line with a length of d and characteristic impedance of Z_0 , its $ABCD$ parameters are:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cosh(\gamma \cdot d) & Z_0 \cdot \sinh(\gamma \cdot d) \\ \frac{1}{Z_0} \cdot \sinh(\gamma \cdot d) & \cosh(\gamma \cdot d) \end{bmatrix}.$$

The line insertion and reflection losses can be calculated respectively by [40]:

$$S_{21} = \frac{2}{A + \frac{B}{Z_0} + Z_0 \cdot C + D}, \quad S_{11} = \frac{A + \frac{B}{Z_0} - Z_0 \cdot C - D}{A + \frac{B}{Z_0} + Z_0 \cdot C + D}.$$

2.2 Physical Realization of NLTLs

The NLTLs as monolithic integrated circuits are fabricated on semiconductor substrates. NLTLs can be designed for a variety of applications, but all suffer from parasitic effects of lossy transmission lines, parasitic series resistance of diodes or

varactors, and other parasitics that depend on the layout. What follows is a discussion of the physical nature of the NLTLs without considering the various parasitic effects. Reasons are given for choosing particular types of transmission lines, varactors, and layout structures.

In order to make an NLTL work at high-GHz frequencies, a low loss transmission line is required. Currently, there are several types of monolithically integrable transmission lines: microstrip, CPW, stripline, suspended substrate line, coplanar strips, and slotline [40]. Slotline has very low characteristic impedance, so it would be a poor choice for 50- Ω systems since diode loading reduces the NLTL impedance. Coplanar strips have very high characteristic impedance, but these lines require a balanced signal. In high frequency systems, most circuits require an unbalanced signal referenced to a common ground. Striplines require symmetric dielectric material on both sides of the conductor; hence they are inappropriate for monolithic integration. Suspended substrate lines are very similar to microstrips but with lower dielectric losses. The problem with this structure is that it requires the conductors to be suspended in a grounded box, good for packaging but unsuitable for on-wafer interconnects. Therefore, only microstrips and CPWs are commonly used in monolithic integrated circuits.

Microstrips have been around for a long time [41], and there are tremendous resources available, by way of mathematical models and simulation equivalents for

different sizes of the lines and discontinuities, and so on. Microstrip is essentially a strip of metal on an insulator that has a back plane of ground metal. This structure is ideal for inserting devices in series with the line, but in order to place an element in shunt, a via-hole must be drilled through the substrate to let the devices make a connection with the ground. Via-holes introduce parasitic inductance, fabrication difficulties, and layout problems. Even though microstrip is still the industry standard for microwave and millimeter wave circuits, CPWs have become more popular in the last decades.

Coplanar waveguide (CPW) offers convenient shunt and series element equivalents, allows a reasonable range of impedance and has relatively low losses. In this research work, CPWs are chosen as the NLTLs' linear components. Unfortunately, because there are few and inadequate CPW discontinuity models available, intensive electromagnetic simulations are necessary when designing such CPWs.

NLTLs' nonlinear behaviour comes from capacitive nonlinearity of diodes or varactors. Diode loss reduces NLTL nonlinear performance, as pointed out in [20]. Good diodes for NLTL applications should have low series resistances. This can be achieved by choosing proper diode or varactor structures and doping profiles. In this study, NMOS varactors are used as the NLTL nonlinear components since they are standard components in the CMOS process. Different structures of the NMOS varactors are discussed in Chapter 4.

In high-frequency operation, it is also important to minimize discontinuities of interconnections and to reduce sources of parasitic effects associated with NLTL layouts. Figure 2-7 shows layout-produced parasitic inductances and capacitances for a NLTL cell, where the series inductance is produced by via-connections between varactors and CPWs, and the shunt capacitance is caused by metal fins used to connect the varactors to the CPW lines. These parasitics can affect NLTL functioning; they can couple energy from desired the CPW mode into other undesired modes. There are many ways to produce a good layout. Increasing via density is one way to reduce the parasitic inductance. Making a perfectly symmetric layout structure on both sides of the CPW signal line can avoid the coplanar-strip-mode (the two CPW ground planes at different potentials). Reducing unnecessary metal overhanging can reduce the parasitic capacitance. However, the layout parasitics cannot be totally eliminated. It is necessary to take them into account when doing the designing.

Figure 2-7 Diagram of layout-produced parasitic inductances and capacitances for a NLTL cell.

2.3 Pulse-Compression Technology

Transmission lines loaded with diodes or varactors can function as pulse-compression components. Rodwell [20] contributed classical analyses of Schottky-diode CPW lines. If dispersion and losses are neglected, one need only consider the LC equivalent circuit operating well below the Bragg frequency. For such an NLTL, the small-signal propagation delay decreases with increasing reverse bias voltage. For an input pulse with fall time T_{f_in} over a voltage swing from V_{low} to V_{high} , this predicts the output pulse fall time to be

$$T_{f_out} = T_{f_in} - l \left| \sqrt{LC(V_{high})} - \sqrt{LC(V_{low})} \right| \quad (\text{s}), \quad (2.12)$$

where l is the line length, and L and $C(V)$ are the inductance and voltage-dependent capacitance per unit length respectively. Theoretically, the output pulse fall time can reach to zero.

Rodwell defined the effective loading capacitance of the diode over the pulse's voltage swing (and therefore also the effective NLTL impedance.) to make it a constant. This results in the so called *large-signal* capacitance

$$C_{ls} = \frac{1}{V_{high} - V_{low}} \int_{V_{high}}^{V_{low}} C_j(V) dV \quad (F) \quad (2.13)$$

and characteristic impedance

$$Z_{ls} = \sqrt{\frac{L}{C_{ls} + C_{line}}} \quad (\Omega) \quad (2.14)$$

where $C_j(V)$ is the diode capacitance, L is the CPW per unit length inductance, and C_{line} is the CPW per unit length capacitance.

Following Figure 2-3, where the diode equivalent circuit is R_v in series with $C(V)$, the varactor cutoff frequency is defined as

$$f_c = \frac{1}{2\pi R_v C(V)} \quad (Hz) \quad (2.15)$$

while the Bragg cutoff frequency is

$$f_B = \frac{1}{\pi d \sqrt{L(C_0 + C(V)/d)}} \quad (Hz) \quad (2.16)$$

Here we assume that L and C_0 are the per-unit-length inductance and capacitance of the CPW. In large-signal operation, $C(V)$ is replaced by C_{ls} .

According to [20], when f_c/f_B is $4 \sim 6$, the NLTL is in a weak dispersion mode, and can produce shock waves. In other words, if we input a pulse wave, the rising or fall edge of the pulse wave at the output will be sharpened. If the f_c/f_B ratio is less than 4, the compression mode will not be exhibited by the NLTL. If the ratio is much higher than 6, the NLTL becomes more dispersive and solitons can be generated in the line. The soliton case will not be discussed in this thesis.

There are three things that can limit the NLTL pulse compression efficiency: the substrate or dielectric loss, the diode resistance, and the CPW attenuation, including the frequency-dependent skin-effect loss. The substrate loss depends on the process technique. The diode resistance loss depends on the diode's structure. The CPW loss depends on the CPW dimensions. For example, CPW skin loss can be reduced by increasing the centre conductor width of the CPW but at the cost of increasing the radiation loss. In general, the longer, the NLTL, the larger the attenuation. The use of the tapering technique can reduce the NLTL attenuation. Because of the nonlinear properties of the NLTLs, when a pulse signal propagates along the line, the waveform which is closer to the input end has a narrower bandwidth, with fewer harmonics, than that of the waveform further down along the line. One can use a low Bragg cutoff

frequency at the input end and a high Bragg cutoff frequency at the output end to reduce the total length and size of the NLTL, thereby reducing its attenuation. There are two ways to taper an NLTL. One way is to taper the width of the CPW signal line, making it widest at the input end and gradually narrowing it down to the output end, as mentioned in [61]. The disadvantage of this method is that it creates CPW discontinuity and makes both the design and layout complicated. Another way is to taper the NLTL section length, that is, placing the longest section-length at the input end and gradually reducing the lengths of the sections towards the output end, as presented in [20]. This work used the second tapering method. No matter which method is used, the width of the varactors used in the NLTL has to be tapered too. One reason is to keep the right ratio between the Bragg cutoff frequency and the varactor cutoff frequency at the each section. A second reason is to keep the impedance at the each section at $50\text{-}\Omega$. More details on the tapering technique will be given in the later chapters.

2.4 Summary

This chapter gives an overview of NLTL pulse-compression technology. The three main characteristics of the NLTL, dispersion, dissipation, and nonlinearity, are fully described. The challenges of designing all-silicon pulse-compression NLTLs are pointed out. Solutions for overcome these challenges are suggested. The framework of physically realizing the NLTLs is also described.

Chapter 3 Silicon CPW Performance

3.1 Characterization of Silicon CPWs

CPWs have become the most popular components among planar microwave transmission lines during last two decades. Figure 3-1 shows the structure of a silicon CPW (top) and its electromagnetic field distribution (bottom). A CPW line consists of one signal line and two symmetrical ground planes. This balanced structure makes it easy to fabricate CPWs on semiconductor substrates and to integrate them with other MMIC or RFIC components. As shown in Figure 3-1, some of the electric field lines produced by the signal line terminate at the ground planes, the remainder terminate at the metal ground beneath the substrate. The magnetic field lines loop around the signal line, and are always perpendicular to the electric field.

Figure 3-1 Silicon CPW structure and the field distribution.

Figure 3-2 The cross-section view of a silicon CPW with pertinent dimensions.

Figure 3-2 gives a cross-sectional view of the silicon CPW [25]. Because a standard CMOS process is used, the center conductor (signal line of the CPW) is fully encapsulated in layers of dielectric glass, while the ground planes either make ohmic contacts to the underlying Si-substrate or can be left floating.

If the transverse CPW dimensions are much smaller than the shortest wavelength of the applied signals, a quasi-TEM mode can be assumed [25]. This means that the transverse fields carry the majority of the electromagnetic energy. In this study, the largest transverse dimension of the silicon CPWs is around 200 μm . The operation frequency is 5 to 50 GHz; that is, the shortest wavelength is above 3 mm since the

relative effective permittivity is approximately 4. Thus the quasi-TEM criterion is satisfied. At higher frequencies, higher order modes may be induced. The lowest mode above the TEM mode in the CPW line is the TE_{10} mode. Frankel and Hasnain gave more precise characterization of CPW lines without using the quasi-TEM assumption [39], [42]. The details are presented in Appendix A. The formulas in Appendix A may serve only as design guidelines for the implementation of silicon CPWs, since the circuit analyzed was much simpler than a real silicon CPW, the layer of dielectric material under the CPW metal not being included. To predict the behavior of a CPW on a silicon chip without using quasi-TEM assumptions would be very complicated; sophisticated mathematical mapping would be involved. It would be very hard to obtain explicit formulas. Fortunately, powerful numerical methods, such as FEM (Finite Element Method) as used in the HFSS software [50], and Method of Moments (MOM) in Sonnet [43], can help with simulations.

To analyze the CPW of Figure 3-2, the equivalent circuit needs to be defined. Two quasi-TEM equivalent circuits for silicon CPWs have been discussed in [25]. They are shown in Figure 3-3. The circuit in Figure 3-3(a) includes elements from the model of Shibata [29], which was derived from a full-wave computer analysis of a similar metal-insulator-semiconductor (MIS) structure. The circuit in Figure 3-3(b) is modified version of the model used by Zaghloul's group [25]. This group did intensive simulation and on-chip testing, and proved that quasi-TEM equivalent circuits can predict the measured silicon CPW parameters over a broad frequency range from 0.1 to 40 GHz, the circuit in Figure 3-3(b) is an improved model. In general, one can use

the circuit of Figure 3-3 (c) as a theoretical model for silicon CPWs, where Z is the total series impedance per unit length, and Y is the total shunt admittance per unit length.

Figure 3-3 Quasi-TEM equivalent circuit of a silicon CPW: (a) Shibata's model, (b) modified model, (c) symbolic model.

In Figure 3-3, capacitor C_p represents the capacitance between the signal line of the CPW and the Si substrate. Capacitor C_s represents the capacitance between the signal line and the ground plane of the CPW. The conductance G_t is due to transverse current induced in the Si substrate by the electric field, while C_t similarly represents transverse displacement currents. Both G_t and C_t are determined by the electrical energy stored in the substrate. Elements L and R_s are the inductance and resistance of

the aluminum conductors, respectively. R_l represents the longitudinal current losses in the Si beneath the signal line.

In theory, the complex propagation constant γ and the complex characteristic impedance Z_0 of CPW are determined from Z and Y as

$$\gamma = \alpha + j\beta = \sqrt{Z \cdot Y} \quad \text{and} \quad Z_0 = Z' + jZ'' = \sqrt{\frac{Z}{Y}}.$$

In practice, by measuring the S-parameters of the CPW lines, and using the ABCD cascade matrix, the propagation constant and characteristic impedance can be determined [38]. With the propagation constant and characteristic impedance known, the frequency-dependent equivalent-circuit parameters per unit length of the lines (R , L , C , and G) can be calculated from

$$j\omega C + G = \frac{\gamma}{Z_0} \quad \text{and} \quad j\omega L + R = \gamma \cdot Z_0.$$

Also, the attenuation factor α and the normalized phase factor $\beta c / \omega$ can be calculated from the complex propagation constant. Here, c is the speed of light. The normalized phase factor also gives the effective permittivity, since $\beta c / \omega = \sqrt{\epsilon_r}$.

3.2 Silicon CPW Losses

There are three main contributions to the loss in silicon CPWs. The first is the conductor loss which includes metal skin-effect loss. The second is the silicon substrate loss, and the third is the radiation loss. The conductor loss is proportional to $f^{0.5}$ [37], the dielectric loss to f [38], and the radiation loss to f^3 [39], where f is the operating frequency.

Using a wider CPW signal line can reduce the conductor loss but at the expense of increasing the radiation loss [20]. Using a wider gap between the CPW signal line and the ground planes can also reduce the conductor loss but this will increase both substrate and radiation losses. For this research, silicon CPWs of various sizes have been simulated by using a commercial EM simulator ADS [44]. Figure 3-4 shows the attenuations of the CPWs with the signal line width $W = 4 \mu\text{m}$ and the gap between the signal and ground lines $S = 4, 8$ and $10 \mu\text{m}$. Figure 3-5 shows the attenuations of the CPWs with the signal line width $W = 8 \mu\text{m}$ and the gap between the signal and ground lines $S = 8, 12, 16$ and $24 \mu\text{m}$. Figure 3-6 gives the attenuations of the CPWs with $W = 10 \mu\text{m}$ and $S = 10, 18, 30$ and $40 \mu\text{m}$. The simulation assumes a commercial CMOS 0.18- μm process. The simulation parameters, such as the properties of the substrate and layout layers, were based on the data for the CMOS 0.18- μm process.

The simulation assumed that the top metal layer of the process (Metal 6) was used for the CPW lines.

The simulated results in Figure 3-6 are reasonable in view of the results in [25]. In this work, the attenuation values are almost 10 times lower than those in [25]. The reasons are as follows:

(1) In this study, a Si substrate with a resistivity of $10 \Omega \cdot \text{cm}$ was used, whereas in [25], the substrate resistivity was in the $1 \Omega \cdot \text{cm}$ range, which introduced much more substrate loss.

(2) In this work, the top metal layer of the process was used, while in [25], the second and lower layers were used. As the measured and simulated results in [25] indicate, the higher the metal layer is above the substrate, the lower the attenuation, since the substrate losses are the dominant CPW losses.

(3) Wide-gap (larger S) CPW structures were used in this study. In general this reduces the total losses. This point will be discussed in more detail in the next section.

The simulated results show that within a certain range, using wide W and S dimensions can reduce attenuation. Because the conductor loss is proportional to $f^{0.5}$, the dielectric loss to f , and the radiation loss to f^3 , from the shapes of the attenuation

curves we can see that, at frequencies below 50 GHz, the attenuation is due mainly to substrate and metal losses, and dominated by the substrate losses.

Figure 3-4 Attenuation of silicon CPWs ($W = 4 \mu\text{m}$).

Figure 3-5 Attenuation of silicon CPWs ($W = 8 \mu\text{m}$).

Figure 3-6 Attenuation of silicon CPWs ($W = 10 \mu\text{m}$).

3.3 Slow-wave Technique for Substrate Loss Reduction

CPWs on lossy silicon substrates tend to have high attenuations at frequencies above 20 GHz [45], [25]. The loss is caused by parasitic coupling of energy from the CPW signal line to the lossy substrate. Using a CPW with a narrow gap between the signal and the ground lines can reduce the substrate loss but at the expense of

increased conductor loss [20], [45]. Using a wide-gap line, on the other hand, reduces conductor loss but allows more electric field penetration into the silicon dioxide and the substrate, resulting in higher dielectric and substrate losses. Some research groups have suggested new design methodologies to overcome the problem. One of these methods was presented by Yook's group [46]. They introduced a so-called elevated-CPW. In this structure, the CPW signal line is elevated above the CPW ground planes and is further away from the substrate. This reduces the substrate losses. The other method, introduced by Harjani's group [47], uses multiple metal layers under the CPW signal line to shape the electric fields and prevent the energy from getting into the substrate. Yook's structure is difficult to fabricate on a standard CMOS chip since micromachining is required. Harjani's structure makes layout and fabrication much more complex.

A simpler approach is to add a ground shield underneath the silicon dioxide as suggested in [45]. This ground plane isolates the substrate from the signal line, reducing the substrate loss. However, since the return current flows mostly in the ground shield instead of the CPW ground, the skin effect loss is higher, which is due to microstrip-type behavior. Following Cheung and Long's idea [48], floating railway-tie-like metal strips, see Figure 3-7, are used as the ground shield instead of a solid metal plane, thereby improving the CPW performance. The advantages of this new structure, which is called slow-wave CPW, or S-CPW, are that the ground shield

return path is broken and the shield eddy currents are minimized, thereby reducing both skin effect and overall losses.

Figure 3-7 Slow-wave structure and design parameters.

The fundamental equations for the characteristic impedance Z_0 and phase velocity V_p of a lossless transmission line propagating a TEM mode are:

$$Z_0 = \sqrt{L/C} \text{ , and } V_p = 1/\sqrt{LC} \text{ .}$$

Here, L and C are the unit-length inductance and capacitance of the CPW. Using floating ground strips allows the use of wide-gap CPW lines to reduce conductor

losses, and also to achieve high characteristic impedance Z_0 (due to the large L). Since the floating ground strips are very close to the CPW layer, they increase C . Large L and C decrease the propagation velocity V_p , so that the structure operates in a slow-wave mode. Slow-wave propagation results in low loss, as observed by Hasegawa [26]. This loss reduction occurs in two ways. First, the silicon substrate loss is diminished because of the reduced E -field penetration therein. Second, the increased permittivity shortens the physical length of the CPW, further reducing the losses. Figure 3-8 shows the electromagnetic (EM) field distribution for both conventional and slow-wave CPWs.

Figure 3-8 EM field distribution of (a) conventional CPW, (b) slow-wave CPW.

Figure 3-8 shows that for a conventional CPW, most of the E -field flux goes into lossy silicon substrate, whereas for a slow-wave CPW, most of the E -field flux ends at the floating ground.

S-CPWs were first mentioned in 2004 [48], but attracted little attention since no design data were published. Furthermore, no one has built them on standard CMOS chips. To extend the results of [48], Amaya introduced new design parameters, such as H_1 , H_2 , t_2 , W_t , O , and S_T as shown in Figure 3-7 [49]. These parameters were used by the author in fully three-dimensional EM modeling employing the HFSS software [50]. The dependence of the CPW characteristics on the geometry of the floating metal layer was intensively investigated, and the optimum dimensions were determined.

Slow-wave CPW techniques offer more design flexibility than is available with conventional CPWs. In conventional CPW the electrical parameters depend only on the centre conductor width W and the gap S between the centre conductor and the coplanar grounds, see Figure 3-7. S-CPW provides additional design parameters by introducing a floating electrode directly beneath the line's centre conductor. The vertical spacings H_1 , H_2 for these floating electrodes and their thickness t_2 are limited and fixed for a given commercial process. Three extra design parameters for the SCPW are: the width W_T of each floating electrode, the spacing S_T between the floating electrodes, and the overhang projection O of the floating electrode into the

ground plane area, see Figure 3-7. This improved flexibility provides more freedom in the design, such as the tradeoff between wider center conductors with higher radiation loss and narrower ones with higher metallic loss, as well as the tradeoff between wide gaps with higher substrate attenuation and narrow ones with smaller impedances. Figures 3-9 and 3-10 present some simulation results. In these figures, the spatial “Duty Cycle” is $W_T / (W_T + S_T)$.

Figure 3-9 Effect of overhang projection O and spatial duty cycle on line impedance for a slow-wave CPW.

Figure 3-10 Effect of overhang projection O and spatial duty cycle on line attenuation for a slow-wave CPW.

Figures 3-9 and 3-10 show that

- By increasing the overhang O , the slow-wave CPW total capacitance increases, therefore decreasing the characteristic impedance Z_0 of the line.
- As the spatial duty cycle of the slow-wave CPW is increased from 0.25 to 0.5 and 0.67, Z_0 decreases due to the added capacitance (the effective permittivity ϵ_{eff} increases).

- The slow-wave CPW attenuation decreases as the spatial duty cycle is increased beyond 0.5, because the E fields are confined near the surface, reducing dielectric and radiation losses.

Figure 3-11 shows the HFSS setup for simulating a slow-wave CPW.

Figure 3-11 HFSS geometry for a slow-wave CPW simulation. M5 and M6 denote process metal layers.

In this figure L is the length of the transmission line section under simulation. L can be set to a much smaller value than the real length by using ‘Wave Port’ in HFSS, since the simulator assumes that the same structure extends, along the propagation direction,

beyond the simulation area. Using a small L can save the simulation time. The results are then normalized with the real lengths. Here L is set to 50 μm .

As a comparison, the electric field distributions for both conventional and slow-wave CPWs are shown in Figures 3-12 and 3-13, which were produced by using HFSS simulator [50]. It is evident that the slow-wave CPW with added floating ground confines the E -field to the neighborhood of the signal line and therefore reduces the substrate loss.

Figure 3-12 E -field distribution of a conventional CPW.

Figure 3-13 *E*-field distribution of a slow-wave CPW.

3.4 *Silicon CPW Implementation*

Figure 3-14 Technology stack for the CMOS 0.18- μm process.

After simulation and optimization, it was possible to realize a few S-CPWs in a CMOS 0.18- μm process. The process used for this research provides six metal layers, as shown in Figure 3-14. The top layer (M6) is the thickest and is the furthest from the substrate. The CPW was fabricated in this layer to minimize metallic and substrate losses. The floating ground strips were built in the second layer (M5), which is much thinner than the top layer, but thicker than the other four lower layers. Another benefit of making the second layer a ground-shield layer is that it is spaced above the substrate. Even if a small amount of electric field energy leaks through the metal strips, it remains mostly in the much less lossy silicon dioxide region.

Figure 3-15 is a photograph of the $2.0 \times 1.5 \text{ mm}^2$ chip. The second row and the first half of the third row contain the slow-wave CPW and conventional CPW components. The on-chip S_{21} measurements are very promising. Figure 3-16 shows the measured and simulated results for a slow-wave CPW and a conventional CPW. Here a 10- μm -wide signal line and a 30- μm -wide gap are used for both types of CPWs. Both are 200 μm long. For the floating ground of the slow-wave CPW, the strip width is 5 μm , and the gap is 0.5 μm . Figure 3-16 shows that the slow-wave CPW has much lower insertion loss than the conventional CPW, especially at higher frequencies. Notice that, with the same size of CPW, adding the floating ground strips increases C and decreases Z_0 . The measured Z_0 of the conventional CPW is $\sim 85 \Omega$, while it is $\sim 50 \Omega$ for the slow-wave CPW. The conventional CPW has more return loss than that of the S-CPW, which is near zero. Allowing a 20-30% return loss deduction, the overall loss of the conventional CPW is still greater than that of the slow-wave CPW, and increases with frequency more rapidly. Having a very low return loss, the overall loss of the slow-wave CPW line is approximately equal to S_{21} , which is $\sim 0.25 \text{ dB/mm}$ at 40 GHz.

Figure 3-17 shows the measured effective relative permittivity $\epsilon_{r,eff}$ results for the slow-wave CPW and the conventional CPW lines. One can see from this figure that by adding the floating ground, the $\epsilon_{r,eff}$ is increased by ~ 2.5 times. This means that the propagation velocity and wave length are decreased by a factor of ~ 1.5 .

Figure 3-15 CMOS $2.0 \times 1.5 \text{ mm}^2$ chip with CPW, S-CPW and NMOS varactor components.

Figure 3-16 Comparison between the measured (—) and simulated (---) $|S_{21}|$ for a conventional and a slow-wave CPW.

Figure 3-17 Effective relative-permittivity $\epsilon_{r,eff}$ for a conventional and a slow-wave CPW.

The CPW S -parameters were measured with an HP 8722ES vector network analyzer (VAN) at frequencies up to 40 GHz using Picoprobe ground-signal-ground (GSG) probes. The short-open-load-through (SOLT) calibration and de-embedding technique was used to subtract out the pad parasitics [67].

3.5 Conclusion

This chapter mainly concerns the performance of silicon CPWs. One big drawback is their high substrate losses. The lower the substrate resistivity, the larger the loss. The low resistivity of the $10\text{-}\Omega\cdot\text{cm}$ commercial substrate used here requires design efforts to overcome the loss. A new technique, slow-wave CPW, was introduced in this chapter, and the extensive simulation, chip design and chip measurement procedures are described. This research results show that using slow-wave structures in all-silicon CPW design is a very good solution, especially for high frequency operations.

Chapter 4 CMOS Varactor Characterization

MOS varactors are key components in all-silicon NLTLs. The nonlinear $C(V)$ characteristics of the varactors, together with the transmission line parameters, determine the nature of the nonlinear wave propagation along the NLTL. Varactors used in RF applications such as voltage controlled oscillators need a high quality factor Q and a capacitance-voltage characteristic $C(V)$ that provides a large linear tuning range. In contrast, varactors used in pulse-compression nonlinear transmission lines (NLTLs) require a large cutoff frequency f_c for high-order harmonic generation, and a $C(V)$ characteristic specifically optimized for a high degree of nonlinearity. Specific monotonic or non-monotonic $C(V)$ characteristics can be optimized for either single-edged or double-edged pulse compression.

4.1 CMOS Varactor structures

In order to choose the most suitable varactors for use in NLTL pulse compression, six different CMOS transistor-based varactor structures (Figure 4-1) were investigated in this study.

Figure 4-1 Cross-sections of six varactor structures.

In Figure 4-1, each varactor has two terminals: one connects to the gate (terminal 'a'), the other to the bulk (terminal 'b'). The source and the drain are interconnected, and are either connected to the bulk (structures 1, 2, 4, 5, and 6) or left floating (structure 3). Structure 1 has an *N*-type gate, with the *N*-type source and drain formed in an *N*-type well, creating an accumulation-mode MOS (AMOS) varactor with a monotonic $C(V)$ characteristic [51]. Structure 2 is slightly different: its use of a *P*-type

rather than an N -type gate shifts the transistor threshold voltage and displaces the $C(V)$ curve by ~ 1 V along the voltage axis. Structure 3 is a standard NMOS varactor, with its source-drain floating relative to the bulk, making the device work in inversion mode (IMOS). It too has a monotonic $C(V)$ curve. Structure 4 is also a standard NMOS varactor but the source-drain is connected to the bulk, making the device work in all three modes –inversion, depletion and accumulation– producing a non-monotonic $C(V)$. Structure 5 has differently-doped source and drain; one is N -type, the other P -type. We use this structure to supply minority carriers to an AMOS device (structure 1), enabling it to work not only in accumulation but also in depletion and inversion modes. Its $C(V)$ is non-monotonic. Structure 6 is similar to structure 5, but it supplies minority carriers to an IMOS device (structure 3). Structure 6 also has a non-monotonic $C(V)$. Since it has an N -source and a P -drain, it is referred to here as a SnDp device.

4.2 $C(V)$ Characteristics and Cutoff Frequency

To evaluate the advantages and disadvantages of each structure, a commercial device simulator, Synopsis Medici [52], was used. The Y -parameters of the six MOS varactor structures were generated at 0.5 GHz using the DC and AC small-signal analysis features of the simulator. Here, $Y = G + j\omega C$, since Medici assumes a simplified varactor model consisting of a capacitance in parallel with a conductance. Figure 4-2 presents an extracted model of a varactor. The parasitic L 's and R 's can be

removed from all three terminals – DS (drain and source), G (gate) and B (bulk) – by using Koolen's technique [68]. This yields the intrinsic model (shown in the broken-line box in Figure 4-2), for which the input admittance Y_{in} looking into terminal G can be extracted as

$$Y_{in} = j\omega \left[\frac{C_{ox} \cdot C_{dep}}{C_{ox} + C_{dep}} + C_{parasitic} \right] + G_{dep} + G_{parasitic} \quad (4-1)$$

Figure 4-2 CMOS intrinsic varactor model.

In the simulations, each unit-cell varactor structure was given a gate length of 0.5 μm , a gate width of 1.0 μm , a gate oxide thickness of 0.025 μm , a well doping level of $6 \times 10^{17}/\text{cm}^3$, a terminal diffusion doping level of $3 \times 10^{20}/\text{cm}^3$ and a substrate doping level of $10^{15}/\text{cm}^3$. The explanations for these simulation settings are: (1) The longer

the gate length, the bigger the $C_{\max}/C_{\min}$ ratio, but the lower the cutoff frequency [54]. It is found in this study that, for NLTL pulse-compression applications, 0.5- μm gate length can give compromised values for both $C_{\max}/C_{\min}$ ratio and cutoff frequency. (2) Due to its 2-D feature, Medici assumes 1.0- μm gate width for all the devices under the simulation. To obtain the real results, the simulated results need to be characterized by the gate width. (3) The gate oxide thickness and doping level used here are “roadmap” values, they are typical of today’s CMOS technologies [53].

Figure 4-3 presents the $C(V)$ curves of the six structures. As expected, structures 1, 2, and 3 have capacitances that vary monotonically with bias voltage, while structures 4, 5, and 6 have non-monotonic $C(V)$ s. In the first group (Figure 4-3, top), structure 3 (IMOS varactor) shows a slightly higher $C_{\max}/C_{\min}$ ratio (>5) than the other two. The $C(V)$ of structure 3 is very nonlinear with a sharp $C_{\max}$ to $C_{\min}$ transition, favouring high harmonic generation in pulse compression. In the second group (Figure 4-2, bottom), the three structures have similar $C_{\max}/C_{\min}$ ratios. The differences are in their threshold voltages, and therefore in the voltage ranges of their nonlinear $C(V)$ s. For structures 5 and 6 this range is from -3.5 to +0.5 V; for the others it is from -2.0 to +2.0 V.

Figure 4-3 $C(V)$ characteristics of the six CMOS varactor structures.

Figure 4-3 also shows that the $C(V)$ curve of IMOS (structure 3) is similar to that of MOS capacitors. This occurs because the IMOS structure is similar to a MOS capacitor structure. But IMOS varactors have higher $C_{\max}/C_{\min}$ ratios. Also the switching speed of IMOS varactors is much faster than that of MOS capacitors. The IMOS varactor acts as a gate-controlled diode, as Grove pointed out [54]. In the IMOS varactor, at inversion mode, minority carriers can be supplied by an external circuit to the inversion layer through a side of the diffusion region, enabling the gate

capacitance to increase rapidly to the value of the oxide capacitance. But in a pure MOS capacitor, minority carriers must be generated thermally in the bulk, which is an extremely slow process. This explains the difference between IMOS varactors and MOS capacitors [54].

By equating the varactor's calculated admittance $Y = G + j\omega C_p(v)$ with the reciprocal of its series-model impedance $R_v(v) + 1/[j\omega C_v(v)]$, the (bias-dependent) cutoff frequency is found to be

$$f_c = \frac{1}{2\pi R_v C_v(v)} = \frac{\omega^2 C_p(v)}{2\pi G} \quad (\text{Hz}) \quad (4-2)$$

Figure 4-4 gives the $f_c(V)$ curves of the six structures. In the upper group of Fig. 4-4, structure 3 has the highest maximum f_c but the lowest minimum f_c . Structures 2 and 3 have a slightly higher large-signal cutoff frequency (~ 690 GHz) than that of structure 1 (~ 620 GHz). Here, large-signal cutoff frequency, $f_{c,ls}$, is calculated by taking the average f_c value over the gate voltage range. Higher f_c also means a higher Q factor, since $Q = \omega C/G$. In the second group of Figure 4-3, structure 5 has the highest maximum f_c while structures 5 and 6 both have lower minimum f_c than structure 4 since the bottom plate of the capacitor is formed by the inversion-layer, increasing the resistance [55]. Structure 5 has the highest average f_c and structure 6 has the lowest average f_c . In general, when the device resistance is relatively high, that is, when the oxide-silicon interface is depleted, f_c is high. The value of f_c also depends

on the individual varactor structures. This is because of the bias-dependence of the varactor resistance R_v . In practice, setting the bias voltage slightly below the threshold value V_t (around 0.7 V for the CMOS process) will keep the f_c high.

Structure 3 (IMOS) from the first group is chosen as the monotonic-type varactor for single-edge compression because of its highly nonlinear $C(V)$, its compatible f_c and its availability in standard CMOS processes. In some applications the AMOS varactor would be a better candidate because of its higher average $f_{c,ls}$. Structure 4 from the second group is chosen as the non-monotonic-type varactor for double-edge compression, due to its standard-CMOS-process availability, even though structure 5 might be a better candidate. A further reason for choosing NMOS-type varactors for both structures is that their majority carriers are electrons, whose mobility is nearly twice that of holes. These devices are good for high frequency operation. Implementation and measured results for the two NMOS varactors are presented in the next section.

Figure 4-4 $f_c(V)$ characteristics of the six CMOS varactor structures.

4.3 NMOS Varactor Implementation

Structure 3 is named an "SDF" varactor since its source and drain float independently from the bulk, and structure 4 is called an "SDB" varactor since its

source and drain are connected to the bulk. Both varactors were fabricated in a standard 0.18- μm CMOS process.

Figure 4-5 Measured and simulated $C(V)$ curves of (a) an SDF varactor and (b) an SDB varactor.

Both types of varactors were built on a CMOS chip (top row in Figure 3-15). Each varactor has a gate of length of 1.0 μm and width 7.5 μm , and a total of 60 fingers. In Figure 4-5, the ‘Test1’ curves were obtained with an HP 4280A 1-MHz capacitance meter. On-chip pad parasitics were not subtracted. The ‘Test2’ curves were obtained using an HP 8722ES vector network analyzer, two ground-signal-ground (GSG) probes, and two Picosecond 5541A 26-GHz bias tees. In this case, the pad parasitics

were de-embedded and the data were derived from network parameters at 10 GHz. The ‘Simulation’ curves were obtained from ADS simulations, using the varactor models, also at 10 GHz. The difference between curves ‘Test1’ and ‘Test2’ in the vicinity of -2 V is due to pad parasitic effects.

Taking the chip intrinsic parameters into account, it is found that a longer gate length provides a larger $C_{\max}/C_{\min}$ ratio, but at the expense of increased varactor losses and decreased f_c . For the large varactors in Figure 4-5, the measured average $f_{c,ls}$ is ~190 GHz for the SDF type and ~120 GHz for the SDB type. The measured $C_{\max}/C_{\min}$ is ~6 for the SDF type and ~3 for the SDB type. In these NLTL designs, the gate length is scaled down to 0.25 μm , resulting in a fourfold increase in f_c , and ~15% decrease in $C_{\max}/C_{\min}$.

The work on CMOS varactors described up to this point was only the first stage. There was a concern that the varactor models used in the ADS simulation might not extrapolate to high-frequency behaviours. The HP 8722ES vector network analyzer with separated external bias tees was acceptable for this kind of testing but not the best choice, since it makes calibration very difficult. Even through the 26-GHz bias tees were used, the measured data at 20 GHz were found to be noisy. (Only the data below 10 GHz were reasonable - see ‘test 2’ curve in Figure 4-5).

This part of the work provided guidelines for the subsequent CMOS varactor research, which was done using better simulation strategies and better testing

equipment. Also the research covered a higher frequency range. The results are presented in the next section.

4.4 High-Frequency Characteristics of CMOS Varactors

Fast-developing state-of-the-art CMOS technologies, with cutoff frequencies over 200 GHz [56], have made millimeter-wave silicon RF and MMICs a reality [57], [58]. These devices help fill the demand for low-cost, low-power and compact wireless communication products. CMOS varactors, as key components in many RFICs, have received much attention [55], [56], [57], [58], and [59].

Very few publications have discussed the behavior of CMOS varactors above 50 GHz [57], [58]. Reference [55] gives a very good overview of CMOS varactor structures but covers only up to 5 GHz. Reference [59] has a good discussion of CMOS varactors, but the analysis is based on standard foundry-supplied models, which do not normally extend above 20 GHz range. In this work, the CMOS varactor study is extended to 55 GHz. The study focuses on four of the six varactor structures in Figure 4-1: They are the AMOS varactor shown in Figure 4-1(1), the IMOS (or SDF) varactor shown in Figure 4-1(3), the NMOS (or SDB) varactor shown in Figure 4-1(4), and the SnDp varactor (*N*-type source, *P*-type drain) shown in Figure 4-1(6). As pointed before, the first two structures have monotonic $C(V)$ curves, and the last

two have non-monotonic $C(V)$ curves. All these four structures were fabricated on a TSMC CMOS 0.18- μm chip.

The fabricated structures were tested using an HP 4280A 1-MHz capacitance meter for the low-frequency parameters. For GHz-range measurements, an Agilent N5250A Performance Network Analyzer (PNA) with built-in bias tees, a Karl Suss PA 200 probe station, and programmable heads with Picoprobe GSG-67A-100 probes were used. The measurement setup is shown in Figure 4-6. On-wafer measurements were extracted up to 67 GHz but for accuracy, post-measurement calculations were carried only to 55 GHz. A CS-5 ceramic-substrate kit was used to set the measurement reference plane to the tip of the probes. Koolen's technique [68] was used to de-embed the extrinsic parameters due to use of pads and interconnects. The 2-port admittance-matrix parameters of the varactor (DUT in Figure 4-6) were extracted by using “open” and “short” correction structures. More calibration and de-embedding details for high-frequency measurements are given in Appendix B.

Figure 4-6 The CMOS varactor measurement and de-embedding setup.

The four CMOS varactor types are also simulated up to 55 GHz by using the process-oriented device simulator Medici [52]. This simulator is able to predict the electrical characteristics of devices from user-specified structures, including material properties and doping profiles. To examine the high-frequency behaviour of the CMOS varactors, the tested and simulated $C(V)$ and cut-off frequency curves of the four kinds of fabricated CMOS varactors were obtained at 1 MHz, 5 GHz, 20 GHz, and 55 GHz. Figure 4-7 shows the results for the AMOS varactor. The left column contains the simulated data, the right column the measured data. The top row shows the $C(V)$ curves, the bottom row the cutoff frequencies. Similarly, Figure 4-8 shows the results for the IMOS varactor, Figure 4-9 for the NMOS varactor, and Figure 4-10 for the SnDp varactor.

Figure 4-7 Simulated and measured data for the AMOS varactor.

Figure 4-8 Simulated and measured data for the IMOS varactor.

Figure 4-9 Simulated and measured data for the NMOS varactor.

Figure 4-10 Simulated and measured data for the SnDp varactor.

The AMOS varactor data of Figure 4-7 show that on the average, the simulated $C_{\max}/C_{\min}$ ratios are ~ 3.8 while measured values are ~ 3.4 . The simulated cutoff frequencies are ~ 135 GHz while measured values are ~ 130 GHz. The IMOS varactor data of Figure 4-8 show that on the average both the simulated and measured $C_{\max}/C_{\min}$ ratios are ~ 3 . Also, the cutoff frequencies are both ~ 100 GHz. The small differences between simulation and measurement are attributed to discrepancies between simulation settings and real process values. Of particular interest is the frequency-dependence of the $C(V)$ curves. For the AMOS varactor, see Figure 4-7, the simulated $C(V)$ s vary little from 1 MHz to 55 GHz, whereas the measured high-frequency values are slightly higher than those at 1 MHz. This may be due to the 1-MHz measurements being taken using a capacitance meter, whereas the PNA was used for the 5-55 GHz measurements. For the IMOS data of Figure 4-8, the simulated and measured $C(V)$ curves both show decreased nonlinearity as the frequency increases. The reason for this degeneration is that the IMOS varactor has a larger channel resistance than the AMOS varactor, since in the IMOS structure the well makes no contribution to the channel conductivity.

The NMOS and SnDp varactors have non-monotonic $C(V)$ characteristics. Figure 4-9 shows that on the average, the NMOS varactor simulations predict $C_{\max}/C_{\min}$ ratios of ~ 3.6 compared with measured ratios ~ 3.4 . However both simulated and measured

$C(V)$ curves degenerate towards monotonicity as the frequency increases. The simulated average cutoff frequencies are ~ 100 GHz while the measured values are ~ 95 GHz. Figure 4-10, for the SnDp varactor, shows that at lower frequencies, the average simulated $C_{\max}/C_{\min}$ ratios are ~ 3.0 and the measured ratios ~ 2.7 . Both simulated and measured SnDp $C(V)$ curves are flattened in the 20-to-55 GHz range, with $C_{\max}/C_{\min}$ ratios dropping dramatically. In Figure 4-10, the average cutoff frequencies are all ~ 100 GHz.

Comparing Figure 4-5 to Figure 4-7, 4-8, 4-9 and 4-10 it can be seen that using Medici to simulate CMOS varactor is better than using varactor model through ADS. The results from Medici have better agreement with measured data. Also, the VAN with built-in bias tee gives more reliable testing results.

On the chip, the gate of each varactor has a length of $0.5\ \mu\text{m}$, a width of $5.0\ \mu\text{m}$, and a total of 12 fingers. In the simulation setup, the Medici software's default and normalized width is $1.0\ \mu\text{m}$, and we set the length to $0.5\ \mu\text{m}$. Therefore, the simulated capacitance was multiplied by 60 in order to compare it with the measured data. The cutoff frequencies should not be affected by the number of fingers.

The relatively poor high-frequency response of the SnDp structure is probably due to the need for carriers to travel the full length of the channel from the S or D region. In the other structures, carriers need only travel half the length of the channel. To test this contention, we also simulated NMOS and SnDp varactor structures at 55 GHz with the gate length shortened to 0.2 μm . Figure 4-11 compares the 0.2- μm -gate results with those obtained for the same devices with 0.5- μm gates. For this comparison, all capacitances are normalized to the same gate area. Figure 4-11 shows that with short gate length, the SnDp varactor has reduced $C(V)$ degeneration and a much higher cutoff frequency than the NMOS varactor.

Figure 4-11 SnDp varactor versus NMOS varactor at 55 GHz (short-gate = 0.2 μm , long-gate = 0.5 μm).

For a further comparison, Figure 4-12 shows the series resistances for AMOS, SDF and SDB varactors at 5 GHz. A gate area of $0.5 \times 12 \mu\text{m}^2$ was used for all the three varactors. A unit “Ohms” was used to measure the resistances for the entire device because it is improper to normalize the varactor series resistance to the gate area with the complicated MOS varactor models [60]. Figure 4-12 indicates that among these three kinds of varactors, the AMOS type has the lowest series resistance, and the SDF type has the highest series resistance. The low resistance value of the AMOS varactor is reasonable when compared to that in [60]. The explanation is that in the AMOS varactor structure, since n^+ diffusions are in an n -well, there are fewer parasitic junctions and paths to ground, resulting in higher Q and lower resistance. In the case of the SDF (or IMOS) structure, the source and drain, although connected together, are not connected to the well. Because of this, the well makes no contribution to the channel conductivity (as pointed out before), which makes the resistance higher. It is seen in Figure 4-12 that both the SDF and SDB have their resistance peak value around the depletion region. This is because, in the depletion region, the hole-electron combination rate reaches its maximum value, and leaves no free charge carriers in the channel. But with the AMOS varactor, since there are no minority carriers in the channel, the device never reaches a deep depletion mode. Therefore, for AMOS varactors, there is no resistance peak near the depletion region. This is a big advantage for AMOS varactors.

Figure 4-12 Series resistances at 5 GHz for AMOS, SDF and SDB varactors with gate area= $0.5 \times 12 \mu\text{m}^2$.

4.5 Conclusion

In this chapter, six CMOS varactor structures are investigated and discussed. Both measured and simulated data for the six structures are presented and compared. These research results show that both AMOS and IMOS varactors are good for single-edge pulse compression due to their higher $C_{\text{max}}/C_{\text{min}}$ ratios and cutoff frequencies. However, the AMOS varactor has advantages over the IMOS (or SDF) varactor because its $C(V)$ curve does not degenerate at higher GHz frequencies, and its series

resistance is smaller than that of the IMOS varactor. Both NMOS (or SDB) and SnDp varactors are good for double-edge pulse compression due to their non-monotonic $C(V)$ curves. However, the high-frequency degeneration restricts their use to a much lower GHz range. Using a short-gate SnDp varactor can reduce its high-frequency degeneration and improve its performance.

Chapter 5 All-Silicon Pulse-Compression NLTL

Design

5.1 Design Strategies

A critical condition for a correctly-functioning pulse-compression NLTL is that the ratio of the varactor cutoff frequency f_c to the Bragg cutoff frequency f_B should be ~ 5 [20]. For a 50- Ω NLTL, a CPW line with $Z_0 > 50 \Omega$ is needed. This is because the varactors capacitively load the CPW line, thereby reducing the Z_0 . To get sufficient NLTL nonlinearity, the varactors need a large $C_{\max}/C_{\min}$ ratio. This ratio depends on the CMOS varactor structure. Also, $C_{\min}$ must be large enough to overcome the linear capacitance-per-section of the NLTL. Tapered NLTLs are used for loss minimization [61]. When a pulse propagates along an NLTL, its rising and/or falling edges become sharper, so that the signal bandwidth progressively increases. In the so-called tapering technique, the NLTL section length and varactor gate width are both continuously decreased along the transmission line, thereby correspondingly increasing the Bragg cutoff frequency.

The tapering structures can be linear or non-linear. The nonlinear example is an exponential tapering structure [62], in which the diode capacitances C_n and the diode spacings L_n were tapered following an exponential trend as

$$C_n = C_1 \cdot a^{n-1} \quad \text{and} \quad L_n = L_1 \cdot a^{n-1} , \quad (5-1)$$

where C_1 and L_1 are the first capacitor and section length of the NLTL, a is the taper factor, and n is the index. In this work, the tapering function is simplified to a linear type. The tapering equations are presented as

$$L_{n+1} = (1 - n \cdot K_1) \cdot L_n \quad \text{and} \quad W_{n+1} = (1 - n \cdot K_2) \cdot W_n , \quad (5-2)$$

where L is the NLTL section length, W is the varactor gate width, n is the section index, and K_1, K_2 are taper factors.

Another way to explain this is to say that the tapered NLTLs are designed such that the Bragg cutoff frequency f_B increases gradually along the line, thereby coping with the increasing bandwidth of the propagating signal. Also, the NLTL losses are reduced by using progressively shorter CPW section lengths and smaller-size varactors.

In all-silicon NLTL designs, the greatest challenge is the loss in the silicon substrate. Here slow-wave CPWs are used as NLTL linear elements to overcome the substrate loss. The NMOS SDF and SDB varactors are used as NLTL nonlinear

elements to do single-edge or double-edge pulse sharpening, since they are available components in the standard CMOS process.

Figure 5-1 gives a flow chart of the all-silicon NLTL design process. The Synopsys Medici [52], Ansoft HFSS [50], Sonnet [43], Agilent ADS [44] and Cadence software packages were used for the pulse-compression NLTL simulations and designs. These packages are briefly described as follows:

- Medici is a process-oriented device simulator. The simulator is able to predict the electrical characteristics of devices from user-specified structures, including material properties and doping profiles. It is suitable for evaluation and characterization of CMOS varactor structures.

Figure 5-1 Flowchart of the silicon NLTL design process.

- HFSS is an industry-standard high-frequency simulation program. It can be used for S -parameter extraction and 3-D full-wave EM (electromagnetic) field simulation. Also, it can provide high accuracy, large capacity, and good performance when designing high-frequency and high-speed components. In this study, HFSS was used for characterizing and optimizing the silicon CPW and silicon slow-wave CPW lines.
- Sonnet is also a high-frequency electromagnetic program. Similar to HFSS, Sonnet can analyze 3-D EM field configurations and develop RF models with S -, Y -, Z - parameters. However, Sonnet is sometimes called “2.5-D” software because it can only deal with planar circuits. The advantages of this software are that it requires much shorter simulation times than HFSS, and the setups are simpler and more straightforward. In this study, Sonnet was used for the first-stage evaluation and in producing the initial simulation values for HFSS.
- ADS is a powerful electronic design-automation software system. It offers complete design integration for RF and wireless communication networks. ADS can import component S -parameters from Medici, HFSS, and Sonnet. Here, ADS was used to integrate the CPW and varactor components, and to design the NLTL circuits.

- Cadence is a semiconductor design system. It provides front-to-back design tools and a standard interface to the chip manufacturers. The Cadence design tool was used for the NLTL chip layout. The NLTL lines were built on a standard CMOS 0.18- μm microchip through the facilities of CMC Microsystems Inc.

5.2 NLTL Chip Layout

The NLTLs were fabricated on standard 0.18- μm CMOS chips. There are two kinds of components included in the circuits, CMOS varactors and silicon CPWs. The vias on the chips were used to make the connections between the varactors and the CPW lines. Figures 5-2 and 5-3 give simplified top and cross-sectional views of a NLTL layout. Note that the figures are not drawn to scale.

Figure 5-2 Top view of a CMOS NLTL layout (not to scale).

Figure 5-2 is top view of a CMOS NLTL layout. CMOS varactors were connected to a CPW line symmetrically on the both sides of the CPW signal line. This will keep the NLTL from working in coplanar-strip-mode (the two CPW ground planes at different potentials), as pointed out in Chapter 2.

Figure 5-3 Cross-sectional view of the CMOS NLTL shown in Figure 5-2 at plane a-a' (not to scale).

Figure 5-3 is cross-section view of the CMOS NLTL. It was mentioned in Chapter 2 that certain layout strategies can be used to reduce layout-produced parasitics. Increasing the density of the connecting vias can reduce inductive parasitics.

5.3 Simulated NLTL Waveforms

Simulated waveforms for the all-silicon slow-wave pulse-compression NLTL design are shown in Figure 5-4. The simulations were done by using the ADS

simulator. Measured data for the fabricated CMOS varactors and silicon slow-wave CPW lines were used in the simulation. In other words, the simulated results were based on the measured component data.

Figure 5-4 Simulated slow-wave-NLTL waveforms using measured S-CPW and varactor data.

For the first section of the NLTLs, the CPW signal-line width is $10\text{ }\mu\text{m}$ and the gap between the signal and the ground line is $45\text{ }\mu\text{m}$. Each varactor gate finger is $0.25\text{ }\mu\text{m}$ long by $2.5\text{ }\mu\text{m}$ wide, and there are 30 fingers. Since the tapered technique was used, taper factors are selected to get maximum NLTL efficiency. Here, for the SDF-NLTL, the section-length taper factor is 0.06. The same number is used for the

varactor gate width. For the SDB-NLTL, the section-length taper factor is 0.06, and 0.05 for the varactor gate width. The entire lengths of the SDF and SDB NLTLs are 2.7 mm and 3.4 mm respectively. Both NLTLs have 16 sections. The 10-90% rise and fall times of the input were set at 40 ps. Figure 5-4 shows that the output rise time of the SDF NLTL is about 10 ps, which represents 75% compression. While the output rise and fall times of the SDB NLTL are both about 16 ps, which represent 60% compression.

5.4 Summary

This chapter discusses strategies for designing pulse-compression NLTLs, which include using the tapered technique, simulating with various commercial software packages, and experimenting with different layout designs. The simulated results based on the measured CPW and varactor data are also presented in this chapter. These simulated results demonstrate that SDF varactors are especially good for single-edge compression and that SDB varactors are especially good for double-edge compression. The 75% reduction in the rise time for single-edge compression and 60% reduction on both rise and fall times for double-edge compression indicate it should be very possible to build effective pulse-compression NLTLs on standard silicon chips.

Chapter 6 Device Measurements

Based on the analysis, modeling, and simulation, the NLTL circuits, including their linear CPW and nonlinear varactor components, were realized on CMOS ICs. A CMOS 0.18- μm process was used as the study platform. In this study, frequencies of the input signals are in the range of 5 to 10 GHz. The output signals can reach to 50 GHz due to enriched harmonics, which means that the output waveform can have transient times in the picosecond range. This makes the measurements very difficult. The measurements can be divided into two parts: small-signal frequency-domain testing and large-signal time-domain testing.

6.1 Small-signal Frequency-domain Testing

Small-signal frequency-domain testing includes S -parameter measurement. S -parameters are very useful for small-signal modeling in situations where the values of equivalent elements can be verified. They are also very useful for determining NLTL insertion loss, characteristic impedance and cutoff frequency. Vector network analyzers (VNA) are able to measure the amplitude as well as the phase of the S -parameters of the device under test (DUT). A schematic representation of a VNA is shown in Figure 6-1. The VNA includes an RF generator. The ports are measured in turn, when the source is switched from one port to another. When one port of a 2-port

is connected to the generator, the other is switched to a matched load. For example, when switch is in the “I” state, the port on the left is connected to the generator, the other to the matched load. The delivered signal is then split into two parts. The first part, denoted RA or RB, is an image of the incident wave, while the second part propagates toward the DUT. Directional couplers are used to recover the reflected part of the signal (TA) and a part of the signal that came out of the DUT (TB). However, the measured S -parameters are not the real S -parameters of the DUT, since the reference plane of the VNA can be far from the edge of the DUT. The DUT is usually connected to the VNA through long coaxial cables, represented in Figure 6-1 as boxes A and B. As a consequence, the measured S -parameters must be corrected. This operation can be done by using some specific techniques to calibrate the VNA and then de-embed undesired passive data.

Figure 6-1 Principle of operation of a VNA.

The term “on-wafer measurement” refers to the measurement of integrated devices directly on the wafer, immediately after fabrication, before slicing and encapsulation. It is the most reliable method to characterize integrated devices, since a minimum of parasitics is involved. To carry out on-wafer measurements, the wafer is placed on a specific probe station, where mechanical arms allow tips to be connected to the device. The probes are connected to the VNA through coaxial cables, as shown in Figure 6-2. The probes are essentially transitions from a (large) coaxial cable structure to a (small) CPW structure. Figure 6-3 shows an on-wafer ground-signal-ground (GSG) probe and the probe connecting to a CPW line.

Figure 6-2 On-wafer probing station connected to a VNA.

Figure 6-3 (a) An on-wafer GSG probe. (The inset shows a microscopic view of the probe tips) (b) An artist's view of a probe (black) contacting a CPW structure (grey).

In order to perform accurate on-wafer measurements, an on-wafer calibration and de-embedding scheme is required. The technique called “two-stage calibration” [63] can be used for CMOS on-wafer measurement. It is described below.

(1) *On-wafer calibration*

On-wafer calibration is the first stage of the CMOS measurement process. It is used to determine the number of error coefficients caused by connecting cables that must be extracted by calibration. The on-wafer calibration must be done by using standards fabricated on the wafer [63]. There are three commonly used self-calibration methods: Short-Open-Load-Through (SOTL), Through-Reflect-Line (TRL), and Through-Reflect-Match (TRM). Appendix B gives more detail on these three calibration methods. The TRM method is not frequency-limited but this algorithm requires the measurement of *two perfectly matched loads* (one for

each port). In CMOS technologies, it is not possible to achieve the required match between loads. Indeed, variations close to 10% of the DC values of resistances made on a CMOS process have been observed. Therefore, TRM cannot be implemented on a silicon wafer. Instead, a commercial calibration substrate, made on an alumina substrate, having a load which could tolerate variations less than 0.1%, can be used for an accurate calibration up to 40 GHz [63]. For an alternative, the TRL method can be used. However, as the method is frequency-dependent, we have to know the reference impedance. In practice, to minimize the frequency limitation of the TRL, more TRL lines have to be built on a substrate calibration chip, which takes extra space. In this work, based on cascade configurations, a planar standard, known as CS-5, and the SOTL technique are used to estimate the effects of surrounding parasitic networks on a CMOS device, which can be either a CPW line or a CMOS transistor. After this calibration, the reference planes move to the end of the probe tips, and the reference impedance is equal to $50\ \Omega$.

(2) Passive de-embedding

When measuring a passive device, the pads, feeding lines and other components used to facilitate on-wafer probing can corrupt the measurements with unexpected errors. A passive de-embedding process determines and removes the probe-pad parasitics, which are introduced by the pads fabricated on the silicon wafer. There are a few techniques that can be used to de-embed the effects. In many cases, the TRL technique is used. The standards are built on-chip together with the DUT. There are two CPW lines, and a reflective device. The latter can be an open or a

short circuit. The CPW is designed to have a characteristic impedance of nearly $50\ \Omega$. Then the line length is chosen in order to have the most accurate measurement in the frequency band of interest. If the validity of the calibration must be extended to nearly the whole frequency band, multiple lines have to be used. The TRL structure is shown in Appendix B, Figure B-1. After this calibration, the reference planes are shifted to the middle of the through-line, or to the ports of the DUT under measurement.

When the measured devices have small dimensions, as do integrated submicron transistors and CPWs, it is possible to use methods other than the on-wafer TRL approach. One method, called *admittance corrections*, was first proposed by van Wijnjen for bipolar devices [64] and then enhanced by Fraser [65]. Its validity has been recently demonstrated by Vandamme [66] using three-step de-embedding for deep submicron MOSFETs up to 40 GHz. This method is based on an admittance equivalent circuit of the DUT and the probing structure. In order to extract the different parasitic elements, specific de-embedding structures are measured through open and short circuits. All these structures are designed in order to mimic the same parasitics. The different open and short circuits utilized for the calibration are assumed ideal. Therefore, this algorithm cannot be applied to large devices at high operation frequencies. In addition, there is a three-step open-short method introduced by Cho [67], a two-step technique presented by Koolen [68], and a one-step *ABCD* cascaded matrix method created by Heymann [69].

In theory, the three-step de-embedding can provide more accurate results since more parasitic components are considered. However, this method needs more chip space since four de-embedding structures need to be put on the chip together with the DUT. Also, it needs more testing steps and each step can introduce additional errors. On the other hand, Koolen's method may be not very precise but is simpler and more straightforward, and needs less chip space since only two de-embedding structures have to be put on the chip. More details about these de-embedding methods are given in Appendix B.

In this work, the CMOS varactors, CPWs, NLTLs, and their corresponding de-embedding standards were fabricated in a standard CMOS process and characterized up to 40 GHz. The CPW and varactor S -parameters were measured up to 40 GHz using an HP 8722ES vector network analyzer (VNA) and Picoprobe ground-signal-ground (GSG) probes. After investigating and comparing different kinds of calibration and de-embedding techniques, Koolen's technique was used to de-embed the extrinsic parameters caused by pads and interconnects.

6.2 Large-signal Time-domain Testing

Large-signal testing is critical for capturing the nonlinear behavior of NLTLs. Until now, measuring the nonlinear behavior of RF components was not easy. S -parameters were developed to analyze and model the linear behavior of RF devices in

the frequency domain. They have limitations and are defined only for small-signal linear systems. Large-signal time-domain measurement is more complicated than small-signal testing. A high-speed oscilloscope is required to capture fast-transition waveforms. Figure 6-2 shows the measurement setup used for conventional NLTL time-domain testing, where both input and output signals are sent to a high-speed oscilloscope in order to compare their rise and fall times. A 3-dB power splitter was used to equally divide the input power into two paths. The two dc bias-tees were used to set different bias voltages at the NLTL device. The amplitude of the input at the tested device has to be less than 3 volts since the maximum voltage for a CMOS 0.18- μm chip is 3.3 volts. The challenging aspects are producing high-speed pulse inputs and calibrating the signal propagation delays of the test instrument and connecting wires. Even more difficult is to find an oscilloscope for frequencies > 50 GHz, which can capture the rise and fall times, of the order of 10 to 20 picoseconds, of the testing pulse signals.

Figure 6-4 NLTL time-domain measurement setup using an oscilloscope.

How is one to produce a scattering or S -parameter equivalent that can handle large-signal nonlinear circuits? A few years ago, Root and his associates at Agilent Technologies Inc. started work on this problem. They have now developed the X -parameters concept [70]. Mathematically, X -parameters are rigorous supersets of S -parameters. But unlike S -parameters, X -parameters can be used to represent and analyze the nonlinear behavior of RF components in a much more robust and complete manner. X -parameters include the magnitude and phase characteristics of the device-

generated spectral components corresponding to distortion, which can include harmonics and inter-modulation products. Phase-normalized quantities and the harmonic superposition principle are used in X -parameter theory. More details are given in Appendix B.

X -parameters of a nonlinear device now can be measured by using the new Agilent Nonlinear Vector Network Analyzer (NVNA), which is capable of nonlinear calibration and measurements. The measured X -parameters can be characterized and extracted by the Agilent design software ADS2008 [71]. This new version of the ADS simulator includes an auto-configurable nonlinear frequency-domain simulation ‘PHD block’ component, which converts the measured datafile into a device-specific PHD block instance. Hence, it is possible to deduce NLTL large-signal outputs from frequency-domain measurements. The measured results are then converted into time-domain waveforms. The frequency-domain and time-domain responses of the measured nonlinear device can be obtained for further simulation and design. A simplified representation of the NVNA and its interoperability with ADS is shown in Figure 6-3. The related work has been published in [72, 73, 74].

Since nonlinear VNAs such as the Agilent N5242A are now available, large-signal nonlinearity testing has become more feasible and convenient. The nonlinear VNA can not only measure harmonic amplitudes and phases but also harmonic phase coupling information.

Some of the conventional small-signal measurements for the pulse-compression NLTLs have been carried out at a laboratory of the Communications Research Centre (CRC), Ottawa. The same laboratory is contacting Agilent to review a possibility of purchasing an NVNA system. In this work, the large-signal testing was completed at the CRC laboratory, since an N5242A-type NVNA was provided by Agilent Technologies on a “demonstrator” basis.

Figure 6-5 Simplified configuration of an NVNA and ADS workbench for nonlinear measurements.

6.3 NLTL Chip Designs and Measurement Results

In this work, two sets of NLTLs were designed. One set is an early design that neither used slow-wave CPW nor tapered-line techniques. The designed chip is shown in

Figure 6-6. The chip size is $1.0 \times 2.0 \text{ mm}^2$, the process being a standard CMOS $0.18\text{-}\mu\text{m}$ technology. The top snake-like line in Figure 6-6 is an SDB NLTL with a length of 4.02 mm , which includes 35 sections. The bottom snake-like line is an SDF NLTL with a length of 3.45 mm , which also has 35 sections.

Figure 6-6 Early NLTL designs that omit slow-wave CPW and tapered-line techniques.

The latest NLTL chip design is shown in Figure 6-7. The chip size is $1.5 \times 2.0 \text{ mm}^2$, and the process is again a standard CMOS $0.18\text{-}\mu\text{m}$ technology. The left-most line in Figure 6-7 is an SDB NLTL with a length of 1.06 mm . The second-to-left line in the figure is an SDF NLTL with a length of 0.86 mm . Both lines have 16 sections. Since both slow-wave CPW and tapered-line technologies are used, the NLTL lengths in the latest design are much shorter than in the earlier design.

Figure 6-7 The latest NLTL designs, which use both slow-wave CPW and tapered-line technologies.

The large-signal NLTL measurements were done using an NVNA system provided by Agilent Technologies Inc. The maximum frequency capacity of the N5242A system was 26 GHz. The input signal was a 5-GHz sinusoidal wave since the pulse input (or square-wave input) feature is still under the development by the same company. This means that at the output of an NLTL, the highest harmonic we can see is the 5th harmonic at 25 GHz. The maximum input available is 14 dBm. Since the equipment was available only on a “demonstrator” basis, the NLTLs were tested based on whatever features the equipment could provide. Some test results are shown below. In the figures for the NLTL *time-domain* measurement results, A1 indicates the input voltage wave, B2 the output voltage wave of the NLTLs, and B1 the reflected voltage wave. In the figures for the *frequency-domain* results, the markers, Mkr1, Mkr2, Mkr3,, identify the amplitudes of the 1st, 2nd, 3rd... harmonics.

- *Measurements on the most recent NLTL designs*

Figures 6-8 to 6-15 give test results for the latest NLTL designs. Figure 6-8 shows *time-domain* results for the SBD NLTL with a 5-GHz input at 13.0 dBm, which gives an input with 1.3 V-amplitude. Figure 6-9 gives *frequency-domain* results for the same device under the same conditions. The *time-domain* result shows that double-edge compression is happening and that the output B2 is squarer than the input. The *frequency-domain* result shows that more energy is accumulated in the higher harmonic components. For example, the fourth output harmonic is 41.7 dBm higher than the fourth input harmonic, as shown in Figure 6-9. This is first test case.

Figure 6-8 Measured SDB NLTL time-domain input (A1, yellow) and output (B2, blue) waveforms for a 13.0 dBm input at 5 GHz.

Figure 6-9 Measured frequency-domain SDB NLTL spectrum for a 13.0 dBm input at 5 GHz.

The second test case is shown in Figures 6-10 and 6-11. It is for the SDB NLTL at 13.5 dBm and 5 GHz input, which gives an input with 1.35 V-amplitude. The input power is slightly higher than in the first test case, and the double-edge compression effect is slightly stronger. The output is squarer than that of the first case. More important, the second output harmonic is 14 dBm higher than the second input harmonic, as shown in Figure 6-11. This contrasts with in the first case, where the second output harmonic is only 2 dBm higher than the second input harmonic, as seen in Figure 6-9.

Figure 6-10 Measured SDB NLTL input (A1, yellow) and output (B2, blue) time-domain waveforms for a 13.5 dBm input at 5 GHz.

Figure 6-11 Measured frequency-domain SDB NLTL spectrum for a 13.5 dBm input at 5 GHz.

The third test case is also for the SDB NLTL but with a 5-GHz at 13.9 dBm input, which gives a 1.6 V-amplitude input. The results are shown in Figures 6-12 and 6-13. The NLTL performed a better double-edge pulse compression than the first two test cases. The output waveform better resembled a square shape than the two other cases. Compared to the input harmonics, the second output harmonic increases by 14 dBm, and the third one increases by 16 dBm, as shown in Figure 6-13. This case indicates that the higher the bias voltage, the stronger the compression. This may not always be true. The reason is that the bias value determines the nonlinear capacitance value, which in turn decides the varactor cutoff frequency. The best compression happens only when the ratio of the Bragg frequency to the varactor cutoff frequency reaches certain values, as pointed out in Chapter 3.

Figure 6-12 Measured SDB NLTL input (A1, yellow) and output (B2, blue) time-domain waveforms for a 13.9 dBm input at 5 GHz.

Figure 6-13 Measured frequency-domain SDB NLTL spectrum for a 13.9 dBm input at 5 GHz.

Figure 6-14 and 6-15 are for the SDF NLTL. The input is a 3.0 GHz sinusoidal wave at 13.0 dBm. Looking at time-domain result in Figure 6-14, a very minor compression on the rise-edge of the output B2 can be seen. The power level of the higher harmonics also has a slight change, the output second, third, fourth and fifth harmonics are all higher than the input relative ones, as shown in Figure 6-15. Other input settings have also been tried, but, in general, the single-edge compression is not very impressive. This kind of result for the SDF NLTL had been expected. As pointed out in Chapter 4, the SDF (or IMOS) varactor has a larger channel resistance than an AMOS varactor, since in the SDF structure the well makes no contribution to the channel conductivity. The larger channel

resistance results in higher attenuation and the higher loss killed the nonlinearity. A better result might have been obtained if AMOS varactors had been used instead of SDF varactors.

Figure 6-14 Measured SDF NLTL input (A1, yellow), reflected (B1, purple), and output (B2, blue) time-domain waveforms for a 13.0 dBm input at 3 GHz.

Figure 6-15 Measured frequency-domain SDF NLTL spectrum for a 13.0 dBm input at 3 GHz.

- *Measurements on prototype NLTs*

Figure 6-16 and 6-17 are for the early SDF NLTL designs with input at 14.0 dBm and 4.5 GHz. Nothing is really happening with this design since the device attenuation is too high. Similar results were obtained with other input settings and with the early SDB NLTL designs.

Figure 6-16 Measured SDF NLTL input (A1, yellow), reflected (B1, blue), and output (B2, purple) time-domain waveforms for a 14.0 dBm input at 4.5 GHz.

Figure 6-17 Measured frequency-domain SDF NLTL spectrum for a 14.0 dBm input at 4.5 GHz.

To validate the large-signal test results, a few more extra simulations of SDB NLTLs were done using sinusoid-wave inputs instead of square-pulse-wave inputs. The same design data as that of the latest designed tapered slow-wave SDB NLTL were used in the simulations. With the frequency parameter set to 5GHz for all three cases, Figures 6-18, 6-19 and 6-20 show the results with the input sinusoid wave amplitude parameter set to 2.2, 2.3 and 1.9 V, respectively.

Figure 6-18 The SDB NLTL simulated result with 5 GHz sinusoid input, 2.2 V amplitude.

Figure 6-19 The SDB NLTL simulated result with 5 GHz sinusoid input, 2.3 V amplitude.

Figure 6-20 The SDB NLTL simulated result with 5 GHz sinusoid input, 1.9 V amplitude.

Figures 6-18, 6-19 and 6-20 show that strong double-edge compression happens at the 1.9 – 2.3 V input level. Since there was a limitation on input power (< 14 dBm)

during the large-signal testing, to match the test case, the amplitude of the sinusoid input in the simulation was reduced to 1.6 V. The simulated result is shown in Figure 6-21. The double-edge compression is still effective but not as strong as that with the higher input power.

Figure 6-21 The SDB NLTL simulated result with 5 GHz sinusoid input, 1.6 V amplitude.

Figure 6-22 is a re-constructed waveform of the output signal in Figure 6-21 by using only the 5 lowest output harmonics. With 5 GHz input, the highest harmonic the large-signal testing instrument NVNA can cover is the 5th. This part of the work demonstrates that the test results are reasonable, and that the SDB NLTL is indeed compressing the input signal. Figure 6-23 shows the spectrum of the input signal and the 5 lowest output harmonics. The dot-dash line is for input signal and the solid line for output signal. One can see that the waveform in Figure 6-22 is very similar to the

waveform in Figure 6-12, and the spectrum of the first five harmonics in Figure 6-23 is very similar to the spectrum in Figure 6-13.

Figure 6-22 Reconstructed waveform by using only 5 lowest output harmonics.

Figure 6-23 Spectrum of an input signal and 5 output harmonics.

6.4 Conclusion

This chapter discussed both NLTL small-signal and large-signal measurement strategies. For the small-signal testing, both calibrating the testing equipment and de-embedding of on-chip pad-parasitics are required. Different kinds of calibration and de-embedding technologies were presented, evaluated and compared. For the large-signal testing, a recently-developed harmonic-oriented technique equipped with nonlinear VNA was used. Instead of S -parameters, X -parameters were used to describe large-signal nonlinear behaviour. The theory of the X -parameter technique was presented, and the testing method was discussed in this chapter. A brief description of NLTL chip designs was also presented.

Finally, the NLTL large-signal testing results were presented. Simulations using sinusoidal inputs demonstrated that the large-signal testing was giving correct results.

This work strongly indicates that it should be possible to build excellent pulse-compression NLTLs using a low-resistivity commercial CMOS process.

Chapter 7 Summary and Future Directions

This work investigates the design of all-silicon pulse-compression NLTLs. A slow-wave CPW technology was developed to reduce the silicon loss. Two types of NLTLs were introduced. The SDF type, optimized for single-edge pulse sharpening, is able to reduce the rise time by 75%, while the SDB type, suited for double-edge pulse sharpening, is able to reduce both rise and fall times by 60%. The simulations are based on measured data from on-chip CPW and varactor components of the NLTLs. Five CMOS chips were developed and fabricated in this study. Some of them are for CMOS varactor research, some for silicon CPW research, and some for research on de-embedding techniques. More importantly, two sets of pulse-compression NLTLs were developed on the chips. Each set includes two NLTL lines, one for single-edge pulse compression, and one for double-edge pulse compression. One of these two sets of NLTLs was designed in the early stages of the study; in this case, both NLTL lines were more than 3 mm long since they were designed using neither tapered-line nor slow-wave technologies. The last two designs are much better, since their lengths are both shorter than 1 mm. This is because both tapered-line and slow-wave techniques were used. The large-signal behaviour of both sets of NLTLs was investigated and measured using nonlinear VNA instrumentation. Since the new nonlinear VNA instrumentation was still under development, the measurements were restricted to sinusoidal inputs and output frequencies up to only 26 GHz. That is, with 5 GHz input, only up to 5 harmonics could be measured. Even through the testing instrument had limitations, the testing results are still very promising. The most recently designed

double-edge compression NLTL showed a strong ‘squaring’ effect on an input sinusoidal wave. Almost certainly, better results could have been obtained if the higher harmonics could have been measured. The most recently designed single-edge compression NLTL showed very little sign of compression, due to its larger attenuation. Since this kind of NLTL uses the SDF varactor structure, in which the source and drain of the transistor are floating from the bulk, the well makes no contribution to the channel conductivity, resulting in a larger channel resistance. The large attenuation of the SDF structure decreases the nonlinearity in the NLTL circuits. The prototype NLTLs showed hardly any compression due to their longer length with larger attenuation, as was expected.

This work shows the feasibility of building all-silicon pulse-compression NLTLs using commercial silicon foundry processes. This study has built up rich experience and knowledge on the design of low-loss low-cost pulse-compression NLTLs.

This research work has already produced the following publications:

[1] Ming Li, Rony E. Amaya, Jean-Marc Duchamp, Philippe Ferrari, Robert G. Harrison, and N. Garry Tarr, "Low-Loss Low-Cost All-Silicon CMOS NLTLs for Pulse Compression," *IEEE MTT-S International Microwave Symposium*, pp. 449-452, Honolulu, Hawaii, June, 2007.

[2] Ming Li, Robert G. Harrison, Rony E. Amaya, Jean-Marc Duchamp, Philippe Ferrari, and N. Garry Tarr, "CMOS Varactors in NLTL Pulse-Compression Applications," *Proceedings of the 37th European Microwave Conference*, pp. 1405-1408, Munich, Germany, Oct., 2007.

[3] Rony E. Amaya, Ming Li, Robert G. Harrison, and N. Garry Tarr, "Coplanar Waveguides in Silicon with Low Attenuation and Slow Wave Reduction," *Proceedings of the 37th European Microwave Conference*, pp. 508-511, Munich, Germany, Oct., 2007.

[4] Ming Li, Rony E. Amaya, Robert G. Harrison, and N. Garry Tarr, "Investigation of CMOS Varactors for High-GHz-Range Applications," Accepted by the journal *Research Letters in Electronics*, May 2009.

Additional research publications are expected as direct results of this work, especially on the nonlinear results of large-signal measurements.

Further research work following this study could include:

1. Analytical work on the slow-wave propagation properties of all-silicon NLTLs

The present study has shown that the slow-wave CPW technique has better conversion efficiency at higher frequencies. This is because that the floating

ground in the slow-wave structure can reduce the substrate loss, which is important because the substrate loss is proportional to the operating frequencies. It will be very interesting and meaningful to investigate the slow-wave CPW behaviour at higher frequencies, such as over 100 GHz. Another beneficial study on slow-wave CPWs would focus on their secondary Bragg frequencies. Since the floating ground introduces a secondary periodic structure into the system, there will be a secondary Bragg frequency. This secondary Bragg frequency is higher than the primary Bragg frequency, and may have an impact on the dispersion and primary Bragg frequency of an NLTL. It is anticipated that further study will discover the relationship among these propagation properties.

2. *Developing new NLTL structures using AMOS and short-gate SnDp varactors*

The study has shown that the AMOS varactor is better than the IMOS (SDF) varactor for single-edge pulse compression because of its reduced attenuation and its lower degree of $C(V)$ -degeneration at high frequencies. For the same reasons, the short-gate SnDp varactor is better than NMOS (SDB) varactor for double-edge pulse compression. The SDF and SDB varactors were used in the NLTL designs in this study because they were available components in the CMOS process. With the fast-developing pace of CMOS technology, it will be possible in the near future to build AMOS and SnDp NLTLs, and it is expected that they will provide better pulse-compression capabilities.

Appendix A: Propagation Characteristics of CPW Lines

According to [39] and [42], without making any quasi-static assumptions, when a signal represented by

$$v(t, z) = e^{-\alpha z} e^{j(\omega t - \beta z)} \quad (\text{A-1})$$

propagates through the CPW shown in Figure A-1, the phase constant β is given by (A-2).

$$\beta(f) = 2\pi \cdot \frac{f}{c} \cdot \sqrt{\epsilon_{\text{eff}}(f)} \quad (\text{A-2})$$

Figure A-1 Cross-section view of a CPW.

In (A-1) α is the attenuation and in (A-2) c is the vacuum speed of light.

Zero dispersion occurs when the phase constant β is a linear function of frequency. The phase constant is a function of the effective permittivity ϵ_{eff} . If ϵ_{eff} varies with the frequency, there is dispersion in the CPW line. For this structure, ϵ_{eff} is given by (A-3) [42]

$$\sqrt{\epsilon_{eff}(f)} = \sqrt{\epsilon_q} + \frac{\epsilon_r - \epsilon_q}{1 - n \cdot \left(\frac{f}{f_{te}} \right)^{-b}}, \quad (A-3)$$

where $b = 1.8$, ϵ_r is the relative permittivity of the substrate, $\epsilon_q = \frac{\epsilon_r + 1}{2}$ is the quasi-static effective permittivity, and $f_{te} = \frac{c}{4h\sqrt{\epsilon_r - 1}}$ is the surface wave TE₁₀-mode cut-off frequency. The factor n can be computed from [42]

$$\log(n) = u \cdot \log(2a / (b - a)) + v$$

$$u = 0.54 - 0.64 \cdot ql + 0.015 \cdot ql^2$$

$$v = 0.43 - 0.86 \cdot ql + 0.54 \cdot ql^2$$

$$ql = \log(a / h)$$

where a , b , and h are defined in Figure A-1.

The characteristic impedance is given by

$$Z_{cpw} = \frac{30\pi}{\sqrt{\epsilon_{eff}}} \cdot \frac{K(k')}{K(k)} \quad , \quad (\text{A-4})$$

where $K(k)$ is the complete elliptic integral of the first kind, with $k = a/b$ and $k' = \sqrt{1 - k^2}$.

Appendix B: Calibration Methods for CMOS small-signal measurements

(1) Three On-wafer Calibration Methods

The three commonly used on-wafer calibration methods are the “Through-Reflect-Line” (TRL), “Through-Reflect-Match” (TRM), and “Short-Open-Load-Through (SOLT) techniques. They can be used with on-wafer coplanar-waveguide standards to calibrate measurement systems up to the probe tips. Figure B-1 shows the structures of these three calibration standards [69].

Here is brief summary of the three methods [75]:

- *The TRL method*

The *TRL* method uses multiple transmission lines as standards. These transmission lines are built on the same wafer as the DUT. The measured *S*-parameters use the *TRL* line Z_0 as the reference impedance. After calibration, the reference plane is precisely set at the center of the ‘Thru’ line. But this method requires multiple probe spacings, depending on the line impedances, and is therefore not suitable for fixed-spacing probes. The *TRL* method is a frequency-dependent method. If one needs to extend the testing frequency range, multiple lines have to be put on the chip, which takes more chip space.

Figure B-1 Three calibration structures on coplanar waveguide standards, (a) TRL, (b) TRM, and (c) SOLT.

- *The TRM Method*

Different from *TRL* method, the *TRM* method is a frequency-independent method. It is good for broadband calibrations and measurements. This method uses the matching resistance Z_{match} of the *TRM* ‘Match’ structure as the reference impedance. To have a precise measurement, two match standards, one for each DUT port, are required. The *TRM* method is not suitable for CMOS chip, since process variations make it difficult to build two identical match standards. Besides the *S*-parameters of the *TRM* standards, the ‘Thru’ line delay should also be known for the calibration. Fixed probe spacings can be used with the *TRM* calibration.

- The *SOLT* Method

The *SOLT* method is the most popular method and available on virtually every vector network analyzer. This calibration method performs reasonably well as long as accurate models of calibration standards are determined. The *SOLT* standards take less chip space than *TRL* and *TRM* methods, and can be used for broadband measurements. The negative part of this method is that the measurement results are very sensitive to probe placement. Bad probe positions can introduce negative capacitance to the measurement of the ‘Open’ standard, extra inductance to the measurements of the ‘Short’ and ‘Load’ standards, and cause unpredictable measured results.

The calibration algorithms are available in references [63], [76], [77].

(2) Pad Parasitic De-Embedding Methods

(a) Vandamme's three-step de-embedding Method

This de-embedding algorithm is based on an equivalent circuit obtained from RF measurements. The equivalent circuit is shown in Figure B-2 (a). Figure B-2 (b) gives four de-embedding structures which need to be built on the chip together with the device under test (DUT).

Figure B-2 (a) Equivalent circuit of the RF test structure. (b) Equivalent circuits of the open, short1, short2, and through de-embedding structures.

Measuring the four de-embedding structures give four sets of S -parameters which are used to determine the six unknown parasitic components, Z_1 , Z_2 , Z_3 , G_1 , G_2 , and G_3 . The three-step de-embedding process is given in (B-1), (B-2), and (B-3).

$$Y_A = Y_{measur} - \begin{bmatrix} G_1 & 0 \\ 0 & G_2 \end{bmatrix} \quad (B-1)$$

$$Z_B = Z_A - \begin{bmatrix} Z_1 + Z_3 & Z_3 \\ Z_3 & Z_2 + Z_3 \end{bmatrix} \quad (B-2)$$

$$Y_{DUT} = Y_B - \begin{bmatrix} G_3 & -G_3 \\ -G_3 & G_3 \end{bmatrix} \quad (B-3)$$

The first step corrects for the influence of the admittances, G_1 and G_2 . The second step de-embeds the effects of the series impedances, Z_1 , Z_2 , and Z_3 . The third step removes the coupling G_3 between the two ports [66].

(b) Koolen's one-step de-embedding Method

The equivalent circuit of Koolen's de-embedding method is shown in Figure B-3.

Figure B-3 Koolen's equivalent circuit of the RF device test structure.

In this method, only two de-embedding structures need to be built on the chip with the DUT: one is an open circuit; the other is a short circuit. The one-step de-embedding algorithm is given in (B-4). We can see that there is no need to determine the parasitic components, Z_1 , Z_2 , G_1 , G_2 , and G_3 . The (B-4) formula only deals with Y -parameters directly from the measurements [68].

$$Y_{DUT} = \frac{(Y_{total} - Y_{open}) \cdot (Y_{short} - Y_{open})}{Y_{short} - Y_{total}} \quad (\text{B-4})$$

Appendix C: X-parameters

The theory of X -parameters is based on scattering functions, as are the S -parameters. But X -parameters are very different from S -parameters since they describe a nonlinear relationship in the frequency domain. The word ‘scattering’ refers to the fact that traveling-wave quantities are used in the theory. Here, ‘traveling voltage waves’ are represented by linear combinations of the signal port voltages and currents as defined in equation (C-1),

$$\begin{pmatrix} A \\ B \end{pmatrix} = \begin{pmatrix} \frac{V + ZI}{2} \\ \frac{V - ZI}{2} \end{pmatrix}. \quad (\text{C-1})$$

The incident wave is represented by a capital A and the scattered or reflected wave by a capital B . V refers to the port voltage and I to the port current, and Z to a chosen characteristic impedance.

Figure C-1 Scattering notation for X-parameters.

Figure C-1 gives more explanation of the theory. For the incident and reflected wave quantities, A_{1k} , A_{2k} , B_{1k} , and B_{2k} , the first subscript refers to the signal port and the second subscript to the harmonic index. Note that a harmonic index of 1 represents the fundamental component, a harmonic index of 2 refers the second harmonic, and so on. Using a general frequency-domain black-box modeling technique, the output spectral components, B_{1k} , and B_{2k} , correlate with the input spectral components, A_{1k} and A_{2k} , and can be represented by the functions below:

$$B_{1k} = F_{1k}(A_{11}, A_{12}, \dots, A_{21}, A_{22}, \dots) \quad (C-2)$$

$$B_{2k} = F_{2k}(A_{11}, A_{12}, \dots, A_{21}, A_{22}, \dots) \quad (C-3)$$

where the letter F denotes a correlation function.

In order to simplify the mathematics, phase-normalized quantities are used. That is, A_{11} (the incident fundamental) is used as the phase-reference component. Since A_{11} is the dominant large-signal input component, all other input components (at the harmonic frequencies) are relatively small compared to this dominant component. The mathematics of the phase normalization is quite simple. First, a component P is defined such as that it has a phase equal to the phase of A_{11} and unit amplitude of a unit:

$$P = e^{jp(A_{11})} \quad (\text{C-4})$$

where $p(A_{11})$ represents the phase of A_{11} .

The normalized quantities, denoted by the superscript N are then calculated by multiplying the wave quantities by the reciprocal of P raised to the power k , which corresponds to the harmonic index of that component.

$$A_{mk}^N = A_{mk} P^{-k} \quad \text{and} \quad B_{mk}^N = B_{mk} P^{-k} \quad (\text{C-5})$$

Note that the phase-normalized quantity A_{11}^N is a positive real number, that is $A_{11}^N = |A_{11}|$.

In general, the superposition principle and linear relationships are inapplicable to large-signal nonlinear systems. However, in many practical cases (e.g. narrow-band power amplifiers) there is only one dominant large-signal input component (A_{11}) whereas all other input components are relatively small. In that case, it is valid to use the superposition principle for the relatively small phase-normalized input components. This is called the ‘harmonic superposition’ principle.

The harmonic superposition principle is graphically illustrated in Figure C-2.

Figure C-2 The harmonic superposition principle.

In order to keep the graph simple only A_1 and B_1 are considered. First, assuming that A_{11} is not zero, there will be a significant harmonic contribution to B_2 . Now suppose that a relatively small harmonic component A_{12} is present at the input. This will result in a change in the output spectrum B_2 at the second harmonic. The same principle holds for the third and fourth harmonics, and so on. The overall deviation of the output spectrum B_2 is the superposition of all the individual deviations. The ‘harmonic superposition principle’ is the key to the definition of the X -parameters. This principle leads to a very simple mathematical formulation of X -parameters, as shown in (C-6) [70].

$$B_{mk}^N = \sum_{nk} X_{mknk} (A_{11}^N) A_{nk}^N + X'_{mknk} (A_{11}^N) A_{nk}^{N*} \quad (\text{C-6})$$

The phase-normalized B^N waves are the output quantities. They are written as a linear combination of the phase-normalized A^N waves, the input quantities, and their conjugates (denoted by the superscript $*$). Each X -function is defined by four subscripts: the first two denote the port and harmonic index of the associated output

component B , and the latter two describe the port and harmonic index of the associated input component A .

The X -functions have some unique features compared to S -parameters. First, they relate input and output spectral components together at different frequencies. Second, they describe the fact that the ratio between the scattered B^N and incident A^N waves depends on the phase of the A components relative to that of A_{11} . In general, X -parameters are nonlinear functions of the input signal power and can map incident signal at specific frequencies to other frequencies. Since X -parameters include many cross-frequency dependencies, there are more X -parameters than S -parameters. More information on X -parameters and the related large-signal testing are available in references [72], [73], and [74].

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