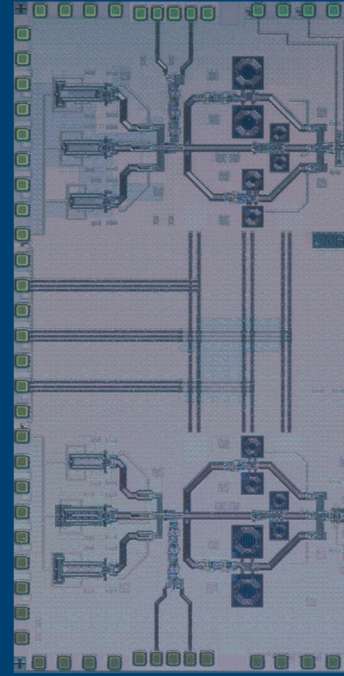
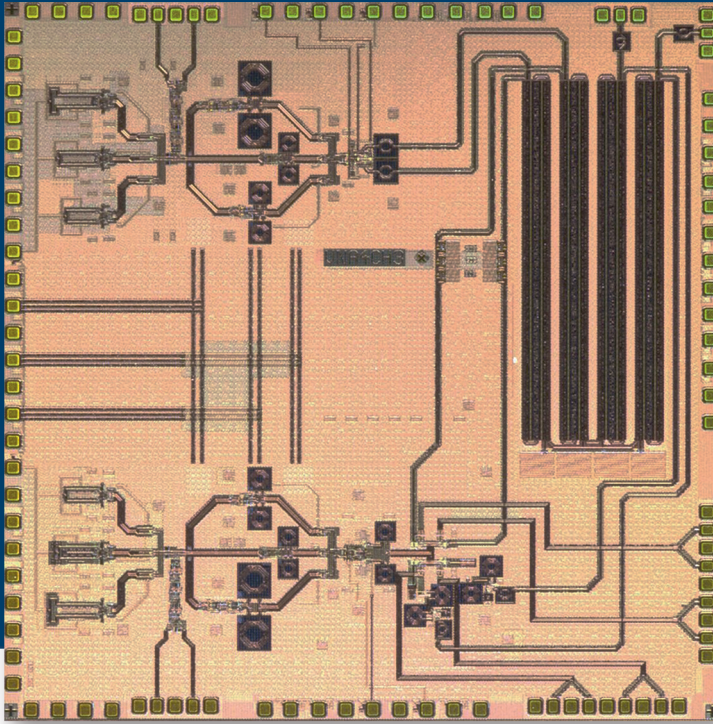




FAU Forschungen, Reihe B, Medizin, Naturwissenschaft, Technik 15

Johannes Nehring

Highly Integrated Microwave Vector Network
Analysis Circuits and Systems
for Instrumentation and Sensing Applications

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Highly Integrated Microwave Vector Network
Analysis Circuits and Systems for
Instrumentation and Sensing Applications

Mikrowellschaltungen und -systeme für die
hochintegrierte vektorielle Netzwerkanalyse
zur Anwendung in der
Messtechnik und Sensorik

Der Technischen Fakultät
der Friedrich-Alexander-Universität Erlangen-Nürnberg
zur Erlangung des Grades
DOKTOR-INGENIEUR

vorgelegt von
Johannes Nehring
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To Steffi.

Kurzfassung

Die vorliegende Arbeit befasst sich mit dem Thema der integrierten vektoriellen Netzwerkanalyse. Das Messverfahren der Netzwerkanalyse ermöglichte nicht nur in der Vergangenheit bedeutende technologische Fortschritte in der Hochfrequenztechnik sondern wird auch zukünftig aufgrund seines hohen Potenzials in messtechnischen und biomedizinischen Applikationen von zentraler Bedeutung sein. Allerdings ist die Anwendung von Netzwerkanalysatoren in vielen Szenarien aufgrund der Größe, des Gewichts und der Kosten der Geräte nur eingeschränkt möglich. Aus diesem Grund werden im Rahmen dieser Arbeit die Machbarkeit und das Potenzial des Konzepts *VNA-on-chip* untersucht. Zu diesem Zweck werden theoretische und praktische Studien anhand schmalbandiger, vektorieller Reflektometer mit unterschiedlichen Graden der Integration durchgeführt. Nach der erfolgreichen Bestätigung des Konzepts wird die Entwicklung eines Chipsatzes für einen integrierten 2-Tor-Netzwerkanalysator von 4 GHz bis 32 GHz in einer SiGe-Technologie erläutert. Dieser Chipsatz dient der Realisierung eines heterodyn Netzwerkanalysators und besteht aus einem Signalgenerator (Stimulus) mit integriertem Schalter und einem 4-Kanal-Empfänger. Im Weiteren wird ein voll integriertes RF-Frontend inklusive der integrierten Richtkoppeler präsentiert. In Kombination mit einem modularen Demonstratorsystem dient der Aufbau als ein miniaturisierter, vektorieller 2-Tor Netzwerkanalysator mit einer Bandbreite von 4 GHz bis 32 GHz. Die damit gemessene 2×2 -Streumatrix stimmt hervorragend mit Messungen überein, die mit kommerziell erhältlichen Geräten ermittelt wurden. Die Messungen sind stabil wiederholbar und der transiente Drift ist vernachlässigbar. Darüber hinaus wurden hoch integrierte Netzwerkanalysatoren in biomedizinisch relevanten Szenarien eingesetzt. Zu diesem Zweck wurde ein auf Reflexionsmessung basierender Analysator mit einem externen Sensor und ein auf Transmission basierendes Instrument mit einem integrierten Sensor implementiert. Beide Ansätze werden mit einem alternativen Messsystem verglichen.

Abstract

This thesis addresses the topic of highly integrated vector network analysis. Due to its significant contribution to technological advances in the past and its presently high potential regarding instrumentation and biomedical sensing applications, the technique of network analysis plays a key role in microwave and wireless engineering. However, the application of network analyzers is frequently limited due to the size, weight and cost of the instruments. For this reason, the potential and feasibility of the VNA-on-chip is evaluated. Therefore, theoretical and practical studies were performed by means of narrow-band microwave reflectometers with different degrees of integration. After a successful proof of concept, a chipset dedicated to a highly integrated 4-32 GHz 2-port vector network analyzer is developed in silicon germanium (SiGe) technology. The chipset can be utilized for the implementation of a heterodyne network analyzer and features a stimulus signal generator with integrated switch and a four-channel receiver chip. In a further step, a fully integrated RF frontend is demonstrated, including on-chip directive elements. In combination with a modular hardware demonstrator, the assembly serves as a miniaturized 2-port vector network analyzer from 4 GHz to 32 GHz. The measured 2×2 S-matrices show an excellent agreement with the results obtained by commercial measurement equipment. The measurements are highly repeatable and the drift over time is negligible. Furthermore, highly integrated network analyzers were applied to biomedical scenarios. For this purpose, a reflection based analyzer with external probe as well as a transmission based one with an on-chip sensor are presented and compared to an alternative sensing system.

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Acronyms

ADC	analog digital converter
DAC	digital analog converter
DDS	direct digital synthesis
DUT	device under test
EVM	error vector magnitude
FFT	Fast Fourier transform
IC	integrated circuit
IF	intermediate frequency
IL	insertion loss
LNA	low noise amplifier
LO	local oscillator
MUT	material under test
PCB	printed circuit board
PLL	phase locked loop
RBW	resolution bandwidth
RF	radio frequency
SiGe	silicon germanium
SNR	signal to noise ratio
SPDT	single pole double throw
VCO	voltage controlled oscillator
VGA	variable gain amplifier
VNA	vector network analyzer
VSWR	voltage standing wave ratio

Chapter 1

Introduction

1.1 Motivation

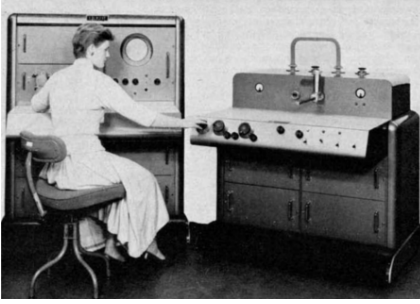
Without any doubt, the measurement technique of vector network analysis played a significant role in the evolution of radio frequency (RF), microwave and wireless industries. Since the 1950s, advances of the network analyzer technology go along with massive breakthroughs in the design of active and passive high frequency devices and circuits. The importance of vector network analyzers (VNAs) in RF and microwave engineering is indisputable. Nowadays, network analyzers are essential instruments of research and development laboratories dealing with high frequencies. In addition, they are key instruments for manufacturing and test. Beyond this, VNAs are increasingly coming into the focus of microwave based sensing applications in biomedical scenarios.

The importance of network analysis can be pointed out by means of its historical evolution, illustrated by the timeline in Fig. 1.1. It shows that the steady advancements are driven by the ambition to create more useful and more efficient instruments that would simplify the process of designing high frequency components and systems. The earliest approaches to measure reflection coefficients made use of slotted lines which allowed the determination of a voltage standing wave ratio (VSWR). Evidently, this technique demanded tedious manual measurements and calculations. The first device allowing the direct measurement of S-parameters without manual calculations was the Z-g-diagraph invented by Rohde & Schwarz in 1950. It was the first instrument, which can be regarded as a VNA even though the term

- 1930s • VSWR measurements by means of slotted lines
- 1950 • Dr. Rohde and Dr. Schwarz invented the Z-g-diagraph
- 1958 • Elliot Automatic Swept-Frequency Impedance Meter (A.S.F.I.M.)
- 1960s • Scattering parameters became popular
- 1965 • Wiltron 310 network analyzer
- 1966 • HP 8405 vector voltmeter
- 1967 • HP 8410 VNA up to 12.4 GHz
- 1968 • HP 8450 experimental system including error correction
- 1970s • First VNAs with included source (GR 1710, HP8505)
- 1974 • Invention of the 12-term calibration model
- 1980s • Introduction of computer aided design techniques
- 1984 • HP 8510 including error correction, time domain and pulsed measurements
- 1990s • Wireless boom: VNAs became essential for manufacturing and test
- 2001 • 4-port VNA (E5071A) for differential measurements
- 2000s • Electronic calibration, large signal and noise measurements, etc.
- 2020s • VNA on a chip?

Figure 1.1: Historical milestones in the development of network analyzers.

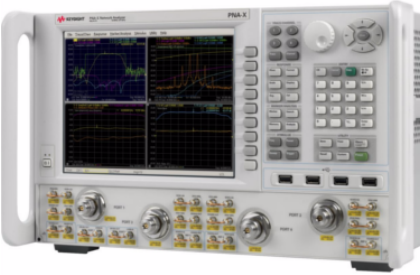
did not exist at that time. From this point on, the trend was towards further measurement automation as with the 1958 presented Automatic Swept-Frequency Impedance Meter (A.S.F.I.M.) by Elliot Brothers Ltd. This extensively large instrument facilitated the measurement of reflection and transmission properties of a test object [3]. Despite the fact that the concept of scattering parameters was already known since 1947 [4], their usage became popular in the 1960s [5]. In 1965, the Wiltron 310 network analyzer firstly introduced real-time operation over octave bands and provided the highest achievable measurement accuracy. However, the concept of error correction, which is nowadays commonly referred to as calibration, was not yet invented. The Hewlett-Packard HP 8405 vector voltmeter (1966) was fundamental for later VNAs. In combination with directional couplers and an external signal source it could be utilized for S-parameter measurements up to 1 GHz. In 1967, the HP 8410 featuring a polar display extended the frequency range up to 12.4 GHz. Compared to previous systems it showed a higher degree of integration but still it was composed by three separate functional boxes. In 1970, the HP 8542 covering a frequency range up to 18 GHz was introduced. It was a bulky three-rack measurement instrument and made use of the emerging computer technology. At that time, the 12-term calibration model was published as well [6]. In the 1970s, a more compact design was targeted, which



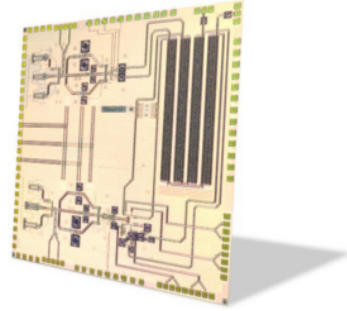
(a) 1958. [1]



(b) 1970. [1]



(c) 2016. [2]



(d) Future vision.

Figure 1.2: Evolution of vector network analyzers: Impedance meter from 1958 (a), HP8542 from 1970 (b), Keysight PNA-X (c), Future vision: VNA on a chip (d).

was accomplished by integrating the signal sources into the analyzer. An important milestone was set by the introduction of the famous HP 8510 vector network analyzer reaching up to 26.5 GHz. It was the first instrument system with included error correction, time domain reflectometry and the ability to perform pulsed measurements. Thus, it became an industry standard for a long time. In the 1980s, the measured scattering parameters served as input for emerging computer aided design software. This novel design flow pushed microwave technology further ahead and made the VNA an even more powerful instrument. The wireless boom in the 1990s made network analyzers to essential instruments not only in research and development but also in manufacturing and test. At that time, the requirements for measurement speed became much more stringent. In 2001, the first 4-port analyzer (E50701A) came to the market facilitating differential characterization of devices under test (DUTs). Later on, the performance of VNAs was steadily improved and the functionality was vastly extended. Modern analyzers offer multiple ports, support noise measurements and can be applied for the

characterization of nonlinear devices. With the aid of frequency extension modules, a bandwidth of far more than 100 GHz is achieved. Error vector calibration is often accomplished by an electronic calibration module allowing a fast and comfortable procedure. [1, 7, 8]

Despite the impressive evolution of network analyzers since the 1950s, there is still a huge potential for further improvements and developments. Even though, the VNA is a universal measurement instrument, its application is frequently limited to stationary scenarios by its bulky size, heavy weight and its extensive cost. A more compact, cheap and portable device with comparable measurement accuracy and functionality would be the enabler of numerous innovative applications. Not only could a miniaturized VNA be used for classical instrumentation purposes, but it could also pave the way towards mass-market applications of emerging microwave biomedical sensors. In the latter, VNAs are frequently used for the characterization of novel sensor types. This doctoral thesis addresses the further miniaturization and the integration of the VNA on a chip. The evolution of network analyzers and the future vision of the VNA on a chip is illustrated in Fig. 1.2.

1.2 Goal and Structure of this Work

The goal of the research conducted in the context of this thesis is to demonstrate the feasibility and potential of highly integrated, multiport, microwave vector network analyzers with multi-octave bandwidth. In addition, the application of a miniaturized VNA to instrumentation as well as to biomedical sensing scenarios shall be demonstrated. For this purpose, theoretical and practical studies are conducted. In distinction to several other research activities in this field, the measurement of true scattering parameters, a high degree of integration and self-sustained operation are within the main scope of this work.

This thesis is structured as follows. In Chapter 2, the basic concepts of vector network analysis are summarized and VNA topologies as well as the required components are discussed regarding their suitability for integration in silicon. Chapter 3 is dedicated to the realization of narrow-band microwave reflectometers. Being the simplest form of VNAs, fundamental studies are performed by means of the presented circuits leading to important conclusions for the design of a 2-port multi-octave bandwidth microwave VNA. Chapter 4 is dedicated to the circuit and system design of a chipset for a 4-32 GHz 2-port network analyzer in SiGe technology, which is fundamental for the implementation of the highly integrated VNA described in Chapter 5. The latter includes a hardware demonstrator, calibration, characterization

of several DUTs by their scattering parameters and a detailed performance investigation. Chapter 6 describes microwave dielectric sensing systems for biomedical scenarios, which are based on integrated network analyzers. In addition, an alternative approach relying on a permittivity sensitive phase locked loop (PLL) is presented. Finally, the thesis is concluded and an outlook for future work is given.

The single chapters of this thesis are arranged in a manner, so that they can be studied independently of each other. Consequently, this occasionally leads to minor repetitions of content. Nevertheless, the chapters build up on each other consecutively and the author recommends to enjoy this thesis from the first to the last page.

1.3 State of the Art

Hereinafter, the state of the art regarding miniaturized VNAs will be described. The current status of commercially available benchtop instruments is explicitly outside the scope of the following.

Up to now, a limited number of publications was addressing the topic of miniaturized network analysis and only few of them followed an integrated approach. The six-port receiver architecture was utilized to implement vector reflectometers (1-port VNA) for millimeter waves [9, 10]. Integrated realizations [11, 12] and discrete realizations [13] with either relative or absolute reflection coefficient measurements were presented. Furthermore, phase-shifter based sampled standing wave reflectometers were shown as printed circuit board (PCB) assemblies [14, 15] and based on integrated circuits [16]. A direct-conversion concept with separate phase and magnitude detection was proposed in [17]. Quasi-optical 2-port VNAs with antennas acting as measurement ports were used for the characterization of frequency selective surfaces [18]. The technique of nonlinear transmission lines was exploited in order to realize network analyzer testsets with multi-octave bandwidth using external signal generators [19–21] in order to extend the frequency range of conventional VNAs. Inherently broadband approaches based on pseudo-noise devices were proposed in [22, 23]. A single-channel CMOS receiver utilized for the measurement of one scattering parameter up to 3 GHz is presented in [24]. An extended version incorporating a transceiver for biomedical sensor characterization up to 2.4 GHz is described in [25]. One of the first integrated heterodyne 1-port VNAs for the determination of scattering parameters was shown in [26, 27] by the author. The first chipset for a highly integrated 2-port heterodyne VNA with multi-octave bandwidth 4 GHz–32 GHz was proposed by the author [28]. In [29], a millimeter wave

reflectometer with octave bandwidth and a fully integrated multiport analyzer for on-chip sensor characterization was proposed under contribution of the author.

To the best of the author's knowledge, this thesis presents the first full 2-port highly integrated VNA covering a multi-octave bandwidth from 4 GHz to 32 GHz. Up to now, the highest degree of integration is achieved by a fully integrated VNA RF frontend including the signal source, receiver and the testset. The first broadband directional coupler on a chip based on cascaded quarter-wavelength coupled line segments is presented. The measurement of absolute scattering parameters based on a valid error correction is novel in the combination with the covered bandwidth, number of ports and degree of integration of a miniaturized VNA.

Chapter 2

Basics of Integrated Vector Network Analyzers

In the following, the basic concepts of vector network analysis shall be summarized briefly. In addition, network analyzer topologies and components are discussed regarding their suitability for integration in silicon. Finally, the semiconductor technology used for this thesis is described.

2.1 Scattering Parameters

At low frequencies, impedance (Z) or admittance (Y) parameters are used in order to describe N-port networks. The impedance matrix Z represents the relation between the voltages ($V_1 \dots V_N$) and currents ($I_1 \dots I_N$) at the terminals of the device:

$$\begin{pmatrix} V_1 \\ V_2 \\ \vdots \\ V_N \end{pmatrix} = \begin{pmatrix} Z_{11} & Z_{12} & \cdots & Z_{1N} \\ Z_{21} & Z_{22} & \cdots & Z_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ Z_{N1} & Z_{N2} & \cdots & Z_{NN} \end{pmatrix} \cdot \begin{pmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{pmatrix} \quad (2.1)$$

Admittance matrices are defined accordingly. Consequently, the single elements Z_{ij} with $i, j = 1 \dots N$ are given by the ratio of the voltage V_i measured at port i and the current I_j injected into port j under the condition that all other ports are left open ($I_k = 0$ for $k \neq j$):

$$Z_{ij} = \left. \frac{V_i}{I_j} \right|_{I_k=0 \text{ for } k \neq j} \quad (2.2)$$

The determination of the admittance parameters is performed in the according manner, resulting in the condition $V_k = 0$ for $k \neq j$, which stands for short circuits at all ports besides the stimulating one. At low frequencies, ideal open and short circuits can be realized so that Z- and Y-parameters can be determined directly by measurements. At high frequencies and especially in the microwave region, ideal short and open standards do not exist. Whereas a short circuit shows significant series inductances, an open circuit is shunted by parasitic capacitances. Thus, another concept for the measurement of network parameters has to be applied.

For this reason, the scattering parameters (S-parameters) were introduced. In contrast to Z-parameters, the S-parameters describe the relation between forward ($V_{1...N}^+$) and backward ($V_{1...N}^-$) voltage waves of a N-port network:

$$\begin{pmatrix} V_1^- \\ V_2^- \\ \vdots \\ V_N^- \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} & \cdots & S_{1N} \\ S_{21} & S_{22} & \cdots & S_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1} & S_{N2} & \cdots & S_{NN} \end{pmatrix} \cdot \begin{pmatrix} V_1^+ \\ V_2^+ \\ \vdots \\ V_N^+ \end{pmatrix} \quad (2.3)$$

Consequently, the single elements of the scattering matrix are given by the ratio of the outgoing voltage wave V_i^- at port i and the incoming voltage wave V_j^+ at port j under the condition that all other ports are terminated and there are no incoming waves ($V_k^+ = 0$ for $k \neq j$):

$$S_{ij} = \left. \frac{V_i^-}{V_j^+} \right|_{V_k^+ = 0 \text{ for } k \neq j} \quad (2.4)$$

This definition of S-parameters assumes the termination of all ports with equal impedances, in general the system impedance Z_0 . In most practical cases, this holds true. However, a more generalized definition of scattering parameters is more common and widely spread. It introduces the wave parameters

$$\begin{aligned} a_n &= \frac{V_n^+}{\sqrt{Z_{0n}}} \\ b_n &= \frac{V_n^-}{\sqrt{Z_{0n}}} \end{aligned} \quad (2.5)$$

for incident (a_n) and reflected waves (b_n), which represent the ratio of the voltage waves and the termination impedances Z_{0n} at the corresponding ports. The generalized scattering matrix relates the incident and reflected

waves

$$\begin{pmatrix} b_1 \\ b_2 \\ \vdots \\ b_N \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} & \cdots & S_{1N} \\ S_{21} & S_{22} & \cdots & S_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ S_{N1} & S_{N2} & \cdots & S_{NN} \end{pmatrix} \cdot \begin{pmatrix} a_1 \\ a_2 \\ \vdots \\ a_N \end{pmatrix} \quad (2.6)$$

whereas the single matrix elements can be determined by

$$S_{ij} = \left. \frac{b_i}{a_j} \right|_{a_k=0 \text{ for } k \neq j} \quad (2.7)$$

under the condition that all ports are terminated with Z_{0n} and that there are no incident waves besides a_j . The typical port impedance of 50Ω can be realized even at microwave frequencies so that the direct measurement is possible. The measurement itself is usually performed by a VNA. [30]

2.2 Integrated Network Analyzer Building Blocks

In the following, the functional principle of a VNA will be described by means of its three main components. These are a stimulus, a testset and a receiver. The VNA functionality is illustrated in Fig. 2.1 with the aid of the latter components.

2.2.1 Stimulus

In contrast to many other instrumentation devices, VNAs are equipped with a built-in signal source acting as a stimulus for the DUT. This stimulus is required in order to characterize a DUT at a certain frequency or over a bandwidth, so that its frequency response can be determined. In most cases, the

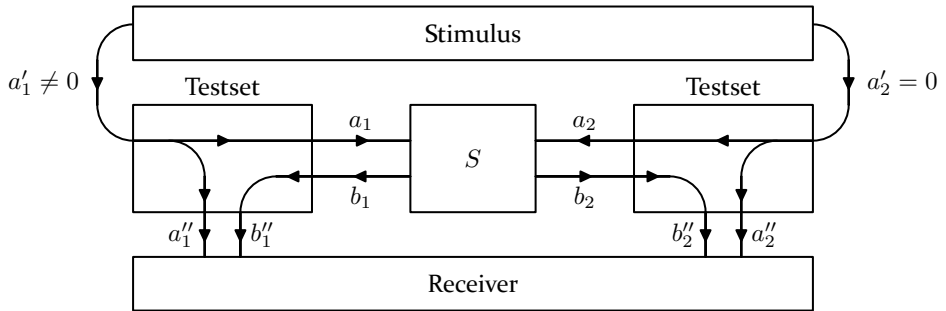


Figure 2.1: Functional principle and building blocks of an ideal 2-port VNA.

stimulus is a swept single frequency source, however it can also be carried out as an inherently wideband generator. Nevertheless, this thesis focuses on single frequency stimulation.

2.2.2 Testset

The testset is the central component of each VNA as it facilitates the actual measurement of reflection coefficients and scattering parameters in general. It fulfills two important tasks: Firstly, it facilitates the separation of incident and reflected waves. Secondly, it is used to derive a reference channel, which is required in order to determine scattering parameters by building the ratio of incident and reflected wave portions.

The functional principle of an ideal testset in a 2-port VNA can also be seen from the block diagram in Fig. 2.1 for the case of a stimulation at port 1. An ideal testset would forward the stimulating wave a'_1 to the DUT as well as to the reference receiver channel so that $a''_1 = a'_1 = a_1$ and would separate b_1 from a_1 so that $b''_1 = b_1$. In this way, the S-parameters of the test object with respect to stimulation at port 1 are given by $S_{11} = b_1/a_1 = b_1/a'_1$ and $S_{21} = b_2/a_1 = b_2/a'_1$. The scattering matrix of an ideal testset is

$$S = \begin{pmatrix} 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 1 & 0 & 0 & 0 \end{pmatrix} \quad (2.8)$$

with the ports 1 to 4 representing the interfaces to the stimulus, DUT, measurement channel receiver and the reference channel receiver, respectively. However, such testsets are not realizable.

Directional Coupler

Testsets can be implemented using any type of directional couplers. There are several topologies such as the Lange, branchline and rat-race (ring hybrid) couplers which can be realized in microstrip technology. However, the coupled line directional coupler is the most promising approach with regard to multi-octave bandwidth and integration on silicon. It is based on a parallel pair of transmission lines and thus, can be laid out efficiently. Its bandwidth can be extended by cascading multiple segments with varying coupling coefficients.

In general, the excitation of coupled lines can be described as a superposition of even and odd mode. While the even mode builds up for equal current directions, the odd mode is created by opposite current directions within the

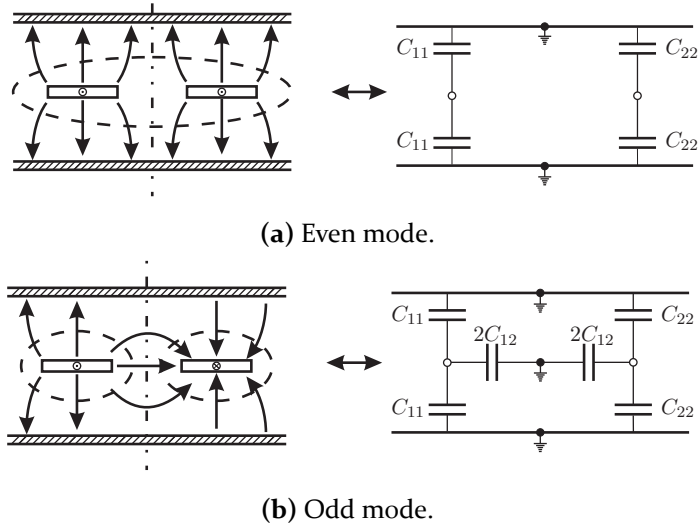


Figure 2.2: Electric and magnetic field distribution and equivalent circuit diagram of coupled striplines for even mode and odd mode excitation.

coupled conductors. The field distribution of electric and magnetic fields as well as the resulting equivalent circuits are illustrated in Fig. 2.2 for both cases. Under the assumption of symmetric shapes, the capacitances of both modes (C_e , C_o) are given by $C_e = C_{11} = C_{22}$ and $C_o = C_{11} + 2C_{12}$, respectively. Consequently, the characteristic impedances are given by

$$\begin{aligned} Z_{0e} &= \frac{1}{v_p C'_e} = \frac{\sqrt{\epsilon_r}}{c C'_e} \\ Z_{0o} &= \frac{1}{v_p C'_o} = \frac{\sqrt{\epsilon_r}}{c C'_o} \end{aligned} \quad (2.9)$$

with the phase velocity $v_p = c/\epsilon_r$, the speed of light c and the relative dielectric constant ϵ_r of the material. It can be shown, that the voltage coupling coefficient k can be expressed in terms of the characteristic impedances of even and odd mode:

$$k = \frac{Z_{0e} - Z_{0o}}{Z_{0e} + Z_{0o}} \quad (2.10)$$

The scattering parameters of ideal coupled lines can be derived by means of an even-odd mode analysis and are given by

$$S_{12} = S_{21} = S_{34} = S_{43} = \frac{\sqrt{1 - k^2}}{\sqrt{1 - k^2} \cos \theta + j \sin \theta} \quad (2.11)$$

for the direct path and by

$$S_{13} = S_{31} = S_{24} = S_{42} = \frac{jk \tan \theta}{\sqrt{1 - k^2} + j \tan \theta} \quad (2.12)$$

for the coupled path with the electrical line length θ . Ideal couplers provide ideally matched ports ($S_{ii} = 0$, $i = 1 \dots 4$) and an infinite isolation ($S_{14} = S_{41} = S_{23} = S_{32} = 0$). In real implementations, however, the input matching is strongly affected by the layout and the isolation is degraded by unequal propagation constants of even mode and odd mode. The directivity is a measure of quality for directional couplers and is defined as the ratio between coupled and isolated path. [30]

Topology

The testset can be realized by directional couplers in two ways (see Fig. 2.3). The simplest way is to utilize a single coupler for the derivation of a reference channel as well as for the wave separation. Despite being area efficient, this topology suffers from a crosstalk between reference and measurement channel caused by the imperfect matching of the receiver channels (indicated by gray dashed lines in the illustration). A better receiver decoupling is achieved by the dual directional coupler testset. In this constellation, the reflections originating from the receiver inputs are dissipated at the terminated ports of each coupler. Indeed, the parasitic crosstalk can be taken into account with an appropriate calibration. Nevertheless, the dual coupler topology is superior and leads to a more stable calibration.

2.2.3 Receiver

The main function of a receiver is to demodulate a RF signal. In a VNA application, this means the determination of amplitude and phase. As there is plenty of literature about the topic of radio frequency receivers [31, 32], the basics shall be repeated only briefly in the following. The selection is limited to homodyne and heterodyne receivers, as they can be integrated in silicon. Alternative approaches like the six-port receiver [33], which are not an option for an integrated VNA, will not be discussed. In the most parts of literature, receiver types are evaluated regarding their suitability for wireless communication systems, whereas this thesis focuses on an instrumentation application. Thus, the requirements for the receiver are slightly different for two main reasons. First, a VNA usually covers a significantly higher bandwidth exceeding several octaves. Second, the frequency components within a network analyzer are usually known leading to different requirements for filtering and band selection.

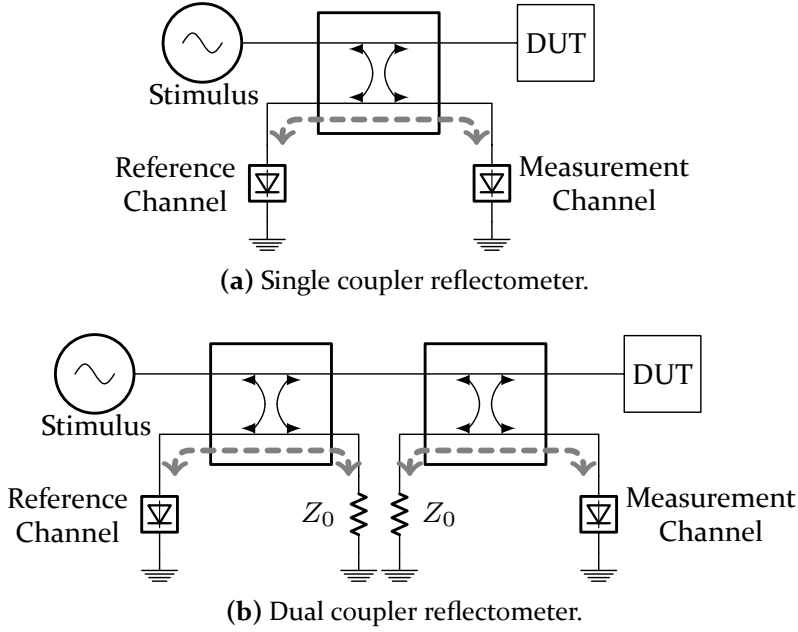


Figure 2.3: Realization of the testset as single (a) and dual (b) coupler reflectometer with the respective crosstalk paths (dashed gray lines).

Homodyne

The principle of the homodyne receiver is to convert a RF signal with the frequency f_{RF} down to an intermediate frequency (IF) of $f_{IF} = 0$ using a mixer driven with a local oscillator (LO) frequency of $f_{LO} = f_{RF}$. For this reason, it is also referred to as zero-IF or direct conversion receiver. A typical implementation as used in communication systems is illustrated in Fig. 2.4. A coarse bandpass filter prevents the receiver from being driven into compression by strong out-of band or interfering signals. A low noise amplifier (LNA) reduces the overall noise figure. In order to maintain phase information during the direct conversion, quadrature mixing is necessary. Therefore, the signal is split and fed to two mixers, of which one is driven by an in-phase (I) and the other by a quadrature (Q) LO signal. Channel selection can be performed by a lowpass filter at the mixer output. A drawback of the homodyne receiver is the parasitic DC offset at the mixer output due to self-mixing effects caused by leakage of the LO to the mixer input. In addition, flicker noise significantly degrades the lower limit of the dynamic range [31]. With regard to the applicability in an integrated, wideband VNA the homodyne type suffers from additional drawbacks. First, the 90° phase shifter

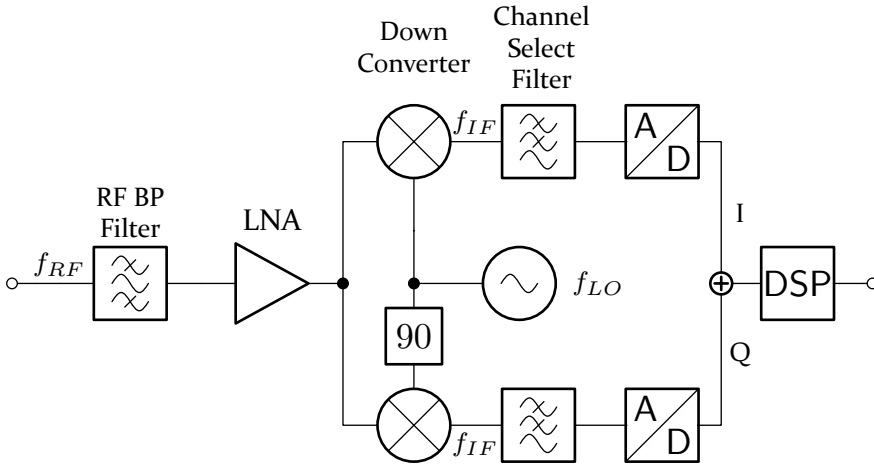


Figure 2.4: Block diagram of a homodyne receiver with I/Q demodulation as used in communication systems.

required for IQ-demodulation is hardly realizable for a multi-octave bandwidth. Second, in the presence of harmonics, the desired down-converted signal containing the phase and magnitude information about the DUT is superposed by unwanted signal portions. The frequency spectrum at the mixer output is given by $f_{IF} = n f_{RF} \pm m f_{LO}$ with n and m being the harmonic index of the stimulus and the LO, respectively. A graphical illustration is shown in Fig. 2.5 under consideration of three harmonic frequencies. In the realistic case of both signals including same order harmonics ($n = m$), several combinations contribute to the zero-IF component. This parasitic superposition corrupts the desired signal amount. While this mechanism is irrelevant for most communication systems, it takes severe effect for measurement devices covering a bandwidth of more than one octave. This is the case for most VNAs, which makes the homodyne receiver a disadvantageous option.

Heterodyne

The principle of the heterodyne receiver is to convert a RF signal with the frequency f_{RF} down to a lower IF at f_{IF} . Therefore, it utilizes a mixer which is driven by a LO at a frequency f_{LO} . The intermediate frequency is given by $f_{IF} = f_{LO} \pm f_{RF}$ and consists of an upper and a lower sideband, of which the lower one is of interest and the upper one is rejected by filters. A real implementation such as a receiver for communication systems requires further components and is shown in Fig. 2.6. A coarse bandpass filter rejects strong

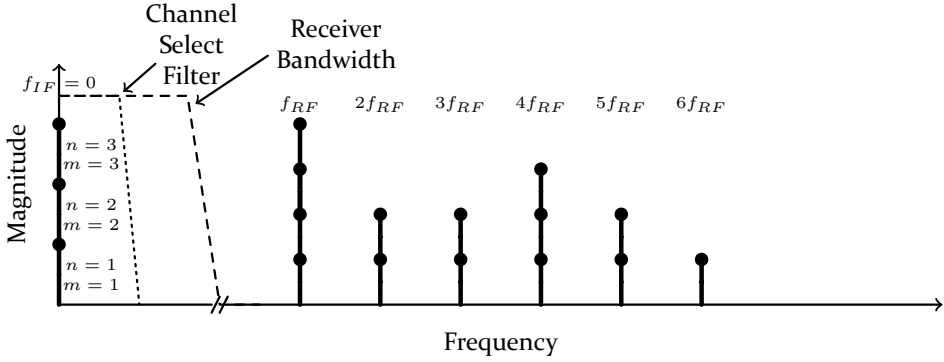


Figure 2.5: Intermediate frequency (IF) spectrum of a homodyne VNA under consideration of harmonics.

interferers in the received signal and a LNA minimizes the noise figure of the overall system. Another bandpass filter is necessary in order to remove so-called image frequency, which is located at $f_{LO} - f_{IF}$ ($f_{LO} < f_{RF}$) or $f_{LO} + f_{IF}$ ($f_{LO} > f_{RF}$). This unwanted frequency would be superposed to the frequency f_{RF} at the IF band and consequently corrupt the desired signal. Thus, the image reject filter needs to suppress the image frequency whereas the desired frequency should not be attenuated. It becomes clear, that such filters can be realized more easily when the offset ($2f_{IF}$) between those two frequencies is large. This corresponds to a high IF. On the other side, the requirements for channel select filters are more relaxed when the IF is low. This problem can be circumvented by a super-heterodyne implementation with two mixing stages where image rejection is performed at the first one and channel selection at the second one. However, the requirements for the receiver are slightly different in a VNA application. In contrast to a communication system, which is fed by an antenna, a VNA is a closed system. Thus, interferers are usually very weak or, for the most part, completely absent. Hence, the coarse RF bandpass filter as well as the image reject filter are irrelevant. Nevertheless, the channel select filter affects the frequency resolution, noise level and dynamic range of the system.

In a swept VNA the RF stimulus and the LO are usually operated with a fixed frequency offset so that all IF components can be designed for narrow-band operation. It is feasible to design the remaining receiver components such as the LO and the LNA for multi-octave bandwidth operation, which is typical for instrumentation devices. Even though interferers are usually absent, the harmonic frequencies of stimulus and LO have to be taken into account due

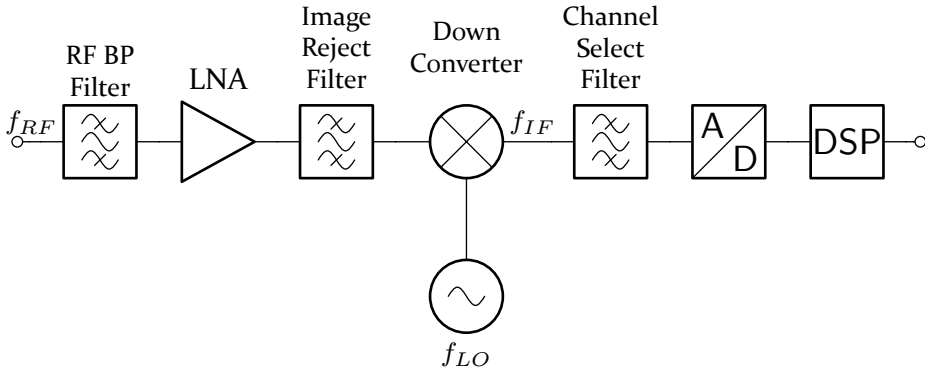


Figure 2.6: Block diagram of a heterodyne receiver with single down-conversion as used in communication systems.

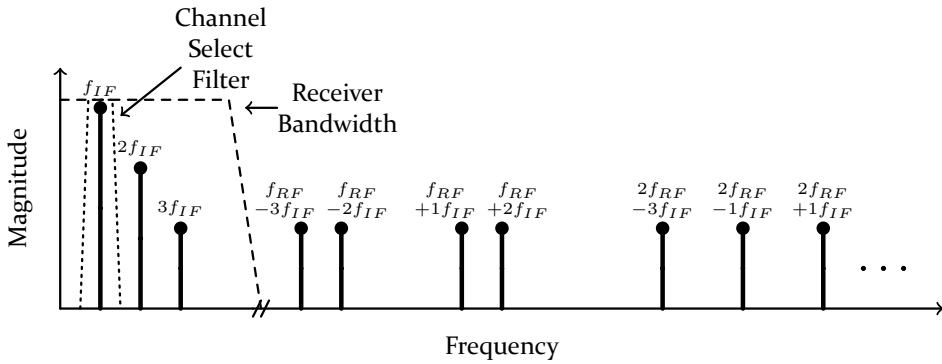


Figure 2.7: Excerpt of the IF spectrum of a heterodyne VNA under consideration of harmonics ($f_{RF} > f_{LO}$).

to the significantly higher bandwidth of a measurement device compared to a communications system. The IF spectrum of a heterodyne VNA under consideration of harmonics is given by $f_{IF} = n f_{RF} \pm m f_{LO}$ with n and m being the harmonic index of the stimulus and the LO, respectively. The frequency allocation for $f_{RF} > f_{LO}$ is illustrated in Fig. 2.7 considering the first three harmonics ($n, m = 1 \dots 3$) and an ideal mixer with infinite isolation. While the mixing products for $n = m$ lead to low frequency components $n f_{IF}$, which are usually within the IF band, a set of higher frequencies appears around f_{RF} and $2f_{RF}$. However, these are usually rejected as they are out of the IF circuit bandwidth. The desired fundamental f_{IF} is free of parasitic superpositions and can be selected by either an analog or a digital filter.

2.3 Measurement Uncertainty

As every measurement device, network analyzers are subject to some measurement uncertainty. The most central effects shall be described in the following.

2.3.1 Random Errors

Various random error sources can degrade the measurements of a VNA. Several components, in particular active circuits, show a dependency of temperature. Thus, their parameters such as gain, noise figure and phase transfer function can alter with a changing ambient temperature or device temperature. For this reason, integrated as well as discretely assembled VNAs should be used only after having reached their operation temperature. In addition, a climatically stable environment contributes to a reduction of measurement uncertainty caused by thermal drift.

Furthermore, stable measurements presume the repeatability of the contributing components. At microwave frequencies, minor concussions or poor mechanical contacts can significantly degrade the measurement accuracy. Thus, connectors and cables used for interfacing the DUT need to be of high quality. Additionally, a mechanically fixed measurement setup is recommended.

Every physical measurement device is affected by noise. The resulting random error affects the measurements the more, the smaller the signal to noise ratio (SNR) is. On the other hand, the noise contribution can be neglected in the case of a high SNR. Consequently, stimulus signal power needs to be chosen high enough, especially in the case of high loss DUTs. Furthermore, a low noise receiver with a low IF bandwidth reduces the system noise floor.

2.3.2 Nonlinear Errors

Scattering parameters describe a test object in a linear manner, for active devices this corresponds to the small signal behavior. Resulting from that, it has to be guaranteed that both the DUT and the VNA itself are operated within their linear dynamic range, of which the lower limit is given by the system noise floor and the upper limit by compression effects, as illustrated in Fig. 2.8.

For the DUT this requirement can be met by choosing an appropriate stimulus output power. As a network analyzer is usually comprised of several measurement channels, it has to be ensured that all of them operate linearly. However, especially during the characterization of DUTs with either

high loss or gain the signal power delivered to the single receivers can deviate strongly. In such scenarios, variable attenuators help to equalize the power levels so that linear operation is guaranteed.

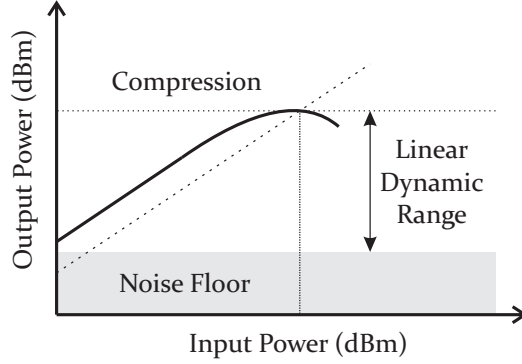


Figure 2.8: Illustration of the linear dynamic range limited by compression and noise floor.

2.3.3 Linear Errors

In contrast to the previous effects, which are of random or nonlinear nature, there are linear effects which contribute to the measurement accuracy. Given that the superposed signal is known in both magnitude and phase, it can be removed by post processing in the digital domain. This procedure is generally referred to as calibration or error vector correction.

The major part of linear errors are caused by imperfect VNA component properties such as source match, receiver match, testset directivity as well as insertion loss in general. These lead to multiple reflections, attenuation and parasitic superposition of the desired measurement signal. In general, these effects are modeled by an error network, which is located between an ideal VNA and the DUT, as depicted in Fig. 2.9 for an N-port VNA. Depending on the network analyzer type and architecture, the error model contains 3 to 22 terms, which can be determined by measurements of well known standards. Once the error model is known, the raw measurement results can be corrected in order to retrieve the test object's true scattering parameters. For further information the reader is referred to [34].

2.4 SiGe Technology

The circuits and systems described in this thesis were fabricated on the SiGe bipolar technology B7HF200 by Infineon Technologies [36]. Despite targeting millimeter wave applications such as automotive radar, the process is

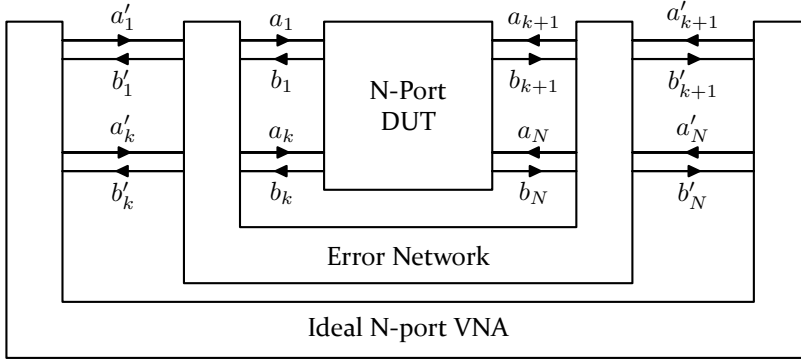


Figure 2.9: General error model for the calibration of an N-port VNA, based on [35].

very well suited for the design of circuits and systems for integrated network analyzers. The technology provides a minimum feature size of $0.35\ \mu\text{m}$ resulting in a minimum effective emitter width of $0.18\ \mu\text{m}$. Three different npn transistors optimized for high voltage (HV), high speed (HS) and ultra high speed (UHS) as well as lateral pnp transistors are available. The collector-substrate capacitance is minimized by both deep and shallow trench isolation. The process provides a transit frequency (f_t) and a maximum oscillation frequency (f_{max}) of 200 GHz and 250 GHz, respectively. The transistor performance is summarized in Table 2.1 including the maximum collector current density ($j_{c,max}$) as well as the collector-emitter breakdown voltages with open (BV_{CEo}) and shorted base (BV_{CEs}). Two polysilicon resistors ($150\ \Omega/\text{sq}$, $1000\ \Omega/\text{sq}$), a high accuracy TaN resistor ($20\ \Omega/\text{sq}$) and MIM capacitors ($1.4\ \text{fF}/\mu\text{m}^2$), which are realized by Al_2O_3 insulation in between two metal layers, can be used. A high performance differential varactor diode can be applied to voltage controlled circuits such as oscillators [37]. Four copper metal layers with thicknesses of 600 nm, 600 nm, 1200 nm and 2500 nm can be utilized for the realization of interconnects and passive microwave components such as inductors, transformers and directional couplers. The upper-most layer is best suited for such components as high quality factors

	$f_t(\text{GHz})$	$f_{max}(\text{GHz})$	$j_{c,max}(\frac{\text{mA}}{\text{m}})$	$BV_{CEo}(\text{V})$	$BV_{CEs}(\text{V})$
HV	35	120	0.5	4	14.5
HS	170	200	5	1.7	6.5
UHS	200	250	6.5	1.5	5.8

Table 2.1: Summarized npn transistor parameters in B7HF200.

can be achieved. Laser trimmable fuses are available in an additional aluminum layer which serves also as interconnect to the contacting pads. The latter are implemented in a surface layer made of gold. For research purposes, this layer can also be utilized for the realization of sensing structures.

Chapter 3

Narrow-band Microwave Vector Reflectometers

3.1 Introduction

A vector reflectometer is the simplest form of a vector network analyzer. It consists of only one measurement port and provides the capability to determine a single scattering parameter: the complex reflection coefficient S_{11} or Γ . Thus, vector reflectometers are also referred as 1-port VNAs. Compared to wideband multiport analyzers, the complexity of a narrow-band vector reflectometer is low as it is comprised of less components with lower requirements. For this reason, the realization of narrow-band reflectometers is an appropriate feasibility study for a highly integrated multi-octave bandwidth 2-port VNA. Basic investigations on such reflectometers serve as a foundation for the chipset and the realization of a wideband 2-port VNA, which will be described in the Chapters 4 and 5, respectively. In the following, a hybrid integrated and a fully integrated approach for 1-port VNA RF frontends are presented. Both realizations are operated utilizing an appropriate IF backend.

3.2 Calibration

An appropriate calibration procedure is essential for the measurement of reflection coefficients with a real 1-port VNA. The non-ideal properties of the reflectometer are represented by an error 2-port, as illustrated in Fig. 3.1. In this way, the VNA can be considered as ideal. The reflection coefficient of the DUT $\Gamma_{DUT} = a_{DUT}/b_{DUT}$ is transformed by the error 2-port leading to

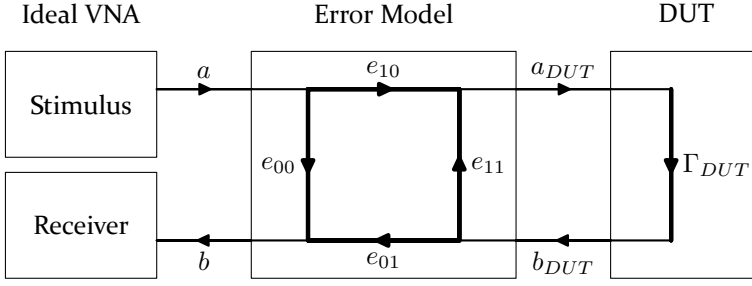


Figure 3.1: Error model of a 1-port VNA.

the actually measured value $M_{DUT} = a/b$. The error parameters e_{11} and e_{00} model the source match and the directivity of the wave-separating element, respectively, whereas e_{10} and e_{01} stand for the insertion loss of the measurement system. One of the latter parameters can be set to unity as they appear exclusively as a product of each other. Consequently, the measurement of only three calibration standards is necessary in order to determine the error 2-port completely. The standards open (O), short (S) and match (M) are frequently used for this purpose, which is why the procedure is also referred to as OSM-calibration. The reflection coefficients of the open Γ_O and the short Γ_S need to be known whereas the match standard is supposed to be ideal ($\Gamma_M = 0$). With the measured values of the standards open M_O , short M_S and match M_M and with $e_{01} = 1$ the error terms are:

$$e_{00} = M_M$$

$$e_{10} = \frac{(\Gamma_O - \Gamma_S)(M_O - M_M)(M_S - M_M)}{\Gamma_O \Gamma_S (M_O - M_S)}$$

$$e_{11} = \frac{\Gamma_S (M_O - M_M) - \Gamma_O (M_S - M_M)}{\Gamma_O \Gamma_S (M_O - M_S)}$$

Once the error terms are determined, the reflection coefficient Γ_{DUT} of an arbitrary DUT can be calculated from the measured value M_{DUT} [35].

$$\Gamma_{DUT} = \frac{M_{DUT} - e_{00}}{e_{10} + e_{11}(M_{DUT} - e_{00})}$$

3.3 14-18 GHz Hybrid Integrated Reflectometer

The 1-port VNA described first is based on a hybrid integrated circuit. The RF frontend is composed of integrated circuits mounted on a RF PCB, which

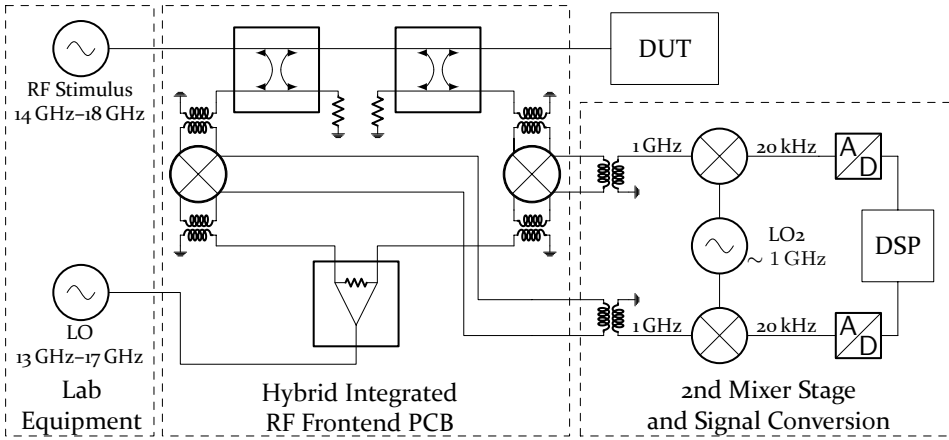


Figure 3.2: Block diagram of the reflectometer setup based on a hybrid integrated RF frontend.

carries the required passive components. The low frequency backend is realized on a separate board. Additional external laboratory equipment completes the hybrid integrated reflectometer operating in a frequency range from 14 GHz to 18 GHz.

3.3.1 System Architecture

The 1-port VNA is realized in a super-heterodyne architecture as illustrated by the block diagram in Fig. 3.2. Two external signal generators serve as stimulus (14 GHz to 18 GHz) and LO (13 GHz to 17 GHz). A set of two directional couplers is used, of which the first one derives the reference channel from the stimulus signal and the second one separates the incident wave from the reflected wave. The isolated port of each coupler is terminated with the reference impedance of $50\ \Omega$. In principle, the same functionality is given by a setup with one coupler only. However, using two couplers the reflected signals caused by the mixer input mismatch are mainly terminated instead of being superposed to the neighboring channel's measurement signal. As a last consequence this leads to a reduced error term magnitude and thus to a more stable calibration. For the same reason, all other components are optimized for minimum return loss. The first down-conversion mixer stage is realized by two integrated Gilbert-cells. Due to the fully differential architecture these active mixers provide an improved isolation between the ports and a high conversion gain compared to unbalanced mixer architectures. The circuits were designed in the SiGe HBT technology described in

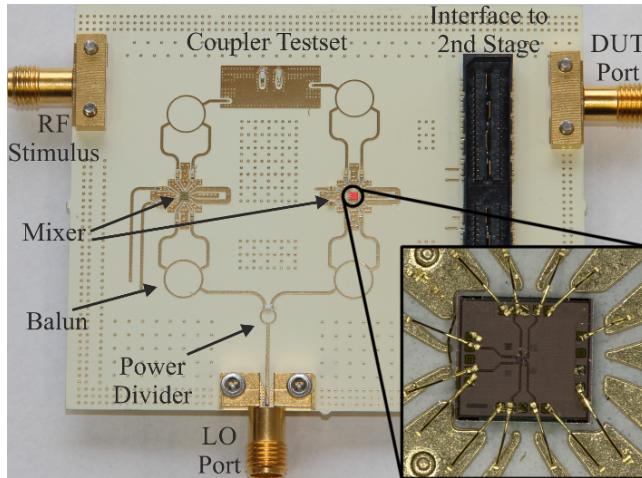


Figure 3.3: RF frontend of the hybrid integrated 1-port VNA. [27] (Copyright 2014, IEEE)

section 2.4. The chips were bonded onto the PCB via wedge-wedge wires. Baluns are used in order to embed the fully differential mixer circuits into the single-ended system. Ring hybrids were implemented for the RF and the LO interface whereas a transformer balun is used for the lower IF frequency of 1 GHz. The LO distribution is carried out as a Wilkinson power divider. The DUT can be connected via a coaxial 2.92 mm connector. The fabricated and assembled RF frontend including a close-up view of the mounted integrated circuits (ICs) is shown in Fig. 3.3.

In a second mixer stage, commercially available mixers are used to down-convert the measurement signals to frequencies as low as 20 kHz so that low-cost ADCs can be utilized. The second IF is sampled with a rate of 600 kHz. The samples are transferred to a 400 MHz digital signal processor that computes a Fast Fourier transform (FFT) of the measurement signals or forwards them via a serial interface for further processing and analysis.

3.3.2 Components

Directional Couplers

As mentioned previously, directional couplers fulfill two important tasks in the microwave reflectometry. The first coupler derives a reference signal, which is proportional to the stimulus in both magnitude and phase. The second coupler separates the incident wave from the reflected wave. The

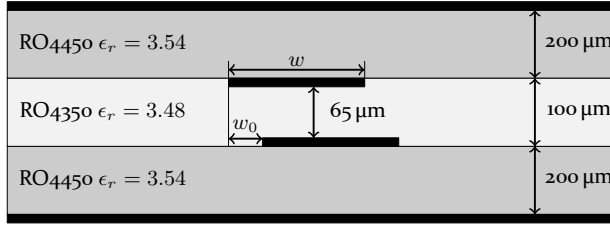


Figure 3.4: Coupled line structure in a quasi-homogeneous four-layer stackup for high directivity directional couplers. [27] (Copyright 2014, IEEE)

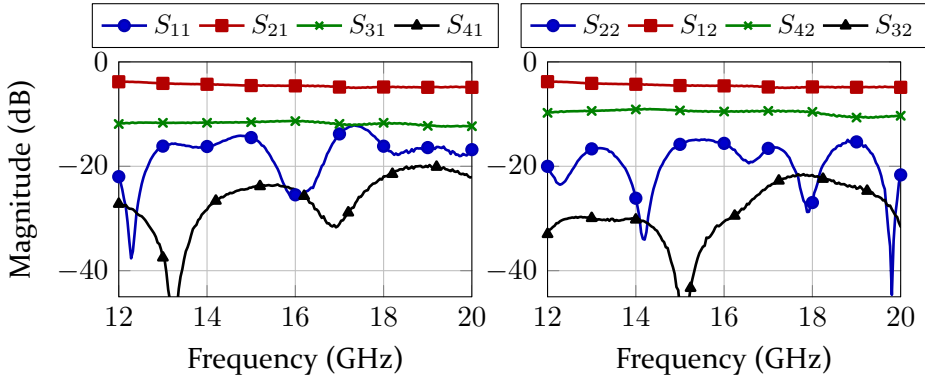


Figure 3.5: Measured S-parameters of the testset on a separately fabricated PCB. Port 1 represents the RF stimulus, port 2 the DUT interface, port 3 the reference channel and port 4 the reflection measurement channel. [27] (Copyright 2014, IEEE)

reference coupler was designed to provide weaker coupling than the second coupler in order to prevent a limitation of the dynamic range by driving the reference channel into compression. Both couplers require high directivity in order to minimize error terms and thus make calibrations more stable. However, conventional quarter-wavelength directional line couplers suffer from decreased directivity due to unequal propagation constants of even and odd mode in inhomogeneous PCB stackups. For this reason, the quasi-homogeneous four-layer PCB stackup shown in Fig. 3.4 was chosen. The dual directional coupler testset was fabricated as a stand-alone component for characterization purposes. The measured transmission (S_{21} , S_{12}), the coupling of the reference channel (S_{31}) and of the reflection measurement channel (S_{42}) as well as the isolation of reference and reflection channels (S_{41} , S_{32}) are shown in Fig. 3.5. A return loss of better than 12.1 dB and 14.9 dB is achieved at ports 1 and 2. The average coupling is measured to be -11.8 dB and -9.6 dB whereas the isolation is better than 19.6 dB and 21.5 dB. The average insertion loss is 4.5 dB for both paths.

Down-Converters

Two integrated mixers are utilized in order to convert the high frequency measurement signals to an intermediate frequency, while maintaining phase and amplitude in a linear manner. The circuits were implemented as double balanced Gilbert cell mixers [38] with dual emitter followers serving as output buffers. This topology was chosen because of its high performance regarding gain, bandwidth and isolation [31]. Nevertheless, baluns are required at every port in order to embed the fully differential circuit into a single-ended system. The in-fixture conversion gain including the hybrid baluns for a LO power of 0 dBm is illustrated in Fig. 3.6 and shows an average of 16.1 dB from 14 GHz to 18 GHz. Linear operation is guaranteed up to input power levels of -20 dBm. Each mixer circuit draws a current of 20 mA from a 3.3 V supply voltage and is placed on a chip with the size of 1 mm².

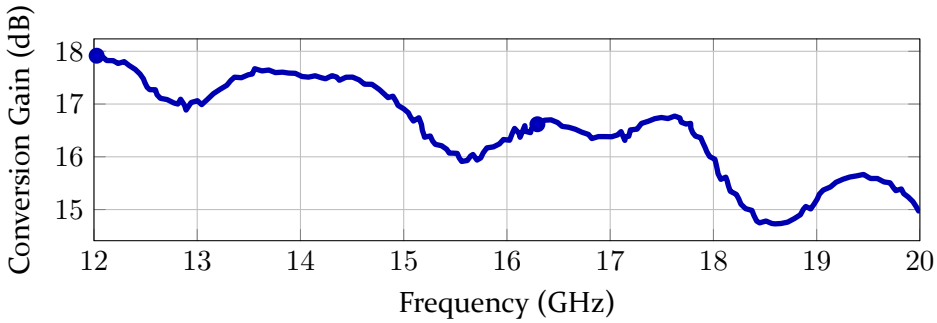


Figure 3.6: Measured conversion gain of the fabricated frequency mixer in a test fixture including parasitics of the hybrid baluns and bond wires for an LO power of 0 dBm. [27] (Copyright 2014, IEEE)

3.3.3 Reflection Coefficient Measurement

In advance of true reflection coefficient measurements, the 1-port VNA needs to be calibrated. For this purpose, the previously described OSM calibration was applied using typical coaxial standards. The calibration algorithm is verified by re-measurements of the three known standards. The corrected measurement results are illustrated in Fig. 3.7. While the open and short standards are shifted around the unit circle versus frequency, the match provides a return loss better than 43.7 dB and appears consequently at the origin of the polar plane. While the results of the open and the short standard are mapped to the true standard values, the measurement of the uncharacterized match standard proves a precise calibration over a frequency range

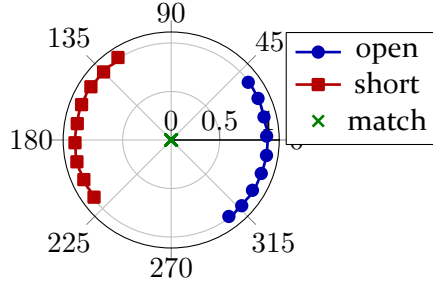


Figure 3.7: Re-measured calibration standards from 14 GHz to 18 GHz. [27] (Copyright 2014, IEEE)

from 14 GHz to 18 GHz. Following this, the complex reflection coefficient of an unknown DUT was measured and compared to the results of a commercial vector network analyzer. The results as well as the deviation between the two measured traces are illustrated in Fig. 3.8. The mean deviation over the measured frequency range is 0.77 dB in magnitude and 5.51° in phase. In total, a good agreement between the results obtained with both systems is achieved.

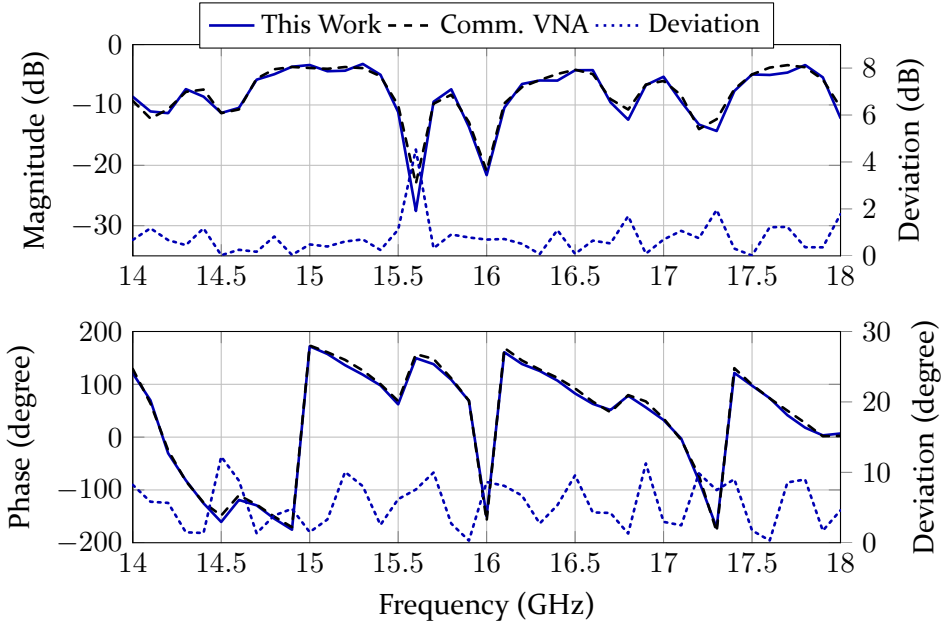


Figure 3.8: Magnitude and phase of the reflection coefficient of a DUT measured with the hybrid reflectometer compared to a commercial VNA. [27] (Copyright 2014, IEEE)

3.4 17–21 GHz Highly Integrated Reflectometer

The second reflectometer is a highly integrated variant. In addition to the down-conversion mixers, also the signal sources for stimulus and LO as well as the directive elements were realized on a SiGe chip. Thus, a fully integrated RF frontend of a 1-port VNA is realized.

3.4.1 System Architecture

The reflectometer architecture is based on the previously described hybrid integrated setup. The block diagram is shown in Fig. 3.9. However, the higher degree of integration leads to major modifications. The signal sources of the heterodyne 1-port VNA are realized by two integrated voltage controlled oscillators (VCOs), each of which is embedded into a PLL. Therefore, the VCO output signals are scaled down by integrated frequency dividers with a factor of eight. The remaining parts of the PLL are realized on an external control PCB utilizing commercially available components. Both loops are phase-locked to the same reference crystal oscillator. The two VCOs run with a fixed frequency offset from each other, which is identical to the desired intermediate frequency of both the reference and the measurement channel. Setting the PLLs in the fractional mode, intermediate frequencies as low as 1 MHz to 10 MHz can be used. The stimulus output power can be regulated by a variable gain amplifier (VGA). In this way, the total output

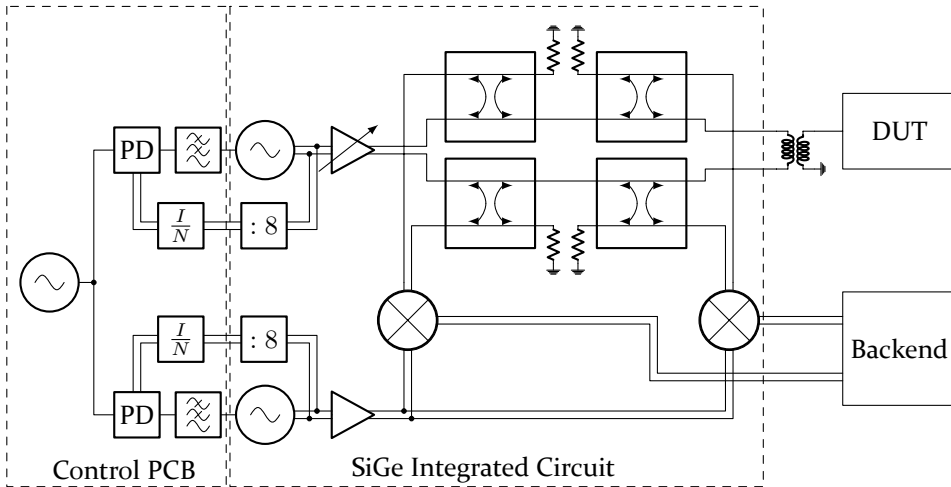


Figure 3.9: Block diagram of the reflectometer based on a fully integrated RF frontend.

power can be adapted to the investigated DUT. Due to the advantages of the differential circuit design, all circuits were realized in a balanced manner. Consequently, the testset was implemented in a dual topology, consisting of two parallel devices. This solution is advantageous versus using multiple baluns as additional return and insertion loss is avoided. The coupler layout is electrically symmetrical in order to prevent phase and magnitude imbalance between the differential paths. Again, the testset was realized by a set of two dual couplers instead of one, so that the calibration error terms are minimized. Two double balanced Gilbert-cell mixers are used as frequency conversion elements. The mixer RF inputs are equipped with matching networks covering the required frequency band from 17 GHz to 21 GHz. The same technique was applied to the stimulus VGA output. Due to the short physical dimensions on a chip, the LO distribution was not implemented by a power divider. Instead, the distribution network was laid out electrically short. The DUT interface is converted from differential to single-ended by an external ring hybrid balun on a RF PCB. Hence, the advantages of balanced integrated circuit design and of unbalanced off-chip signal handling (e.g. regarding calibration) are combined. The system was designed to be operated with a flexible backend so that either a second down-conversion stage can be used or the IF signals are digitized directly. The latter case requires the integrated VCOs to oscillate with a very little frequency offset. Due to on-chip coupling mechanisms, the preconditions for parasitic injection locking can be fulfilled. This would cause one VCO's oscillation frequency to be pulled to the other one's. However, a heterodyne operation requires a frequency offset of the two signal sources. For this reason, the opportunity of a super-heterodyne operation was provided as described in section 3.3.

A micrograph of the reflectometer is shown in Fig. 3.10. The total chip area including pads is $3006\text{ }\mu\text{m} \times 3006\text{ }\mu\text{m}$. The total power consumption of the chip is 469 mW, of which 171 mW, 152 mW and 146 mW are consumed by the VCOs, VGAs and the mixers, respectively.

3.4.2 Components

Integrated Circuits

The two integrated VCOs are realized in a differential cross-coupled topology. The tunable LC tanks are composed of several differential varactor diodes as capacitances and shorted transmission lines as inductances. Interconnects are taken into account as additional parasitic inductances. The VCO core is followed by a buffer amplifier realized as a cascode, which isolates the cores from the following circuits and from the DUT. In this way, load pulling is prevented.

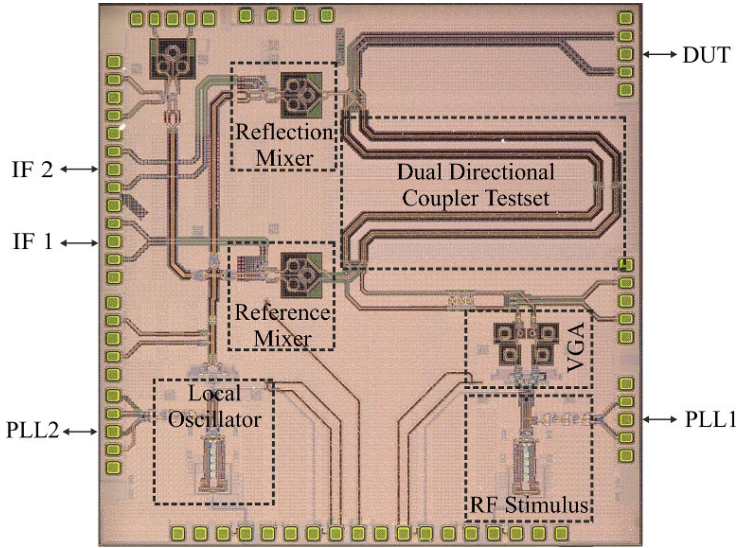
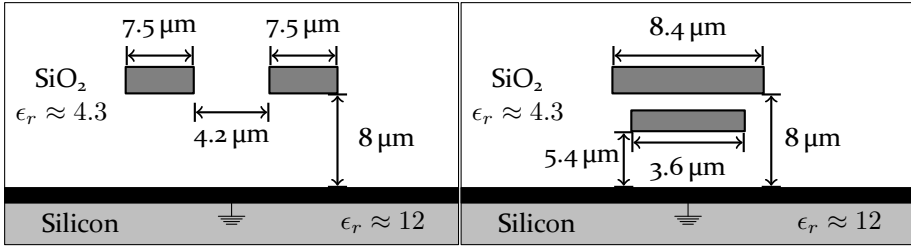


Figure 3.10: Micrograph of the integrated vector reflectometer. [26] (Copyright 2014, IEEE)

The stimulus path of the integrated VNA frontend is equipped with a VGA, which is realized as a cascode with a twin common base stage. The amplifier gain can be controlled by current steering between these two stages. The frequency prescalers are implemented as fed back latches. The down-converting Gilbert cells are based on the circuits used in section 3.3. In addition, the RF inputs were matched to the differential system impedance of $100\ \Omega$.

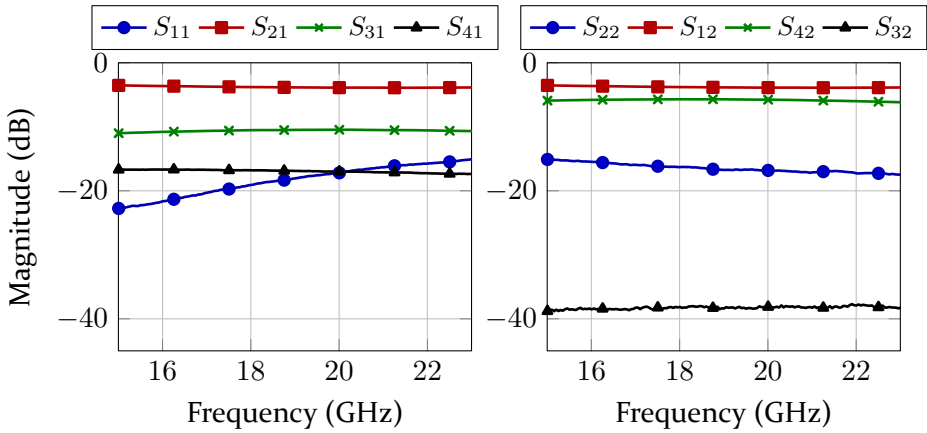
Directional Couplers

As with the hybrid setup, the integrated testset consists of a loosely coupled reference coupler and a tightly coupled reflection coupler. Two different geometries were chosen to realize this constellation, as shown in Fig. 3.11. Parallel coupled lines are utilized for weak coupling, whereas vertically stacked broadside coupled lines are used for the reflection coupler. The coupler layout is a trade-off between area efficiency and performance. At microwave frequencies meandered coupled lines lead to a significant decrease of performance. The measured S-parameters of a separately fabricated testset including the parasitic pad capacitance are illustrated in Fig. 3.12. The mean insertion loss was measured to be 3.8 dB for forward and reverse transmission. The return loss is better than 15.1 dB at both ports. The average coupling



(a) Reference coupler.

(b) Reflection coupler.

Figure 3.11: Cross section of the integrated directional couplers.

Figure 3.12: Measured scattering parameters of the integrated testset. Port 1 represents the RF stimulus, port 2 the DUT interface, port 3 the reference channel and port 4 the reflection measurement channel.

factors of the reference and the reflection couplers are -10.6 dB and -5.9 dB, respectively. The mean isolation between the measurement port and the reference channel is significantly better (38.2 dB) than between the stimulus and the reflection channel (16.9 dB). This effect is due to the reduced isolation of the broadside coupled lines.

3.4.3 Reflection Coefficient Measurement

The 1-port VNA based on the integrated RF frontend is used for the measurement of the complex reflection coefficient as well. Therefore, the previously described OSM-calibration procedure is applied and verified. The measurements were performed at intermediate frequencies below 10 MHz and no

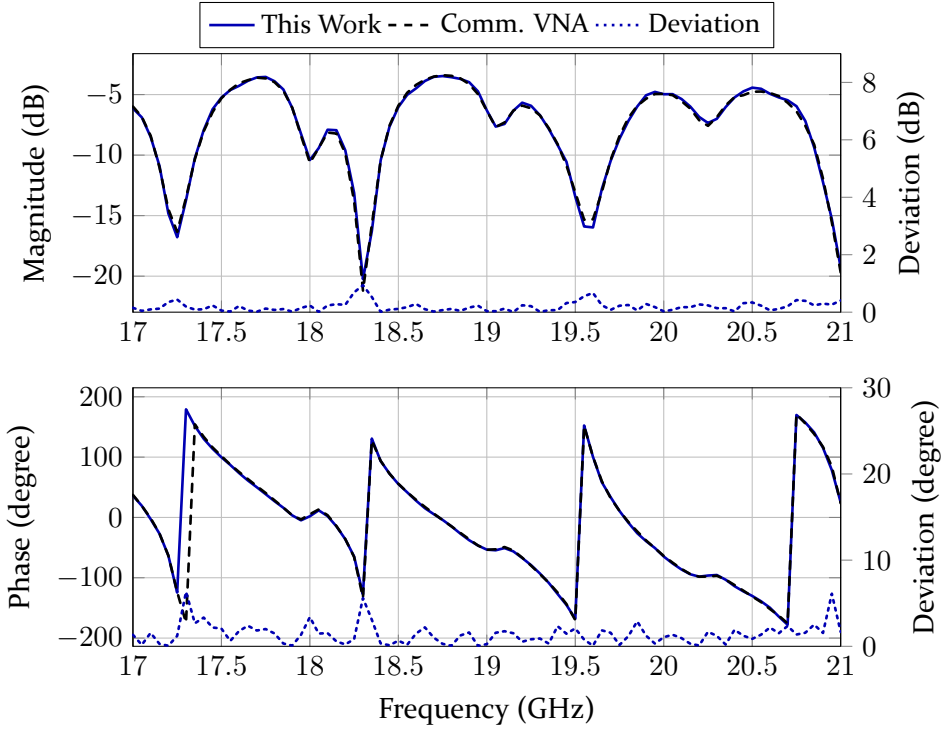


Figure 3.13: Magnitude and phase of the reflection coefficient of a DUT measured with the integrated reflectometer compared to a commercial VNA. [27] (Copyright 2014, IEEE)

parasitic injection locking could be observed. Finally, the complex reflection coefficient of the same DUT as in section 3.3 is determined. The results are compared to measurements performed with commercial equipment and show an excellent agreement, as illustrated in Fig. 3.13. The mean difference between both measurements was as low as 0.2 dB in magnitude and 1.4° in phase over a frequency range from 17 GHz to 21 GHz.

3.5 Conclusion

Two different narrow-band implementations of a vector reflectometer were presented. The first one is a measurement setup consisting of both bulky laboratory equipment as well as of a hybrid integrated RF frontend. The second one is based on a fully integrated frontend combined with external components on PCBs. Both systems were successfully calibrated using a 3-term error model and facilitate the absolute measurement of complex reflection

coefficients. The measured results are in excellent agreement with the measurements conducted with commercial measurement equipment. Thereby, the highly integrated variant showed minor deviation from the reference measurements.

The experiments described in this chapter deliver important perceptions for the implementation of a multi-octave bandwidth multi-port VNA. Hence, it was shown, that parasitic injection locking is not a critical point for practical frequency offsets. Also the low phase noise performance of integrated VCOs compared to external signal generators is not a limiting factor. Two different topologies regarding balanced and unbalanced signal handling were tested. However, a scheme applicable to multi-octave bandwidths has to be found. In total, the concept of a highly integrated microwave VNA was proven to be functional.

Chapter 4

4-32 GHz 2-Port VNA Chipset

4.1 Introduction

Vector network analysis is one of the most important and powerful measurement techniques in the discipline of microwave engineering. It facilitates the analysis of various network types such as active circuitry, passive components or sensors. Modern VNAs are versatile instruments with a wide range of functionalities [35]. Because of the high requirements for accuracy, repeatability and dynamic range VNAs are built from high-performance discrete components, which result in high cost, large physical dimensions and weight of such instruments. Consequently, most devices are not suitable for portable and mass-market applications in industrial or medical environments. However, the potential of VNA based microwave sensors is huge [39–42]. Due to the necessity of a dedicated VNA, these applications are not yet ready for a market introduction. A highly integrated VNA with a small form factor paves the way for the application of network analysis techniques in mass-market applications. For this reason, a VNA chipset will be presented in this chapter [28]. To the best of the author's knowledge, the first chipset for a heterodyne full 2-port vector network analyzer with a multi-octave bandwidth from 4 GHz to 32 GHz is presented. Due to the high integration factor it replaces cost and area intensive measurement instruments and consequently is very well suited for portable mass-market applications in medical or industrial scenarios.

4.2 Wideband Integrated VNA Architecture

In principle, a vector network analyzer consists of a signal source, a directive testset for wave separation and a multi-channel receiver. The signal source stimulates the DUT at a certain frequency or over a bandwidth. The testset is composed of directional couplers or measurement bridges and separates the incident wave from the wave, which is reflected from the DUT and can be used for the derivation of reference channels. A multi-channel receiver is used to evaluate the DUT's complex frequency response and to convert it down to low frequencies for further processing (e.g. calibration). The presented chipset was designed for a heterodyne 2-port VNA with single-ended measurement ports and the corresponding 4-channel receiver as shown in Fig. 4.1. However, different topologies were investigated in advance.

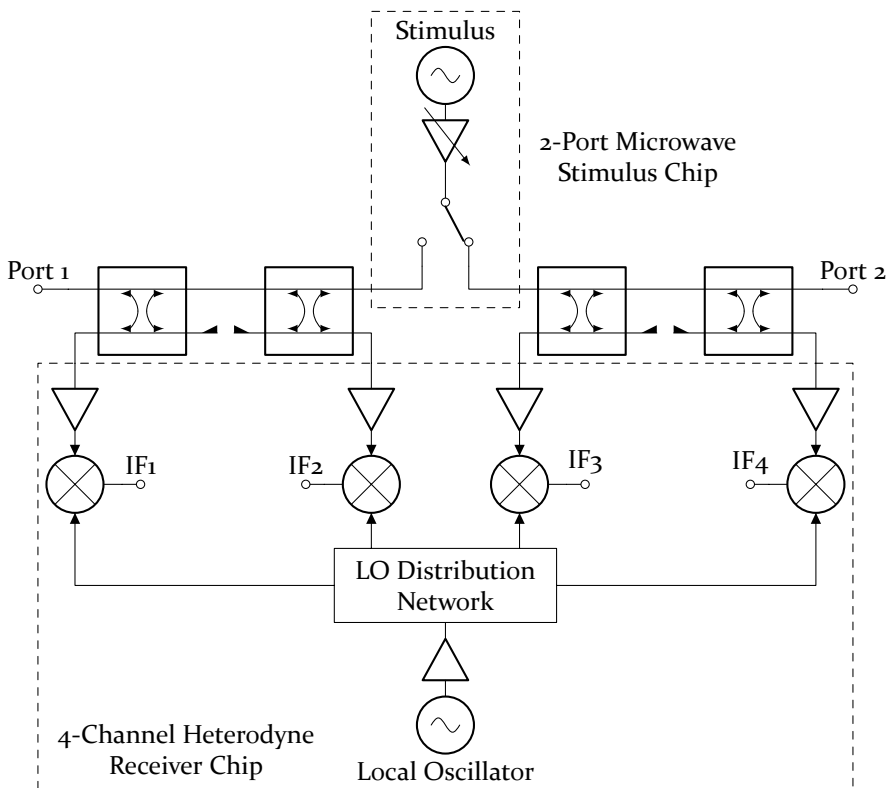


Figure 4.1: Block diagram of a 2-port heterodyne VNA utilizing the proposed chipset. The dashed lines indicate the boundaries of the integrated circuits. [28] (Copyright 2016, IEEE)

4.2.1 Receiver Topologies

Different receiver architectures were investigated regarding their suitability for integrated wideband VNAs. The homodyne topology has the benefit that a single integrated signal source can fulfill the tasks of a stimulus and of a local oscillator in the receiver simultaneously. However, the signal source has to provide quadrature signals in the receiver, which is hard to realize for multi-octave bandwidths. Wideband quadrature signal generation techniques such as quadrature VCOs, polyphase networks and broadside coupled-line couplers were investigated in [43]. Nevertheless, none of the proposed techniques is applicable to a multi-octave bandwidth. Frequency dividers are inherently broadband quadrature signal generators but require the double input frequency, which results in a significantly increased complexity [32]. Furthermore, the requirements for spectral purity are very high in the case of homodyne VNAs as the presence of harmonics in both stimulus and receiver LO inevitably leads to a superposition of the DUT's scattering parameters at the stimulating frequency and its harmonics, as described in section 2.2.3.

A VNA implementation using six-port technology is significantly based on a large number of passive components. Those large physical dimensions are even increased with the need for wideband directional couplers, which are chip area extensive. Despite the low complexity, the six-port technology is not an option for an integrated VNA.

The heterodyne architecture was chosen even though the system complexity is increased as two separate signal sources are required. Due to component re-use possibilities the extra design effort remains low. Therefore, the quadrature signal generation can be omitted so that the VNA bandwidth is from now on only limited by the signal source and the testset bandwidth. In addition, the requirements for the system's spectral purity are relaxed as different stimulus harmonics do not interfere with the fundamental frequency. Also, the heterodyne topology provides the highest dynamic range of the investigated architectures.

4.2.2 Testset Topologies

Vector network analyzers can be built from different testset topologies and the corresponding receivers. 2-port VNAs can be realized with both 3-channel and 4-channel receivers. In the case of a 3-channel receiver the reference channel is derived between the stimulus and the switch by a coupler. Two other couplers are used to generate the measurement channels. With the

4-channel architecture a total of four couplers is required for the derivation of two reference channels and two measurement channels. Clearly, the 3-channel topology is less complex. However, the reduced effort is traded against calibration possibilities as it is not possible to apply a complete calibration to the 3-channel VNA. The imperfection of the switch can not be taken into account, which results in very high requirements for the switch. For this reason, the 4-channel architecture is advantageous as the switch properties can be represented with the common calibration models. In principle, a realization with four channels and two couplers is possible but a significant increase of calibration error parameters has to be taken into account.

4.2.3 VNA Port Interface

The design of the interface to the measurement ports of an integrated wideband VNA is of central importance. On the one hand, the advantages of fully differential integrated circuit design are well known. On the other hand most of the measurement objects which are characterized by VNAs are single-ended. Furthermore, differential off-chip signal routing via coaxial cables creates difficulties as phase and amplitude imbalances between the two differential signal paths are caused by inevitably unequal cable lengths. Also the effort for calibration to an off-chip reference plane is significantly increased as the required number of calibration standards is doubled. In addition, there are no appropriate topologies for baluns, which cover the required bandwidth. Passive approaches are too area intensive [44] and active methods introduce imbalances and additional noise [45, 46]. Consequently, the interface to the DUT was implemented in a single-ended way, whereas all other circuits are realized differentially. Therefore, a dedicated differential to single-ended converting output stage with switch capabilities for 2-port excitation as well as a single-ended to differential converting receiver front-end architecture was chosen. The corresponding circuits will be described in the following section.

4.3 System and Circuit Design

4.3.1 Wideband Frequency Synthesizer

A signal source is a central component of the VNA chipset. An appropriate synthesizer architecture is required in order to cover the desired multi-octave bandwidth. Due to its inherently high bandwidth, the concept of the

direct digital synthesis (DDS) has to be considered [47]. This approach requires a good ramp-to-sine converter in order to provide high spectral purity. However, this element is difficult to realize with both a high performance and a high bandwidth. Furthermore, an additional high frequency clock is required, which determines the upper frequency limit of the system. Thus, the concept of the DDS is especially suitable for the lower microwave region down to frequencies of a few kHz or MHz. Another approach for the octave-wide frequency synthesis is based on a single VCO core and the single sideband mixing of its fundamental frequency with its fourth subharmonics [48]. Due to the imperfection of the required components, the second sideband image can not be suppressed completely and consequently leads to an in-band spurious frequency. Additionally, the active and passive components are area exhaustive, especially if this concept is applied to the microwave region.

In contrast to the described synthesizers, in this thesis a VCO array followed by a bandwidth extension stage is utilized. The block diagram is shown in Fig. 4.2. The three VCOs of which only one is running at a time provide sequential tuning ranges from 16 GHz to 32 GHz. The VCO outputs are combined by a multiplexing amplifier. The octave-wide bandwidth of the VCO array allows a seamless increase of bandwidth by the following extension stage, which consists of three paths. One path feeds the fundamental VCO frequencies through a buffer for increased isolation and an optional filter for a highly linear output. The other two paths incorporate prescalers with the factors of four and two, respectively. In this way, the synthesizer bandwidth

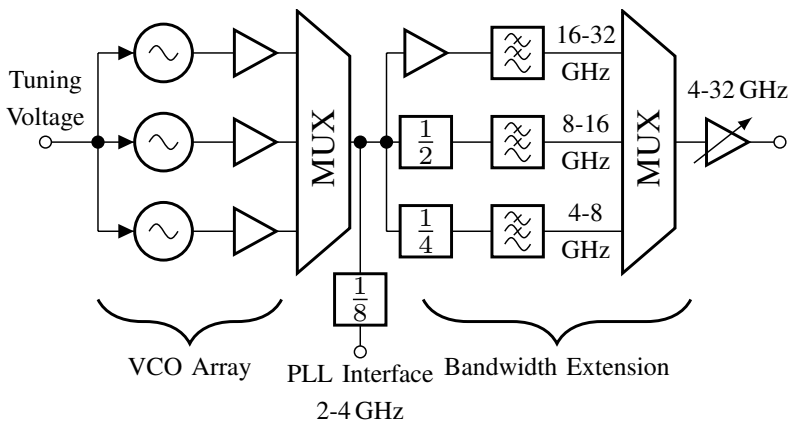


Figure 4.2: Block diagram of the proposed frequency synthesizer covering a bandwidth of three octaves. [28] (Copyright 2016, IEEE)

is extended by two more octaves. Both paths include third-order Chebyshev lowpass filters, which attenuate the harmonic frequencies caused by the prescalers. Of course, more octaves can be generated by adding further or programmable divider stages, however the chip area consumed by the filters is significantly increased at lower frequencies. A second multiplexing stage combines the three paths providing a bandwidth of three octaves at a single output. A wideband variable gain amplifier at the output is extremely useful for many VNA measurement tasks as it increases the dynamic range.

VCO Array

The VCOs were implemented in a fully differential cross-coupled architecture with sequential tuning ranges from 16 GHz to 32 GHz. In the used technology the cross-coupled topology showed better results within the targeted frequency range than Colpitts oscillators, which are frequently used at mm-wave frequencies. Furthermore, a scalable VCO tank model could be developed, which is useful for the design of oscillator arrays. The average, relative tuning range TR_{vco} of the VCOs required to guarantee an octave-wide bandwidth of an array consisting of N oscillators is given by

$$TR_{vco} = 2 \frac{\sqrt[N]{\frac{f_{max}}{f_{min}}} - 1}{\sqrt[N]{\frac{f_{max}}{f_{min}}} + 1} \quad (4.1)$$

where f_{max} and f_{min} are the array's maximum and minimum frequencies, respectively [49]. The desired octave-wide bandwidth requires $\frac{f_{max}}{f_{min}} = 2$, so that a number of $N = 2, 3, 4$ leads to relative tuning ranges (TR_{vco}) of 34 %, 23 %, and 17 %. Even though single VCOs with tuning ranges of more than 30 % have been successfully demonstrated (e.g. [50, 51]), it is not advisable to realize an array with $N = 2$ as process deviations, as well as other parasitics, could easily cause a gap in the array's output spectrum. For this reason, three oscillators with slightly overlapping tuning ranges were designed.

The schematic of VCO-1 and the following buffer are shown in Fig. 4.3. The oscillator core consists of an active part and a passive LC resonator. The active part is formed by the transistors T_1 , which are cross-coupled by feedback capacitors and, thus, provide an input impedance with a negative real part to the tank. The tank resonator consists of purely passive components and is inevitably lossy. It is the task of the active device to compensate energy losses within the tank and to guarantee a stable steady-state oscillation.

One main design challenge results from the physical dimensions and the

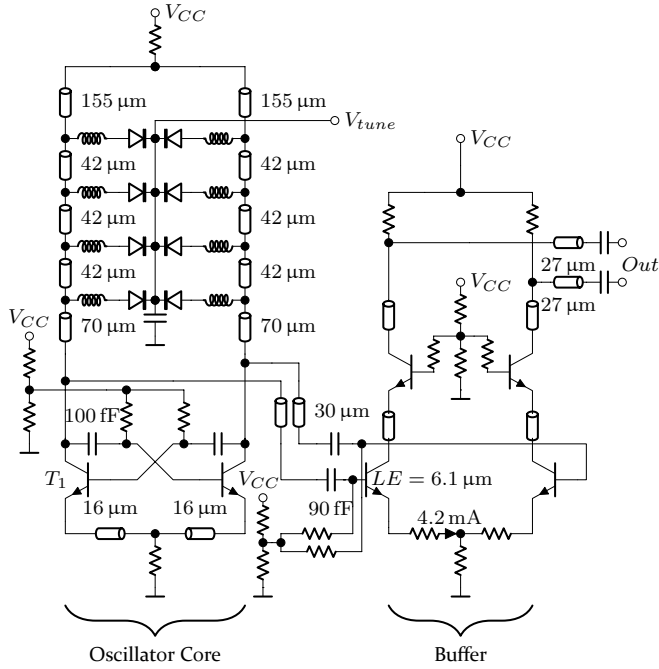


Figure 4.3: Schematic of the oscillator core VCO-1 and the following buffering amplifier. [28] (Copyright 2016, IEEE)

required number of the varactor diodes, which are applied to tune the oscillation frequency. Several varactor pairs have to be used in order to set the oscillation frequency within the upper microwave frequency range. Additionally, the length and width of the components reaches dimensions, which are not negligible in comparison to the wavelength in silicon. Because of this, the interconnects between the varactors have to be taken into account by careful modeling. For this purpose, a scalable tank model including the varactors and its interconnects was developed and applied to the three designed VCOs. The resulting schematics of VCO-2 and VCO-3 are illustrated in Fig. 4.4. While lengthier interconnects are realized as transmission lines, short and thin connections towards and between the varactors are modeled as inductances. In addition to the inductive interconnects, tank transmission lines were used in order to set the oscillator center frequency to the required value. The highly distributed character of all passive tank components required extensive modeling of transmission lines as well as of vertical and horizontal connections by 2.5D electromagnetic field simulations. This leads to a highly accurate design procedure for resonant tanks of oscillators with subsequent tuning ranges.

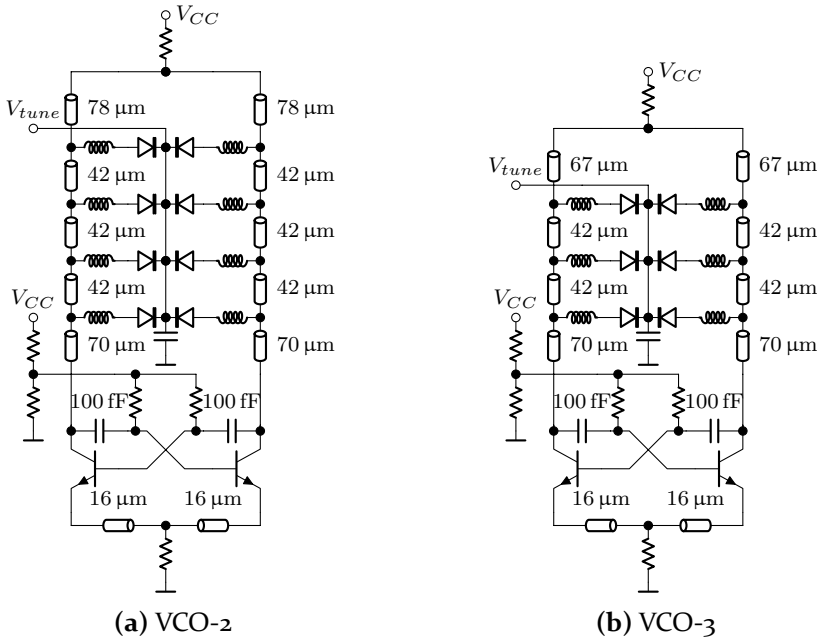


Figure 4.4: Schematics of the other two realized oscillator cores. [28] (Copyright 2016, IEEE)

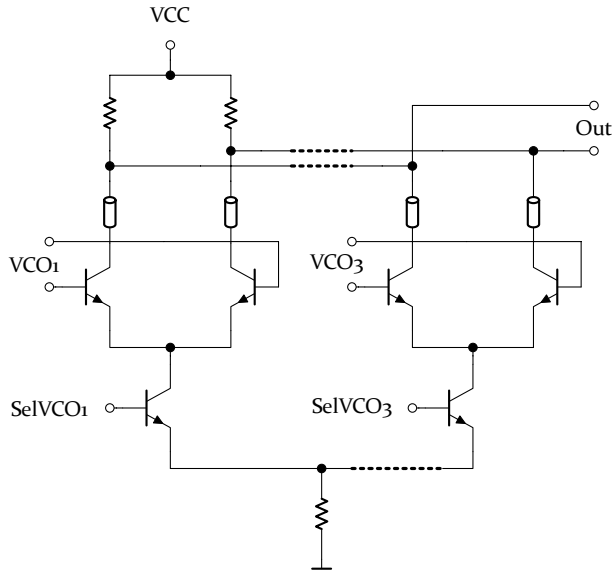


Figure 4.5: Excerpt of the schematic of the designed multiplexing amplifier showing two of three identical cells. [28] (Copyright 2016, IEEE)

The oscillator cores are isolated from the following components by a buffer cascode amplifier, which is also shown in Fig. 4.3. As the oscillator output node is located between the active part and the tank, the phase noise performance as well as the tunability are strongly affected by the loading circuit. For this reason, the cascode was optimized regarding its input impedance. Due to a low bias current of 4.2 mA per side, small transistor sizes (emitter length: 6.1 μm) could be utilized so that the tuning range was not reduced by large static capacitances. On the other hand, a sufficient output power was maintained. As a result of the high backward isolation of the cascode the oscillator cores are robust versus load pulling mechanisms.

The three differential oscillator outputs are combined to a single-output by a 3-to-1 multiplexing amplifier, which is shown in excerpts in Fig. 4.5. The circuit consists of a triplet of differential pairs as input stages with a shared collector load at the output. The channel selection is applied by current sources which are directly coupled to the oscillator supply so that only one channel and one VCO are selected at a time. This leads to a reduced power consumption and additionally parasitic injection locking as well as unwanted mixing products caused by a finite multiplexer isolation are prevented.

Post-Synthesis Bandwidth Extension

The octave-wide bandwidth of the VCO array allows a seamless bandwidth extension by frequency division. Each division by multiple factors of two delivers another octave of bandwidth. Consequently, the realized post-synthesis bandwidth extension stage consists of three paths of which one can be selected to generate the desired frequency band. Along with the three VCOs, nine subsequent frequency ranges can be synthesized. The frequency dividers were realized with two latches in a feedback loop. The latches themselves were realized in current steering technology containing a differential and a regenerative pair, as shown in [32]. The low-pass filters were realized in a third order Chebyshev topology containing two capacitors and one spiral inductor. Due to the differential circuit design the requirements for the filter steepness are relaxed as only odd harmonics have to be removed.

The combining multiplexing amplifier was realized in a similar manner as shown in Fig. 4.5. Due to the presence of frequencies as low as 4 GHz, a broadband biasing scheme has to be applied from the multiplexer towards the synthesizer output. Instead of series capacitors, level shifters are utilized in order to set the desired bias points. As a result, special care has to be taken regarding the circuit stability. The applied design technique will be described in section 4.3.3.

Wideband Variable Gain Amplifier

The synthesizer output stage was realized as a wideband variable gain amplifier. Especially for the network analysis of sensitive active devices it is necessary to reduce the stimulus output power. The schematic of the amplifier is shown in Fig. 4.6 and can be divided in several functional blocks. The first stage is built by a highly linear differential pair (T_1, T_2), which acts as a broadband level shifter to provide an appropriate biasing to the following stage of common collectors. Because of the appearance of low frequencies, series capacitors would reach dimensions which are hardly realizable on a chip. The cascade of two dual common collector stages (T_3-T_6) fulfills an important task within the amplifier as it compensates the low-pass transfer functions of the level shifter and the following cascode amplifier (T_7-T_{10}). Due to the capacitive loading of both common collector circuits, a high-pass transfer function is generated [52]. By choosing proper bias currents and transistor sizes a flat gain up to frequencies of 32 GHz and higher can be achieved. Stability investigations are especially important at this point as common collector circuits tend to instability, which can be caused by parasitic inductive interconnects.

The variable gain capability is implemented within the common base of the following cascode. An additional parallel pair of transistors (T_{11}, T_{12}) which is isolated from the circuit's output reduces the amplifier gain by current steering. If a control voltage V_{gain} higher than the bias level is applied to the base nodes of the parallel pair the current through the load is reduced and, consequently, the gain is decreased. The biasing was chosen to provide good output matching at low frequencies. A separate matching network consisting of spiral inductors and differential capacitors provides very good output matching up to the operation frequency of 32 GHz. As poor source matching directly affects the VNA's error parameters, this point is of central interest when designing network analyzer components.

Source Switch

All the previously shown circuits were realized in a fully differential topology. However, a single-ended interface to off-chip DUTs is advantageous as most measurement objects (e.g. biosensors) are implemented single endedly. For this reason, the broadband variable gain output stage shown in Fig. 4.6 was modified in order to provide two separate single-ended output channels, whereas each channel can either be selected for the stimulation of the DUT or turned off, respectively. In this way, the single pole double

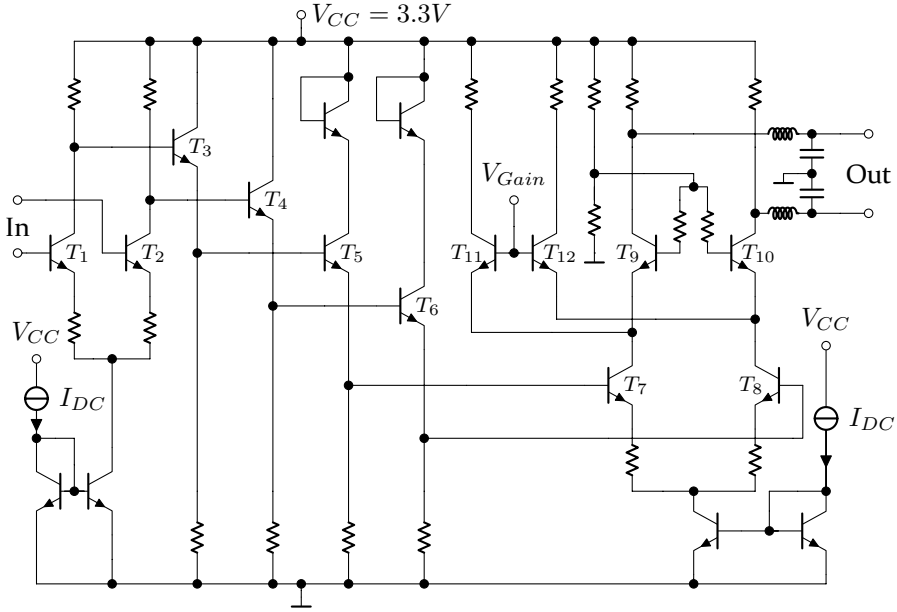


Figure 4.6: Schematic of the designed broadband variable gain amplifier. [28] (Copyright 2016, IEEE)

throw (SPDT) switch, which is required for a 2-port VNA (see Fig. 4.1) can be realized efficiently. The switch topology is based on the previously shown variable gain amplifier. Again, a capacitively loaded two-stage cascade of emitter followers guarantees a broadband behavior. In order to provide the differential to single-ended conversion, as well as a high switch isolation, the following cascode stage was modified as shown in Fig. 4.7. The circuit basically consists of two identical interlaced cascode amplifiers ($T_1 - T_4$, $T_5 - T_8$). Each of these cascodes represents one of the two single-ended output channels (CH A, CH B) and can be selected or turned off with current-steering techniques by applying the supply voltage or ground potential, respectively, to the corresponding nodes (Sel CH A, Sel CH B) in the current sources. During typical VNA operation one can assume that only one channel is selected at a time. Consequently, only one cascode is supplied at a time so that the power consumption is not increased compared to the single VGA with only one cascode. Furthermore, the differential input impedance provided to the emitter follower cascade is independent from the switch state, which is fundamental for the broadband transfer function of the switch. Both cascodes contain one path which is terminated on-chip and one path, which is fed to the output via a matching network to provide a good source matching of

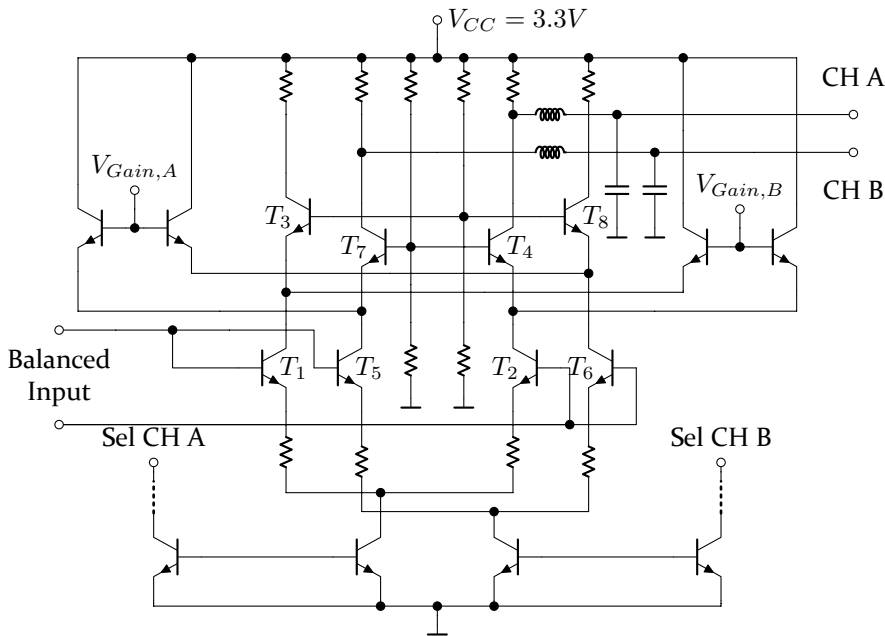


Figure 4.7: Excerpt of the schematic of the designed SPDT source switch. [28] (Copyright 2016, IEEE)

the VNA ports. Furthermore, the gain of both channels can be controlled by applying appropriate voltages ($V_{Gain,A}$, $V_{Gain,B}$) to two additional parallel common base stages.

4.3.2 Wideband 4-Channel Receiver

In the same manner as with the synthesizer the interface to the measurement object was realized single-endedly. In the following subsections, the receiver components will be described.

Low Noise Amplifier

In order to increase the dynamic range of the VNA, a single-ended LNA with multi-octave bandwidth was designed. The amplifier schematic basically consists of a cascode followed by a common collector circuit and is illustrated in Fig. 4.8. Besides the obvious designs goals like high gain and a low noise figure over the full bandwidth, special care was taken to provide a good input matching, as this factor directly contributes to the VNA's error parameters.

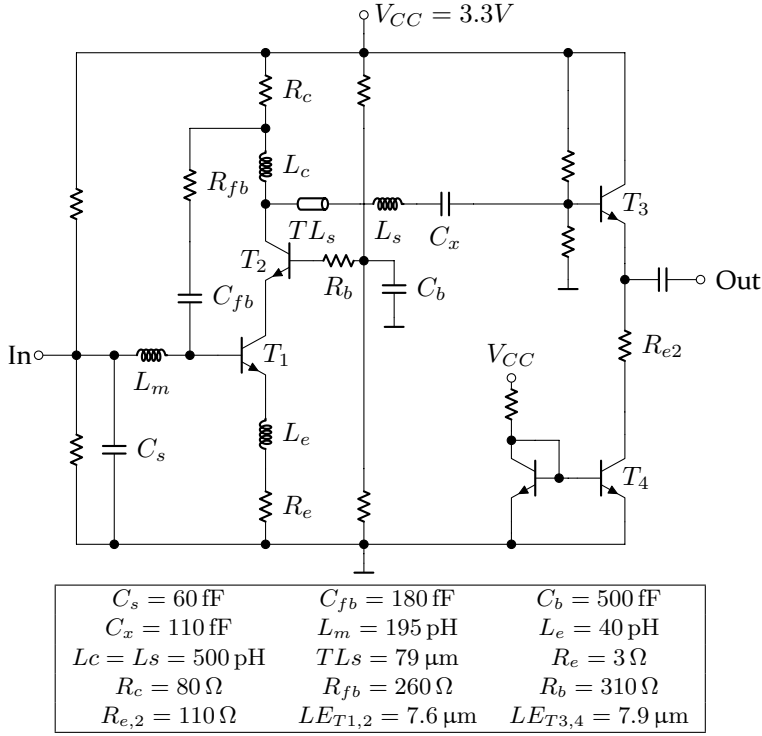


Figure 4.8: Schematic of the designed low noise amplifier. [28] (Copyright 2016, IEEE)

Different measures were taken to eliminate the typical low-pass characteristic of the conventional cascode amplifier. The effect of the applied wideband design techniques is illustrated in Fig. 4.9. This graph shows a comparison of the full LNA compared to the results obtained when one of the applied measures is omitted. One can clearly see, that the shunt-shunt feedback network composed by R_{fb} and C_{fb} contributes to a flat broadband gain as it lowers the gain at low frequencies and shifts it towards high frequencies. The capacitor C_{fb} not only allows a biasing, which is independent from the feedback transfer function, but also improves the input return loss at low frequencies considerably. As the presented amplifier is realized by a cascaded architecture, its noise figure is mainly determined by the gain and noise performance of the first stage, which is the common emitter circuit of the cascode in this case. As the gain is attenuated at lower frequencies and shifted towards higher frequencies, the feedback network slightly increases the noise figure, especially at low frequencies. However, it provides an overall flat noise figure of the LNA. Additionally, both series and shunt peaking

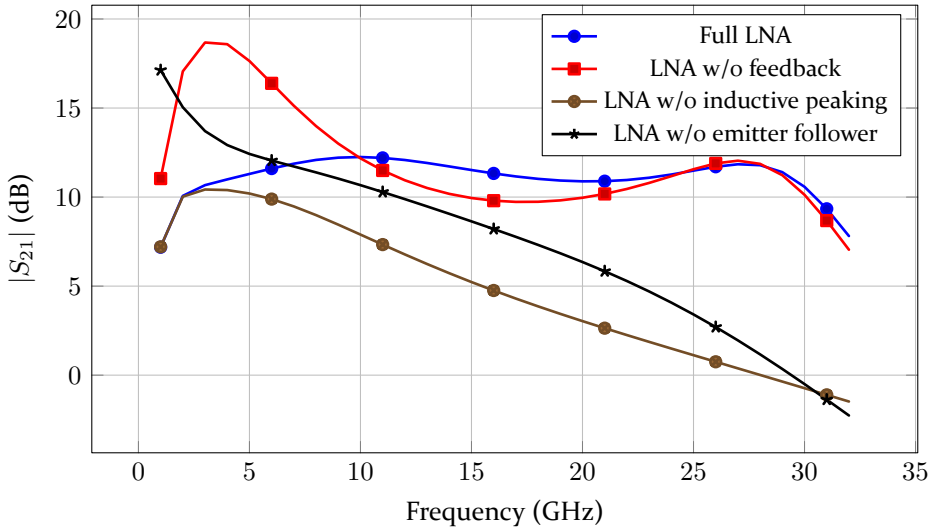


Figure 4.9: Simulated effects of the broadband design techniques applied to the LNA. [28] (Copyright 2016, IEEE)

techniques were applied to contribute to a broadband gain of the amplifier. Therefore, the inductors L_c and L_s , as well as the transmission line TL_s were used whereas the dimensions of the latter were mainly determined by layout restrictions. Both inductances were realized as spiral inductors and designed with the aid of 2.5D electromagnetic field simulations. As the gain of the cascode depends on its output impedance, the shunt inductor L_c is utilized in order to provide a high impedance also at higher frequencies. In this way the low pass characteristic caused by the transistor capacitances can be compensated partially. Furthermore, the series inductor L_s is used for interstage matching and to prevent a simple addition of the subcircuits' input and output capacitances. Both peaking methods significantly increase the gain (Fig. 4.9) and thus also lower the noise figure. The input matching is also improved at lower frequencies. The second stage of the LNA is implemented as an emitter follower. Similarly to the previously shown VGA its high-pass transfer function is exploited to compensate the low-pass characteristic of the cascode. The subcircuit is necessary for the broadband functionality. Despite also improving the input return loss, it adds some noise to the overall circuit. However, this tradeoff can be accepted. For an overall good input matching the network consisting of the spiral inductor L_m and the capacitor C_s was applied. Also, the emitter degeneration components L_e and R_e contribute to a better input matching and higher linearity [53].

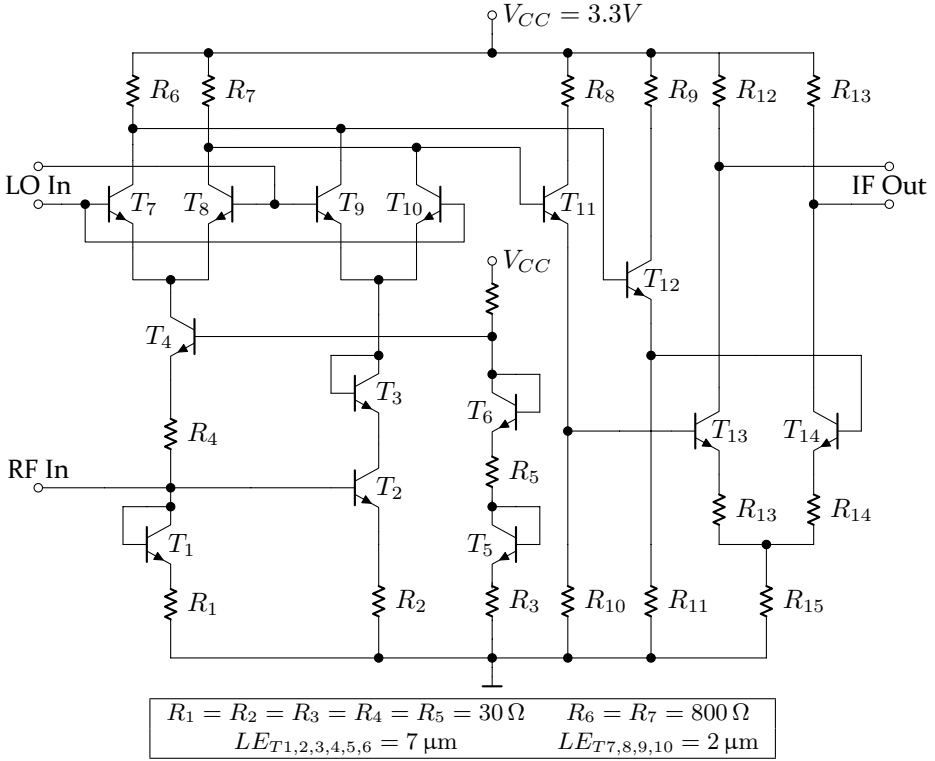


Figure 4.10: Schematic and most important component values of the designed down-conversion mixer. [28] (Copyright 2016, IEEE)

Down-Conversion Mixer

A down-conversion mixer is required in order to convert the frequency response of the DUT to an IF. For this purpose the micromixer shown in Fig. 4.10 was implemented. The micromixer is based on a conventional Gilbert cell, but possesses a different RF input stage. Instead of the differential pair it is equipped with a class A/B translinear input stage [54]. The main advantage in the context of this thesis is the single-ended RF input of this topology. Furthermore, it generally provides a higher linearity and lower gain than the Gilbert cell.

The transistors T_1 - T_4 form a class A/B input stage with a single-ended input and differential output. The four transistors provide identical emitter areas and configurations and are biased with a low current of 460 μA in each path. In this way, the transistors are biased slightly above their cut-off point as long as no RF input signal is present. The transistor T_2 , which is biased

by the transdiode T_1 implements a common emitter stage, which is conductive for the positive half-cycle of the RF input signal and closes during the major part of a negative half-cycle. The common base stage formed by T_4 stays conductive for the negative half-cycle of the input signal and mainly closes during the positive half-cycle. Due to the inverted gain of the common emitter architecture the class A/B stage provides a differential output signal to the switching quad formed by T_7 - T_{10} . The transdiode T_3 is required in order not to exceed the collector-emitter breakdown voltage of T_2 . The input impedance can be adjusted by the resistors R_1 , R_2 and R_4 , whereas all of the transistors show identical values in order to maintain the symmetry of both paths. The switching quad T_7 - T_{10} , which is well known from the Gilbert cell, is loaded with a high resistance (R_6, R_7) followed by an emitter follower (T_{11}, T_{12}) for impedance transformation and, finally, a differential pair (T_{13}, T_{14}) in order to maximize the gain [53].

4-Channel Receiver

The 4-channel receiver required for VNA applications is composed of the described building blocks. The wideband frequency synthesizer acts as a local oscillator and drives the four frequency down conversion elements. An absolutely symmetrical LO distribution was designed and laid out so that an equal power split is guaranteed. Thus, almost identical properties of the four channels can be provided. The presented wideband VGA is utilized as LO buffer.

4.3.3 Design for Wideband Stability

The design of wideband circuits consisting of cascaded transistor stages requires special measures and investigations to ensure internal and external stability. The well-known criteria like the evaluation of the K-factor and the construction of stability circles can be used in order to determine, whether a single stage circuit is stable, conditionally stable or unstable for a wide range of source and load impedances [30]. However, these procedures are only reliable for single stage amplifiers. In principle, one could separate the multi stage amplifiers into several single stage amplifiers and investigate each stage individually. However, especially in combination with the broadband biasing schemes shown before, where one stage is biased by its precedent stage, approximations have to be made because the biasing has to be emulated. In addition, feedback loops between the stages are not taken into account. Because of this, the S-probe introduced in [55] was implemented within the simulator for stability investigations. With the S-probe technique

input impedance $Z_1(\omega)$ and output impedance $Z_2(\omega)$ of any node can be determined without loading or separating the circuit. Based on these results, the reflection coefficients $\Gamma_1(\omega)$ and $\Gamma_2(\omega)$ can be evaluated. Oscillations at the investigated node can only occur, if the product $\Gamma_1(\omega)\Gamma_2(\omega)$ encircles the point (0, 1) in the complex plane in a clockwise direction [55].

In the following, an example of a stability analysis with S-probes will be described by means of the cascaded emitter follower stage of the VGA shown before. On the one hand, it is necessary to load the emitter followers capacitively in order to create a high pass transfer function. On the other hand, emitter followers tend to be unstable with a capacitive load as this can create a negative input resistance at high frequencies. Also, an inductive network at the emitter follower input can create a negative output resistance and consequently lead to parasitic oscillations. For this reason, the interfaces between the cascaded stages were investigated regarding stability. The results for such an analysis of the interface between emitter follower and cascode are shown in Fig. 4.11. For this purpose, the interconnects between the two emitter followers were modeled as inductances, which create a potential for instability of the circuit. In order to determine the maximum tolerable inductance of the interconnect the circuit instability was provoked and monitored via the S-probe technique. The resulting product of the reflection coefficients $\Gamma_1\Gamma_2$ was calculated for different parasitic inductances as illustrated in Fig. 4.11a. One can see that the circuit is stable for the values 201 pH and 301 pH, whereas the critical point (1,0) is encircled clockwise for an inductance of 401 pH at a frequency of 28.1 GHz, which leads to instability. While the S-probe analysis is based on S-parameter simulations, the instability can also be demonstrated by means of transient simulations as shown in Fig. 4.11b-4.11c. The period of about 36 ps of the oscillation in the transient mode corresponds to the frequency for which the stability limit is exceeded in the S-probe analysis. Whereas the transient simulation is time and memory consuming, the S-probe analysis is efficient and significant. Furthermore, it has to be stated that the stability analysis based on the K-factor does not detect the instability and mistakenly leads to the result of an unconditionally stable circuit. The result of the corresponding K-factor simulation is illustrated in Fig. 4.12. Similar scenarios are described in [56].

4.4 Measurement Results

The fabricated chipset consists of a stimulus and a receiver IC and is shown in Fig. 4.13. The chips consume areas of 2870 m × 1870 m and 3006 m × 1870 m, respectively. However, the actual chip area demand of the stimulus is as

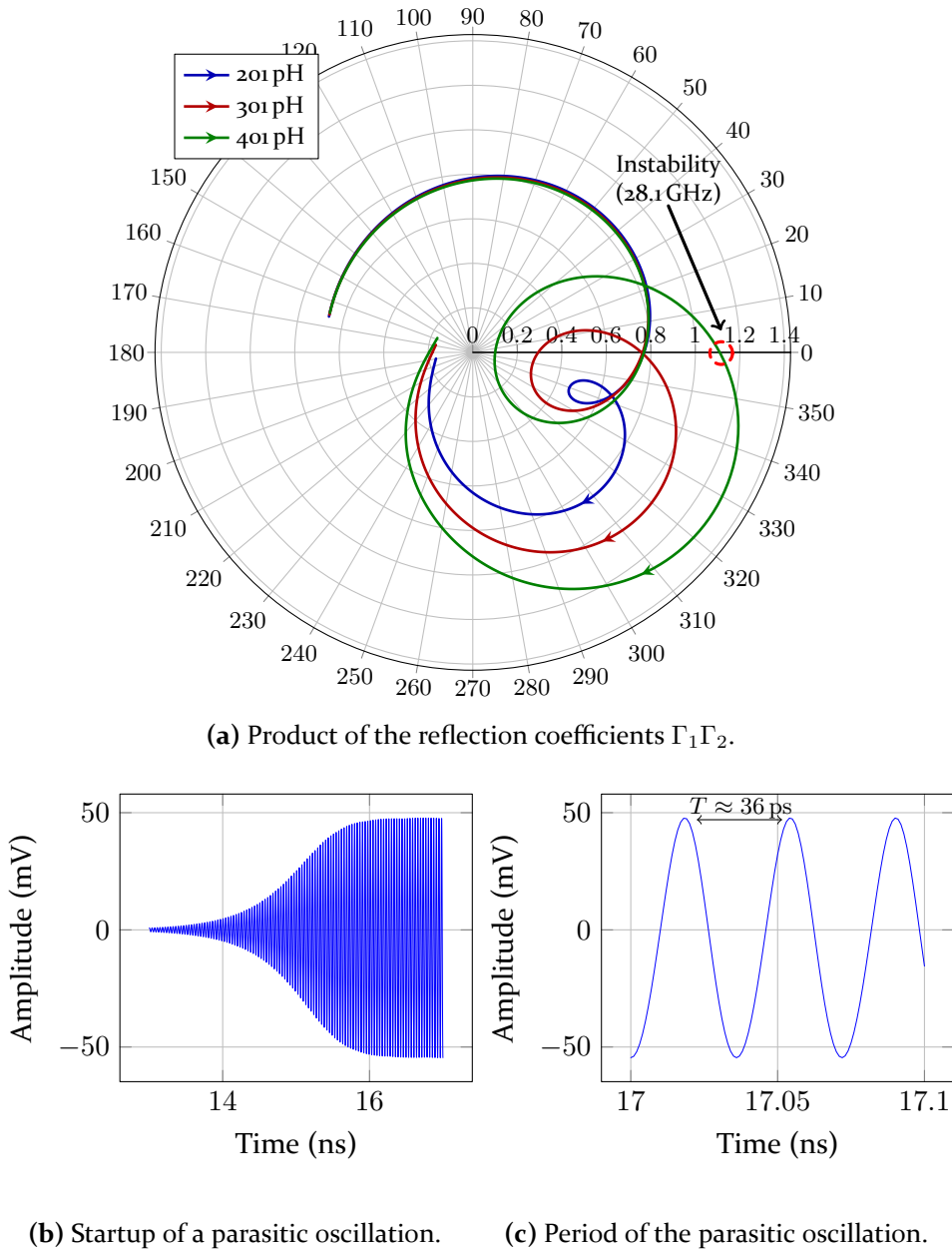


Figure 4.11: Stability analysis of the variable gain amplifier at the interface of emitter follower and cascode based on the product of the reflection coefficients $\Gamma_1\Gamma_2$ from 1 GHz to 110 GHz (Fig. 4.11a) and a transient simulation showing the start-up (Fig. 4.11b) and the period (Fig. 4.11c) of a parasitic oscillation.

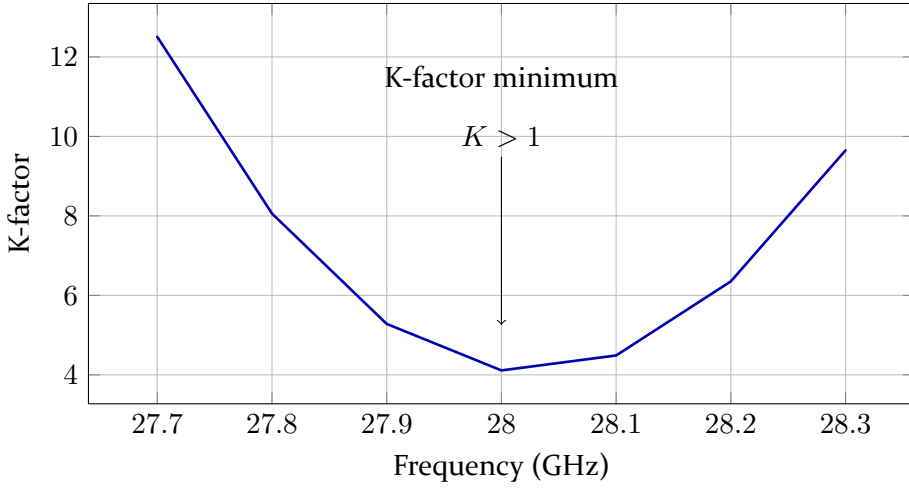
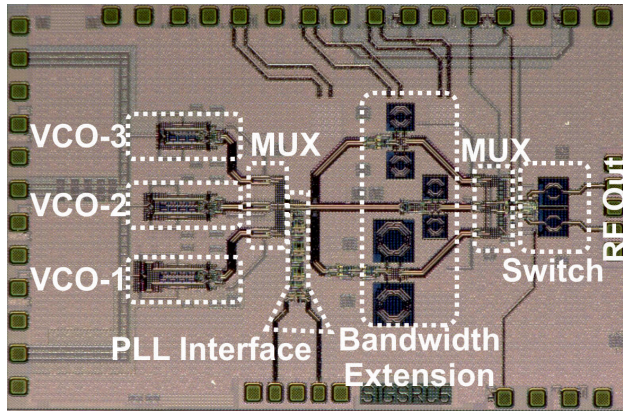


Figure 4.12: Minimum of the K-factor for the unstable VGA within the frequency range from 1 GHz to 110 GHz.

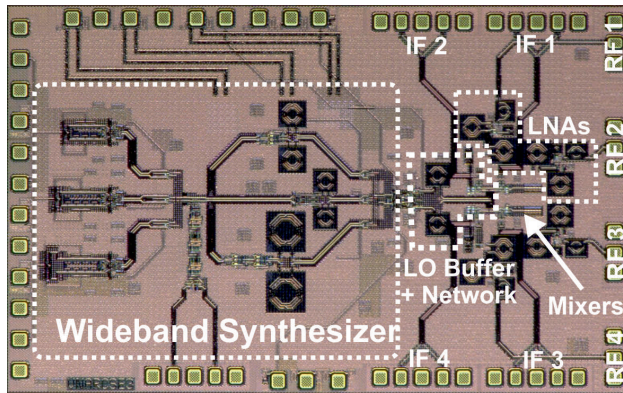
low as $2500 \text{ m} \times 1300 \text{ m}$, but due to chipsize restrictions a larger die was fabricated. Depending on the selected frequency band the stimulus IC draws a current of 209 mA to 218 mA from a 3.3 V supply (690 mW to 719 mW), the receiver draws 351 mA to 360 mA (1.16 W to 1.19 W). In addition, a version of the receiver without the LNAs, but with the mixers only, was realized. The circuits were fabricated in Infineon's B7HF200 SiGe technology described in section 2.4.

4.4.1 Synthesizer

The designed synthesizer was implemented for two purposes. In its function as a LO for the 4-channel receiver it is used with a differential output. In its function as stimulus for the DUT it is used with the presented differential to single-ended converting SPDT switch. The measured tuning characteristic of the VCO array included in both variants is illustrated in Fig. 4.14. The three frequency bands overlap slightly and the relative tuning ranges reach values of 26.6 %, 24.8 % and 28.2 % for tuning voltages from 2.5 V to 8.5 V. The fact that the tuning sensitivity (K_{VCO}) reaches its maximum values at the lower ends of the tuning ranges is caused by the tuning characteristic of the used varactors. The absolute tuning ranges for the lower frequency bands can be derived from the fundamental ones by division with the factors of two and four. In the same manner the K_{VCO} is reduced by the corresponding factor. The phase noise performance of the synthesizer was determined by measure-



(a)



(b)

Figure 4.13: Micrographs of the fabricated chipset: (a) stimulus MMIC, (b) 4-channel receiver MMIC. [28] (Copyright 2016, IEEE)

ments with a signal source analyzer at the low frequency PLL interface. The phase noise at the fundamental VCO frequencies (HIGH) was calculated by adding 18.02 dB. Thus, the effect of the frequency prescalers with the factor of eight is taken into account. In the same manner the phase noise of the frequency bands MED and LOW was determined by adding 12.04 dB and 6.02 dB, respectively. The measured single-sideband phase noise at 1 MHz offset from the carrier frequency is illustrated in Fig. 4.15 for the selectable frequency bands. The phase noise increases towards higher frequencies over all bands and VCOs, whereas the characteristic curve of each VCO is mainly affected by the varactor properties. Within the fundamental frequency band (HIGH) the phase noise is between -96.3 dBc/Hz and -103.2 dBc/Hz. The

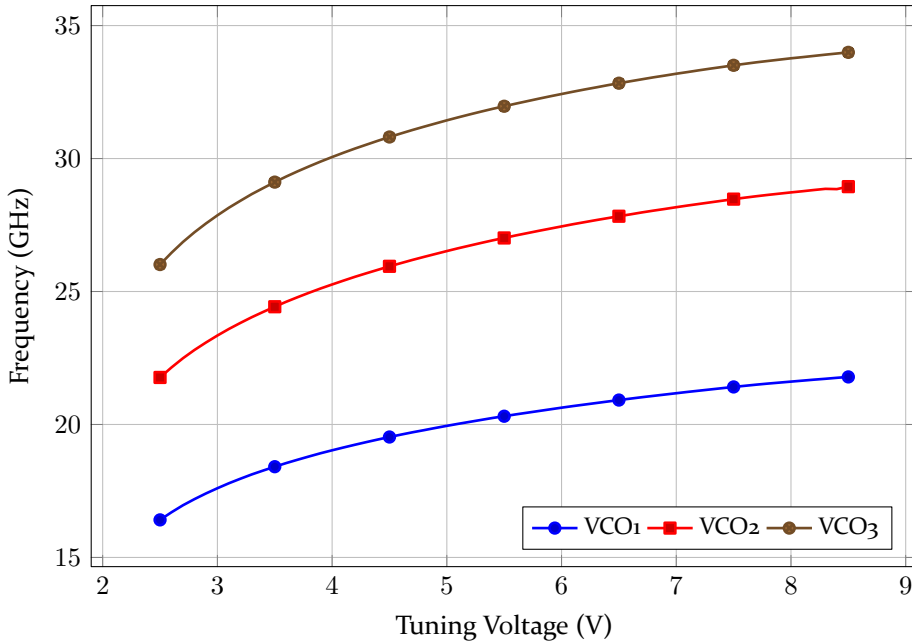


Figure 4.14: Measured tuning characteristic of the three designed VCOs. [28] (Copyright 2016, IEEE)

overall minimum is -115.3 dBc/Hz at a frequency of 4.95 GHz.

The output power over the full bandwidth was measured using GSGSG probes with one port connected to a spectrum analyzer and one port terminated with the system impedance of $50\ \Omega$. The measured values were corrected by adding 3 dB in order to obtain the differential output power. In addition, the losses of the measurement setup including coaxial cables, adapters, etc. were characterized and taken into account over the full measurement bandwidth. The control voltage of the VGA was adjusted in order to achieve the maximum available output power. The measurement results are shown in Fig. 4.16. The synthesizer provides an average output power of 2.5 dBm over the full bandwidth with a minimum of -0.7 dBm at 17.0 GHz and a maximum of 4.8 dBm at 27.0 GHz. The ripples in the measured output power are caused by the third-order Chebyshev filters implemented in the post-synthesis bandwidth extension stage.

Variable Gain Amplifier

The multistage VGA was fabricated additionally as a separate circuit in order to verify its functionality and the applied design procedure. In its function

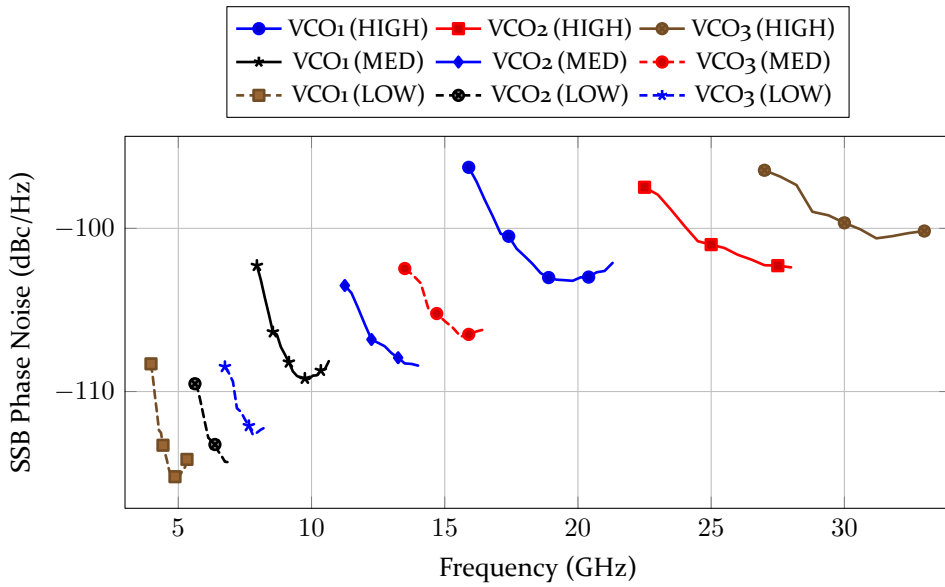


Figure 4.15: Measured single-sideband phase noise at 1 MHz offset from the carrier versus oscillation frequency for the nine selectable frequency bands. [28] (Copyright 2016, IEEE)

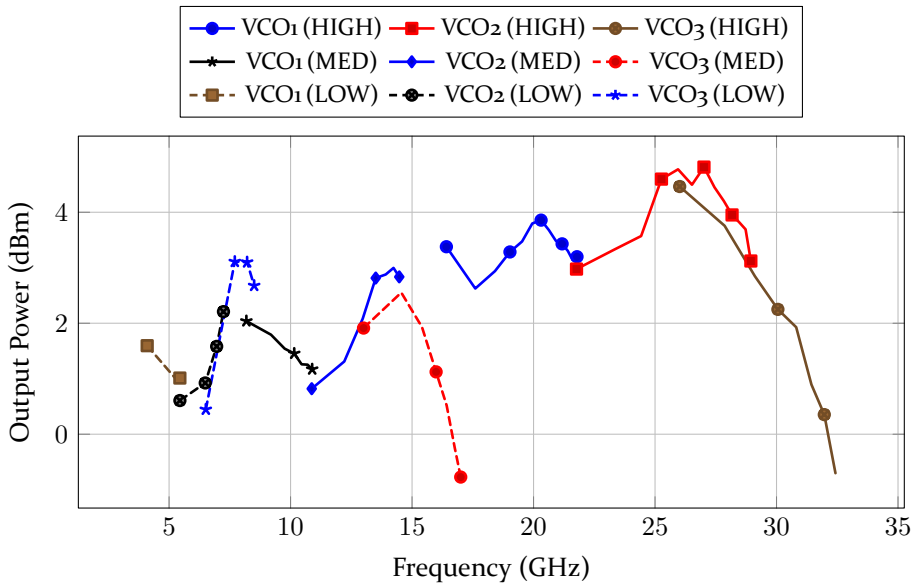


Figure 4.16: Measured differential synthesizer output power for the nine selectable frequency bands. [28] (Copyright 2016, IEEE)

as a differential amplifier it is used as a final stage of the synthesizer in the receiver in order to drive the down-conversion mixers. The measured and simulated S-parameters are shown in Fig. 4.17 for differential operation and a gain control voltage of $V_{GC} = 0$ V. Up to 33.8 GHz, the amplifier maintains a gain of more than 10 dB, whereas it reaches a peak gain of 15.1 dB at 27.1 GHz. The output matching is better than 20 dB for frequencies lower than 25.3 GHz and exceeds 10 dB below 30 GHz. The input was matched to the output of the precedent stage, which differs from the system impedance of $50\ \Omega$. The variable gain capability is demonstrated by the measurements shown in Fig. 4.18. The gain can be diminished by minimally 10.3 dB at 4 GHz and by maximally 16.1 dB at 27.7 GHz by applying control voltages between 1.85 V and 2.3 V. The amplifier's dynamic range at the upper design frequency of 32 GHz is 13.4 dB. Furthermore, the amplifier provides a maximum 1 dB compression point referred to the output of 8.4 dBm (12 GHz) and a minimum of 0.9 dBm at 32 GHz, as shown in Fig. 4.19.

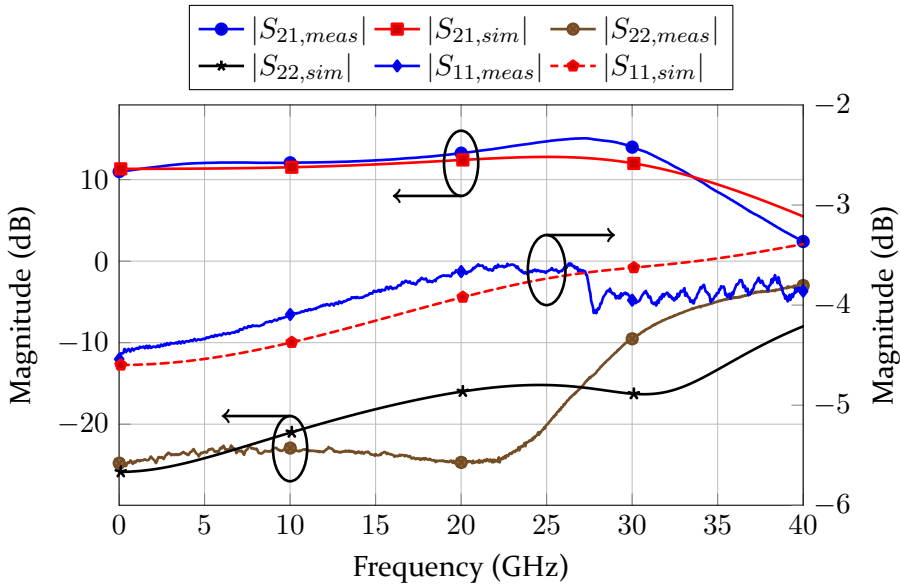


Figure 4.17: Measured and simulated scattering parameters of the broadband variable gain output stage in differential mode with the gain control voltage $V_{GC} = 0$ V. [28] (Copyright 2016, IEEE)

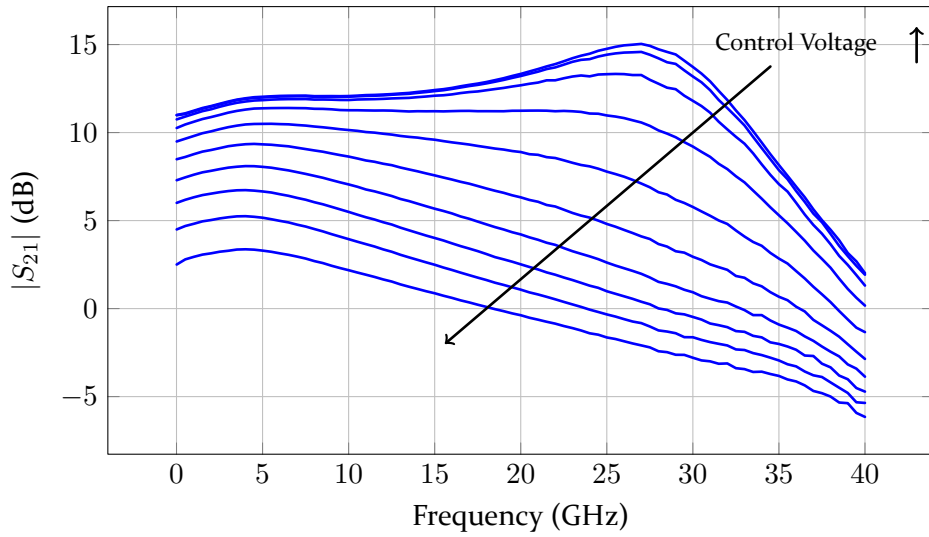


Figure 4.18: Measured gain control functionality of the broadband variable gain stage in differential operation for control voltages from 1.85 V to 2.3 V in steps of 50 mV. [28] (Copyright 2016, IEEE)

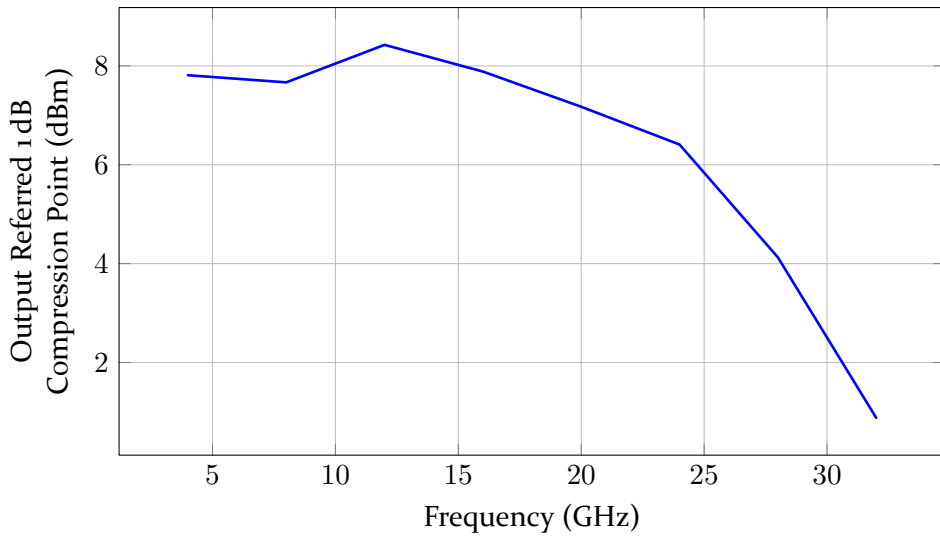


Figure 4.19: Measured output referred 1 dB compression point versus input frequency of the differential variable gain amplifier. [28] (Copyright 2016, IEEE)

Source Switch

The variant of the VGA with differential-to-single-ended conversion and switch capabilities was fabricated on a separate die as well. It was characterized as a 3-port device with one differential input (Port 1) and two single-ended outputs of which one port was selected (Port 2) and one port was deactivated (Port 3). The measurements were performed with a 4-port VNA with two ports set to the true stimulus mode to provide a true differential excitation of the input. The resulting S-parameters which were obtained with the switch set to maximum gain are illustrated in Fig. 4.20. The single-ended gain of the SPDT VGA is between 9.0 dB at 4 GHz and 12.8 dB at 29.7 GHz. Due to the high attenuation of the deactivated port of -43.3 dB at 4 GHz and -10.8 dB at 29.7 GHz a minimum isolation of 23.6 dB can be guaranteed. At lower frequencies the isolation between the two output ports is as high as 52.2 dB. As the source switch mainly determines the VNA's source match, the single-ended output impedance of both ports is of central importance. The activated port provides an output matching better than 20 dB up to 24.1 GHz and better than 10 dB up to 33.4 GHz. The deactivated port is matched better than 20 dB up to 22.7 GHz and better than 10 dB up to 29.7 GHz. Both parameters contribute to a low magnitude of the error parameters of the VNA chipset.

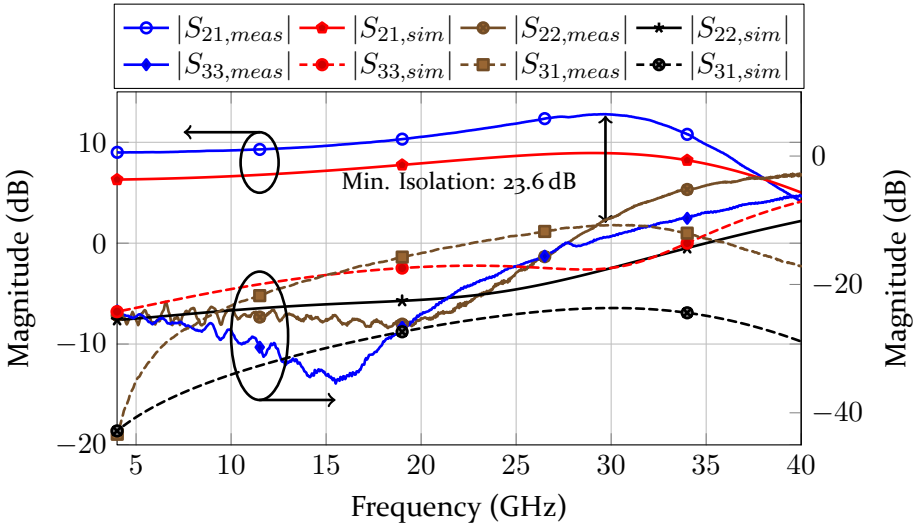


Figure 4.20: Measured unbalanced gain (S_{21}) of the activated port, unbalanced attenuation (S_{31}) of the deactivated port and unbalanced input matching (S_{22} , S_{33}) of the SPDT VGA with one channel turned on and one channel off. [28] (Copyright 2016, IEEE)

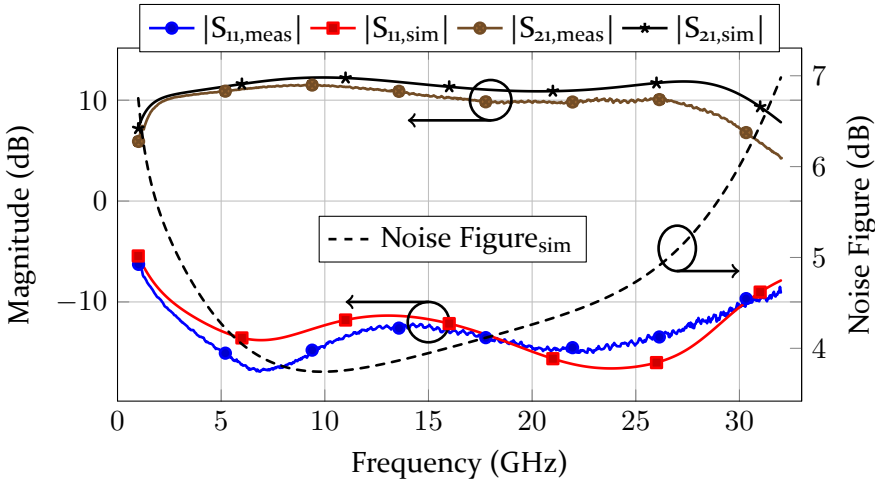


Figure 4.21: Measured and simulated S-parameters and noise figure of the designed low noise amplifier. [28] (Copyright 2016, IEEE)

4.4.2 Receiver

Three different receiver components were fabricated and characterized separately. Besides the wideband LNA, two 4-channel receiver circuits including the designed wideband frequency synthesizer were realized. A high linearity version contains the mixer and the LO only, whereas the low noise variant contains the LNA as a first receiver stage.

Low Noise Amplifier

The designed wideband LNA was fabricated as a stand-alone component for verification purposes. The circuit draws a current of 23 mA from a 3.3 V supply voltage, of which 14 mA are consumed by the cascode stage. The simulated and measured scattering parameters show good agreement, as illustrated in Fig. 4.21. Due to the flat gain as well as the good input impedance matching, the amplifier is very well suited as a first stage of the receiver for the integrated VNA. The input matching is better than 10 dB from 2.2 GHz to 30.6 GHz and better than 9 dB over a multi-octave bandwidth. Especially good input matching of 16.9 dB and 15.0 dB is achieved at 6.9 GHz and 22.7 GHz, respectively. The measured gain is higher than 10 dB from 2.4 GHz to 26.4 GHz. Over the frequency range from 4 GHz to 32 GHz, an average gain of 9.97 dB is achieved. The simulated noise figure reaches its minimum of 3.7 dB at 9.9 GHz and increases up to 7.0 dB at 32 GHz.

4-Channel Receiver

The receivers contain four identical channels with identical circuits and layouts. The conversion gain for an intermediate frequency of 500 MHz was measured with a signal generator at the RF input and a spectrum analyzer at the low frequency output. The losses within the measurement setup were determined by calibration measurements and were taken into account for the measured results. The wideband frequency synthesizer was tuned to a fixed offset of 500 MHz from the signal generator's frequency. The conversion gain over the full bandwidth is shown in Fig. 4.22 for both fabricated variants.

The effect of the LNA clearly leads to a higher conversion gain of the receiver circuit. Both variants show a decreasing gain versus the input frequency, which is caused by the low-pass transfer function of the micromixer circuit. The measured gain ranges from 18.0 dB at 5 GHz to 5.5 dB at 32 GHz and from 9.7 dB at 3 GHz to 2.8 dB at 31 GHz for the high gain and the low gain variant, respectively. The corresponding average gain over the input frequency is 14.4 dB and 7.7 dB. Thus, receivers with high gain containing a LNA and a mixer are more suitable for the measurement channels, whereas the version with a mixer only is appropriate for the reference channel due to a higher linear input range. The measured conversion gain versus the intermediate frequency at a fixed input frequency of 15 GHz is shown in Fig. 4.23. The high gain variant provides conversion gains from 16.3 dB at an IF of 100 MHz to 10.6 dB for an IF of 3 GHz. The low gain variant ranges between 8.2 dB and 2 dB. Not only because of the higher conversion gain, but also due to the relaxed requirements for the following analog digital converter (ADC) low intermediate frequencies are advantageous.

The linearity of the circuits was measured in terms of the input referred 1 dB compression point, as the operation within the linear range is essential for VNA applications (see Fig. 4.24). The high gain receiver allows a maximum input power of -18.9 dBm, -15.8 dBm and -14.0 dBm at 4 GHz, 15 GHz and 28 GHz. At the same frequencies the low gain variant provides higher 1 dB compression points of -9.3 dB, -8.4 dB and -6.5 dB. Consequently, the mixer only version is a good choice for high linearity applications and for the characterization of DUTs with high output powers such as amplifiers. Furthermore, they can be utilized as reference channel receivers in order to prevent compression in these channels.

In the same manner as the output matching of the stimulus, the input matching of the receivers directly affects the VNA error parameters. The measured

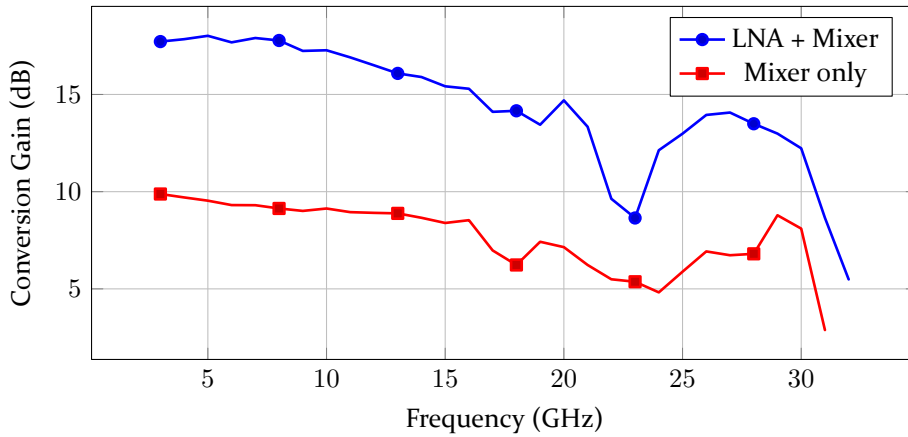


Figure 4.22: Measured conversion gain of the fabricated receiver circuits for an IF of 500 MHz. [28] (Copyright 2016, IEEE)

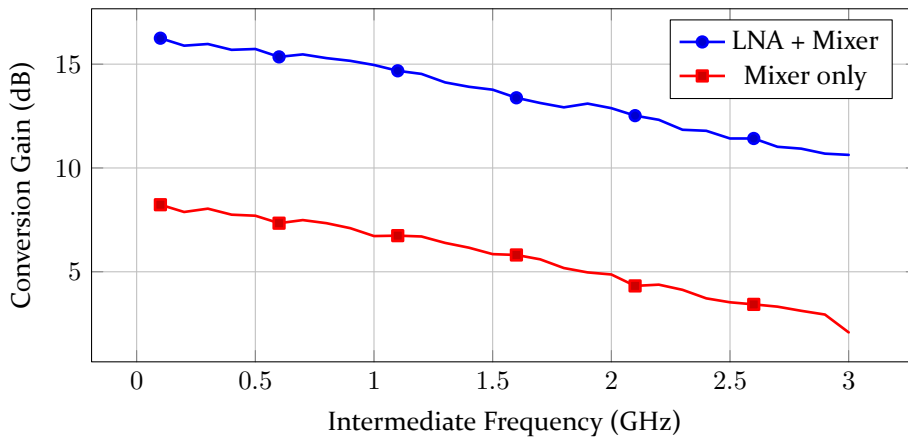
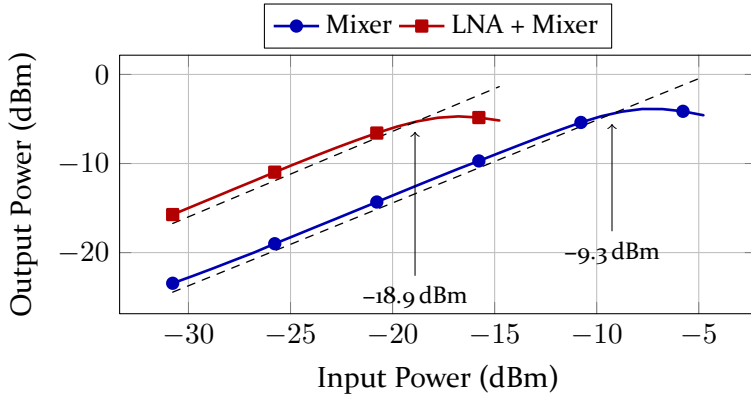
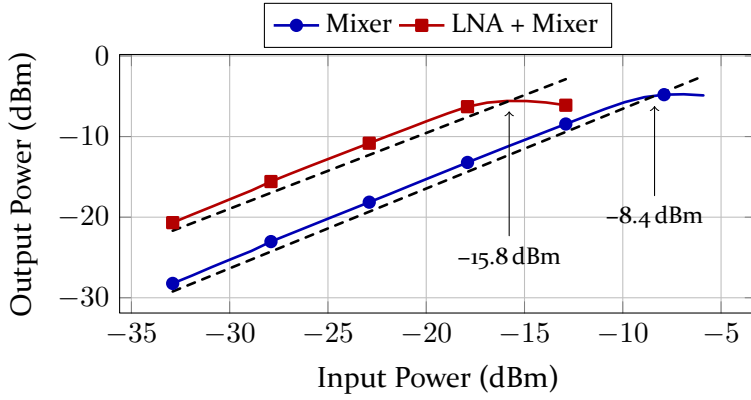


Figure 4.23: Measured receiver conversion gain versus the intermediate frequency at an input frequency of 15 GHz. [28] (Copyright 2016, IEEE)

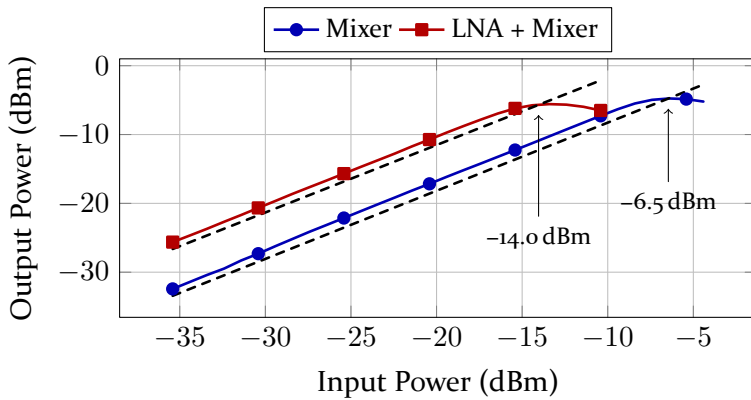
input matching of both variants is shown in Fig. 4.25 for the four channel receivers. Minor deviations between channel 1/4 and channel 2/3 are caused by slightly different transmission line lengths at the inputs. The mixer only version provides an input matching of better than 18 dB up to 32 GHz. The LNA version shows a slightly decreased input matching of 10 dB from 2.6 GHz to 26.7 GHz. Consequently, both versions contribute to a calibration with low error parameters using the proposed chipset.



(a) RF Input Frequency: 4 GHz

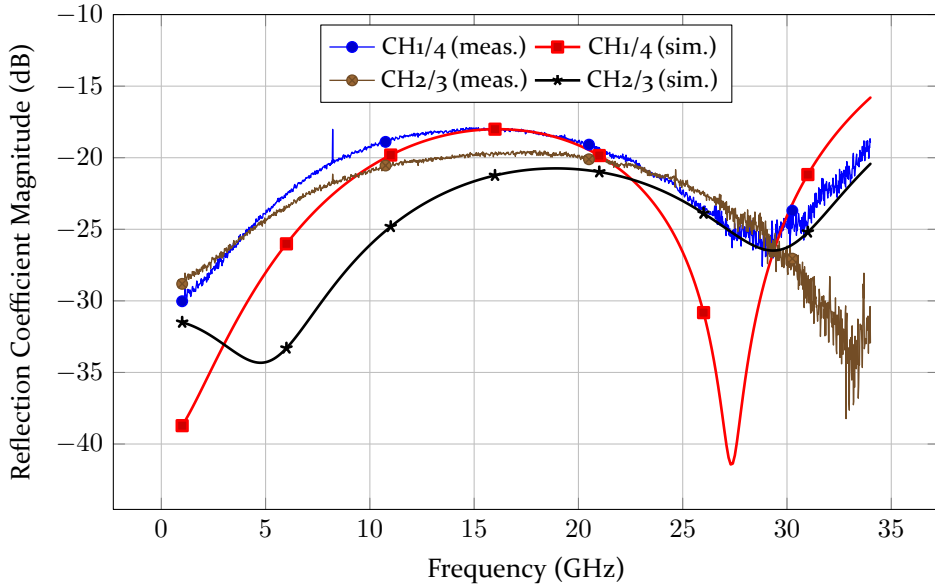


(b) RF Input Frequency: 15 GHz

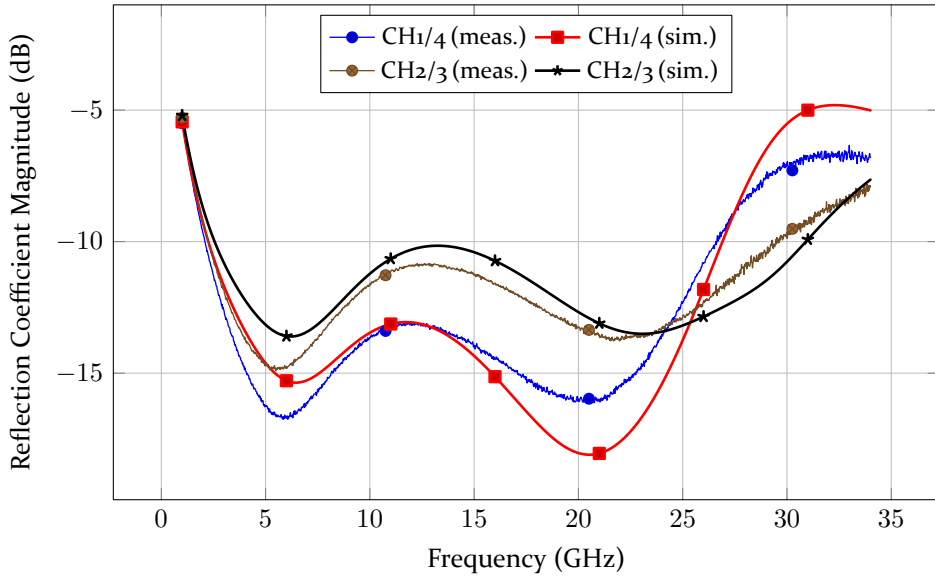


(c) RF Input Frequency: 28 GHz

Figure 4.24: Measured gain compression of the two fabricated receiver circuits at 4 GHz, 15 GHz and 28 GHz. [28] (Copyright 2016, IEEE)



(a) Configuration: Micromixer.



(b) Configuration: Micromixer and LNA.

Figure 4.25: Measured and simulated reflection coefficient of the 4-channel receivers. [28] (Copyright 2016, IEEE)

4.5 Hybrid Integrated VNA Demonstrator Circuit

A fully functional VNA was implemented according to the architecture shown in Fig. 4.1. The realization in hardware is illustrated in Fig. 4.26. In addition to the proposed chipset, wideband directional couplers are required for the wave separation and for the derivation of a reference channel. The design of directional couplers covering the aimed multi-octave bandwidth is challenging and requires special techniques, which are based on a non-uniform characteristic of the coupling coefficient versus the coupler length of multiples of a quarter-wavelength [57]. The resulting geometries show large physical dimensions and can not be integrated due to their length of 40.4 mm. Consequently, the couplers were fabricated on a multi-layer microwave PCB onto which the chipset was mounted and connected via bond-wires. The coupled lines are vertically stacked so that the stimulus IC and the receiver IC were placed on opposite sides of the board. One port of each coupler was terminated by high frequency capable $50\ \Omega$ SMD resistors. The measurement ports can be interfaced by 2.4 mm coaxial connectors so that arbitrary devices can be characterized. The intermediate frequencies are fed via SMA connectors to the subsequent ADCs and the signal processing unit. The two integrated signal sources were phase-locked to the same reference frequency by two separate PLLs.

The setup can be utilized for 2-port measurements by using a digital back-end with a sufficient number of coherently sampled ADC channels. In addition, the 2-port calibration algorithms have to be implemented and the

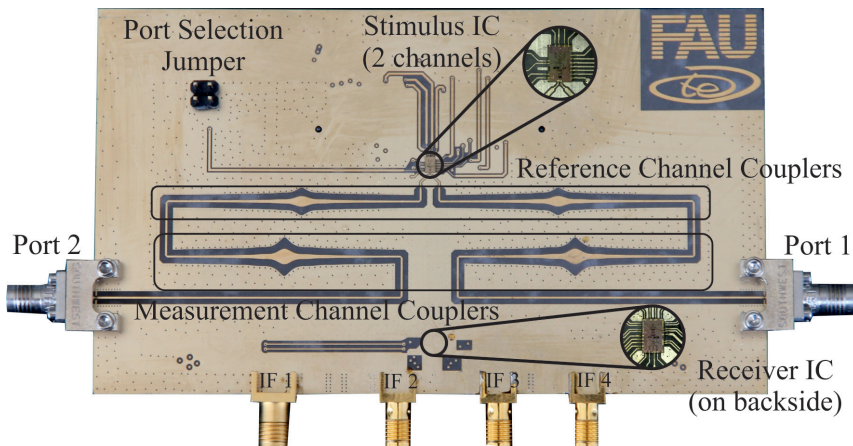
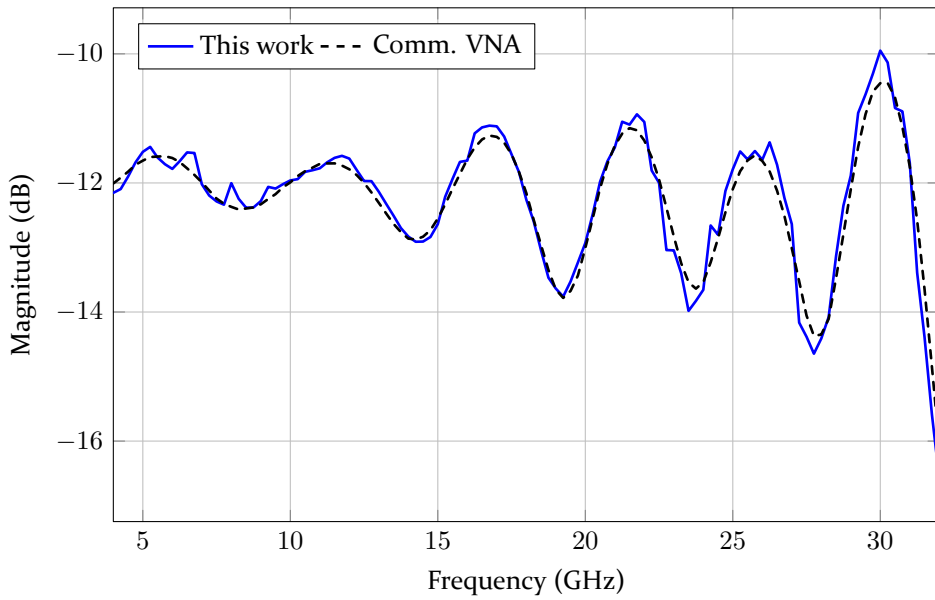
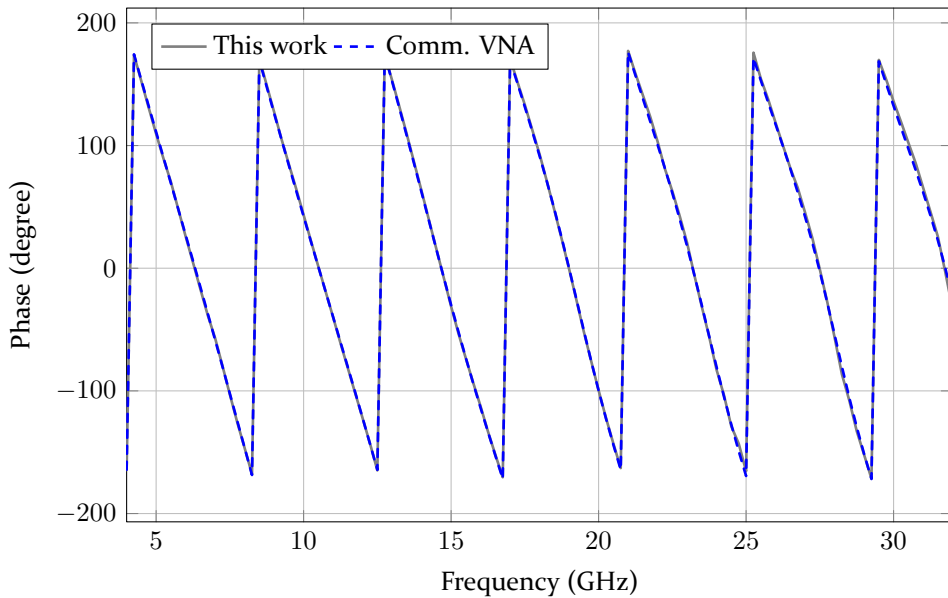


Figure 4.26: Hybrid integrated VNA incorporating the proposed chipset and a test-set of wideband couplers. [28] (Copyright 2016, IEEE)



(a) Magnitude of reflection coefficient.



(b) Phase of reflection coefficient.

Figure 4.27: Measured complex reflection coefficient of an arbitrary DUT compared to the results from a commercial VNA. [28] (Copyright 2016, IEEE)

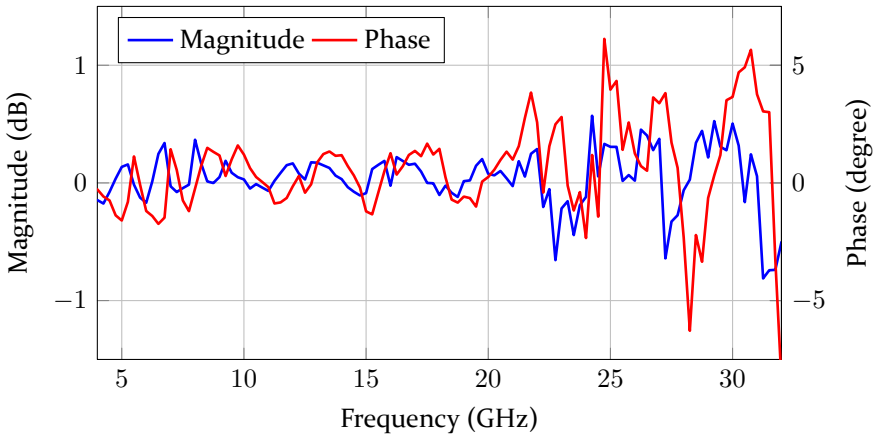


Figure 4.28: Ratio between the measurement results of the miniaturized and the commercial VNA. [28] (Copyright 2016, IEEE)

corresponding types and numbers of mechanical calibration standards have to be available. However, in this case the system was utilized as a 1-port vector network analyzer as a proof-of-concept. From a functional point of view, the main difference consists in the evaluation of two more differential measurement channels. Thus, coaxial calibration standards were used to apply a 3-term error calibration and to set the reference plane to the end of a coaxial cable. Finally, the complex reflection coefficient of an arbitrary DUT was determined and compared to the results, which were obtained with a commercial vector network analyzer (see Fig. 4.27). The ratio between the measurement results of the miniaturized and the commercial VNA was calculated and illustrated in Fig 4.28. The average deviation between the two measurements is 0.18 dB in magnitude and 1.8° in phase. These deviations are within a range, which can be attributed to the repeatability of coaxial connectors, the phase variation of the cables and minor deviations between the used calibration standards. Consequently, an excellent agreement between the measurements of two independent measurement systems can be stated, which is a validation of the concept of integrated vector network analyzers.

4.6 Conclusion

A chipset for a highly integrated full 2-port vector network analyzer from 4 GHz to 32 GHz was presented. This chipset can be used for various self-sustained sensing applications in medical and industrial environments. The presented heterodyne topology was shown to be the ideal architecture for the integration of the VNA with single-ended measurement ports. A VCO

array based wideband frequency synthesizer with an average output power of 2.5 dBm and a single-sideband phase noise between -96.3 dBc/Hz and -115.3 dBc/Hz was utilized as local oscillator and as a stimulating signal source. The octave-wide VCO array is comprised by three cross-coupled oscillators with tuning ranges of 26.6 %, 24.8 % and 28.2 %. Design techniques for a wideband variable gain amplifier with a gain of more than 10 dB as well as a SPDT source switch with a minimum isolation of 23.6 dB were applied. Circuits for a single-ended wideband LNA with an average gain of 9.97 dB and a micromixer circuit were shown and used for the composition of a 4-channel receiver. A design technique for wideband stability based on S-probes was demonstrated and applied to all circuits. The concept of integrated VNAs was validated by a hybrid integrated demonstrator. The presented chipset was combined with wideband directional couplers and acts as a fully functional vector network analyzer. The complex reflection coefficient of an arbitrary DUT was measured and showed an excellent agreement with the measurements obtained with a commercial vector network analyzer. The average difference is 0.18 dB in magnitude and 1.8° in phase.

Chapter 5

Highly Integrated 4-32 GHz 2-Port VNA

5.1 Introduction

Most commonly, VNAs are used as instrumentation devices for research and development or test purposes. Due to their size, weight and cost the devices are generally not applicable to portable applications. Targeting a highly integrated 2-port network analyzer, a 4-32 GHz chipset incorporating a switch integrated stimulus and a four-channel receiver was presented in Chapter 4. The successful demonstration for reflection coefficient measurements was demonstrated by means of a hybrid assembly utilizing wideband directional couplers implemented on a multilayer RF PCB. However, additional external components are required for a self-sustained operation of the miniaturized network analyzer. In the following, a self-sustained, wideband, highly integrated 2-port VNA will be described. In addition to the chipset, phase lock techniques for the stimulus and LO as well as an IF backend are necessary. Further miniaturization of the instrument can be achieved by the integration of a compact, wideband testset.

5.2 Fully Integrated VNA Frontend

The challenge in the design of an integrated RF frontend for a VNA is the integration of the following three core components: 1. signal generator, 2. directive element, 3. multi-channel receiver. The signal generator is required to provide a stimulus in order to excite a DUT over a certain frequency range. The directive element separates the waves which travel towards the DUT

from the reflected waves. In addition, it can be applied for the derivation of reference channels. A multi-channel receiver is needed in order to determine the magnitude and the phase of measurement and reference channels. In particular, the ambition to cover a bandwidth of three octaves from 4 GHz to 32 GHz demands innovative technical approaches.

The design and verification of an integrated wideband signal generator including a source switch and the corresponding 4-channel receiver was described in detail in Chapter 4. To the best of the author's knowledge, the first fully integrated 2-port VNA system on chip (SoC) is presented in the following. This includes the design of the first integrated multi-section directional couplers.

5.2.1 System-On-Chip

A block diagram of the fully integrated 4 GHz to 32 GHz 2-port VNA frontend is illustrated in Fig. 5.1. The system on chip is based on the building blocks described in Chapter 4 and was realized in the SiGe bipolar technology B7HF200 provided by Infineon Technologies, described in section 2.4. The VNA frontend chip is comprised of the essential building blocks signal generation, wave separation by directive elements and a 4-channel receiver. Due to the application of the balanced-to-unbalanced converting source switch and the unbalanced RF inputs of the micromixer, an unbalanced VNA port interface could be realized. This is advantageous for several reasons. Not only most external measurement objects are unbalanced, but also differential off-chip routing via coaxial cables introduces massive phase imbalances due to unequal physical lengths. In addition, an unbalanced measurement interface saves chip area, as the number of required directive elements is halved. Both the stimulus signal generator and the local oscillator rely on VCO arrays followed by bandwidth extension stages. Both signal sources feature a divide-by-eight prescaler output, as well as shared tuning voltages for embedding in an external PLL. The frequency bands can be selected by applying the appropriate control voltages to the multiplexer circuits. The variable gain amplifiers, as well as the source switch can be controlled by external analog tuning voltages. Four broadband directional couplers were realized as a cascade of three coupled line segments. The couplers C_1 and C_3 are used in order to derive a reference channel from the stimulus whereas C_2 and C_4 serve as directional elements for the separation of stimulating and reflected waves. One port of each coupler is terminated with the system impedance of $50\ \Omega$. A large series capacitor is utilized as a broadband DC block. Instead of placing it in series with the couplers' direct

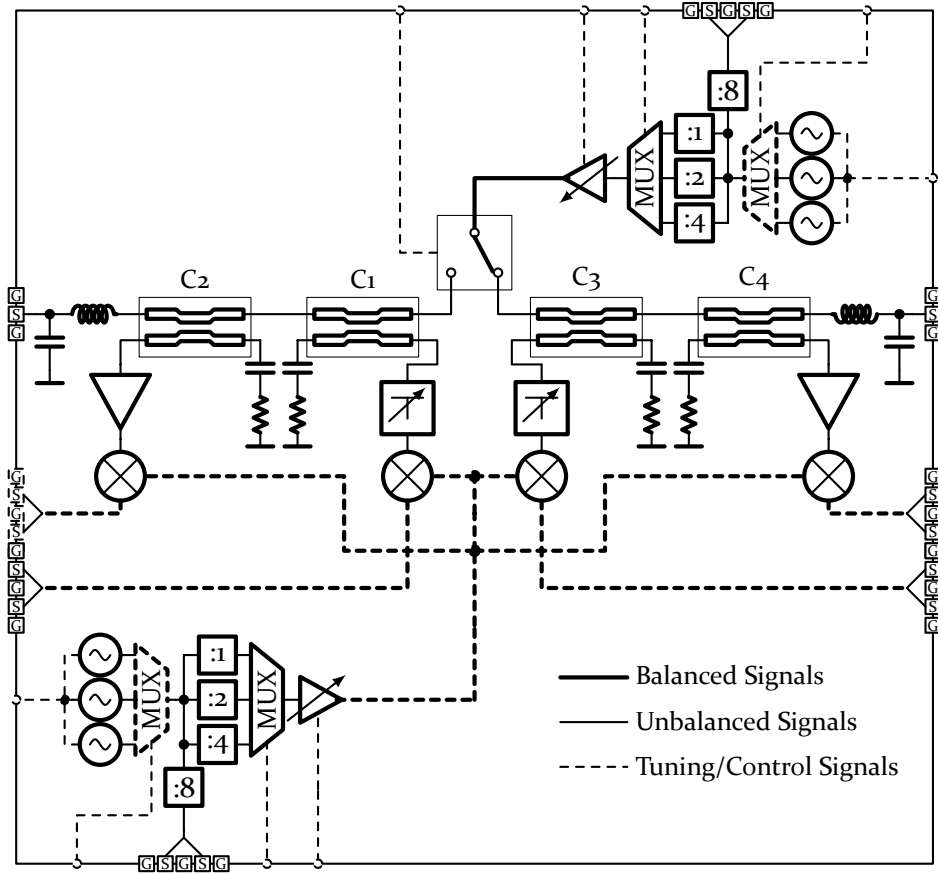


Figure 5.1: Simplified block diagram of the fully integrated 4 GHz to 32 GHz 2-port VNA frontend IC including the essential interfaces. [58] (Copyright 2016, IEEE)

or coupled path, it was placed at the terminated ports in order to prevent a high-pass transfer function of the measured signals. Nevertheless, the capacitor was chosen large enough so that the VNA port matching is not degraded significantly. The measurement channels are comprised of a wideband LNA as well as a down-conversion micromixer in order to guarantee a high sensitivity regarding the measurement signals, which are attenuated by lossy off-chip components. On the contrary, the reference channels are equipped with variable attenuators instead of the LNAs in order to secure the linear operation of the down-conversion micromixers even at high stimulus output powers. The attenuators were realized as a set of resistive T-networks with attenuations of 3 dB, 6 dB and 9 dB. They are adjustable by aluminum laser

fuses and can be set to values from 0 dB to 18 dB. All receiver channels are driven differentially by a wideband local oscillator, which is realized in the previously described architecture. The local oscillator distribution network was designed to be electrically short and absolutely symmetrical in order to guarantee an equal power split between the micromixers. The four differential intermediate frequencies can be connected via two GSGSGSGSG pads with a pitch of 100 μm . The two unbalanced VNA measurement ports can be interfaced via two GSG pads with a pitch of 100 μm as well. In order to account for the mismatch, which is created by the pad capacitance and the sum of minor mismatches of the stimulus, the receiver and the couplers, a lowpass matching network is utilized. In this way, a good VNA port matching is guaranteed, which is essential for low VNA error parameters.

A photograph of the fabricated system on chip is shown in Fig. 5.2. The chip covers an area of 4006 $\mu\text{m} \times 4006 \mu\text{m}$. However, the effectively occupied area is lower. A larger size was chosen in order to meet the chip size standards. The system on chip draws a current between 565 mA and 575 mA from a 3.3 V supply voltage (1.86 W to 1.90 W), depending on the selected frequency band.

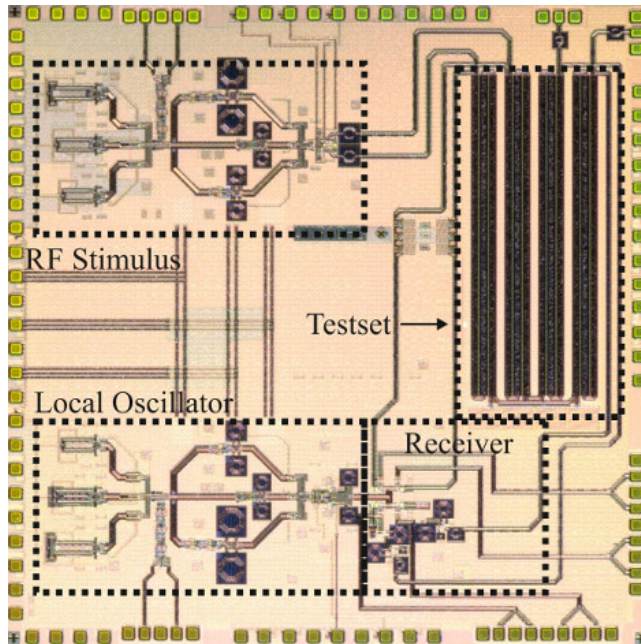


Figure 5.2: Photograph of the fabricated 4 GHz to 32 GHz 2-port VNA frontend. The system on chip consumes an area of 4006 $\mu\text{m} \times 4006 \mu\text{m}$. [58] (Copyright 2016, IEEE)

5.2.2 Wideband Directive Element

Directive elements are key components of each VNA. Besides providing an appropriate coupling factor, a high directivity is required in order to minimize the VNA error terms. In the context of this work there are the additional challenges of a multi-octave bandwidth and a small form factor for integration on a chip.

Multisection Couplers

The voltage coupling function $S_{31}^i(f)$ of the i -th quarter-wavelength coupled line section with the coupling coefficient k_i and the center frequency f_c is given by

$$S_{31}^i(f) = \frac{jk_i \tan(\frac{\pi}{2}f/f_c)}{\sqrt{1 - k_i^2 + j \tan(\frac{\pi}{2}f/f_c)}} \quad (5.1)$$

with $S_{31}^i(f) = S_{13}^i(f) = S_{24}^i(f) = S_{42}^i(f)$ whereas the transmission $S_{21}^i(f)$ is

$$S_{21}^i(f) = \frac{\sqrt{1 - k_i^2}}{\sqrt{1 - k_i^2 \cos(\frac{\pi}{2}f/f_c) + j \sin(\frac{\pi}{2}f/f_c)}} \quad (5.2)$$

with $S_{21}^i(f) = S_{12}^i(f) = S_{34}^i(f) = S_{43}^i(f)$. Due to the quarter-wavelength geometry the coupler is inherently bandwidth limited. For instance, the relative bandwidth of a -10 dB coupled-line coupler is as low as 100.3 % referred to its center frequency. The bandwidth can be increased by cascading N coupled-line segments with different coupling coefficients k_i , as shown in Fig. 5.3a for three segments ($N = 3$). In general, multi-section couplers are built up symmetrically with tight coupling at the center and loose coupling towards the outer segments. The signal flow chart in Fig. 5.3b illustrates that the coupling function for the cascade $S_{31}^c = b_0/a_0$ results from numerous superposed paths with several feedback loops. With port 2 being terminated, $a_3 = 0$ can be assumed so that the wave a_2 can be expressed as:

$$a_2 = b_1 \frac{S_{31}^3 S_{21}^2}{1 - S_{31}^2 S_{31}^3} \quad (5.3)$$

With (5.3) it can be shown that the incident wave portion b_1 is:

$$b_1 = a_0 \frac{S_{21}^1}{1 - S_{31}^2 S_{31}^3 - \frac{S_{31}^3 S_{21}^2 S_{31}^3 S_{21}^2}{1 - S_{31}^2 S_{31}^3}} \quad (5.4)$$

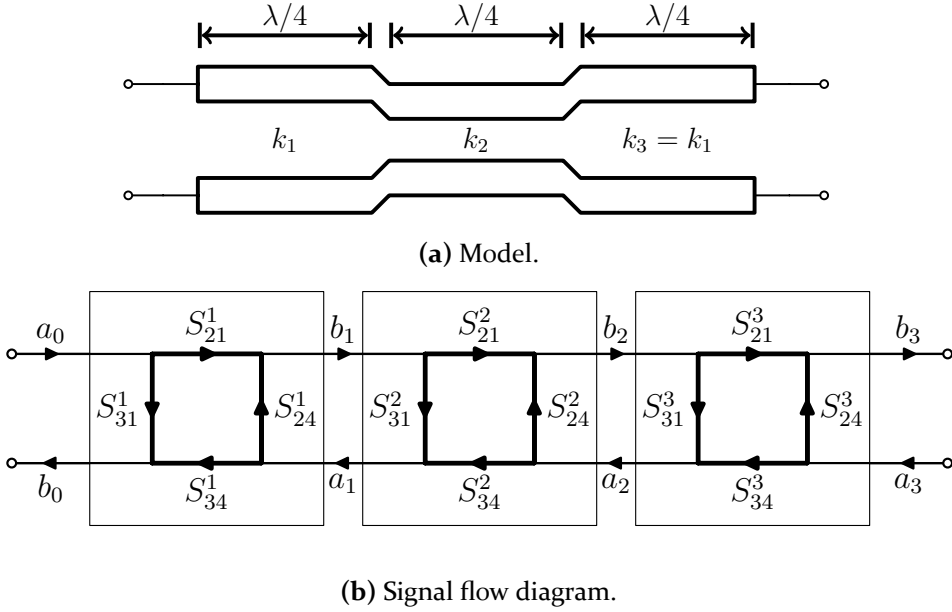


Figure 5.3: Model and signal flow diagram of a wideband directional coupler consisting of three coupled quarter-wavelength coupled lines. [58] (Copyright 2016, IEEE)

In combination with

$$a_1 = b_1 S_{31}^2 + a_2 S_{21}^2 \quad (5.5)$$

the voltage coupling function of the three-stage directional coupler can be given in a closed form

$$S_{31} = \frac{b_0}{a_0} = S_{31}^1 S_{31}^2 b_1 + S_{21}^1 S_{21}^2 a_2 \quad (5.6)$$

as a function of the transmission and coupling coefficients of the single sections. Ideally matched and lossless components are assumed for this derivation. By choosing the N coupling coefficients k_i appropriately, wideband coupling functions with different transfer characteristics can be achieved [59]. The resulting transfer functions for -10 dB couplers with $N = 1, 3, 5, 7, 9$ sections are illustrated in Fig. 5.4. The coupling functions of a single section ($N = 1$) and a three-stage coupler ($N = 3$) are given by (5.1) and (5.6), respectively. The coupling of longer cascades is obtained by evaluating the corresponding signal flow charts and the resulting equations. For the sake of simplicity the derivations are omitted here. It is shown that the bandwidth of multi-section couplers increases with the number of segments. However,

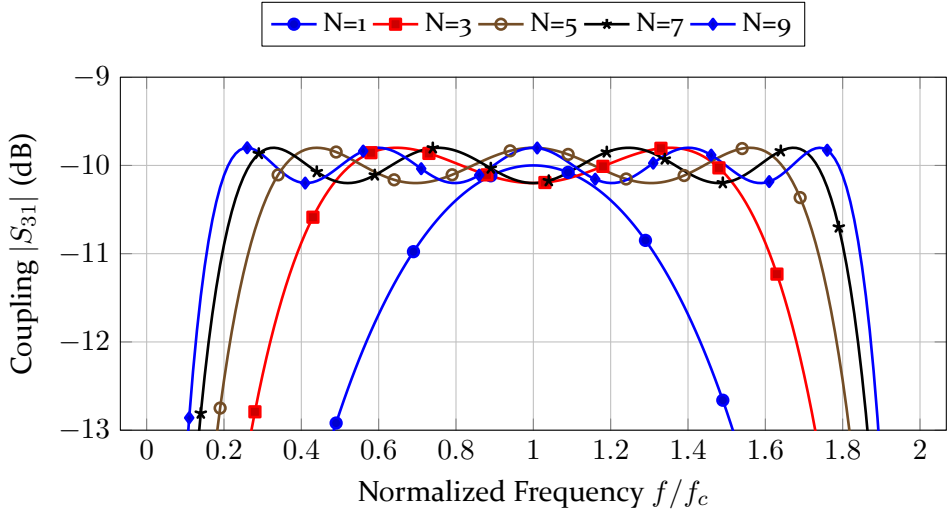


Figure 5.4: Coupling function of -10 dB couplers with different cascade lengths. [58] (Copyright 2016, IEEE)

the most significant increase is observed between the single-section ($N = 1$) and the three-stage coupler ($N = 3$). Due to the limited chip area, the three-section coupler represents a good trade-off between covered chip area and bandwidth.

Design and Layout

In order to realize a three-stage directional coupler with an average coupling $|S_{31}| = -10$ dB, two quarter-wavelength coupled line sections were designed. The central section provides a strong coupling of $k_2 = 0.45$, whereas the outer segments are coupled loosely with $k_1 = k_3 = 0.14$. The design center frequency of all sections is $f_c = 18$ GHz, which leads to a physical length of the quarter-wavelength sections of $2000 \mu\text{m}$ in silicon-dioxide. A cross section view of the designed segments is illustrated in Fig. 5.5. Both variants consist of two parallel coupled lines. In order to realize the strong coupling of the inner segment, the lines are separated by the minimum allowed distance of $2.4 \mu\text{m}$. Additionally, a defected ground plane is utilized to provide the maximally achievable coupling of parallel coupled lines in the used technology in a 50Ω -environment. The line width was determined by 2.5D electromagnetic field simulations of the even and odd mode impedances. The outer segment is composed of two parallel coupled lines with a spacing of $9.6 \mu\text{m}$. The interconnects between the segments were kept short in order

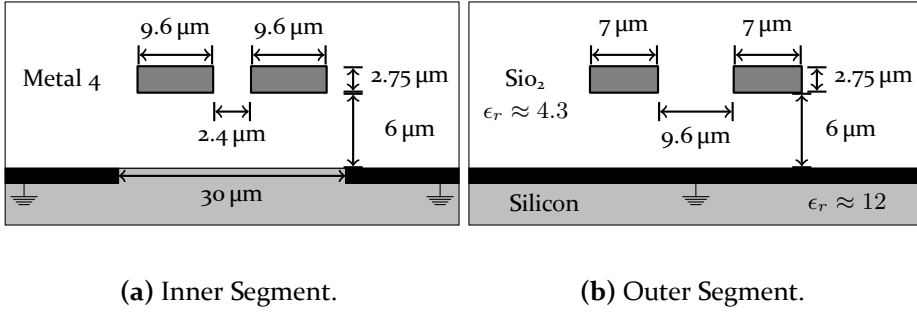


Figure 5.5: Cross section of the designed quarter-wavelength coupled line segments. [58] (Copyright 2016, IEEE)

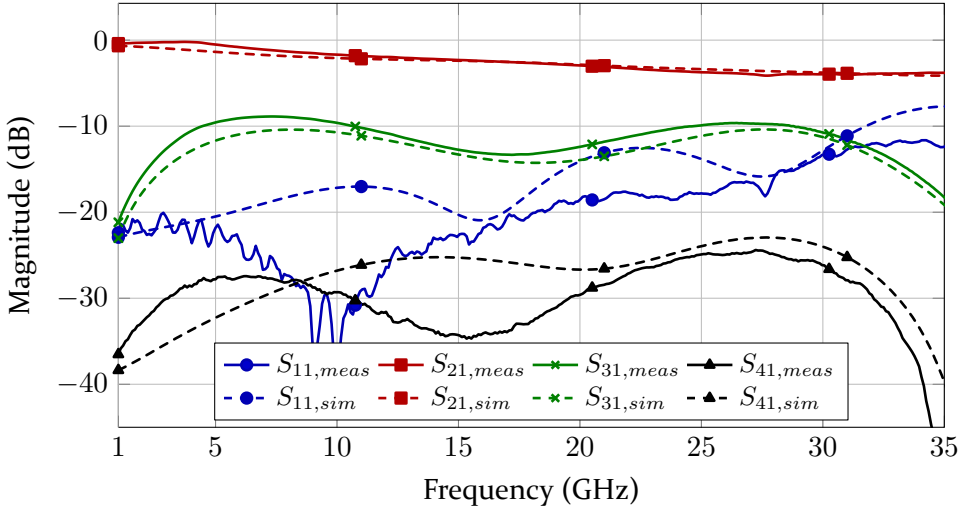


Figure 5.6: Measured and simulated S-parameters of the fabricated wideband directional coupler consisting of three cascaded coupled line segments. [58] (Copyright 2016, IEEE)

to avoid additional phase shift which significantly degrades the coupler performance. The multi-section coupler was fabricated as circuit breakout and characterized by an on-wafer S-parameter measurement. The results show very good agreement with the simulation and are illustrated in Fig. 5.6. The return loss is better than 15 dB up to 28.9 GHz and better than 12 dB over the full operational bandwidth. The insertion loss (IL) increases with frequency due to substrate and conductor loss mechanisms and reaches 3.9 dB at 32 GHz. A broadband coupling function with an average of -10.8 dB was achieved. The absolute and relative 3 dB-bandwidth is 30.2 GHz and 168 %, respectively.

respectively. This is an improvement of 18.7 GHz and 67.7 % compared to the single-section -10 dB coupler. The ripples are the result of the chosen coupling coefficients and are acceptable for VNA applications. Furthermore, a high isolation of 24 dB is maintained over the full bandwidth, which leads to a minimum directivity of 14.7 dB.

5.2.3 Double Directional Coupler Testset

As shown in Fig. 5.1, the VNA testset used for the derivation of reference and measurement channels is composed of two three-stage directional couplers. In general, a single directive element can fulfill the task. However, the double directional coupler topology provides a stronger isolation between reference and measurement channel as reflections from the receiver inputs are mainly absorbed by the port terminations instead of being scattered towards the other channel. The layout of the testset is illustrated schematically in Fig. 5.7. The single sections were realized with straight coupled lines instead of meandered lines, as meandering reduces the length of the sections only, but increases the total consumed chip area. The single coupled segments are placed in parallel, which results in the minimum achievable occupied chip area. The interconnects between the sections were kept electrically short whereas the interconnect between the couplers is not critical and phase shifts can be tolerated. One port of each coupler was terminated and blocked for direct current. The measurement results are shown in Fig. 5.8 and are in good agreement with the simulations. The return loss of the complete structure is better than 14.4 dB up to 32 GHz whereas the insertion loss is 8.0 dB at 32 GHz due to the long physical length of the cascaded elements. The average coupling over the bandwidth from 4 GHz to 32 GHz is -11.4 dB, which leads to absolute and relative bandwidths of 30.4 GHz and 170 %, respectively. Because of the termination of two ports an isolation of better

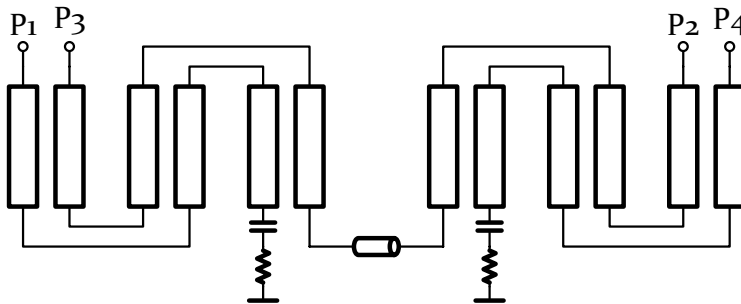


Figure 5.7: Schematic layout of the double directional coupler VNA testset. [58] (Copyright 2016, IEEE)

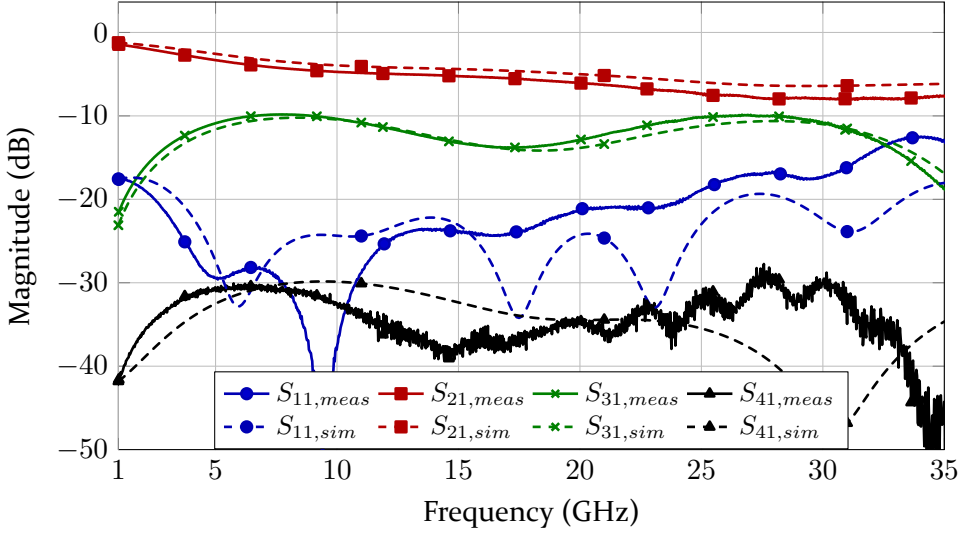


Figure 5.8: Measured and simulated S-parameters of the fabricated wideband test-set consisting of two 3-section couplers. [58] (Copyright 2016, IEEE)

than 27.8 dB can be achieved, which corresponds to a minimum directivity of 18.6 dB.

5.3 2-Port VNA System Realization

The miniaturized 2-port VNA system was realized by five different modules leading to a self-sustained measurement device covering a bandwidth from 4 GHz to 32 GHz. A simplified block diagram is illustrated in Fig. 5.9 and a photograph of the physical implementation is shown in Fig. 5.10. The VNA frontend chip is mounted on a high frequency compatible substrate (RF PCB) via wire bonds. The integrated circuit was placed in a cavity, which leads to a minimized return loss due to short and even bond wires. The measurement ports are routed via shielded coplanar transmission lines to 2.4 mm coaxial connectors. The RF PCB is connected via a high frequency multipin connector with the Backend-/PLL-Board. Phase-lock of the integrated VCOs is guaranteed by routing the prescaled outputs (2 GHz to 4 GHz) of the two integrated wideband frequency synthesizers to two external ICs containing fractional dividers and phase-detectors. Active loop filters consisting of two stages are required to map the phase detector output voltage range (0 V to 5 V) to the tuning voltage range of the used integrated VCO array (2.5 V to 8.5 V), which was described in section 4.3.1. The phase-locked loops are completed by a single crystal oscillator with a frequency of 50 MHz, which acts as

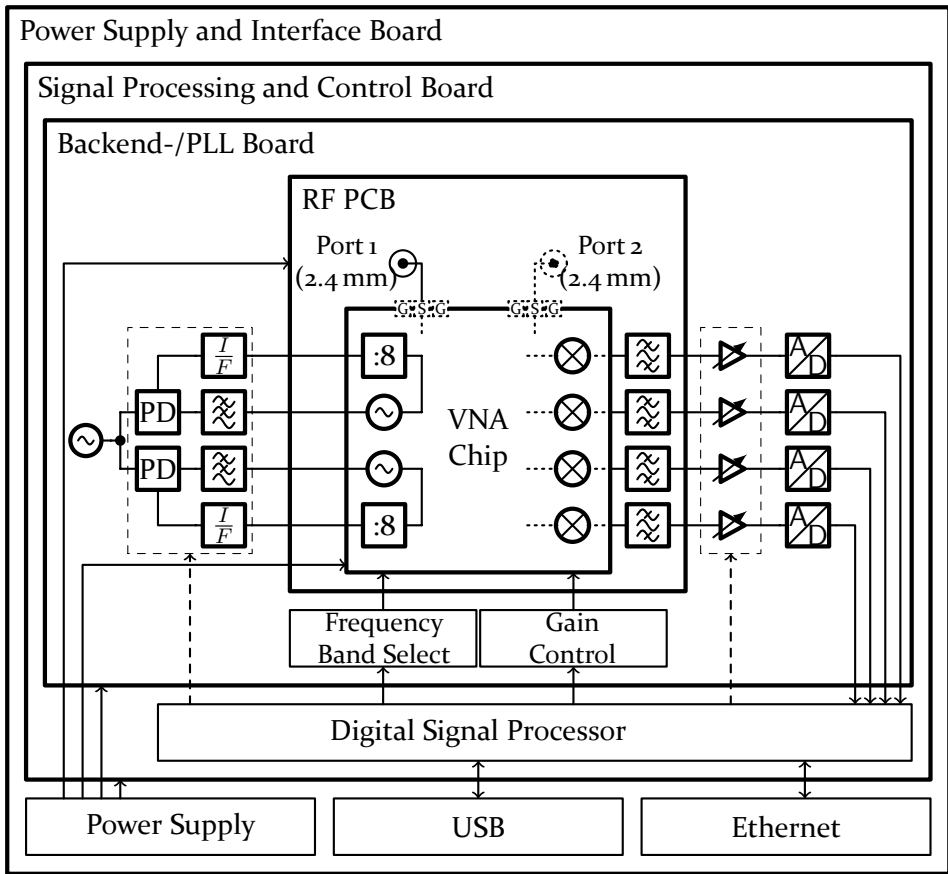


Figure 5.9: System overview of the modularly implemented miniaturized 2-port vector network analyzer. [58] (Copyright 2016, IEEE)

a reference oscillator for both loops. In this way, it is guaranteed that the RF stimulus and the local oscillator are synchronized to each other. Due to the heterodyne realization of the VNA both loops will be set with a frequency offset corresponding to the intermediate frequency of the measurement signals. Besides the PLLs a digital backend is realized on the board. Variable gain amplifiers are used to scale the measurement signals to the maximum scale of the ADCs so that quantization noise is minimized. The ADCs provide a resolution of 14 bit at a sampling rate of 600 kHz and an analog input bandwidth of 80 MHz. Digital down-conversion effects via subsampling can be exploited as aliasing can be neglected for the most part due to the single-frequency operation of the VNA. A matrix of analog switches is used to select

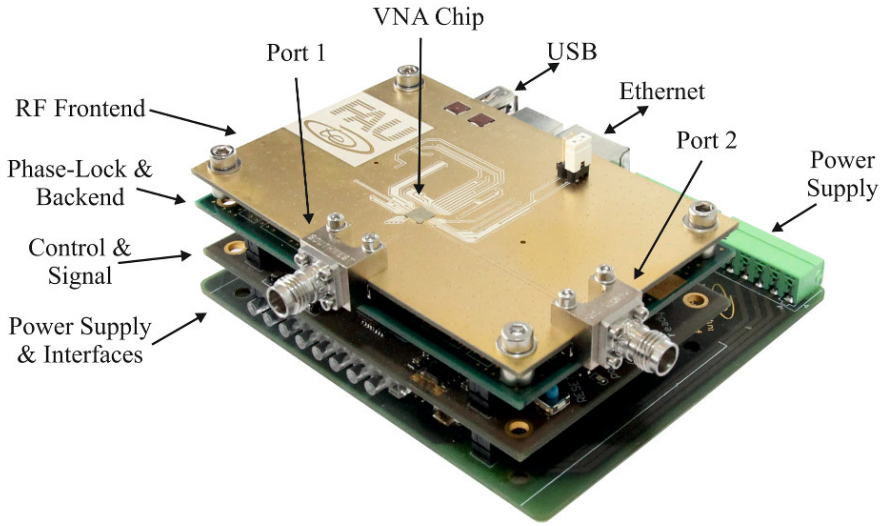


Figure 5.10: Photograph of the miniaturized 2-port VNA system realization. [58]
(Copyright 2016, IEEE)

the desired frequency band of the integrated synthesizers. The output power of the RF stimulus and the local oscillator can be controlled via DACs. A digital signal processor (DSP) is mounted on a separate board and controls the PLLs, the frequency band select and the output powers, as well as the variable gain IF amplifiers. In addition, it reads the digitized samples of the four measurement channels. It can also be used for IQ-demodulation of the IF signals, which is fundamental for the calibration of the VNA. The processor itself can be controlled via USB and Ethernet interfaces, which are realized on a separate board. Furthermore, the DSP provides the measurement results to arbitrary display units via these interfaces. Additionally, power supply units for all boards and its components including the VNA chip are realized. The PCBs can be vertically stacked so that a compact form factor is achieved. The resulting physical dimensions of the 2-port multi-octave VNA are 100 mm×90 mm×38 mm.

5.4 Calibration

The 2-port VNA was calibrated using the 16-term error model illustrated in Fig. 5.11. The error box is a 2-port network containing 16 unknown terms, one of which can be set to unity so that 15 terms have to be determined by calibration measurements. The used model is complete as it also considers crosstalk paths which are dependent on the DUT [34]. The correlation

between the true wave portions interacting with the DUT ($\vec{a}, \vec{b}$) and the measured waves ($\vec{m}$) is given by the 4×4 -matrix $\vec{C}$:

$$\begin{pmatrix} b_1 \\ b_2 \\ a_1 \\ a_2 \end{pmatrix} = \vec{C} \cdot \begin{pmatrix} m_1 \\ m_4 \\ m_2 \\ m_3 \end{pmatrix} \quad (5.7)$$

By splitting $\vec{C}$ into four 2×2 -matrices ($\vec{E}, \vec{F}, \vec{G}, \vec{H}$)

$$\vec{C} = \begin{pmatrix} \vec{G} & \vec{E} \\ \vec{F} & \vec{H} \end{pmatrix} \quad (5.8)$$

the correlation is given by a set of two vector equations:

$$\begin{aligned} \vec{G} \cdot \begin{pmatrix} m_1 \\ m_4 \end{pmatrix} + \vec{E} \cdot \begin{pmatrix} m_2 \\ m_3 \end{pmatrix} &= \begin{pmatrix} b_1 \\ b_2 \end{pmatrix} \\ \vec{F} \cdot \begin{pmatrix} m_1 \\ m_4 \end{pmatrix} + \vec{H} \cdot \begin{pmatrix} m_2 \\ m_3 \end{pmatrix} &= \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \end{aligned} \quad (5.9)$$

In combination with the DUT's scattering parameters $\vec{b} = \vec{S}\vec{a}$, this leads to:

$$\vec{G} \cdot \begin{pmatrix} m_1 \\ m_4 \end{pmatrix} + \vec{E} \cdot \begin{pmatrix} m_2 \\ m_3 \end{pmatrix} = \vec{S} \cdot \left\{ \vec{F} \cdot \begin{pmatrix} m_1 \\ m_4 \end{pmatrix} + \vec{H} \cdot \begin{pmatrix} m_2 \\ m_3 \end{pmatrix} \right\} \quad (5.10)$$

Equation (5.10) is valid for a single switch position. By introducing the measurement matrix

$$\vec{M} = \begin{pmatrix} m'_1 & m''_1 \\ m'_4 & m''_4 \end{pmatrix} \cdot \begin{pmatrix} m'_2 & m''_2 \\ m'_3 & m''_3 \end{pmatrix}^{-1} \quad (5.11)$$

(5.10) is extended so that both switch positions are represented whereas $m'_1 \dots m'_4$ and $m''_1 \dots m''_4$ correspond to the switch set to port 1 and port 2, respectively. Finally, the correlation between the 16 error terms of $\vec{C}$, the measured values $\vec{M}$ and the scattering parameters of the DUT is given by:

$$\vec{G} \cdot \vec{M} + \vec{E} = \vec{S} \cdot \left\{ \vec{H} \cdot \vec{M} + \vec{F} \right\} \quad (5.12)$$

Equation (5.12) can be used to determine the error terms. The measurement of each 2-port calibration standard with two switch positions delivers four equations. However, five calibration measurements are required in order

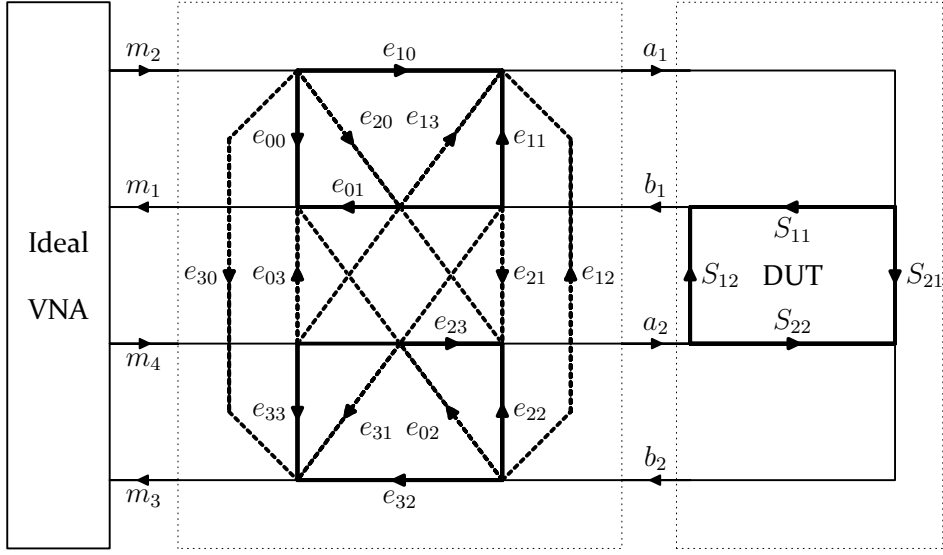


Figure 5.11: Graphical representation of the implemented 16-term calibration model. [58] (Copyright 2016, IEEE)

to solve the equation system because the resulting equations contain linear combinations. Once all error parameters are determined, the scattering parameters of the DUT are given by:

$$\vec{S} = \left\{ \vec{G} \cdot \vec{M} + \vec{E} \right\} \left\{ \vec{H} \cdot \vec{M} + \vec{F} \right\}^{-1} \quad (5.13)$$

In the following, the calibration will be performed by measuring one thru standard (T) and four combinations of the one port standards open (O), short (S) and match (M): MO, MS, OS, OM.

5.5 2-port VNA Measurements

In order to verify the functionality of the miniaturized VNA, different 2-port DUTs were characterized. Therefore, the complex 2×2 scattering matrix was determined utilizing the previously described 16-term calibration. A typical measurement setup including coaxial cables is illustrated in Fig. 5.12. In addition, each DUT was characterized by measurements with a commercially available VNA. Both devices were calibrated using the same set of coaxial calibration standards and the measured scattering parameters were compared to each other as a benchmark for the developed miniaturized VNA. Two different parameters are used to evaluate the quality of the miniaturized VNA.

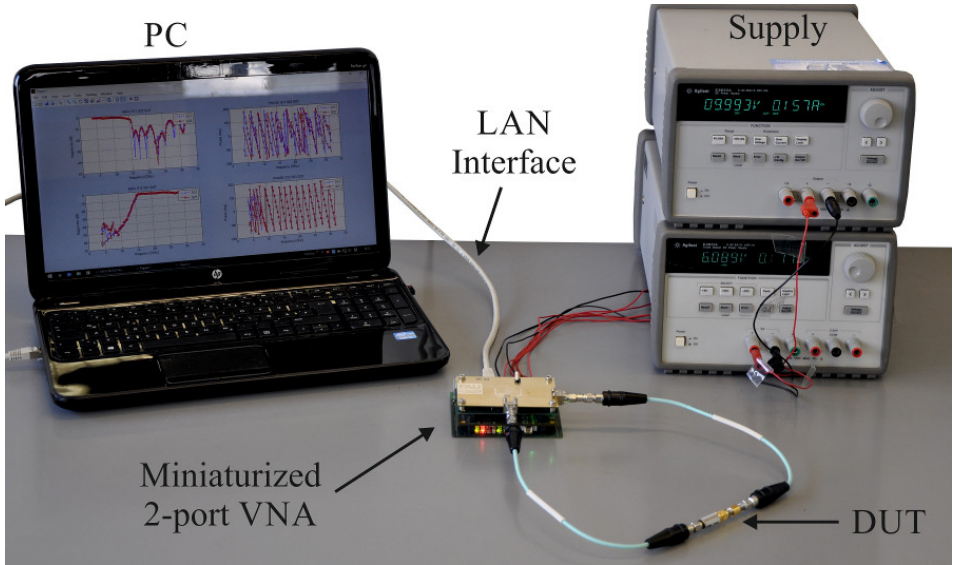


Figure 5.12: Measurement setup of the miniaturized 2-port VNA including a personal computer (PC) for the visualization of the measured results. [58] (Copyright 2016, IEEE)

Firstly, the scattering parameters based on their difference in logarithmic magnitude and their phase difference in degree. Mathematically, this corresponds to a normalization of the elements of measured scattering matrix S to the elements of the reference matrix S_{REF} in a linear scale. The resulting normalized matrix will be used in a logarithmic scale in order to compare the two measured scattering matrices and is given by:

$$\Delta S|_{dB} = 20 \log(S) - 20 \log(S_{REF}) \quad (5.14)$$

The matrix ΔS will be denoted as deviation from now on. From an engineering point of view, ΔS is an illustrative measure for the comparison of scattering parameters. However, it is not the true difference between the S-matrices. The latter is represented by the error vector, which is the complex difference between the two measurement results. It is the most appropriate and correct way to represent the similarity of two scattering matrices. On the other side, it is not as illustrative as the normalized logarithmic matrix ΔS . Consequently, both parameters are used within this thesis. The error vector matrix is given by:

$$EV = S - S_{REF} \quad (5.15)$$

5.5.1 DUT 1: Substrate Integrated Waveguide

The first characterized DUT is a substrate integrated waveguide section embedded between two microstrip transmission line sections on a PCB with coaxial connectors. The structure acts as a high-pass filter because the wave propagation within the waveguide is possible only above the cut-off frequency, which is defined by the waveguide geometry. Frequencies below the cut-off are reflected at the transition points between the waveguide and the microstrip lines. Due to the DUT's high dynamic range in both reflection and transmission coefficients, it was chosen in order to benchmark the proposed miniaturized VNA.

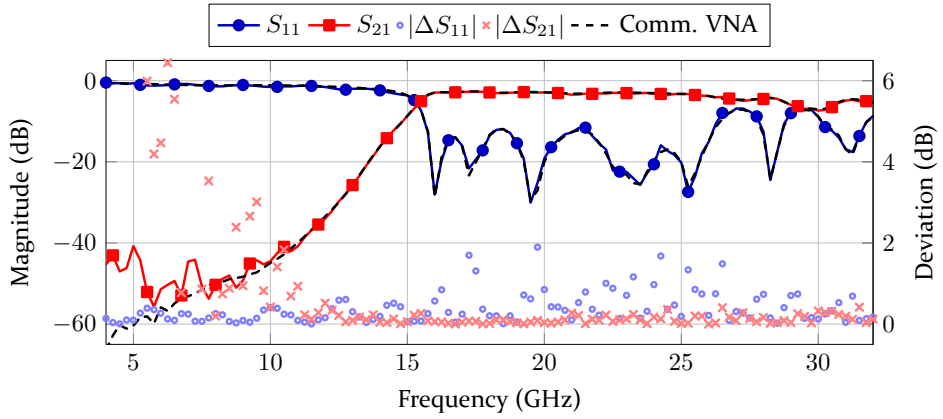
The measurement results for the complex 2×2 scattering matrix from 4 GHz to 32 GHz are illustrated in Fig. 5.13. In the interest of better readability the measured phase was plotted for the transmission parameters (S_{12} , S_{21}) only. In addition, the scattering parameters obtained with a commercial vector network analyzer, as well as the deviation between the two measured scattering matrices are shown.

The characterization of the DUT based on the measured scattering matrix S leads to almost identical DUT properties as with the reference measurements. For instance, the -3 dB corner frequency was determined to be 15.37 GHz for forward and 15.36 GHz for reverse transmission while the reference measurements lead to corner frequencies of 15.38 GHz for forward and 15.37 GHz for reverse transmission. Linear interpolation between the frequency points was applied for these investigations. Furthermore, the mean pass-band (16 GHz to 26 GHz) insertion loss was measured to be 3.08 dB for forward and 3.09 dB for reverse transmission. In contrast to this, the reference measurements result in a pass-band insertion loss value of 3.00 dB for both forward and reverse transmission. A more detailed overview of DUT properties including minimum and maximum values of the measured pass-band insertion loss is given in Table 5.1. In total, a very good agreement between the measurements can be observed. The mean deviation ΔS was determined to be

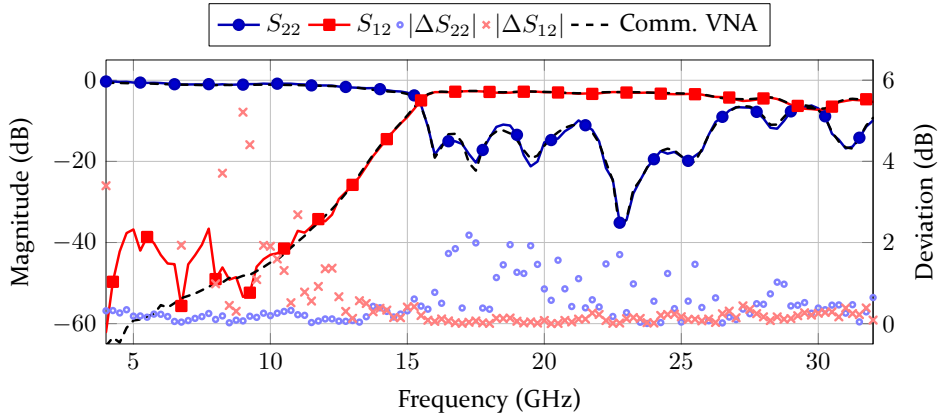
$$\begin{aligned} \overline{\Delta S_{11,4...32 \text{ GHz}}} &= 0.56 \text{ dB} / 4.69 \\ \overline{\Delta S_{22,4...32 \text{ GHz}}} &= 0.81 \text{ dB} / 7.62 \end{aligned} \quad (5.16)$$

for the reflection coefficients over the full bandwidth from 4 GHz to 32 GHz and

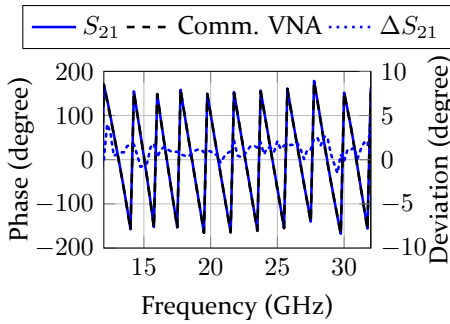
$$\begin{aligned} \overline{\Delta S_{12,10...32 \text{ GHz}}} &= 0.30 \text{ dB} / 1.83 \\ \overline{\Delta S_{21,10...32 \text{ GHz}}} &= 0.18 \text{ dB} / 1.28 \end{aligned} \quad (5.17)$$



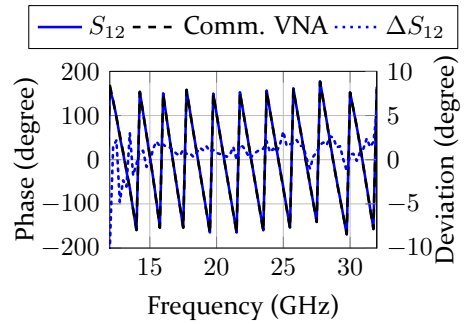
(a) Stimulation at port 1.



(b) Stimulation at port 2.



(c) Stimulation at port 1.



(d) Stimulation at port 2.

Figure 5.13: Measured complex reflection (S_{11} , S_{22}) and transmission (S_{12} , S_{21}) of a MSL-SIW transition and deviation from the reference measurement. [58] (Copyright 2016, IEEE)

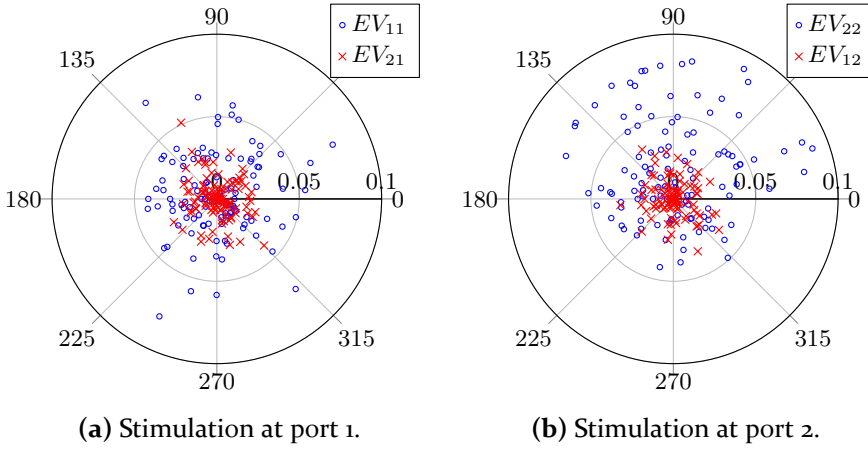


Figure 5.14: Measured error vectors for DUT₁ from 4 GHz to 32 GHz. [58] (Copyright 2016, IEEE)

for the transmission coefficients over a frequency range from 10 GHz to 32 GHz. A limited bandwidth was used for determining the mean transmission values, as the transmission factors of less than -40 dB below 10 GHz do not lead to significant average values. It can be seen, that the average deviation over the investigated bandwidth is less than 0.30 dB for transmission and less than 0.81 dB for reflection coefficients. Major deviations from the reference occur for a DUT insertion loss of more than 40 dB. Below this value, the higher dynamic range of the commercial equipment has only minor impact on the measurement results.

The error vector is a measure, which is independent of the scattering parameters' magnitude. The complete error vector matrix EV is illustrated in two polar diagrams in Fig. 5.14 and proves an excellent agreement between the results obtained with the miniaturized VNA and the commercial VNA. The maximum error vector magnitudes are given by the matrix $|EV_{max}|$:

$$|EV_{max}| = \begin{pmatrix} 0.08 & 0.04 \\ 0.05 & 0.09 \end{pmatrix}$$

Both, the evaluation of the error vectors and the deviation show that both measured scattering matrices match very well. The commercial measurement setup leads to more accurate results at very low transmission values of less than -40 dB. However, in general an excellent agreement can be observed and also the DUT characterization based on typical properties leads to almost identical parameters.

	Corner Freq. (GHz)		Fwd. Passband IL (dB)			Rev. Passband IL (dB)		
	Fwd.	Rev.	Avg	Max	Min	Avg	Max	Min
min. VNA	15.37	15.36	3.08	3.88	2.68	3.09	3.78	2.70
Reference	15.38	15.37	3.00	3.76	2.63	3.00	3.77	2.63

Table 5.1: Characterization of DUT₁ with the miniaturized commercial VNA.

5.5.2 DUT 2: Fixed Attenuators

The previous measurements showed an excellent agreement with commercially available measurement equipment and proof the successful characterization of arbitrary microwave 2-port devices. However, this performance investigation is based on a limited range of transmission and reflection coefficients, which are provided by the DUT. In order to investigate the functionality of the miniaturized VNA over a wide range of transmission coefficients and the full bandwidth, further measurements were performed. For this purpose, the scattering parameters of a set of six fixed coaxial attenuators with values ranging from 3 dB to 40 dB were determined. In this way, the behavior of the miniaturized VNA regarding high transmission attenuation is investigated over a large bandwidth. The measured magnitudes of the transmission coefficients are illustrated in Fig. 5.15 (3 dB, 6 dB, 9 dB) and in Fig. 5.16 (20 dB, 30 dB, 40 dB). Reference measurements were performed using a commercial VNA. In addition, the magnitude of the deviation $|\Delta S|$ is shown. It should be noted that the investigated 30 dB and 40 dB attenuators are specified for frequencies up to 20 GHz (SMA) while the remaining attenuators are specified up to 40 GHz (K). In total, the measurement results match very well with the reference measurements obtained by commercial equipment. The deviation from the nominal attenuation value increases with frequency due to the losses of the attenuators themselves and series connected gender converters. The average deviation from the reference measurements is less than 0.18 dB and 3.73° for attenuation values up to 30 dB and is 0.52 dB and 3.23° for an attenuation of 40 dB. The maximum deviations were determined to be 0.78 dB and 8.64° up to 30 dB attenuation and 2.09 dB and 10.70° for an insertion loss of 40 dB. A detailed listing of the measured deviation is given in Table 5.2. Additionally, the error vector magnitude (EVM) was determined and is illustrated in Fig. 5.17. The measurements illustrated in this section prove that the miniaturized VNA shows a comparable performance for a transmission attenuation up to 30 dB to that of commercial laboratory equipment and, thus, is very well suited for numerous wideband measurement applications.

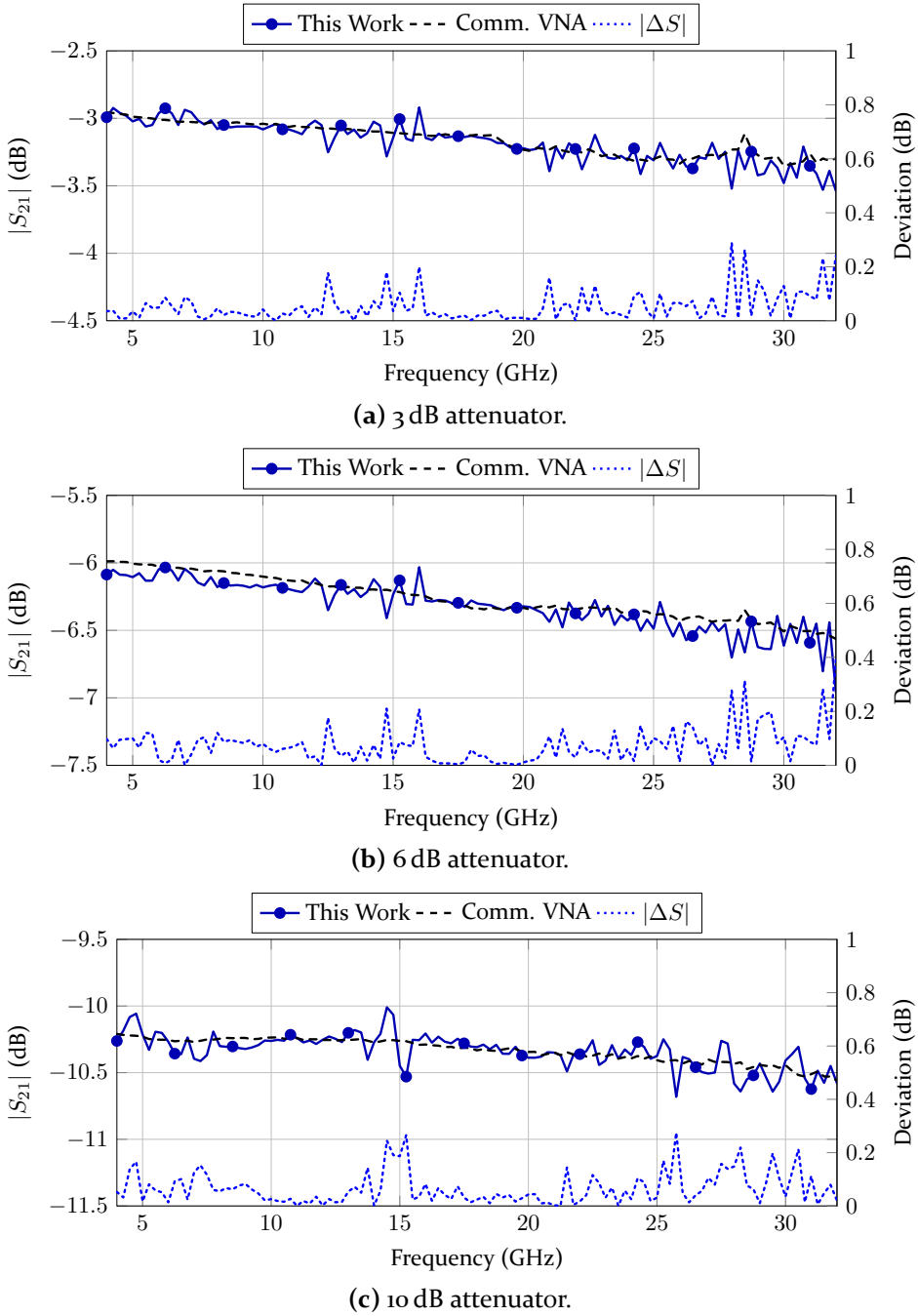
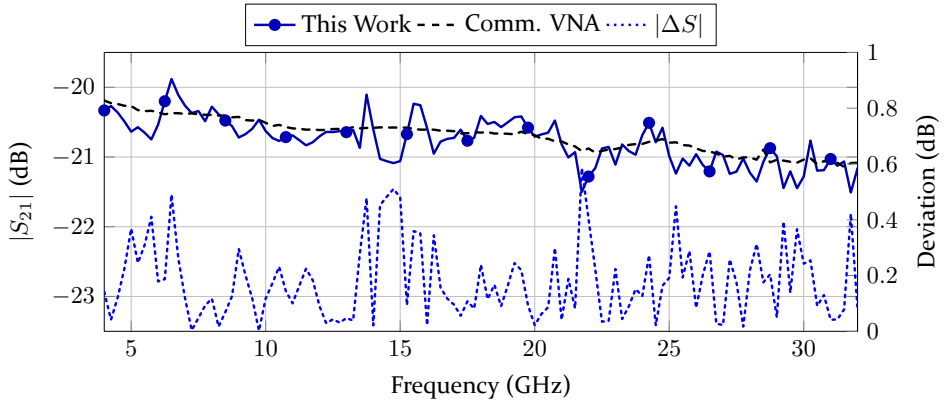
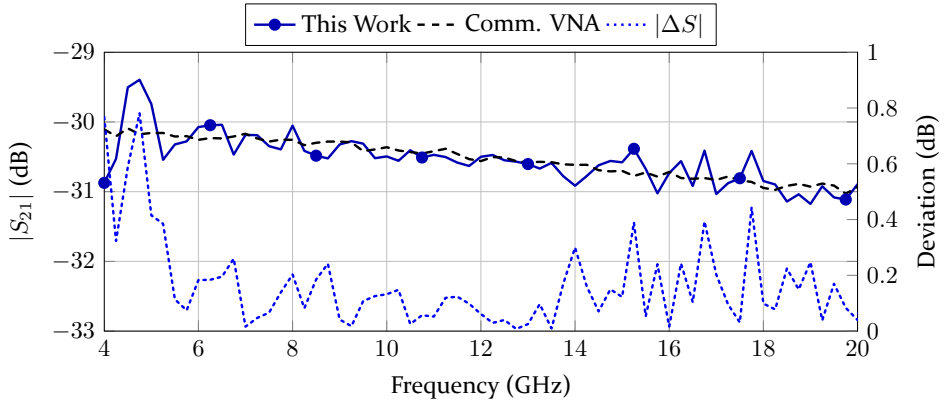


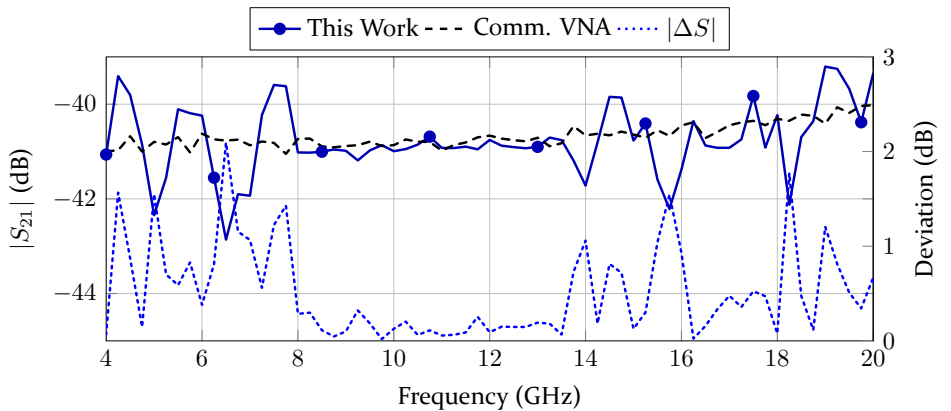
Figure 5.15: Measured transmission magnitude ($|S_{21}|$) of a 3 dB, 6 dB and 10 dB attenuator compared to the results obtained with a commercial VNA. [58] (Copyright 2016, IEEE)



(a) 20 dB attenuator.



(b) 30 dB attenuator.



(c) 40 dB attenuator.

Figure 5.16: Measured transmission magnitude ($|S_{21}|$) of a 20 dB, 30 dB and 40 dB attenuator compared to the results obtained with a commercial VNA. [58] (Copyright 2016, IEEE)

DUT Attenuation	$ \Delta S_{21} $		$ \angle \Delta S_{21} $	
	Mean	Max.	Mean	Max.
3 dB	0.05 dB	0.29 dB	1.83°	4.20°
6 dB	0.08 dB	0.39 dB	2.97°	7.44°
10 dB	0.07 dB	0.28 dB	3.73°	8.64°
20 dB	0.18 dB	0.59 dB	3.58°	9.50°
30 dB	0.17 dB	0.78 dB	1.29°	7.07°
40 dB	0.52 dB	2.09 dB	3.23°	10.70°

Table 5.2: Average and maximum deviation of the complex forward transmission coefficient (S_{21}) from the reference measurement.

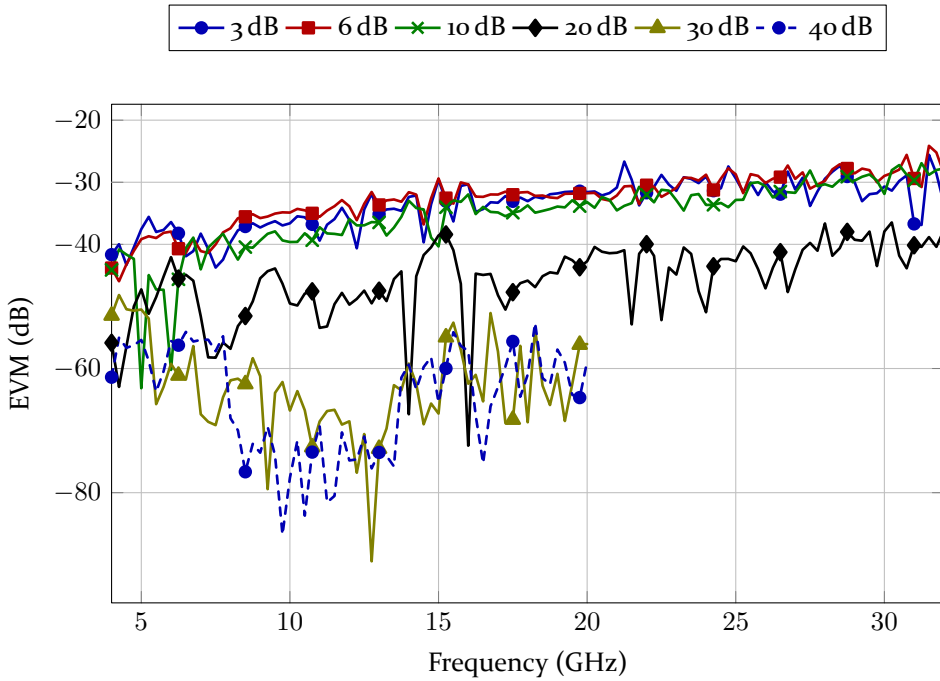


Figure 5.17: Error vector magnitude for the set of fixed coaxial attenuators. The 30 dB and 40 dB attenuators are specified up to 20 GHz only (SMA). [58] (Copyright 2016, IEEE)

5.6 System Performance

5.6.1 Linear Dynamic Range

The linear dynamic range specifies the range of the input power levels for which the VNA operates linearly. It is an important measure as S-parameter measurements require linear operation of the VNA. The lower end of the dynamic range corresponds to the system noise floor, which is the limiting factor for the detection of low power signals. The upper limits depend on the dynamic range definition. The receiver dynamic range is expressed as the difference between the 1 dB compression points of the receiver channels and the noise floor. It is the maximum achievable dynamic range of a VNA. The effective dynamic range can be lower if the stimulus output power is lower than the input referred compression points and can not drive the receiving circuits into compression. Thus, the system dynamic range is defined as the difference of the maximum available stimulus output power and the noise floor. This measure is mainly valid for the characterization of passive DUTs, while the maximum receiver dynamic range can be reached when investigating active DUTs, which add gain to the stimulus output power. The measured dynamic ranges and the essential parameters are illustrated in Fig. 5.18. The power levels of noise floor, the compression points and the output power were measured at the IF output of the miniaturized VNA at a coaxial test point with a spectrum analyzer and a resolution bandwidth (RBW)

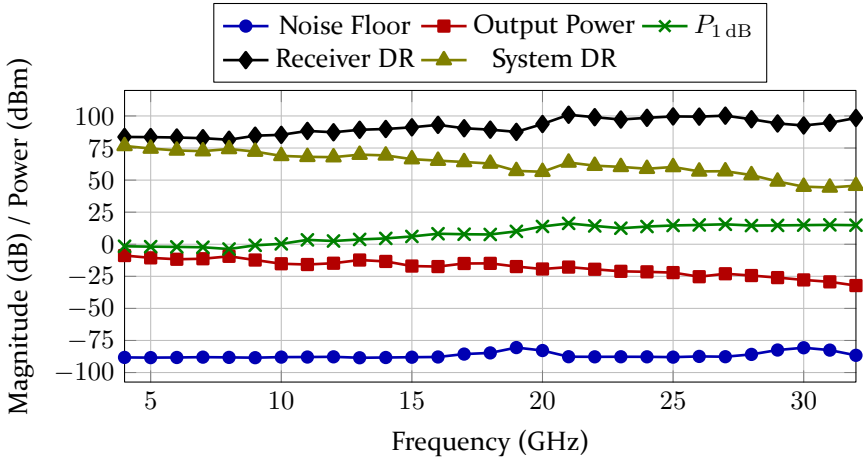


Figure 5.18: Dynamic range, noise floor, compression points and output power of the miniaturized VNA (RBW = 100 kHz). [58] (Copyright 2016, IEEE)

of 100 kHz. The input power sweeps were performed using an external signal generator representing the reflected power of an arbitrary DUT. The noise level was determined to be between -88.5 dBm and -80.5 dBm. The compression points were determined to be between -3.7 dBm and 16.3 dBm. The output power decreases over frequency from -8.8 dBm to -32.2 dBm. Whereas the compression points are fairly high compared to the pure chipset, the stimulus output power is low compared to the output power of 0.4 dBm at 32 GHz of the stand-alone frequency synthesizer. Both effects are caused by the insertion loss of the on-chip transmission lines, the cascaded directional couplers, the bond wires, the coplanar transmission lines on the RF PCB and the transition to the coaxial measurement port. The resulting dynamic ranges of the system and the receiver were determined to be from 44.3 dB to 76.6 dB and from 81.5 dB to 100.9 dB (100 kHz RBW), respectively.

5.6.2 Measurement Repeatability

In order to evaluate the measurement repeatability of the miniaturized VNA, a set of $N = 1000$ measurements per frequency point was performed. A coaxial waveguide served as DUT. The standard deviation σ was determined and used as an indication for the repeatability. The results are illustrated in Fig. 5.19 for the magnitude and the phase of reflection and transmission coefficients. The standard deviation for the reflection is lower than $9.2 \cdot 10^{-4}$ in magnitude and lower than 0.0027° in phase. For the transmission maximum values of $1.2 \cdot 10^{-3}$ and 0.012° were determined.

5.6.3 Drift over Time

In addition, the drift of the measurement results over time was determined. For this purpose, a set of five frequency sweeps with offsets of two hours was performed. As before, a coaxial waveguide was utilized as DUT and its reflection and transmission coefficient was determined. In the interest of better readability all measurements are normalized to the measurements obtained at the time t_0 . Consequently, the curves for the later measurements represent the deviation in decibels and degree, as illustrated in Fig. 5.20. The experiment was performed at room temperature. A maximum drift of 0.21 dB and 0.037° regarding reflection and 0.14 dB and 0.041° regarding transmission was observed over a duration of eight hours. Thus, stable measurements over long time periods are guaranteed.

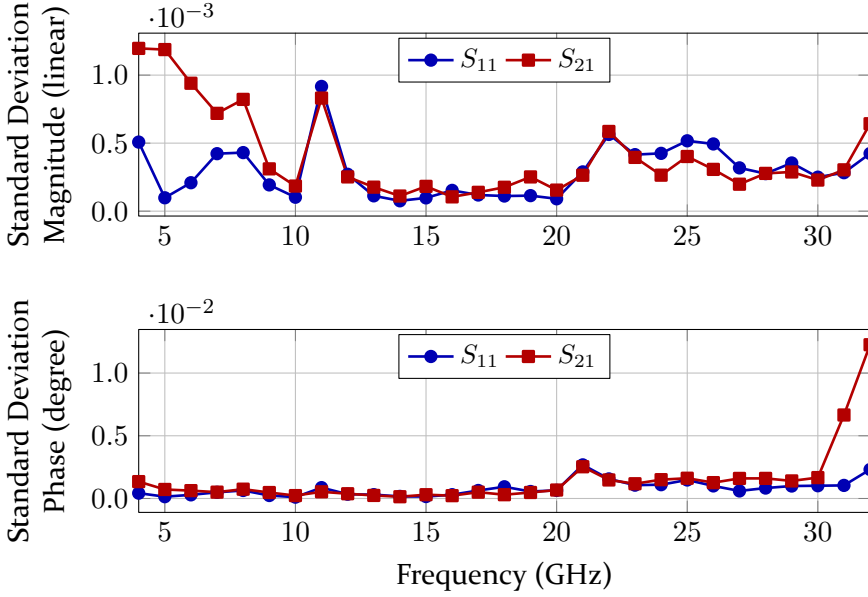


Figure 5.19: Measured standard deviation of uncalibrated reflection and transmission coefficient. [58] (Copyright 2016, IEEE)

5.7 Conclusion

In this chapter, the operation of a miniaturized 2-port VNA with a bandwidth from 4 GHz to 32 GHz was successfully demonstrated. The analyzer is based on the chipset described in chapter 4 and was supplemented with an integrated wideband testset. To the best of the author's knowledge the utilized directive elements are the first fully integrated couplers incorporating such high bandwidth. The fully integrated VNA RF frontend is part of a self-sustained hardware demonstrator consisting of four vertically stacked PCBs. A 16-term 2-port calibration procedure was implemented and applied to the analyzer so that the full S-matrix of several 2-port DUTs could be determined successfully. The measured scattering parameters are compared to the results obtained with a commercial VNA and show excellent agreement.

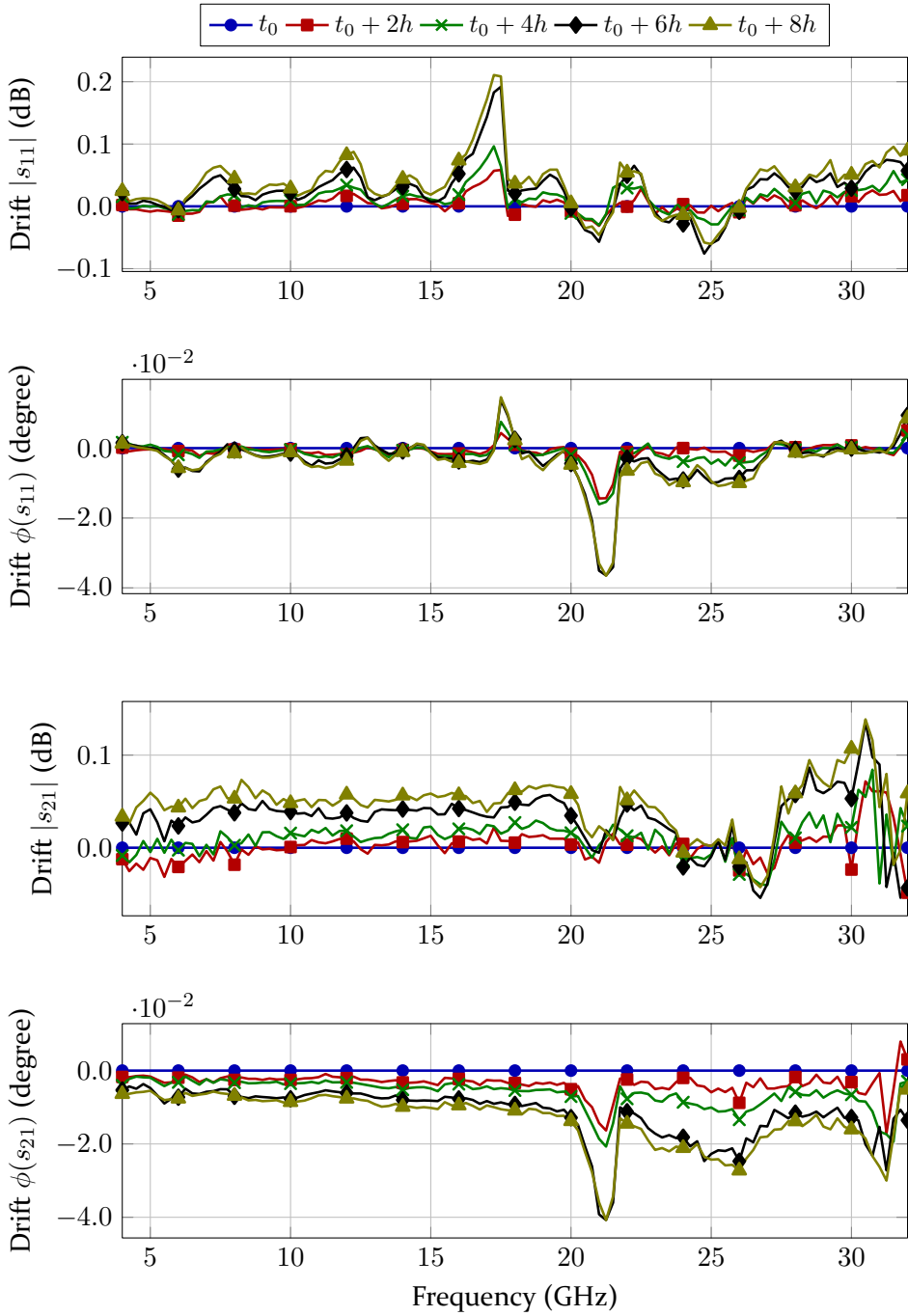


Figure 5.20: Measured drift of a reflection and transmission measurement in magnitude and phase over a time of 8 hours. [58] (Copyright 2016, IEEE)

Chapter 6

Biomedical Microwave Dielectric Sensing Systems

6.1 Introduction

For the most part, VNAs are used in the disciplines of instrumentation and measurement in order to characterize microwave components. The application of miniaturized analyzers for such tasks was shown in the chapters 3 and 5 of this thesis. However, being a universal measurement device, network analyzers are increasingly utilized for microwave material characterization in biomedical scenarios. Numerous promising approaches such as broadband dielectric spectroscopy based on microfluidic channels were published recently. These attempts address not only the determination of the relative permittivity of biological tissues [60] and liquids [41] but also the distinction between living and dead chinese hamster ovary (CHO) cells [61] and the detection of human cancer cells [39]. These recent advances prove a huge potential of microwave technology in biological sensing. However, most of the microwave sensors require a commercial VNA for the scattering parameter measurement.

This chapter shows, how miniaturized measurement equipment can be used for the characterization of biologically relevant liquids instead of costly and bulky instrumentation devices. For this purpose, three different microwave permittivity sensitive systems are presented. The systems combine sensors with read-out circuits, which extract the information about the material under test (MUT). Thus, all of them are self-sustained and do not need additional laboratory equipment. This results in low financial efforts and small

form factors, which are essential parameters for commercialization of microwave electronics. The first system is comprised of the highly integrated VNA described in chapter 5 and a coaxial probe which interacts with the MUT. The second system is an integrated VNA combined with on-chip sensors for transmission based material characterization. The third system is an alternative approach and is significantly different from the VNA based approaches. Instead, it is a permittivity sensitive PLL integrating the sensing device into the stimulating circuit.

The three systems are utilized in a biomedical scenario for the characterization of binary solutions consisting of ethanol and methanol. Despite having similar physical properties regarding color and smell, the two liquids have different effects on the human body. From a biomedical point of view it is safe to consume low amounts of ethanol whereas methanol is highly poisonous and should be avoided. In the following, pure ethanol and pure methanol will be denoted as E100:Mo and E0:M100, respectively. The mixture ratios of both liquids are marked accordingly.

6.2 Reflection based Analyzer with Off-Chip Probe

The material characterization with the aid of open ended coaxial probes is a well known technique and is widely spread due to the availability of commercial products [62,63]. The coaxial probe interacts with the MUT and provides a material dependent reflection coefficient to a read-out circuit. In general, a commercial VNA is used to extract the information about the MUT based on the measured reflection coefficient. In combination with additional measurements of known materials the complex relative permittivity can be determined. The probes have to be machined at a high precision and require appropriate sealing in order to guarantee repeatable measurement circumstances [64]. Nevertheless, the required commercial VNA is the dominating factor in terms of cost, complexity and required space of such setups. Thus, a sensing system based on the previously described miniaturized VNA will be demonstrated in the following.

6.2.1 Measurement Setup

The used measurement setup is illustrated in Fig. 6.1. Whereas one port of the miniaturized VNA is left open, the other port is connected to the coaxial probe via a standard coaxial cable. The probe itself is immersed into a test tube containing the MUT. In order to minimize parasitic effects of the setup, the probe was mounted in a stiff fixture and was not moved during

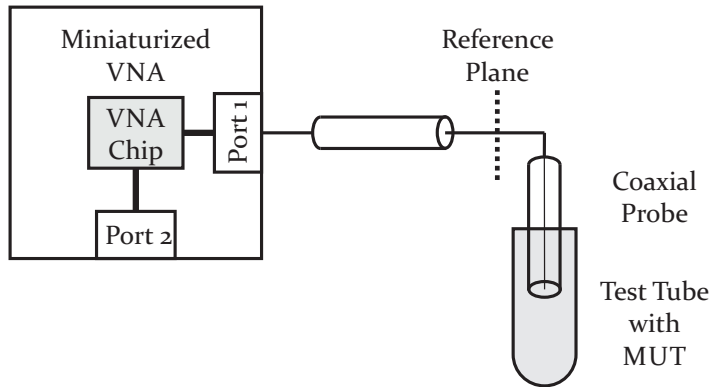


Figure 6.1: Measurement setup utilizing the miniaturized VNA as a reflectometer and a coaxial probe as a permittivity sensor. [58] (Copyright 2016, IEEE)

the experiment. In this way, bending of the cables and concussions of the setup could be avoided. Instead, the test tube was positioned on a height-adjustable stage so that the probe could be immersed appropriately into the investigated test tube.

6.2.2 Coaxial Probe

The used probe is part of the Agilent 85070E Dielectric Probe Kit. Its cross section is illustrated in Fig. 6.2. In principle, it is an open-ended coaxial waveguide with an impedance of $50\ \Omega$. A sealing prevents the characterized MUT from being soaked into the inner of the probe. The sensing capability is given by the interaction of the electromagnetic field with the material under investigation at the probe's open end.

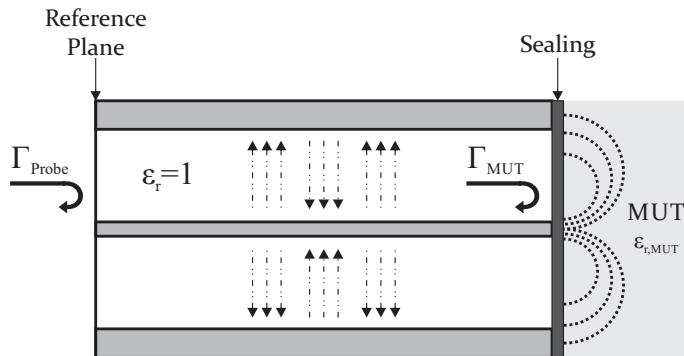


Figure 6.2: Cross sectional view of the used coaxial probe. [58] (Copyright 2016, IEEE)

6.2.3 Characterization of Ethanol-Methanol Solutions

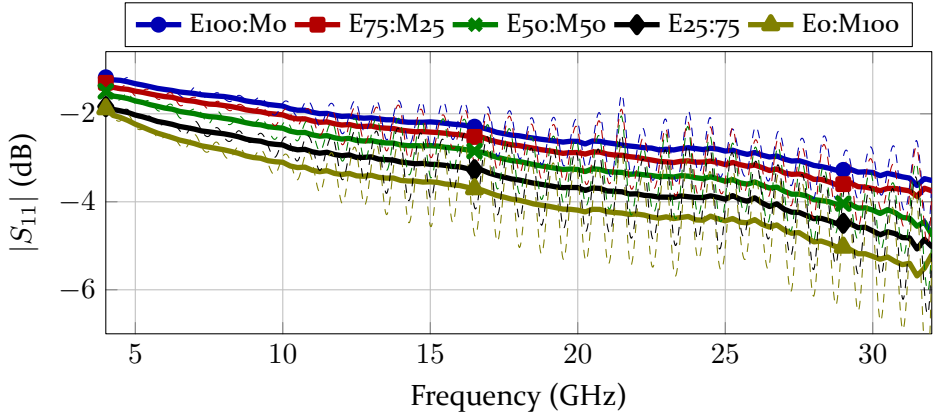
The described sensing system was used in order to characterize solutions containing different amounts of ethanol and methanol. Five test tubes with different mixture ratios with concentration steps of 25 % ranging from pure ethanol to pure methanol were prepared. The 1-port analyzer was calibrated by means of an OSM calibration procedure so that the reference plane is located between the coaxial cable and the probe as shown in Fig. 6.1. Consequently, the probe's reflection coefficient S_{11} is evaluated in order to retrieve information about the investigated material. The measurements were performed over the full bandwidth from 4 GHz to 32 GHz with a frequency resolution of 250 MHz. A spline interpolation was applied to the measured curves in order to represent ripples of the reflected signals more clearly, which are caused by multiple reflections. The measured reflection coefficients with the probe immersed into the different liquids at a constant ambient temperature of 18 °C are illustrated in Fig. 6.3, whereas the measured and interpolated data is indicated in dashed lines. In the interest of better readability, the moving average over $N = 10$ frequency points is plotted in solid lines, additionally. The magnitude of all measured reflection coefficients decreases with frequency due to the increasing losses in both the probe itself and in the investigated solution. One can see from the results that an increasing amount of methanol leads to an increment of loss as well. The five solutions show significant deviations from each other in both magnitude and phase. Thus, the mixture ratios can be distinguished from each other clearly with the proposed sensing system. The reflection coefficients of the probe loaded with pure ethanol and pure methanol deviate by 2.0 dB and 8.6° at the center frequency of 18 GHz. Despite the fact that the investigated concentrations are equally spaced in steps of 25 %, the resulting curves show unequal spacing. This effect is due to the nonlinearly correlation of the permittivity of binary solutions [65]. Consequently, the sensitivity, which is

$f = 18 \text{ GHz}$	Magnitude Sensitivity (dB)	Phase Sensitivity (degree)
$S_{11,E100:M0}/S_{11,E75:M25}$	0.32	1.93
$S_{11,E75:M25}/S_{11,E50:M50}$	0.47	1.86
$S_{11,E50:M50}/S_{11,E25:M75}$	0.57	2.22
$S_{11,E25:M75}/S_{11,E0:M100}$	0.66	2.58

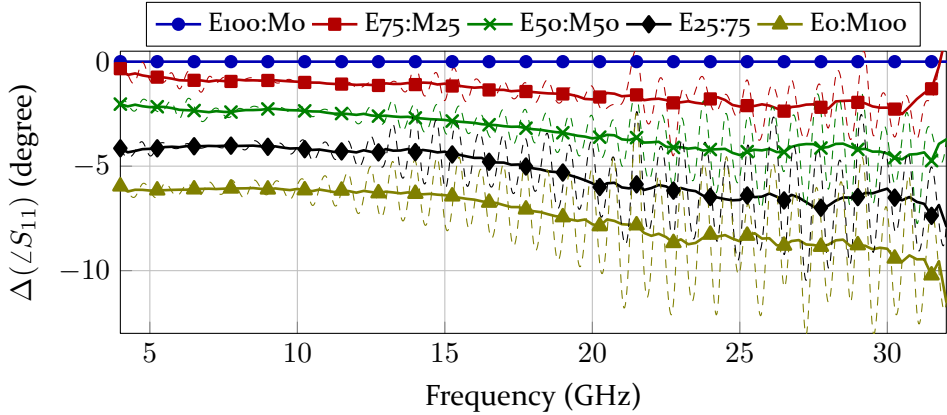
Table 6.1: Sensitivity of the reflection based sensing system for different concentration steps at the center frequency of 18 GHz.

defined as the change of the reflection coefficient caused by 25% concentration deviation, behaves nonlinear as well. The resulting sensitivities are summarized in Table 6.1 for the center frequency of 18 GHz.

The described setup can be applied to determine the relative dielectric constant ϵ_r of materials. For this purpose, an appropriate calibration algorithm (e.g. as in [66]) needs to be applied. However, this topic is outside the scope of this thesis and was investigated intensively in [67].



(a) Magnitude of the probe's measured reflection coefficient.



(b) Phase of the probe's measured reflection coefficient referenced to pure ethanol (E100-Mo). [58] (Copyright 2016, IEEE)

Figure 6.3: Measured reflection coefficient of the coaxial probe immersed into different mixtures of ethanol (E) and methanol (M). The solid lines represent the moving average ($N = 10$, $\Delta f = 250$ MHz) over the measured values (dashed lines).

6.3 Transmission based Analyzer with On-Chip Sensor

In contrast to the previously described permittivity sensing system, which is based on an external probe, the following section is dedicated to a highly integrated permittivity sensitive analyzer. The system is based on a reduced VNA architecture combined with an on-chip sensing element.

6.3.1 Measurement Setup

The system architecture of the proposed analyzer is illustrated in Fig. 6.4. As with the previously described systems, the analyzer incorporates a heterodyne receiver with two integrated wideband signal sources. These are embedded in a dual PLL architecture with a single reference oscillator. It also contains two IF output ports for the measurement signals. Thus, the chip is compatible to the hardware demonstrator presented in section 5.3. In contrast to the previously described systems, it relies on the measurement of the transmission coefficient of an on-chip sensor, which is affected by an external MUT. The sensor is implemented as a differential transmission line on the surface layer of the chip. This layer is well suited for sensor structures as it is not covered by any passivation, which leads to a maximum interaction with the MUT. The line geometry was optimized for maximum sensitivity at a differential line impedance of $100\ \Omega$ by electromagnetic field

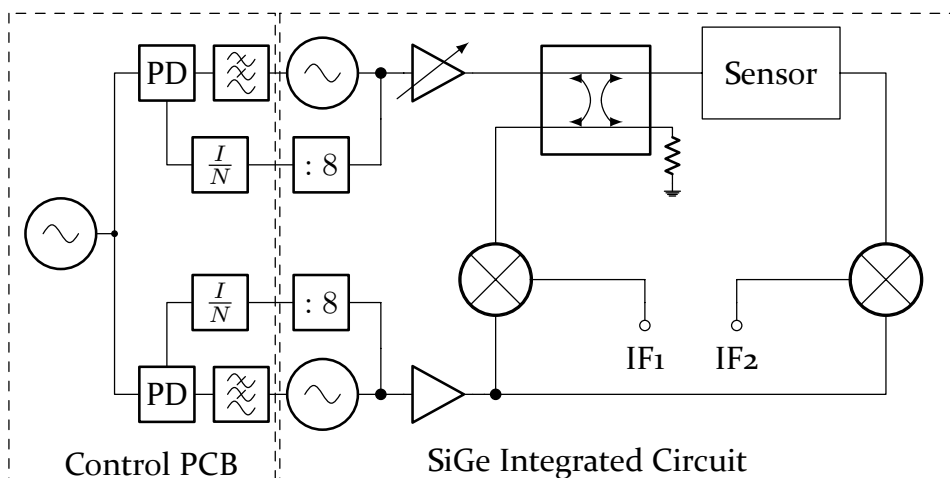


Figure 6.4: Architecture of the fully integrated permittivity sensitive system based on a reduced network analyzer topology. [58] (Copyright 2016, IEEE)

simulations. As the sensitivity of transmission line sensors correlates with the line length (low losses assumed), the sensor was designed with a length of $2400\text{ }\mu\text{m}$. The RF stimulus signal is fed to the sensing element via a directive element. This coupler is used to derive a reference channel which is proportional to the stimulus in both magnitude and phase and additionally isolates the reference signal from wave portions reflected by the sensor and other system components. Both, the reference signal and the transmitted portion of the stimulating wave are converted down to an IF by two Gilbert cells. The uncalibrated transmission coefficient can be determined by building the ratio between the measurement and the reference signal in the following signal processing stages. A micrograph of the fabricated circuit is shown in Fig. 6.5. The chip was mounted on a carrier PCB, which provides the required peripheral interfaces. The MUT was filled into an appropriate container so that the sensing chip could be immersed into the liquid. The penetration depth was monitored by a microscope so that the on-chip sensors are covered equally during the single measurements. All test objects were prepared with identical volumes so that heating due to the power dissipation of the analyzer chip affects all measurements in the same way. The experiment was performed at a constant ambient temperature of $20\text{ }^{\circ}\text{C}$. The measurement setup is depicted in Fig. 6.6.

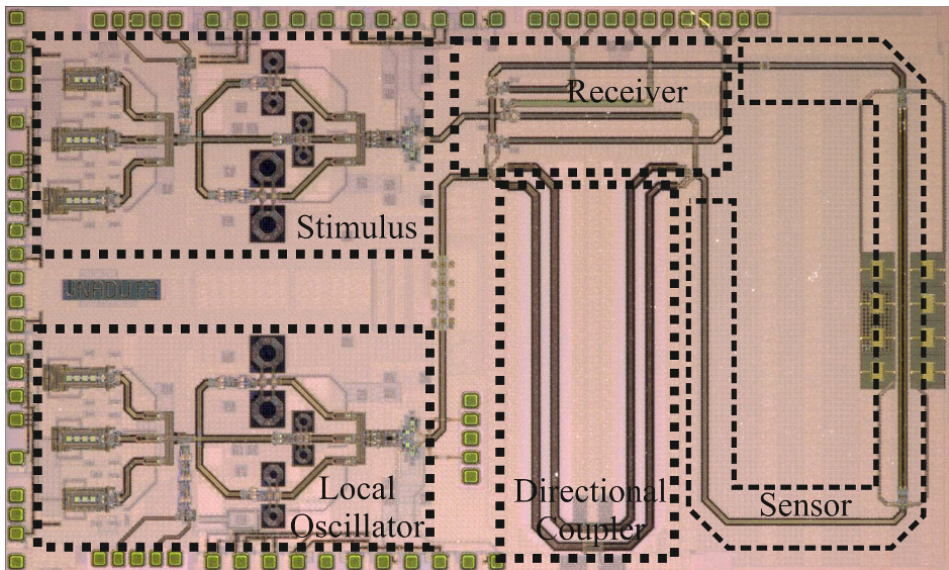


Figure 6.5: Photograph of the fabricated transmission based network analyzer with on-chip sensors. [58] (Copyright 2016, IEEE)

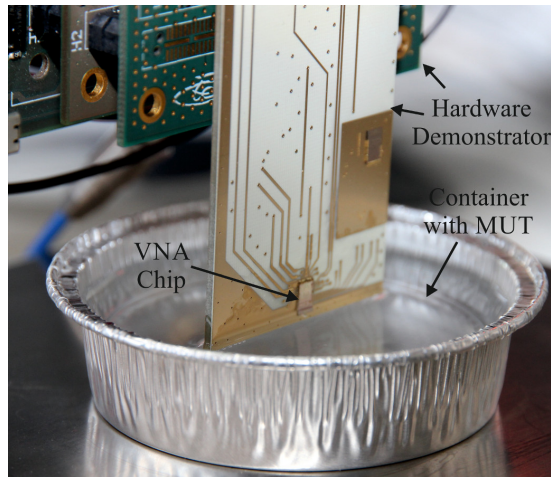
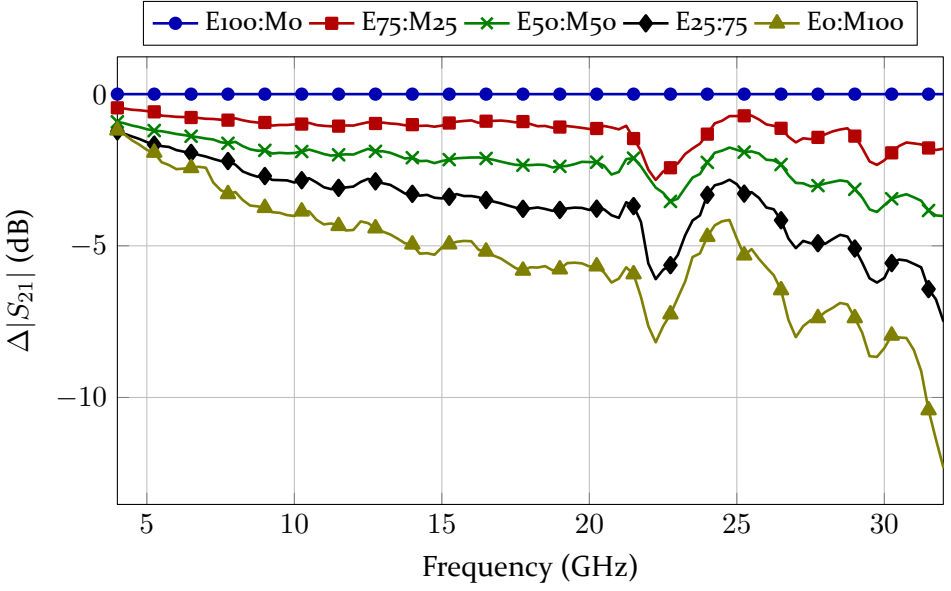


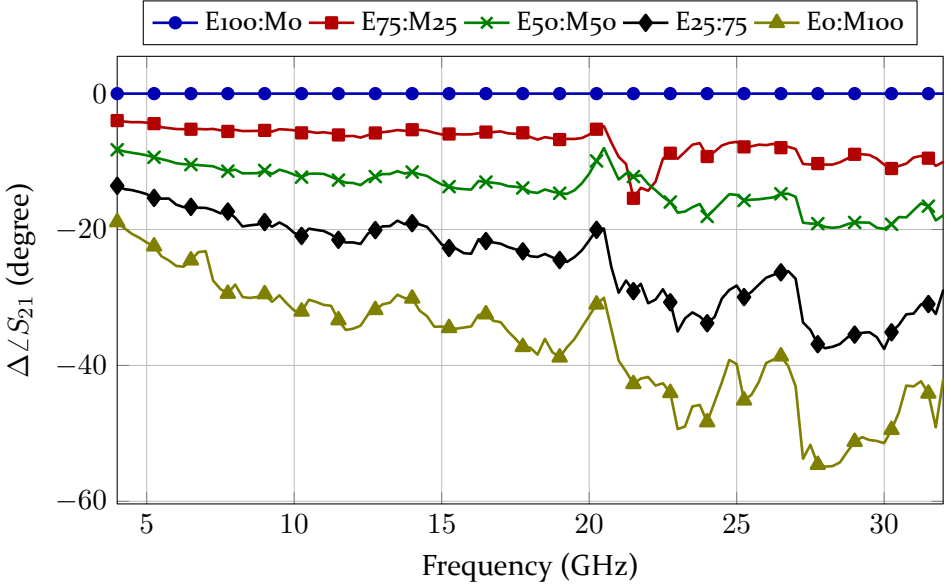
Figure 6.6: Measurement setup: The on-chip sensors are immersed into the MUT. [58] (Copyright 2016, IEEE)

6.3.2 Characterization of Ethanol-Methanol Solutions

The described network analyzer with integrated sensing elements was used for the characterization of binary solutions consisting of ethanol and methanol. Like in the previous experiment, five solutions with concentrations varying in steps of 25 % were prepared and the sensor was immersed into the solution. A calibration procedure was omitted, as it is not required for the determination of relative transmission coefficient changes caused by different MUT permittivities. The measurements were performed over the bandwidth from 4 GHz to 32 GHz in steps of 250 MHz. The measured forward transmission coefficient S_{21} for the five investigated solutions is illustrated in Fig. 6.7. In the interest of better readability, all results were referenced to pure ethanol (E100:Mo). A significant deviation between the measurements of the investigated solutions can be observed in both magnitude and phase. In accordance with the reflection based analyzer, the transmission coefficient magnitude decreases with increasing methanol proportions, reflected by higher dielectric losses in methanol compared to ethanol. The magnitude of the forward transmission coefficient of pure ethanol and pure methanol deviate by 5.7 dB and 37.4° at the center frequency of 18 GHz. Again, the sensitivity depends on the mixture ratio and increases with the amount of methanol. An overview over the sensitivity at the center frequency of 18 GHz is given in Table 6.2. In general, the transmission based approach shows a significantly higher sensitivity than the coaxial probe based setup.



(a) Relative forward transmission magnitude $\Delta|S_{21}|$.



(b) Relative forward transmission phase $\Delta\angle S_{21}$.

Figure 6.7: Measured forward transmission ΔS_{21} of the integrated sensor referenced to pure ethanol (E100:Mo). [58] (Copyright 2016, IEEE)

$f = 18 \text{ GHz}$	Magnitude Sensitivity (dB)	Phase Sensitivity (deg)
$S_{21,E100:M0}/S_{21,E75:M25}$	0.99	6.39
$S_{21,E75:M25}/S_{21,E50:M50}$	1.35	8.07
$S_{21,E50:M50}/S_{21,E25:M75}$	1.43	9.46
$S_{21,E25:M75}/S_{21,E0:M100}$	1.94	13.46

Table 6.2: Sensitivity of the transmission based sensing system for different concentration steps at the center frequency of 18 GHz.

6.4 Permittivity Sensitive PLL

The previously described sensing systems rely on a dedicated signal source, which stimulates a sensor. By evaluating the sensor's response by means of appropriate receiver circuits, one can extract information about the MUT. This leads to a comparably high complexity, a large number of components and consequently to a high power consumption. For this reason, the following subsection is dedicated to a simpler approach, which combines the RF stimulus and the sensing element in a single component. The concept relies on PLLs based on permittivity sensitive oscillators and was initially introduced for chemical sensing in [68, 69]. The following section describes a stand-alone microwave sensor for medical and biological MUTs utilizing an interdigital capacitor in the uppermost gold layer of a silicon-germanium process [70]. A standard microcontroller is used to configure the sensor as well as to read the sensor information.

6.4.1 Sensor Concept

The sensor concept is a stand-alone microwave permittivity sensing system with a low-frequency sensor-readout. A block diagram is shown in Fig. 6.8. The material under test (MUT) is sensed at microwave frequencies (19.2 GHz to 20.8 GHz) and the sensor response can be determined at DC with a simple voltage measurement. The actual sensing interface is an interdigital capacitor on a chip whose capacitance is influenced by the material's permittivity. As this interdigital sensor is embedded into the LC tank of a voltage controlled oscillator, the oscillation frequency depends on both the control voltage and the material, which surrounds the sensor. A change in the MUT permittivity leads to a frequency shift. Embedding the VCO into a PLL gives the opportunity to characterize the MUT at a defined frequency, so that the material properties are mapped to the resulting tuning voltage. Thus, an increase in permittivity leads to an increased sensor capacitance resulting in

lower oscillation frequencies of the VCO. Assuming tuning characteristics with positive slopes, higher tuning voltages are required to set the oscillator to the desired frequency. In this way, the high frequency material properties are directly converted down to DC so that no costly microwave measurement equipment containing frequency converters is required. Furthermore, the read-out is directly integrated into the sensor stimulus reducing the total number of components and the complexity of the system.

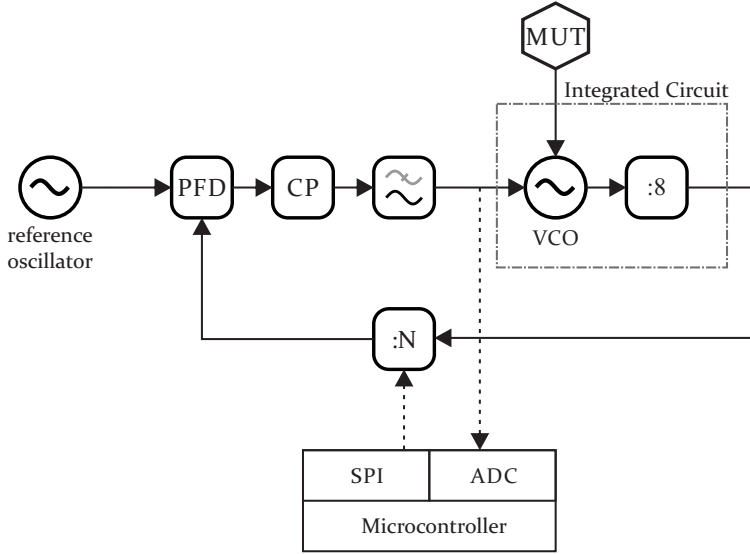


Figure 6.8: Block diagram of the permittivity sensitive phase-locked loop. [70] (Copyright 2015, IEEE)

6.4.2 Integrated Circuit Design

Voltage Controlled Oscillator

The voltage controlled oscillator is based on the circuits described in chapter 4. The simplified schematic is shown in Fig. 6.9. The active negative resistance device was realized by two transistors, which are cross-coupled by a feedback capacitor. The inductive part of the resonator is composed of transmission lines, which are used as interconnects between the single components and as tank line inductances. The capacitance consists of two different elements. Voltage controlled varactors are used to provide a tunability of the device, which is required for embedding the oscillator into a

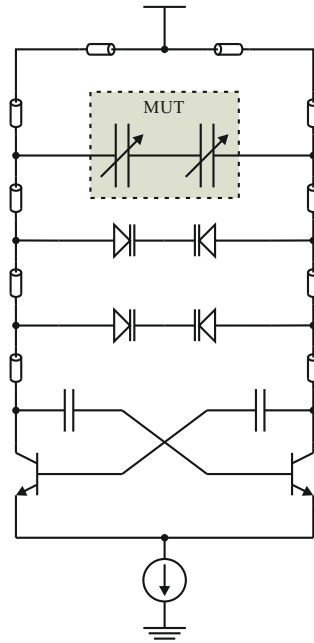


Figure 6.9: Simplified schematic of the cross-coupled VCO containing the material sensitive capacitors. [70] (Copyright 2015, IEEE)

PLL. The sensing capability of the VCO is given by two interdigital capacitors with a capacitance, which is material dependent. In this way, the VCO can be tuned by both a control voltage and the material under test, which surrounds the integrated circuit.

Interdigital Sensor

An interdigital capacitor acts as a sensing element. Its capacitance is determined by a constant part and a material dependent part. The constant part is given by the permittivity of the chip's silicon-dioxide layers. Additionally, the capacitance is influenced by the material, which is located above the integrated circuit. The interdigital geometry was optimized for a high sensitivity. Therefore, the sensor was fabricated in the upper-most available conducting layer, which is usually reserved for pads. Thus, there is no passivation layer above the sensor, which increases the sensitivity regarding the MUT. Furthermore, an optimum trade-off regarding the top layer design rules, sensitivity, as well as the required overall capacitance had to be found. The resulting geometry is shown as a micrograph of the fabricated integrated circuit in Fig. 6.10.

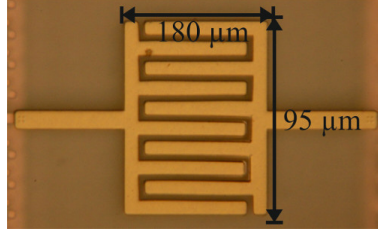
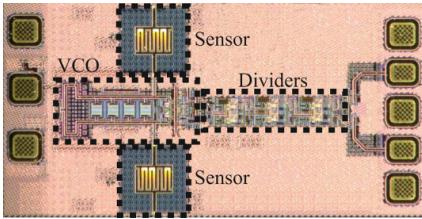


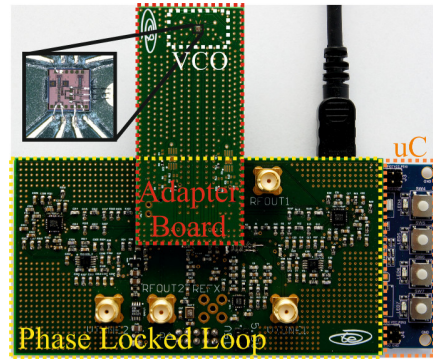
Figure 6.10: Micrograph of the interdigital sensor capacitor fabricated in the up-permost pad layer of the SiGe process. [70] (Copyright 2015, IEEE)

6.4.3 System Implementation

A micrograph of the fabricated voltage controlled oscillator containing the material sensitive interdigital capacitor is shown in Fig. 6.11a. The integrated circuit consumes a chip area of $1128 \mu\text{m} \times 554 \mu\text{m}$ and was bonded on an adapter PCB, which connects it to the main board containing the phase-locked loop and its components. The system frequency as well as all other loop configurations can be applied by a microcontroller via SPI. Furthermore, the tuning voltage can be measured using the microcontroller's analog-to-digital converters or standard voltmeters. Due to the on-chip frequency prescalers, the output frequency is as low as 2.5 GHz so that standard commercially available components can be used for all configuration and measurement tasks. A photograph of the PCB stack is shown in Fig. 6.11b. The measurement data can be obtained from the microcontroller via USB on a serial interface.



(a) Integrated circuit.



(b) System overview.

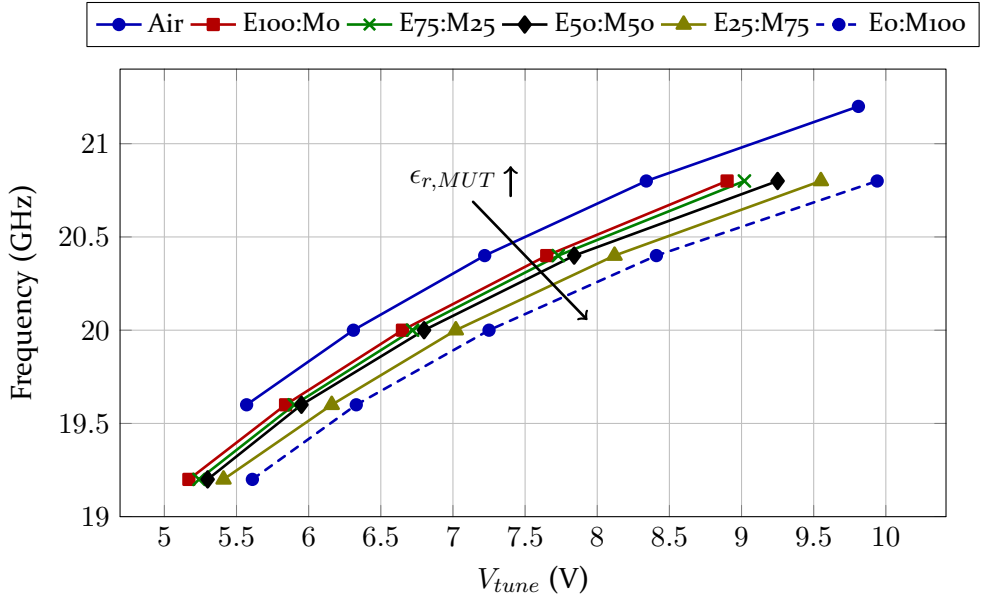
Figure 6.11: Detailed view of the fabricated SiGe integrated circuit (a) and overview over the modular stand-alone sensing system (b). [70] (Copyright 2015, IEEE)

6.4.4 Characterization of Ethanol-Methanol Solutions

As a proof of concept, the system was used in a biomedical application. As both liquids are not conductive at low frequencies, the sensor IC could be immersed into the liquids so that the interdigital sensing element is covered by the MUT completely. Five different mixture ratios of ethanol and methanol were produced in steps of 25 %, ranging from pure ethanol to pure methanol. The sensor was contacted with the MUTs and the tuning voltages (V_{tune}) over the VCO frequencies from 19.2 GHz to 20.8 GHz were measured. The resulting tuning characteristics are illustrated in Fig. 6.12a. In addition, the results for the sensor loaded with air ($\epsilon_r = 1$) are shown. It can be observed that the oscillation frequencies are lowered with increasing relative dielectric constants of the MUTs. For the sake of better readability, the measured tuning voltages were referenced to the results obtained with pure ethanol (E100-M0) and are shown in Fig. 6.12b. The results show that there is a difference in tuning voltage of 1.04 V at 20.8 GHz between pure ethanol and pure methanol. Thus, concentration differences in steps of 25 % can be distinguished clearly by a simple DC voltage measurement or low-cost digital analog converters (DACs) of a microcontroller. In general, it can be observed, that an increase of permittivity (e.g. higher methanol part) leads to an increased tuning voltage at a given oscillation frequency. This is due to the fact, that a higher permittivity of the MUT increases the interdigital sensor's capacitance and thus lowers the oscillation frequency of the VCO at a fixed frequency. Assuming positive slope varactors, the tuning voltage must be increased in order to set the VCO to a desired frequency. The system sensitivity depends on the mixture ratio and is summarized in Table 6.3.

$f = 20.8 \text{ GHz}$	Sensitivity (V)
$V_{tune,E100:M0} - V_{tune,E75:M25}$	0.12
$V_{tune,E75:M25} - V_{tune,E50:M50}$	0.23
$V_{tune,E50:M50} - V_{tune,E25:M75}$	0.30
$V_{tune,E25:M75} - V_{tune,E0:M100}$	0.39

Table 6.3: Sensitivity of the permittivity sensing PLL versus concentration variation ($f = 20.8 \text{ GHz}$).



(a) Tuning characteristic.

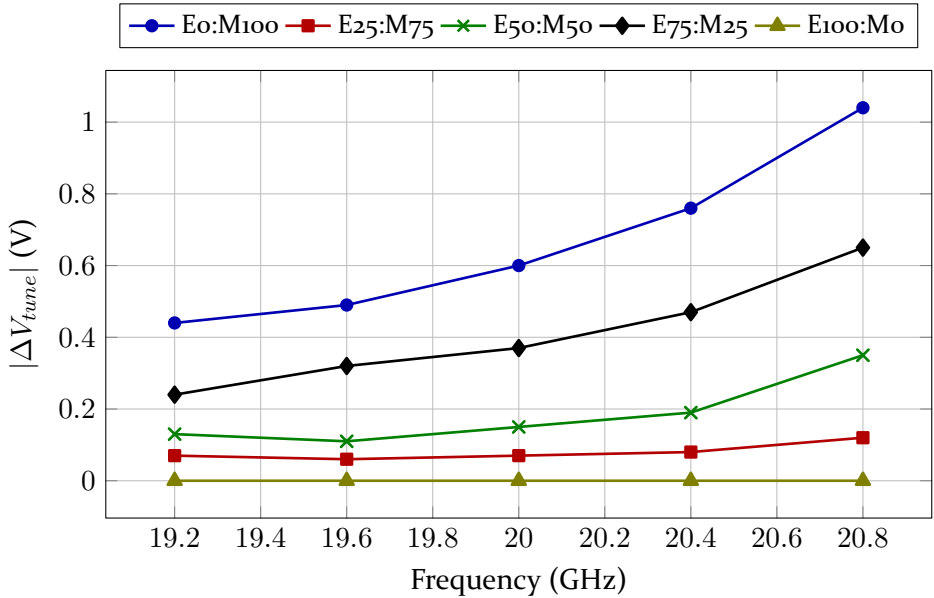
(b) Variation of tuning voltage (V_{tune}).

Figure 6.12: Measured tuning characteristics (a) for the VCO loaded with air and binary solutions of ethanol and methanol and the resulting tuning voltage variation (b) referenced to pure ethanol (E100-Mo).

6.5 Conclusion

In this section three permittivity sensing systems were described. Two of them rely on a heterodyne VNA architecture with a reflection based external coaxial probe and an on-chip transmission based sensor, respectively. Furthermore, an alternative approach was realized in terms of a permittivity sensing PLL. The main advantage of the analyzer based systems is the higher amount of information, which is retrieved about the MUT. While the absolute relative permittivity mainly affects the phase of the measured scattering parameters, the material loss is represented by the magnitude. The transmission based approach showed superior sensitivity regarding binary solutions of ethanol and methanol. One upside of this technique is the scalability of the sensitivity with the sensor's length. On the contrary, the geometry of a coaxial sensor can be modified only to a certain extent under the premise of a $50\ \Omega$ environment. The permittivity sensing PLL shows a high sensitivity as well. However, one cannot gain information about the complex dielectric constant as the system's output parameter is a scalar DC voltage. On the other hand, the complexity is low compared to the analyzers, which makes this sensor a good choice for low-cost applications (e.g. concentration monitoring). Due to the high degree of integration the systems relying on the on-chip transmission line and the interdigital capacitor are advantageous over the coaxial probe based system. The analyzer based sensing systems can retrieve a higher amount of information about the investigated materials due to their multi-octave bandwidth. Concluding, it can be stated that the transmission based highly integrated sensing system is very well suited for precise measurements whereas the permittivity sensitive PLL is the preferred choice for low-cost scenarios. The comparison of the described sensing system is summarized in Table 6.4.

	Reflection Based Analyzer	Transmission Based Analyzer	Permittivity Sensing PLL
Sensitivity	o	+	-
Complexity	o	o	+
Size	-	+	+
Bandwidth	+	+	o

Table 6.4: Comparison of the described microwave dielectric sensing systems (+: plus, o: neutral, -: minus).

Chapter 7

Conclusion and Outlook

This thesis is dedicated to the investigation, realization and application of highly integrated VNAs covering a multi-octave microwave bandwidth from 4 GHz to 32 GHz. All fabricated circuits and systems were laid out for heterodyne receiver topologies, which are advantageous with regard to both feasibility and measurement accuracy. Initial practical implementations of narrow-band vector reflectometers with different degrees of integration served as experimental studies investigating the realization and performance of such systems. The concept of the VNA-on-chip was proven to be valid by a very good agreement of the gained measurements with the results obtained by commercial measurement equipment.

Based on these incipient successful results, a higher degree of integration, more bandwidth and a higher number of measurement ports was targeted. Therefore, a chipset for a 4-32 GHz 2-port VNA was developed in SiGe technology. The chipset is comprised of a wideband signal source featuring an integrated switch for 2-port stimulation and the corresponding heterodyne receiver. Both chips are based on an octave bandwidth VCO array followed by a bandwidth extension stage utilizing frequency dividers. The receiver contains four identical channels so that a complete error vector correction can be applied to the analyzer. Both chips feature single-ended RF interfaces, which facilitate the interconnect of most DUTs and calibration standards. The chipset's functionality was successfully demonstrated by a hybrid assembly containing broadband directional couplers on a PCB.

A self-sustained 2-port VNA was realized based on a fully integrated RF frontend, which is comprised of the above mentioned chipset and an on-chip wideband testset. The latter consists of a set of four three-section directional couplers covering the required bandwidth of 4 GHz to 32 GHz. The frontend

is embedded into a modular hardware demonstrator providing the required peripheral functions such as phase lock, signal conversion, processing and general control, as well as power supply. A 16-term calibration model was applied and facilitates the measurement of 2×2 S-matrices of arbitrary test objects. A very good agreement with the results obtained by a commercial VNA was observed over the full bandwidth. The measurements were proven to be highly repeatable and the drift over time is negligible.

Due to the reduced size, miniaturized VNAs are ideally suited for sensing applications. Therefore, highly integrated analyzers were utilized for the characterization of binary solutions of ethanol and methanol with a reflection based external probe and a transmission based on-chip sensor. It was proven that solutions with concentration differences of 25 % could be distinguished from each other clearly. In addition, an alternative approach based on a permittivity sensitive PLL was demonstrated.

In conclusion, it can be stated that the concept of the VNA-on-chip was successfully introduced. The results pave the way for future activities in the area of integrated network analyzers. In this context, calibration procedures for setting the reference plane on the chip represent beneficial extensions and open the door towards enhancements of built-in test and characterization. Moreover, a further increase of bandwidth can be achieved by harmonic mixing and an inherently wideband, albeit lossy, resistive directional element. Besides the latter enhancements, a huge potential of the integrated VNA arises in the combination of on-chip sensing devices in biomedical scenarios. Future research activities should address the symbiosis of the herein presented highly integrated VNAs with the recent advances of miniaturized biomedical sensors.

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Workshop:

- J. Nehring, M. Hofmann, G. Fischer, D. Kissinger, and R. Weigel "Design und Synthese breitbandiger Richtkoppler für die vektorielle Netzwerkanalyse," in *RADCOM*, Hamburg, Germany, May 2013.

Awards:

- J. Nehring, M. Bartels, R. Weigel, and D. Kissinger, *First Place Student Paper Competition, MTT-S Biomed. Wireless Technol. Netw. Sens. Syst.*, San Diego (CA), USA, Jan. 2015.
- J. Nehring, S. Lindner, M. Hofmann, and G. Vinci, *First Prize European Microwave Week Student Challenge*, Amsterdam, The Netherlands, Oct. 2012.

This thesis addresses the topic of highly integrated vector network analysis. Due to its significant contribution to technological advances in the past and its presently high potential regarding instrumentation and biomedical sensing applications, the technique of network analysis plays a key role in microwave and wireless engineering. However, the application of network analyzers is frequently limited due to the size, weight and cost of the instruments. For this reason, the potential and feasibility of the VNA-on-chip is evaluated. Therefore, theoretical and practical studies were performed by means of narrow-band microwave reflectometers with different degrees of integration. After a successful proof of concept, a chipset dedicated to a highly integrated 4-32 GHz 2-port vector network analyzer is developed in silicon germanium (SiGe) technology. The chipset can be utilized for the implementation of a heterodyne network analyzer and features a stimulus signal generator with integrated switch and a four-channel receiver chip. In a further step, a fully integrated RF frontend is demonstrated, including onchip directive elements. In combination with a modular hardware demonstrator, the assembly serves as a miniaturized 2-port vector network analyzer from 4 GHz to 32 GHz. The measured 2×2 S-matrices show an excellent agreement with the results obtained by commercial measurement equipment. The measurements are highly repeatable and the drift over time is negligible. Furthermore, highly integrated network analyzers were applied to biomedical scenarios. For this purpose, a reflection based analyzer with external probe as well as a transmission based one with an on-chip sensor are presented and compared to an alternative sensing system.

