Helsinki University of Technology Radio Laboratory Publications
Teknillisen korkeakoulun Radiolaboratorion julkaisuja
Espoo, April 2005 REPORT S 269

RECEIVER FRONT-END CIRCUITS AND COMPONENTS FOR MILLIMETRE AND SUBMILLIMETRE WAVELENGTHS

Ville S. Möttönen

Dissertation for the degree of Doctor of Science in Technology to be presented with due permission for public examination and debate in Auditorium S4 at Helsinki University of Technology (Espoo, Finland) on the 6th of May at 12 o'clock noon.

Helsinki University of Technology
Department of Electrical and Communications Engineering
Radio Laboratory

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Sähkö- ja tietoliikennetekniikan osasto
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ISBN 951-22-7615-1   (print)
ISBN 951-22-7616-X   (PDF)
ISSN 1456-3835

Otamedia Oy
Espoo 2005

Preface

This thesis work was carried out at Radio Laboratory of Helsinki University of Technology during 1997–2004.

I am deeply grateful to my supervisor Professor Antti Räisänen for the valuable guidance and support during my research work. He provided me the opportunity and facilities to carry out this work. The years have been to me interesting and challenging familiarisation to radio engineering.

Special, warm thanks go to Docent Arto Lehto and Dr. Petri Piironen for instructing my work during these years.

I thank the official reviewers, Dr. Jeffrey Hesler and Dr. Alain Maestrini, for their valuable suggestions and comments. Great compliments also belong to Docent Arto Lehto and Dr. Juha Mallat for proofreading of the dissertation.

I wish to thank all the present and former persons I have worked with in Radio Laboratory. Especially, I want to thank my former workmate Dr. Jussi Säily for cheerful moments like the Green Card test and the chaps in our workshop for any machining needed for the work.

My heartfelt thanks go to all my close-ones and friends for memorable moments spent together. Without my family, my sister Milja, my mother Marja-Liisa, and my father Lauri, I would not be this far. All the friends in the Wonkamiehet group are acknowledged for refreshing events and moments.

Finally and especially, I want to thank my wife Satu for her patience and support during all these years. During writing this thesis, me and my wife got our first child, daughter Venla. It has been memorable time in every way. Surely, future will be too.

This work received financial support from Academy of Finland, European Space Agency (ESA), National Technology Agency (TEKES), Helsinki University of Technology, Graduate School of Electronics, Telecommunications, and Automation (GETA), Jenny and Antti Wihuri Foundation, Emil Aaltonen Foundation, Foundation of Technology, and Foundation of the Finnish Society of Electronics Engineers.

Espoo, March 23, 2005

Ville Möttönen

Abstract

This dissertation focuses on the development of millimetre- and submillimetre-wave receiver front-end circuits and components. Seven scientific articles, written by the author, present this development work. A short introduction to the technology related to the designs of the thesis precedes the articles. The articles comprise several novel structures and techniques intended to further improve the performance of receivers or to provide new ways for receiver circuit implementation, summarised as follows.

1) Novel rectangular waveguide-to-CPW waveguide transition using a probe structure. The measured insertion and return loss of an X-band (8.2−12.4 GHz) back-to-back structure are less than 0.5 dB and more than 17 dB, respectively, over the entire frequency band (fractional bandwidth of > 40 %). The transition is used in a submm-wave mixer.

2) Novel rectangular waveguide-to-CPW transition using a fin-line taper. The measured insertion and return loss of an X-band (8.2−12.4 GHz) back-to-back structure are less than 0.4 dB and more than 16 dB, respectively, over the entire frequency band.

3) Novel tunable waveguide backshort based on a fixed waveguide short and movable dielectric slab. The measured return loss for a W-band backshort is less than 0.21 dB (VSWR > 82) over the entire frequency band of 75−110 GHz.

4) New coaxial bias T. The insertion loss is less than 0.5 dB at 3−16 GHz (fractional bandwidth of 137 %) and 0.1 dB at 5.2−14.1 GHz. In the latter range, the return loss is more than 30 dB. The RF isolation is greater than 30 dB at 1−17 GHz.

5) First millimetre-wave subharmonic waveguide mixer using European quasi-vertical Schottky diodes. The mixer utilises a single diode chip with quartz filters in a four-tuner waveguide housing. A single-sideband noise temperature of 3500 K and conversion loss of 9.2 dB (antenna loss included) have been measured at 215 GHz with an LO power of 3.5 mW.

6) Balanced-type fifth-harmonic submillimetre-wave mixer. It uses two planar Schottky diodes, quartz filters, and a tuner-less in-line waveguide housing with an integrated diagonal horn antenna and new LO transition structure. The designed RF range is 500–700 GHz enabling the use of an LO source at 100−140 GHz. A conversion loss of about 27 dB has been measured at 650 GHz with an LO power of 10 mW. The mixer has been in use in phase locking of a submm-wave signal source.

7) Characterisation procedure of planar Schottky diodes with extensive dc, capacitance, and wide-band (up to 220 GHz) *S*-parameter measurements and parameter extraction. Parameters of a simple diode equivalent circuit and results of extensive measurements are available for designers and diode manufacturers for further use.

Keywords: backshort, bias T, rectangular waveguide-to-coplanar waveguide transition, fifth-harmonic mixer, millimetre-wave, planar Schottky diode, subharmonic mixer, submillimetre-wave, waveguide, wide-band.

List of Publications

This thesis is based on the work presented in the following publications:

[P1] V. S. Möttönen and A. V. Räisänen, "Novel wideband coplanar waveguide-to-rectangular waveguide transition," *IEEE Transactions on Microwave Theory and Techniques*, vol. 52, no. 8, pp. 1836−1842, Aug. 2004.

[P2] V. S. Möttönen, "Wideband coplanar waveguide-to-rectangular waveguide transition using fin-line taper," *IEEE Microwave and Wireless Components Letters*, vol. 15, no. 2, pp. 119−121, Feb. 2005.

[P3] V. S. Möttönen, P. Piironen, and A. V. Räisänen, "Novel tunable waveguide backshort for millimeter and submillimeter wavelengths," *IEEE Microwave and Wireless Components Letters*, vol. 11, no. 9, pp. 370–372, Sep. 2001.

[P4] V. S. Möttönen, P. Piironen, and A. V. Räisänen, "Low-loss wideband microwave coaxial bias T," *Microwave and Optical Technology Letters*, vol. 29, no. 4, pp. 236–238, May 2001.

[P5] V. S. Möttönen, P. Piironen, J. Zhang, A. V. Räisänen, C.-I. Lin, A. Simon, and H. L. Hartnagel, "Subharmonic waveguide mixer at 215 GHz utilizing quasivertical Schottky diodes," *Microwave and Optical Technology Letters*, vol. 27, no. 2, pp. 93–97, Oct. 2000.

[P6] V. S. Möttönen and A. V. Räisänen, "General purpose fifth-harmonic waveguide mixer for 500−700 GHz," in *Proc. 34th European Microwave Conference*, Amsterdam, The Netherlands, 12−14 October, 2004, pp. 1145−1147.

[P7] V. S. Möttönen, J. Mallat, and A. V. Räisänen, "Characterisation of European millimetre-wave planar diodes," in *Proc. 34th European Microwave Conference*, Amsterdam, The Netherlands, 12−14 October, 2004, pp. 921−924.

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Contribution of Author

Publication [P1]	The author's own work; Prof. Antti Räisänen supervised the work.
Publication [P2]	This is entirely the author's own work.
Publication [P3]	The author's own work; Prof. Antti Räisänen and Dr. Petri Piironen supervised the work.
Publication [P4]	The author's own work; Prof. Antti Räisänen and Dr. Petri Piironen supervised the work.
Publication [P5]	The author designed, constructed, and measured the mixer and prepared the publication. Dr. Jian Zhang modelled the diode chip. Prof. Antti Räisänen and Dr. Petri Piironen supervised the work. Dr. Chih-I Lin, Dr. Ansgar Simon, and Prof. Hans Hartnagel from TUD were responsible for diode fabrication.
Publication [P6]	The author's own work; Prof. Antti Räisänen supervised the work.
Publication [P7]	The author's work except that the on-wafer measurements were carried out at VTT (Technical Research Centre of Finland) Information Technology, MilliLab. Prof. Antti Räisänen and Dr. Juha Mallat supervised the work.

Contents

Abbreviations

ADS	Advanced Design System
AM	amplitude modulation
APD	anti-parallel pair of Schottky diodes
BWO	backward wave oscillator
CPS	coplanar strips
CPW	coplanar waveguide
CSL	coupled slotline
dc	direct current
D-plane	diagonal plane
DSB	double-sideband
E-plane	plane of the electric field
ESA	European Space Agency
ESTEC	European Space Research and Technology Centre
F-band	waveguide frequency band from 90 GHz to 140 GHz
FEM	finite-element method
GaAs	gallium arsenide
G-band	waveguide frequency band from 140 GHz to 220 GHz
GETA	Graduate School of Electronics, Telecommunications, and Automation
HFSS	High Frequency Structure Simulator
H-plane	plane of the magnetic field
IF	intermediate frequency
InP	indium phosphide
I-V	current-voltage
Ka-band	waveguide frequency band of 26.5–40 GHz
KASIMIR	Key Advanced Structure Investigations for Mm- and Sub-mm-wave Integrated Receivers
Ku-band	waveguide frequency band of 12.4–18 GHz
LNA	low-noise amplifier
LO	local oscillator
MDS	Microwave Design System
MilliLab	Millimetre Wave Laboratory of Finland – MilliLab
MMIC	monolithic microwave integrated circuit
PHEMT	pseudomorphic high electron mobility transistor
PLL	phase-locked loop
QUID	quartz-substrate upside-down integrated device
RF	radio frequency
SiO_2	silicon dioxide
SMARAD	Smart and Novel Radios Research Unit
S-parameter	scattering parameter
SSB	single-sideband
TE_{10}	transverse electric field mode
TE_{30}	transverse electric field mode
Tekes	National Technology Agency
TKK	Helsinki University of Technology
TRL	thru-reflect-line
TUD	Technical University of Darmstadt
UG-387/U-M	waveguide flange standard
V-band	waveguide frequency band of 50–75 GHz
VNA	vector network analyser

VSWR	voltage standing wave ratio
VTT	Technical Research Centre of Finland
W-band	waveguide frequency band of 75−110 GHz
WR-5	rectangular waveguide size for a frequency band of 140−220 GHz
WR-8	rectangular waveguide size for a frequency band of 90−140 GHz
WR-10	rectangular waveguide size for a frequency band of 75−110 GHz
X-band	waveguide frequency band of 8.2−12.4 GHz
YIG	yttrium iron garnet

Symbols

α	angle of the radial stub
β_1	phase constant in the waveguide
β_2	phase constant in the waveguide section containing a dielectric slab
Γ	reflection coefficient
$\Delta\phi$	shift in the phase of the reflection coefficient
ΔV	change in the junction voltage per decade of the current
ε_r	relative permittivity, dielectric constant
η	ideality factor
λ	wavelength at the centre frequency in the partly substrate-filled waveguide; wavelength in the CPW
λ_g	wavelength in the waveguide; wavelength in the CPW
σ	conductivity
ϕ_{bi}	built-in potential
a	width of the waveguide
a'	width of the narrowed waveguide
b	height of the waveguide
b_1	separation between the filter element and waveguide wall
b_2	separation between the filter element and waveguide wall
C_{ac}	parasitic parallel capacitance
C_{fr}	fringing field parasitic capacitance
C_{j0}	zero-bias junction capacitance
C_p	parasitic parallel capacitance
$C_{p'}$	parasitic parallel capacitance
$C(V_j)$	non-linear junction capacitance
C_1	capacitance of the dc block capacitor
C_2	capacitance of the RF block capacitor
d	CPW ground-to-ground spacing
$\boldsymbol{E}$	electric field
f_c	centre frequency
g	length of the soldering gap
h	height of the substrate
I_0	saturation current
I_1	diode current
I_2	diode current
k	Boltzmann's constant
l	length of the dielectric slab inside the waveguide
l_{rs}	length of the radial stub
l_1	distance from the centre of the transition probe to the ground plane of the CPW
l_2	distance from the centre of the transition probe to the end of the dielectric substrate

l_3 distance from the transition probe to the waveguide wall
L conversion loss
L_f anode finger inductance
L_p parasitic inductance
L_{DSB} double-sideband conversion loss
n number of harmonics
n index number, $n = 1, 2, 3 \ldots$
P_{LO} local oscillator power
q magnitude of electron charge
R_s series resistance
s distance from the centre of the transition probe to the waveguide narrowing
S_{11} scattering parameter
S_{21} scattering parameter
t thickness of metallization; height of the guide channel
$\tan\delta$ loss tangent
T absolute temperature
T_{DSB} double-sideband noise temperature
V_{br} breakdown voltage
V_j junction voltage
V_1 voltage applied to the diode
V_2 voltage applied to the diode
w width of the centre conductor of the CPW; width of the guide channel
w_1 width of the CPW with finite ground planes
Z_{stub} characteristic impedance of the high-impedance stub
Z_0 characteristic impedance of the CPW
Z_{01} characteristic impedance of the transmission line
Z_{02} characteristic impedance of the transmission line
Z_{03} characteristic impedance of the transmission line

For clarity, also abbreviations and symbols used in the dissertation publications are listed above.

1 Introduction

Due to increasing number of millimetre-wave and submillimetre-wave applications and due to increasing commercial interest, receiver front-end techniques are being developed extensively. This concerns both the development of semiconductor technology and circuits based on semiconductor components.

Commonly at millimetre and submillimetre wavelengths (10–0.1 mm, 30–3000 GHz), the radio frequency (RF) signal, received through an antenna, is directly mixed down to the intermediate frequency (IF) with a mixer pumped with a phase-locked local oscillator (LO) source. Figure 1.1 shows the block diagram of a typical receiver front-end. The mixer most often operates as a fundamental or subharmonic mixer, e.g., [1]–[16], although in some applications, harmonic mixers are preferred. Solid-state LO sources required for the mixing process are already available for frequencies up to 1.9 THz [17]–[22]. Furthermore, low-noise amplifiers (LNA) have already been successfully implemented at frequencies up to 215 GHz, e.g., [23]–[29]. Thus, at millimetre wavelengths, amplifiers have become more common as the first stage in the front-end, contrary to Figure 1.1. This thesis work deals with a receiver front-end based on a Schottky-diode waveguide mixer. Some results of the work can be applied well to, e.g., amplifiers and multipliers.

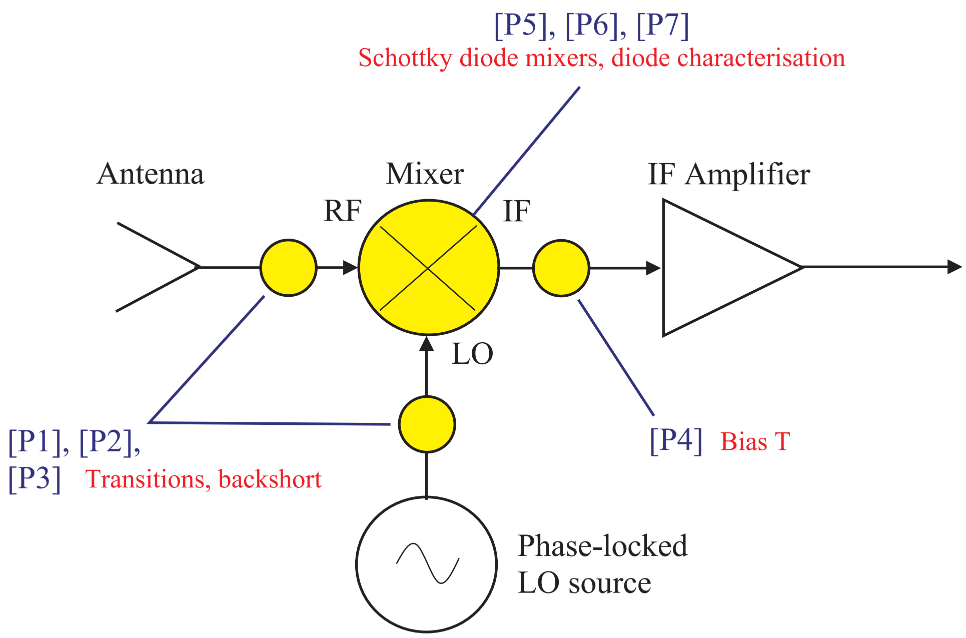

Figure 1.1. *Block diagram of a typical millimetre- or submillimetre-wave receiver front-end showing (marked yellow) also the spots for components or circuits (transitions, backshort, mixers, bias T, diode characterisation), which have been developed in this thesis work, with references to publications [P1]–[P7].*

1.1 Millimetre- and Submillimetre-wave Receiver Front-end Circuits and Components Developed

This dissertation focuses on the development of millimetre-wave and submillimetre-wave receiver front-end circuits and components. Seven publications, written by the author, present this development work. A short introduction to the technology related to the designs of the thesis precedes these publications. The publications comprise several novel structures and techniques intended to further improve the performance of receivers or to provide new ways for receiver circuit implementation.

Figure 1.1 also shows the spots, which this thesis work focuses on. These are referred with the reference numbers of original publications. Furthermore, Figure 1.2, showing a tunable millimetre-wave waveguide mixer, illustrates more clearly the development targets of the work. Although the separate publication works are referred in context of a mixer, a part of them, [P1]–[P3] and [P7], are applicable to other devices as well, e.g., amplifiers, oscillators, and multipliers. Thus, the thesis work has influence in various parts of the receiver front-end. The thesis consists of designs related to waveguide transitions in [P1] and [P2], a waveguide backshort in [P3], a bias T in [P4], Schottky-diode mixers in [P5] and [P6], and planar diode characterisation in [P7].

The dissertation is divided into two parts: a summary and scientific articles. The first part within Chapters 2–6 gives an introductory overview to the technology related to the designs of the thesis. Chapter 7 summarises the work carried out in publications [P1]–[P7]. Final conclusions on the thesis work are drawn in Chapter 8. The second part comprises the seven publications mentioned above. *This division is congruent with the instructions of Dissertation Committee of Helsinki University of Technology.* Thus, the summary contains an explanation of the research subject, the most important results achieved by the candidate, and a list of the articles.

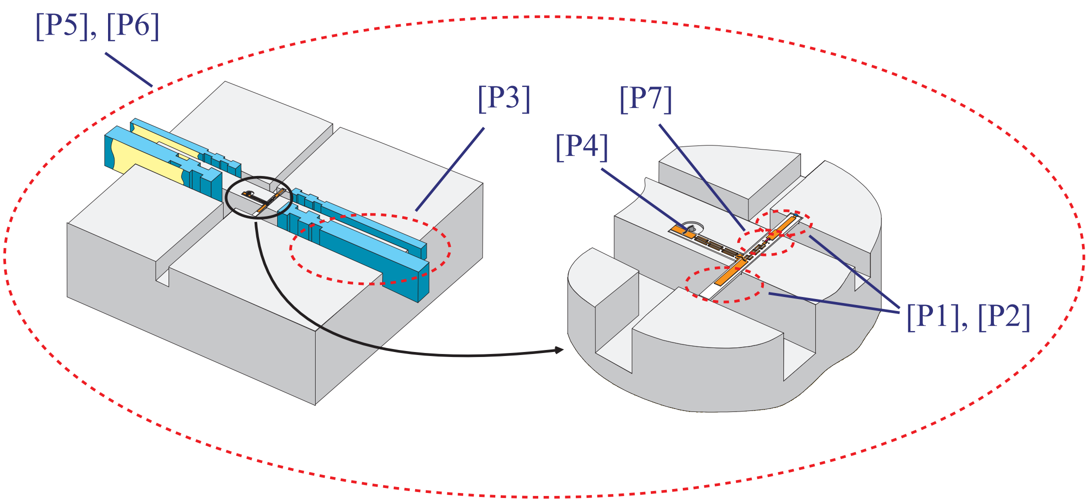

Figure 1.2. *Tunable millimetre-wave waveguide mixer [P5] illustrates the spots, which have been developed in the thesis, with the original publication references [P1]–[P7]. Since this is only an illustration, all the references are not applicable to the mixer shown.*

1.2 New Scientific Results

This thesis has produced the following new scientific results:

1) new coplanar waveguide (CPW)-to-rectangular waveguide transition using a probe structure;
2) new CPW-to-rectangular waveguide transition using a fin-line taper;
3) new tunable waveguide backshort;
4) new coaxial bias T;
5) first millimetre-wave subharmonic waveguide mixer using quasi-vertical Schottky diodes;
6) balanced-type fifth-harmonic submillimetre-wave mixer using two planar single Schottky diodes and a new LO transition structure;
7) characterisation of European planar Schottky diodes with dc, capacitance, and wide-band *S*-parameter measurements.

This thesis work required the invention of new structures, drawing of designs, machining of some structures, extensive simulations, construction of devices, extensive measurements, and preparation of publications.

2 Rectangular Waveguide-to-Planar Transmission Line Transitions

Several different transition structures exist for coupling electromagnetic waves from a rectangular waveguide to a planar circuit. Planar circuits can utilise microstrip lines, striplines, fin-lines, or coplanar waveguides (CPW). Typically, microstrip line and stripline-to-waveguide transitions are realized by bringing a probe, formed by the end of the transmission line, through an aperture in the broad wall of the waveguide as shown in Figure 2.1 (a) [30]–[35]. This enables low-loss operation over the entire waveguide frequency band. However, the width of the circuit cannot be made large without special measures. The probe has typically a rectangular or radial shape. The probe substrate can be placed in the longitudinal direction of the waveguide or perpendicular to that. Transitions in the end of the waveguide use aperture coupling, microstrip tapering, a quasi-Yagi "antenna", a ridged waveguide, or an antipodal fin-line [36]–[44]. Figure 2.1 (b), (c), and (d) show the transitions based on the aperture coupling, quasi-Yagi "antenna", and antipodal fin-line, respectively.

Waveguide-to-CPW transitions often use similar techniques (Figure 2.2), probes through the waveguide broad wall [29], [45]–[47], coupled patches [48], and ridged waveguides [49]–[52].

The transition in Figure 2.2 (c) [49] uses a ridge, which converts smoothly to the centre conductor of a CPW. The measurements of a back-to-back X-band structure (8.2–12.4 GHz) demonstrate an insertion loss of less than 1 dB. The return loss is more than 18 dB at 7.5–11.5 GHz. For a Ka-band (26.5–40 GHz) back-to-back structure the insertion and return loss are less than 1.5 dB and more than 14 dB, respectively. Figure 2.2 (d) shows a different way to use the ridge [50]. In this transition, the end of a cosine-tapered ridge is connected to the end of the centre conductor of a

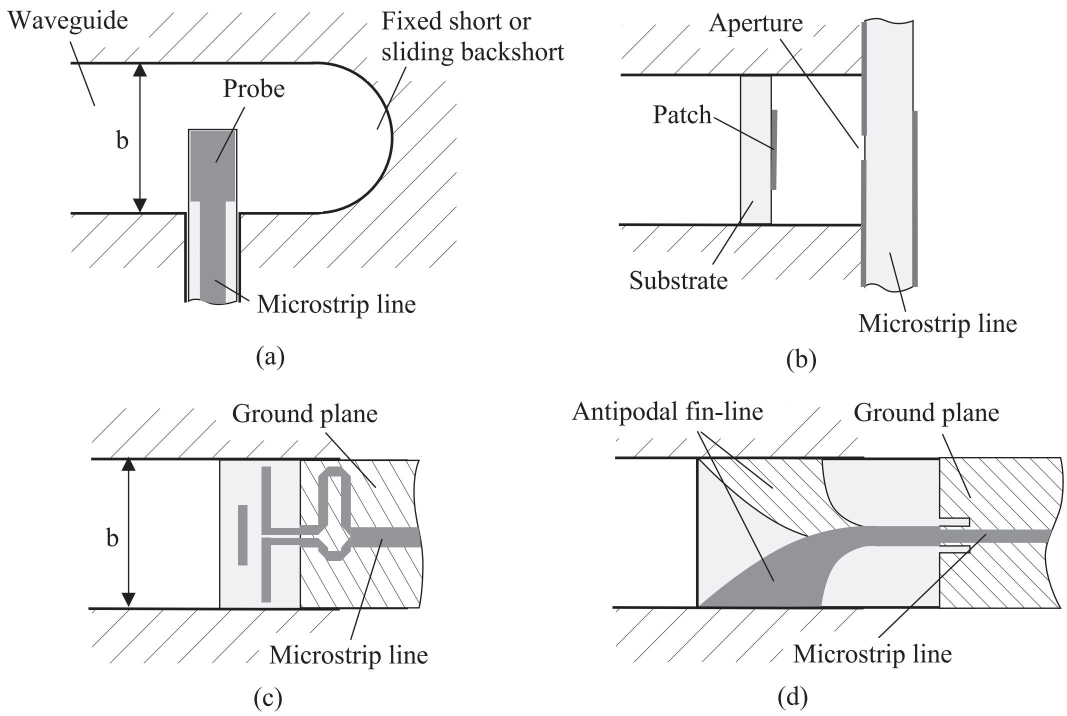

Figure 2.1. *Rectangular waveguide-to-microstrip line transitions based on (a) a probe [32], (b) aperture coupling and a patch [40], (c) a quasi-Yagi antenna [42], and (d) an antipodal fin-line [43].*

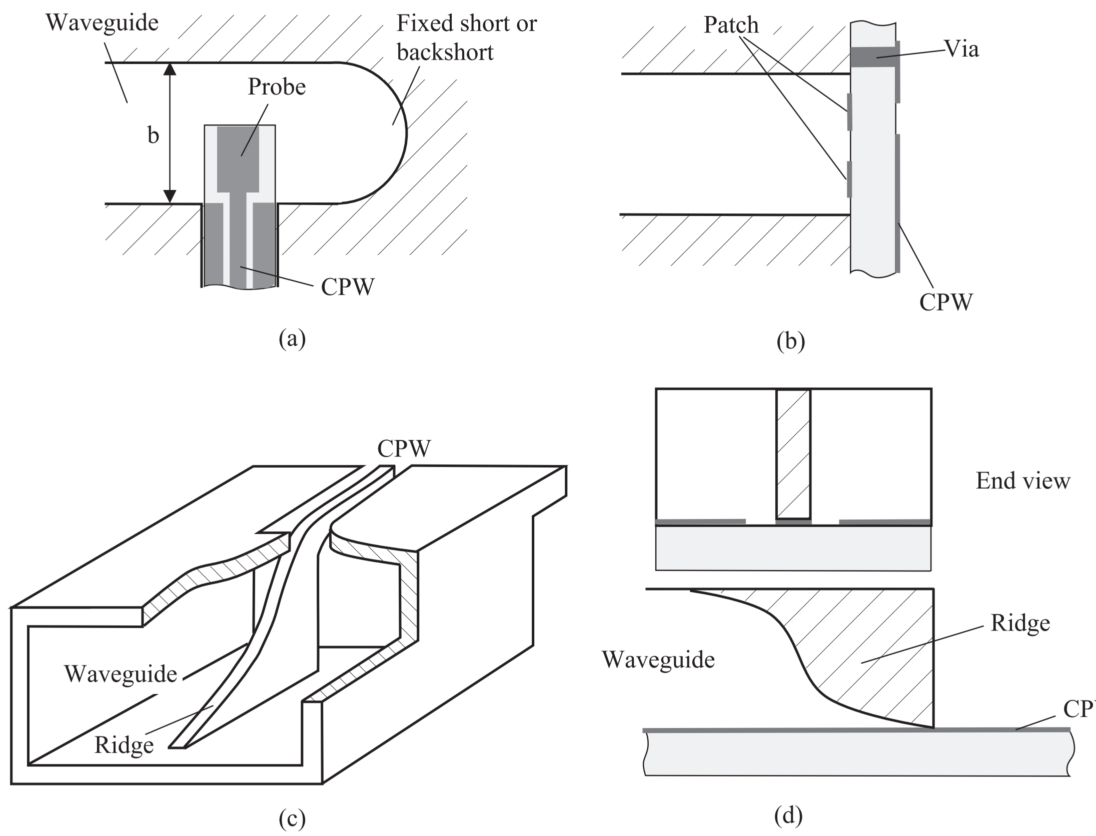

Figure 2.2. *Rectangular waveguide-to-CPW transitions based on (a) a probe [45], (b) coupled patches [48], and (c), (d) ridged waveguides [49], [50].*

CPW providing a smooth transition with wide-band operation. For a K-band (18−26.5 GHz) back-to-back transition (CPW substrate with a relative permittivity $\varepsilon_r = 2.2$) an insertion and a return loss of less than 2 dB and more than 11 dB, respectively, have been measured. A similar type of transition in [51] uses a microstrip line between the ridge waveguide and a conductor-backed CPW. Instead of a smooth ridge taper, a gradual taper with $\lambda/4$-long sections is applied. The microstrip line and conductor-backed CPW are made on a silicon membrane to minimise the dielectric loss. The simulated insertion and return loss for a single transition are less than 0.3 dB and more than 15 dB, respectively, at a frequency range of 27.5−30 GHz. A gradual ridge has also been used in [52] to construct a V-band (50−75 GHz) wafer probe. The ridge waveguide converts to a ridge-through waveguide, which is then connected to the CPW. The ridge-through waveguide suppresses the coupled slotline mode. The measurements of a single V-band probe with an alumina CPW show an insertion and return loss of less than 4 dB and more than 10.7 dB, respectively. Although these ridge-based transitions offer a wide-band operation, their fabrication due to mechanical complexity becomes very difficult or even impossible when approaching submillimetre wave frequencies.

In addition to the ridged waveguide-based designs, there exist other transitions (Figure 2.3) aligned in the longitudinal direction of the waveguide enabling easier construction [53]−[57]. Figure 2.3 (a) shows a transition, which utilises a slotline probe to couple the TE_{10} waveguide mode to a slotline mode [53]. The probe is followed by a quarter-wave slotline transformer and by a slotline-to-CPW transition based on a CPW short stub and air bridges. The unilateral circuit is placed in the centre of the waveguide E-plane. Measurements of a back-to-back transition structure (substrate with ε_r = 2.2) at X-band show an insertion and a return loss of less than 0.75 dB and more than 14 dB, respectively, over a bandwidth of 40 %. The slotline probe is also used in [54] for a transition to a conductor-backed CPW. The transition has a slotline-to-CPW transition different from the one in [53] using also a multi-section slotline. Vias are applied to suppress the coupled slotline and parallel plate mode. The simulated return loss for a back-to-back transition (alumina substrate with $\varepsilon_r = 9.8$) at the V-band is more than 15 dB over a bandwidth of 29 %.

Figure 2.3 (b) shows a quasi-Yagi “antenna”-based waveguide-to-CPW transition [55]. A trenched metal block, which works as a reflector, is used to create a CPW channel. For a coplanar strips (CPS)-to-CPW transition air bridges are required. The measured return loss of a back-to-back structure (substrate with $\varepsilon_r = 10.2$) at X-band is more than 10 dB over a bandwidth of 33 %. Proper operation of the quasi-Yagi “antenna” requires a high-permittivity substrate.

Figure 2.3 (c) illustrates a transition, which uses a unilateral fin-line taper and CPS-to-CPW transition [56]. A disadvantage is that the air bridge required has to be placed precisely in a proper angle. A prototype back-to-back structure (substrate with $\varepsilon_r = 2.22$) has been measured at the W-band. Due to the structure, resonances occur in the waveguide frequency band. The measured minimum insertion loss is 2 dB.

Good performance over a wide band can be obtained with a transition in Figure 2.3 (d) [57]. The transition from the waveguide to the CPW is accomplished through a microstrip line. With an antipodal fin-line the TE_{10} mode converts first to the microstrip line mode and then with a smooth two-sided taper to the CPW mode. The measurements of a Ka-band back-to-back structure (substrate with $\varepsilon_r = 2.22$) give an insertion and a return loss of less than 2.3 dB and more than 9 dB over a frequency range of 28–40 GHz. Due to the tapers the transition is relatively long and it requires two-sided fabrication.

In this thesis, two new waveguide-to-CPW transitions aligned in the longitudinal direction of the waveguide have been designed in [P1] and [P2]. The first one [P1] uses a rectangular probe (Figure 2.4) perpendicular to the waveguide axis. The second one [P2] is based on the use of a fin-line taper and slotline radial stub (Figure 2.5). The designs have been tested with X-band (8.2–12.4 GHz) back-to-back transitions. The measurements have shown the transitions to be low-loss over the full waveguide frequency band. The measured insertion and return loss for the transition in Figure 2.4 (substrate with $\varepsilon_r = 2.33$) are less than 0.5 dB and more than 17 dB (over a 40-% bandwidth), respectively. The transition in Figure 2.5 has been tested with two different substrates ($\varepsilon_r = 2.33$ and 10.8). An insertion and return loss of less than 0.4 dB and more than 16 dB (over a 40-% bandwidth), respectively, are measured for the low-permittivity substrate. Values for the high-permittivity material are less than 1.0 dB and more than 10 dB, respectively, at 8.2–12 GHz (over a 36-% bandwidth). The transition design in [P1] has also been used successfully in the LO transition (90–140 GHz) of a submillimetre-wave mixer [P6]. These transitions provide a practical and reliable alternative for millimetre-wave applications. The simple design, unilateral circuit structure, easy fabrication and suitability for different materials make these convenient for integration of MMIC designs with waveguide systems. Both designs work without air bridges if a proper waveguide housing is used.

Properties of different rectangular waveguide-to-CPW transitions aligned in the longitudinal direction of the waveguide are summarised in Table 2.1.

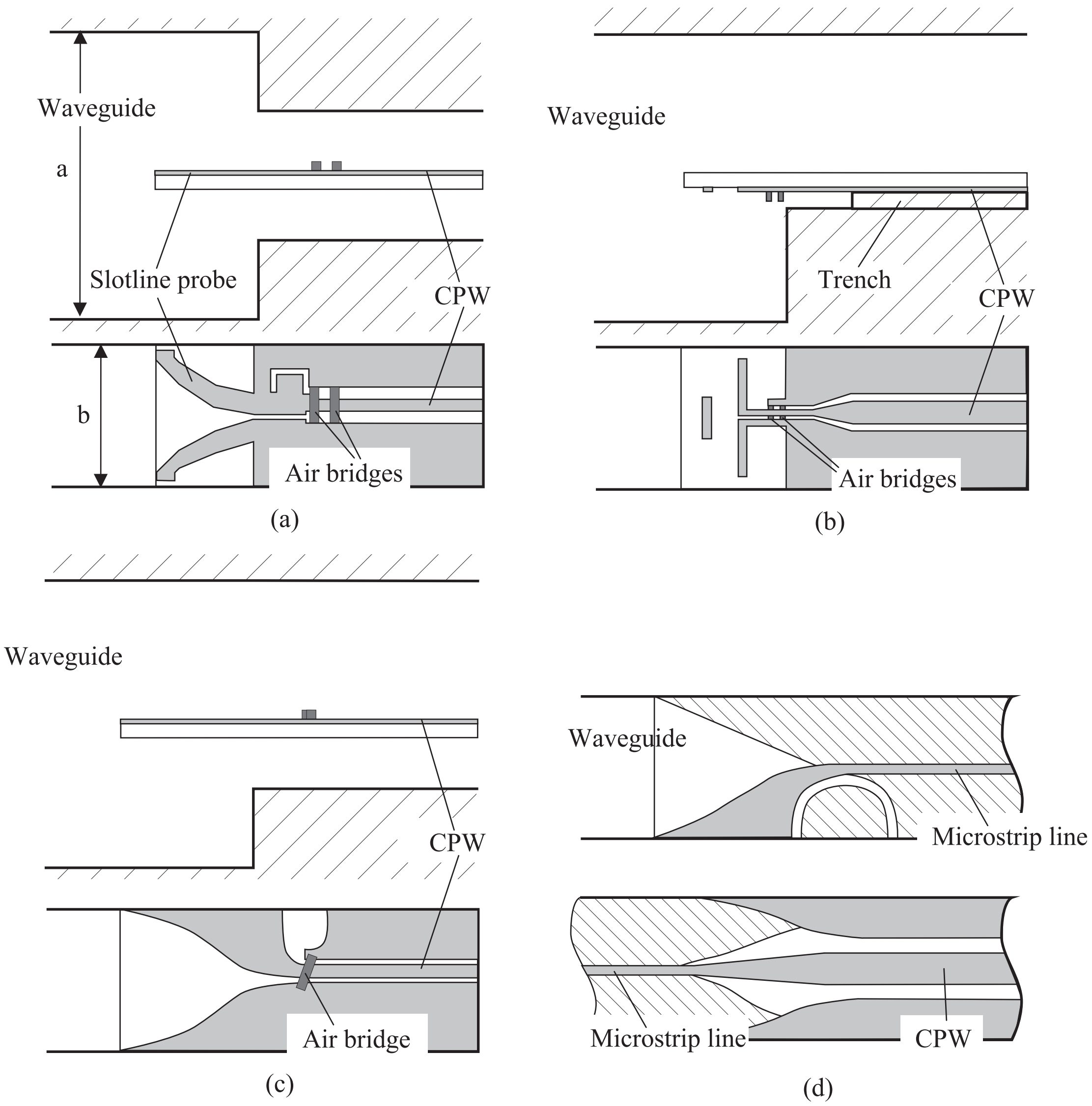

Figure 2.3. *Rectangular waveguide-to-CPW transitions based on (a) a slotline probe [53], (b) a quasi-Yagi antenna [55], (c) a unilateral fin-line [56], and (d) an antipodal fin-line [57].*

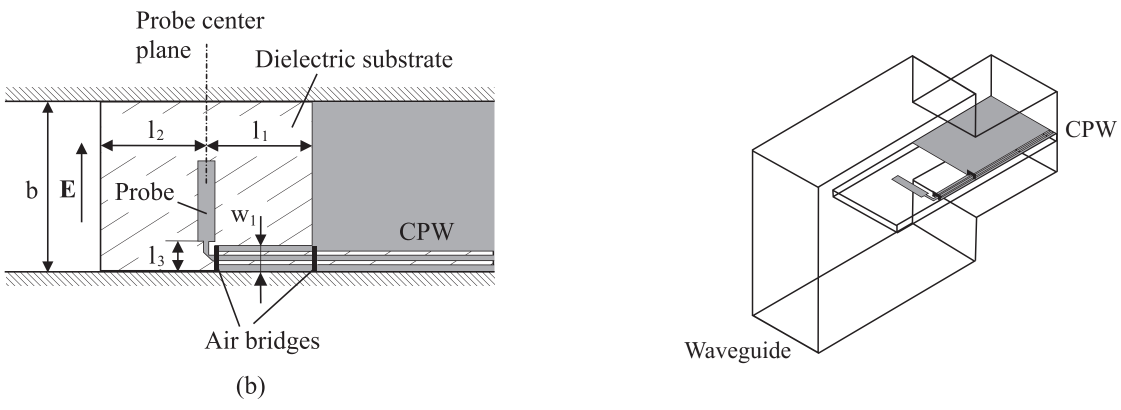

Figure 2.4. *Novel rectangular waveguide-to-CPW transition based on a rectangular probe [P1]: (a) a top view and (b) an overview. An alternative waveguide housing is also shown in [P1].*

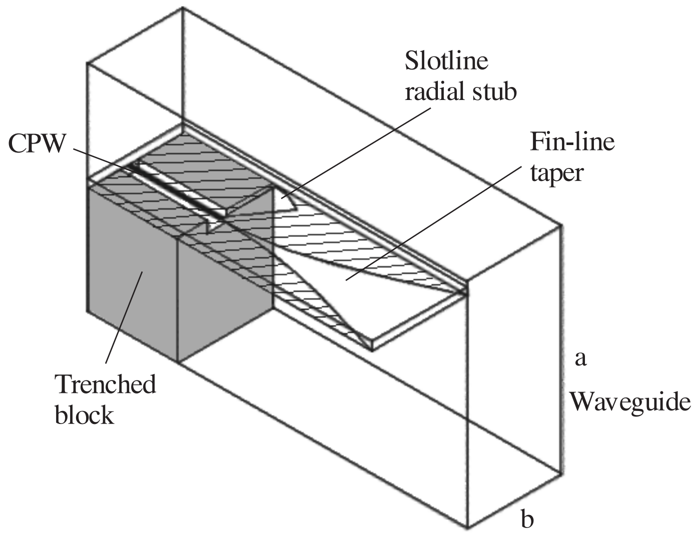

Figure 2.5. *Novel CPW-to-rectangular waveguide transition using a fin-line taper and slotline radial stub [P2].*

Table 2.1. *Properties of different rectangular waveguide-to-CPW transitions aligned in the longitudinal direction of the waveguide.*

Transition	Advantages	Disadvantages
[49], Fig. 2.2 (c), based on a smooth ridge	Wide-band, smooth transition, moderate loss	Mechanically complex, fabrication very difficult at higher millimetre wave frequencies, grounded centre conductor
[50], Fig. 2.2 (d), based on a smooth ridge	Wide-band, smooth transition, moderate loss, simple design	Fabrication difficult at higher millimetre wave frequencies, grounded centre conductor
[51], based on a gradual ridge	Easier fabrication than in [53] and [54], low-loss	Narrower bandwidth than in [53] and [54], fabrication difficult at higher millimetre wave frequencies, grounded centre conductor
[52], based on a gradual ridge	Wide-band, moderate loss	Mechanically complex, fabrication difficult at higher millimetre wave frequencies
[53], Fig. 2.3 (a), based on a slotline probe	Wide-band, low-loss, easy fabrication	Air bridges, grounded centre conductor
[54], based on a slotline probe	Low-loss, easy fabrication	Moderate bandwidth, vias required
[55], Fig. 2.3 (b), based on a quasi-Yagi “antenna”	Wide-band, low-loss, easy fabrication	High-permittivity substrate, air bridges, design complicated
[56], Fig. 2.3 (c), based on a fin-line taper	Easy fabrication	Narrow-band, resonance problems, air bridges, grounded centre conductor
[57], Fig. 2.3 (d), based on antipodal fin-lines	Wide-band, smooth transition, easy fabrication	Moderate loss, two-sided fabrication, grounded centre conductor, design complicated
[P1], Fig. 2.4, based on a rectangular probe	Wide-band, low-loss, easy fabrication, simple design	Air bridges required depending on a waveguide housing in use
[P2], Fig. 2.5, based on a fin-line taper	Wide-band, low-loss, easy fabrication, simple design	Grounded centre conductor

3 Waveguide Backshorts

In tuning of a waveguide-based device, like oscillators, very wide-band multipliers, and harmonic mixers, movable backshorts are used. A backshort, i.e., an adjustable short circuit, which produces a variable reactive load, adjusts the device input or output impedance. For moving the backshort a micrometer drive is often used. In an ideal case, the backshort reflects all the incident power (zero return loss and infinite voltage standing wave ratio, VSWR) and varies the phase of the reflected wave by a desired amount. In practice, a good backshort has a return loss of less than 0.3 dB, i.e., VSWR > 58. Further, if the backshort is not working properly it might produce abrupt changes in the phase.

Tunable backshorts can be divided into contacting and noncontacting ones. The contacting ones have a galvanic contact with the waveguide whereas this is advisedly prevented in case of the noncontacting ones. Figure 3.1 introduces some backshort designs. Figure 3.1 (a) shows a simple contacting backshort: a block or plunger, which slides in the waveguide [58]. The problem with this structure is the erratic contact, which causes the deviation of the electrical short-circuit position and, thus, phase instability of the reflected wave. Also, some power can leak over the backshort. These matters get worse in heavy use due to wear of the surfaces of the block and waveguide. Further, the implementation of a snug fit to the waveguide becomes very difficult at millimetre wavelengths. To obtain a firm contact with the waveguide walls, a cam-based backshort design can be used as presented in [59]. However, this structure is not suitable for millimetre waves.

The performance of the design in Figure 3.1 (a) is improved with the backshort in Figure 3.1 (b), which uses a two-section quarter-wave transformer [58]. The transformer lowers the impedance of the plunger in proportion to the square of the ratio of gaps b_1 and b_2 and, thus, increases the VSWR by $(b_2/b_1)^2$. Unfortunately, due to the quarter-wave sections the structure becomes frequency-dependent. The bandwidth can be increased by using more sections and by using sections of different lengths. At second harmonic frequencies when the transformer sections are close to half-wavelength long, the VSWR decreases significantly which might cause additional losses in multipliers and mixers.

Figures 3.1 (c)–(e) show spring-type (loop- or finger-type springs) backshorts better suitable for smaller waveguides [59], [60]. However, it is difficult to create a uniform contact at submillimetre wavelengths when the waveguide dimensions are fractions of a millimetre. Also, the degradation of contacting areas due to sliding friction still exists which may lead to unreliable operation. Furthermore, one has to take into account the oxidation in material selection (gold coating mostly preferred). When working properly these backshorts provide a short circuit over the entire waveguide frequency band. As a function of frequency they have a small phase variation since the electrical short circuit plane is not in the end of the backshort.

Figure 3.1 (f) presents a noncontacting backshort [61], which uses a similar approach as the one in Figure 3.1 (b). The backshort works as a bandstop filter consisting of several low- and high-impedance sections insulated from the waveguide with a dielectric film. The use of the film, typically Mylar® tape, permits smooth sliding within the waveguide with negligible wear. The absence of electrical contact brings better repeatability compared to the contacting backshorts. The length of the sections can be designed to differ from quarter wavelength (typically $\lambda_g/8$–$\lambda_g/4$, where λ_g is the wavelength in the waveguide) to improve the performance at second harmonic frequencies. For Chebyshev and empirical designs a VSWR of 60–90 (return loss of 0.2–0.3 dB) was measured at a frequency range of 75–116 GHz. Reference [62] presents a similar design with a photolithography-based fabrication process. Dumbbell-type backshorts of Figure 3.1 (g) have been

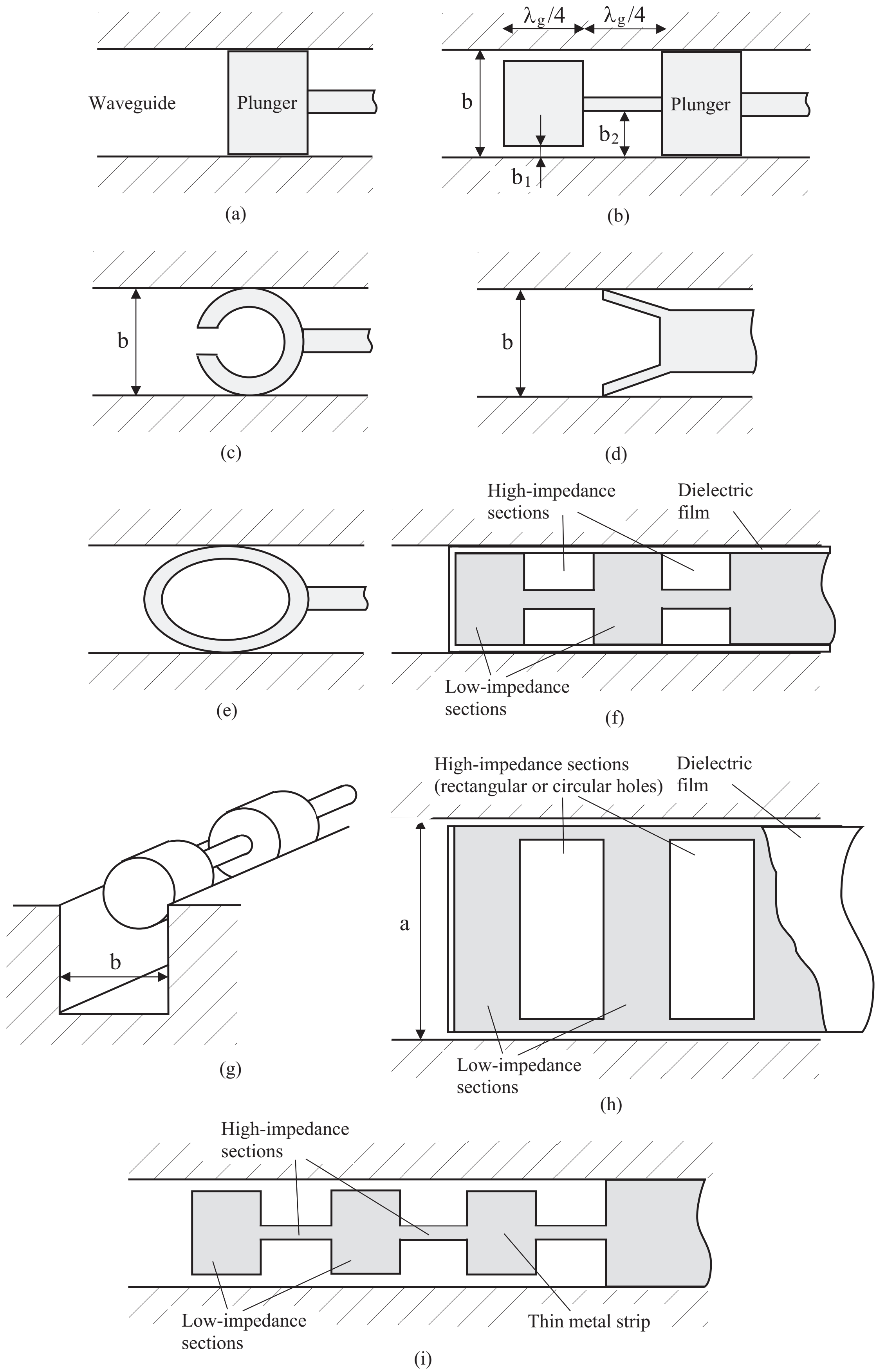

Figure 3.1. *Different rectangular waveguide backshorts. (a)−(e) Contacting backshorts; (f)−(i) noncontacting backshorts. The backshort (i), with a thin metal strip in the center of the waveguide E-plane, is the only planar short circuit.*

applied as well [59], [63], and [64]. In [64], a Ku-band scale model has been used to design a seven-section dumbbell-type backshort for 170–260 GHz. Reference [64] also discusses in-band resonances with noncontacting backshorts having a rectangular or circular cross-section. The resonances are due to undesired modes excited with misaligned plungers. It is shown that the backshort plungers with circular cross-sections are much less prone to the in-band resonances.

When approaching submillimetre wavelengths the high-impedance sections become too thin to be fabricated easily and to be mechanically strong enough for tight sliding in the waveguide. To overcome this, a structure with high-impedance sections made of circular or rectangular holes as shown in Figure 3.1 (h) has been designed [60], [65], [66]. The holes can be fabricated more easily (e.g., using photolithography) and, by using holes, a higher impedance ratio is achieved. With rectangular holes a return loss of less than 0.3 dB (VSWR > 58) over a 30 % bandwidth has been measured at the W-band (75–110 GHz, 38 % bandwidth). Although now the fabrication of the metal bar structure is easier, placing of the thin dielectric tape into a small waveguide is challenging. Instead of using a tape, dielectric layer could be, e.g., sputtered or evaporated on the metal [66], [67]. In [67] a SiO_2 layer is evaporated on the backshort structure similar to the one in Figure 3.1 (f). A VSWR of 96 has been measured at 345 GHz.

Figure 3.1 (i) shows a planar backshort which can be fabricated using photolithography [68]. This enables fabrication at submillimetre wavelengths. It is based on a thin metal strip, which has alternating impedance sections, as in the backshorts above, placed in the center of the waveguide E-plane. The backshort has been tested at the Ku-band (12.4–18 GHz). The measured return loss is less than 0.1 dB (VSWR > 173). The measured phase variation is about 8º over the frequency band. This structure should be aligned accurately in the waveguide in order to maintain the bandwidth.

In the thesis [P3], a new tunable millimetre and submillimetre-wave noncontacting waveguide backshort is designed and tested. Figure 3.2 illustrates the design. It is based on a fixed waveguide short and movable dielectric slab, which is used to adjust the effective phase constant of the wave in a waveguide section. The structure offers a simple design and fabrication. The simplicity makes it also reliable and low-loss. This design is not sensitive to alignment errors. Furthermore, the in-band resonances, which are possible with the noncontacting backshorts as discussed above are avoided in this structure by using a thin quartz slab and narrow guide channel. In this way, only small alignment errors are possible. Measurements of a W-band backshort show a return loss of less than 0.21 dB (VSWR > 82) over the entire waveguide frequency band. This backshort gives most accurate adjustment. For example, at 92.5 GHz (centre frequency of the W-band) a 10-μm tuning distance corresponds to a 1.7º shift in the phase of the reflected wave in case of the other backshorts (Figure 3.1) whereas in the case of the tested new backshort this corresponds to a phase-shift of about 0.37º. This means an almost 4.6 times more accurate tuning capability at this frequency. This is a great advantage for submillimetre-wave applications. A disadvantage, which may limit the use in some applications, is a large and slightly nonlinear phase variation as a function of frequency. For a tuning distance corresponding to a 360º shift in the phase at 92.5 GHz the phase variation over the frequency band is close to 1080º.

Table 3.1 summarises properties of different backshorts.

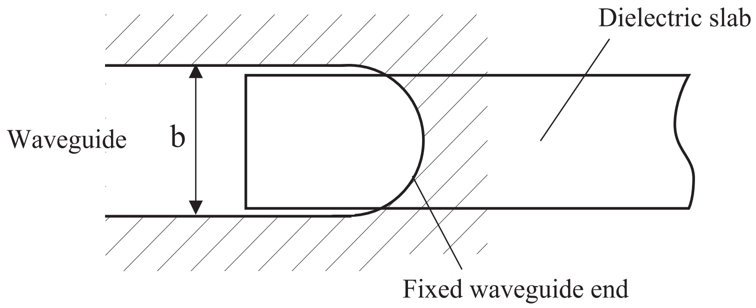

Figure 3.2. *New tunable noncontacting waveguide backshort developed in the thesis [P3]. It is based on a dielectric slab pushed into the waveguide through a hole in a fixed waveguide end. As the slab is moved the electrical length of the waveguide changes and, thus, the phase of the reflected wave changes too.*

Table 3.1. *Properties of different backshort designs discussed above.*

Backshort	Advantages	Disadvantages
Contacting, Fig. 3.1 (a)	Frequency independent	Non-repeatability, phase instability, wear, risk of power leakage, fabrication difficult at millimetre wavelengths
Contacting, Fig. 3.1 (b)	More reliable than (a), better phase stability	Frequency dependence, wear, fabrication difficult at submillimetre wavelengths
Contacting, Fig. 3.1 (c)–(e)	Frequency independent, better contact than in (a) and (b)	Unreliable, phase instability, wear, fabrication difficult at higher frequencies, risk of power leakage
Noncontacting, Fig. 3.1 (f)	Repeatable performance, no risk of wear	Complicated design, fabrication difficult at submillimetre wavelengths, sensitive for alignment
Noncontacting, Fig. 3.1 (g)	Repeatable performance, no risk of wear	Complicated design, fabrication difficult at millimetre wavelengths
Noncontacting, Fig. 3.1 (h)	Repeatable performance, no risk of wear	Complicated design, fabrication difficult at submillimetre wavelengths but easier than that of (f), sensitive for alignment
Noncontacting, Fig. 3.1 (i)	Low-loss, reliable, easier fabrication, no risk of wear, suitable for submillimetre wavelengths	Complicated design, sensitive for alignment
Noncontacting, Fig. 3.2	Low-loss, reliable, no risk of wear, easy design, easy fabrication, suitable for submillimetre wavelengths, most accurate tuning, no risk of in-band resonances	Large and slightly nonlinear phase variation as a function of frequency

4 Bias T – Coupling of Bias to RF Circuit

Devices based on semiconductor components often require dc voltage and for that a biasing network which has a negligible effect on the RF performance of a device. For instance in mixers, a low-loss and wide-band structure is preferred in separating the dc bias and IF signal. The biasing network, bias T, can be either an integrated part of the device or a stand-alone component. Basically, it is a diplexer formed by a low-pass and high-pass filter (Figure 4.1).

Some bias circuits are discussed as follows. In [69], a bias circuit has been designed to be an integrated part of a millimetre-wave Schottky diode mixer. An alumina-based microstrip line circuit with discrete elements provides both the IF match and dc bias separation. The dc bias is connected through a low-pass filter to the middle of an open-circuited $\lambda/2$ stub, which is a part of the IF matching circuit. In this way, the dc bias connection does not affect the IF bandwidth. Integration has also been applied in [70] in a coaxial line structure. The dc bias is coupled through a shorted (at the IF frequency) $\lambda/4$ stub. The matching circuit has an IF VSWR of less than 1.2 (return loss > 20.8 dB) at 1.2–1.8 GHz (bandwidth of 40 %). In [71], the dc bias is applied to a mixer diode through a hybrid IF matching circuit consisting of capacitors and inductors. The circuit has been empirically adjusted for a minimum insertion loss of less than 0.5 dB with an IF bandwidth of 45 %. References [72] and [73] present stand-alone bias T designs based on discrete components and a microstrip line. In these designs, a wide bandwidth has been achieved with a proper design of the low-pass section. In [72], commercially available capacitors and a self-made inductor (copper wire with a 50-μm diameter) have been used on an alumina substrate. The measured insertion loss is less than 2.5 dB up to 20 GHz and less than 0.5 dB at 0.5–7.5 GHz. At the latter frequency range, the return loss is more than 20 dB. In [73], a micromachined inductor is used with commercial capacitors. The measured insertion and return loss are less than 1 dB and more than 15 dB, respectively, at 3–10 GHz. A more sophisticated structure with a band-pass filter (Figure 4.2) has been used in [74]. The sections of the band-pass filter effectively lower the external system impedance near the high-impedance stub. This reduces the shunting effect of the stub enabling a wide-band performance. With this design method a stripline and microstrip line-based bias T have been constructed for frequency ranges of 2–18 GHz and 4.5–45.5 GHz, respectively. For the stripline bias T with a six-element Chebyshev design a return loss of more than 13.5 dB at 2–18 GHz has been measured. The insertion loss increases from 0.4 dB at 2 GHz to 1.4 dB at 18 GHz. In the case of the 4.5–45.5-GHz microstrip line bias T, the insertion and return loss are less than 1 dB and more than 15 dB, respectively, up to 26 GHz. Above 26 GHz the values are less than 1.5 dB and more than 10 dB, respectively. Measurements have been carried out up to 40 GHz. One of the best commercially available bias tees is a Picosecond Pulse Labs bias tee, model 5542 [75]. It has a 3-dB bandwidth from 10 kHz to 50 GHz. At 1–20 GHz, the insertion and return loss are less than 0.8 dB and more than 15 dB, respectively.

Within this thesis work [P4], a coaxial line-based bias T has been realised by using the design adopted from [74]. It has been designed for higher IF frequencies, around 10 GHz. Figure 4.3 shows one half of the bias T. It has two 45-Ω coaxial lines as band-pass filter sections, which was found to give adequate frequency coverage in this case. Unlike in [74], here the dc block capacitor is placed in the first element of the band-pass filter. The bias T has been tested at 1–17 GHz. The use of the coaxial line structure has provided a very low-loss, wide-band operation with a high return loss. The insertion loss is less than 0.5 dB at 3–16 GHz (fractional bandwidth of 137 %). The return loss is more than 20 dB from 4.2 to 15 GHz and more than 30 dB from 5.2 to 14.1 GHz. At the latter range, the insertion loss is about 0.1 dB. The measured RF isolation is more than 30 dB over the whole measurement frequency band. When comparing the results to those (insertion loss of 0.5 dB and return loss of 20 dB at best) of the bias circuits discussed above, the insertion loss is extremely low and the return loss is very high. This bias T can also be designed to provide a proper IF match as an integrated structure.

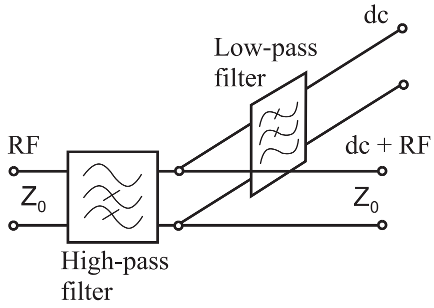

Figure 4.1. *Equivalent circuit of a simple bias T.*

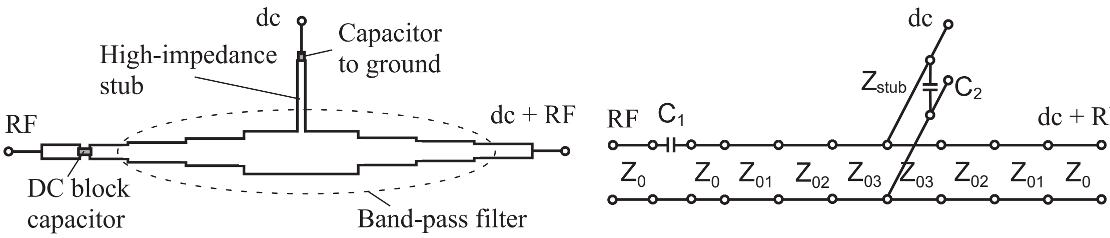

Figure 4.2. *Wide-band bias T structure with six transmission line sections [74]: (a) stripline or microstrip line structure, (b) equivalent circuit (lines and stub $\lambda/4$ in length).*

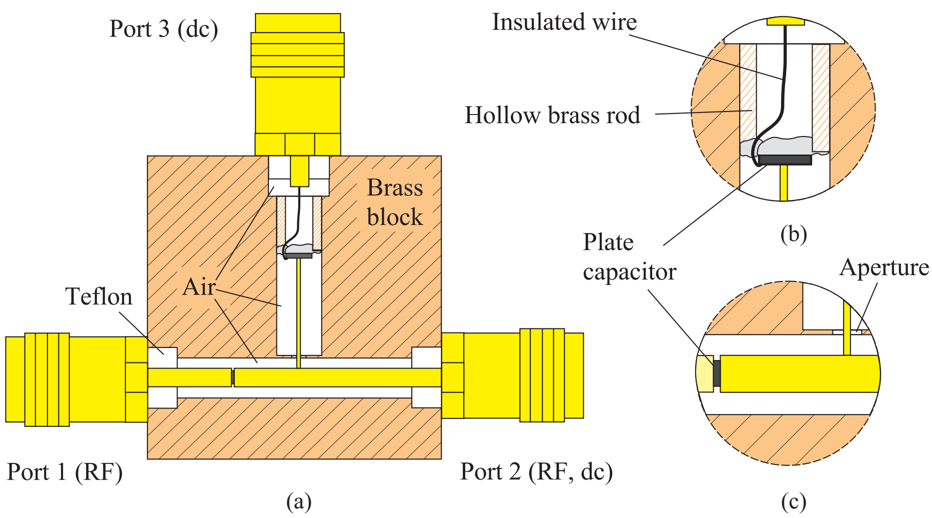

Figure 4.3. *Coaxial line bias T [P4]: (a) cross-section, (b) blow-up of the RF block, (c) blow-up of the dc block and T-junction.*

5 Schottky Diode Waveguide Mixers

As mentioned earlier, the mixer is often the first element in a millimetre- or submillimetre-wave receiver. Furthermore, the most common mixing element is a GaAs-Schottky diode. Main reasons are its ability to provide efficient performance also at room temperature, easy usage, and a well-controlled GaAs fabrication process. GaAs diodes suitable for terahertz applications are already commercially available [76]. For reliability planar diode chips are currently preferred over the whisker-contacted ones. They have also produced comparable mixing performance. This thesis includes the design and construction of two different Schottky diode mixers.

In the mixers constructed, discrete diode chips have been used. The chips have been attached to quartz circuits by flip-chip soldering. For one mixer single-diode chips were soldered in parallel whereas in the other mixer an anti-parallel pair was used. The soldering process had a major role in constructing the mixers. The thesis shows that soldering can be used for constructing efficient mixers. However, there is a tendency toward integration techniques [5], [6], [77]−[80] which eliminates the soldering process and, thus, provides easier, faster, and more reliable construction. Fabrication technologies of the mixer waveguide housing are also being improved for a cheaper and easier direction, e.g., [3], [81].

5.1 Subharmonic Waveguide Mixer at 215 GHz

Subharmonic waveguide mixers using planar Schottky diodes and operating near 200 GHz are presented in [6], [9]−[13], and [15]. Table 5.1 summarises the best results obtained in these. The lowest double-sideband noise temperature (T_{DSB}) of 490 K has been achieved with an integrated structure at 240 GHz [6]. An integrated structure has also provided the lowest double-sideband conversion loss (L_{DSB}) of 4.7 dB [9]. The integrated structures use a so-called quartz-substrate upside-down integrated device (QUID) process, which removes the need for diode flip-chip soldering. In [10], by using multiple layers, the biasing of the anti-parallel diode pair has been enabled. All the mixers using discrete diode chips apply a single GaAs chip with a pair of anti-parallel Schottky diodes. The mixers apply a rectangular waveguide mount with quartz filters. Furhermore, they (except [11], information not available) use a four-tuner waveguide structure.

In this thesis work, a subharmonic waveguide mixer at 215 GHz [P5], has been designed within an international ESA project [82], [83] as a study of possibilities of future millimetre- and submillimetre-wave receivers (both open-structure and waveguide types) based on European Schottky diodes. The main goal was to develop European single and anti-parallel Schottky planar diode chips. The diode development work has been carried out at Technical University of Darmstadt (TUD). These chips have a special quasi-vertical structure [84], [85] which was designed to combine the benefits of earlier planar and whisker diode chip structures. During the design of the subharmonic mixer, based on a chip with an anti-parallel pair of diodes (APD), the diode chip fabrication process was in a strong development phase. Especially, yield and uniformity of the process was not good at the time. This produced additional difficulties to the mixer construction. However when compared to other subharmonic waveguide mixers operating at the same frequency range (Table 5.1) very satisfactory results were obtained. The noise temperature (3500 K, single-sideband) is about 3.6 times higher than the lowest one (490 K, double-sideband). However, the conversion loss (9.2 dB, single-sideband) is comparable to the other ones reported. Also the LO power consumption is similar. The non-uniformity also caused that instead of a two-tuner structure (planned in the early project phase) a four-tuner one [86] was selected for efficient match of the APD chip. This was the first time these diodes were applied in a mixer. The work produced valuable information on the use of this type of diodes for mixer applications. Especially, it showed that these diodes are usable for millimetre- and submillimetre-wave mixers with further improvement of the diodes. Since then the structure of the diode chip and its fabrication process have been further developed by TUD [84], [87], [P7].

Table 5.1. *Performance of subharmonic waveguide mixers using planar Schottky diodes and operating near 200 GHz: double-sideband noise temperature T_{DSB}, double-sideband conversion loss L_{DSB}, LO power P_{LO}.*

Mixer	RF [GHz]	T_{DSB} [K]	L_{DSB} [dB]	P_{LO} [mW]	Other
[6]	240	490	-	3	integrated structure, 2-GHz IF
[9]	200	600	4.7	30	integrated structure, 1.5-GHz IF, high LO power requirement due to resonance problem
[9]	205	700	5	8	integrated structure, 1.5-GHz IF
[10]	210	1420	7.6	6.4	integrated structure, biasable diode pair, unbiased diodes
[10]	210	1640	8.4	3.2	integrated structure, biasable diode pair, biased diodes
[11]	225	900	6.2	1.5	integrated structure
[12]	200	500	6.6	3	discrete diode chip
[13]	230	1900*	9.5*	-	discrete diode chip, 1.5-GHz IF
[15]	205	1590*	8.7*	5.7	discrete diode chip, 1.4-GHz IF
[15]	205	1990*	9.3*	3	discrete diode chip, 1.4-GHz IF
[P5]	215	3500*	9.2*	3.5	discrete diode chip, 1-GHz IF

* for SSB values

5.2 Harmonic Waveguide Mixer

Harmonic mixers are useful devices for measurement instrumentations, e.g., for spectrum analyzers or phase-locking systems. Often in these application, the conversion loss is not critical but other properties like usability, compactness, and wide bandwidth are more important.

A quasi-optical harmonic mixer in [88] uses a corner cube antenna, Schottky diode, and LO waveguide structure. A conversion loss of 40 dB has been achieved at 604.3 GHz for a 6th-harmonic mixing with an LO (~100 GHz) power of 7.2 mW. By using a 25th-harmonic mixing at 2523 GHz, 7 dB signal-to-noise ratio has been measured with an LO (100.9 GHz) power of 9 mW. In [89], a 5th-harmonic mixer at 77 GHz is presented for a phase-locking system. It applies a flip-chip-mounted planar double-diode silicon Schottky diode chip in an alumina microstrip line circuit. The measured conversion loss is 22 dB with a 10-mW LO power and an IF of 1 GHz. By utilising the silicon technology and a double-diode chip in a CPW structure, a conversion loss of 23 dB has been obtained for an 8th-harmonic mixer at 38 GHz with an LO frequency of 4.6 GHz and power of 100 mW [90]. A different quasi-optical approach is applied in [91] where an anti-parallel GaAs diode chip is used with a log-periodic antenna and silicon lens. The mixer operates over a wide frequency range of 30−140 GHz using even harmonic mixing products. The conversion loss is 9.2 dB at 34 GHz and 35 dB at 140 GHz with an IF of 1.4 GHz. The required LO power is 30−80 mW. With the 4th harmonic of the LO the conversion loss is 15−19 dB at 35−60 GHz and with the 6th harmonic it is 21−29 dB at 55−100 GHz. A monolithic 4th-harmonic mixer based on GaAs PHEMT diodes is presented in [92]. By using also PHEMT amplifier stages, a conversion gain of 0.8 dB has been achieved at an LO frequency of 14.5 GHz (12-dBm LO power) and an RF of 60.4 GHz.

Within this thesis work, a 5th-harmonic waveguide mixer has been designed for a frequency range of 500–700 GHz [P6]. It is used for submillimetre-wave test equipment of a hologram-based compact antenna test range [93], [94]. It has been in use in phase locking of a backward wave oscillator operating around 650 GHz. Unlike the subharmonic mixer discussed earlier, this mixer does not include any tuners and, instead of one diode chip, it uses two planar single-diode chips (chips from Virginia Diodes Inc.) in a balanced type configuration on a quartz filter. It also has an integrated diagonal horn antenna for the RF feed. Without tuners and with an LO port access for an InP Gunn oscillator operating at 100–140 GHz it enables an easy usage in antenna test facilities. This is the first time when the balanced type configuration has been used in a harmonic mixer operating at these frequencies. The RF signal sees the diodes in series whereas the LO signal in parallel. In the case of identical diodes and a proper circuit design, even harmonics and mixing products due to even harmonics of the LO do not exist as the CPW mode. However, they can propagate toward waveguide as the TE_{10} mode or as a slotline mode toward the microstrip channel. In the mixer design, the cut-off frequency of the waveguide is 440 GHz. Thus, at frequencies below that the waveguide gives reactive termination. The height of the microstrip channel is 100 µm, which prevents the slotline mode propagation to that direction. The imbalance of the configuration, i.e., a measured open circuit voltage on the IF terminal, is 30–90 mV with an LO power level of 2–10 dBm, respectively, at 130 GHz. Harmonic mixers with similar properties are not readily available. The measured conversion loss of less than 30 dB is low for this type of a mixer. Commercially available harmonic mixers at these frequencies [95], enabling the use of an LO waveguide source at close to 100 GHz, have a quasi-optical structure. For these mixers the conversion loss is 45–50 dB. Table 5.2 summarises the results of different harmonic mixers. By comparing the results, the performance of the 5th-harmonic mixer designed within this thesis work can be considered to be very good.

Table 5.2. *Performance of harmonic mixers using Schottky diodes: conversion loss L, LO harmonic number n, LO power P_{LO}.*

Mixer	RF [GHz]	L [dB]	n	P_{LO} [mW]	Other
[88]	604.3	40	6	7.2	quasi-optical, whisker-contacted diode
[88]	2523	-	25	9	quasi-optical, whisker-contacted diode, 7-dB signal-to-noise ratio
[89]	77	22	5	10	planar double-diode silicon chip, alumina microstrip line circuit
[90]	38	23	8	100	planar double-diode silicon chip, silicon CPW circuit
[91]	30–140	8–35	2,4,6, 8,10	30–80	quasi-optical, planar double-diode GaAs chip
[91]	35–60	15–19	4	-	quasi-optical, planar double-diode GaAs chip
[91]	55–100	21–29	6	-	quasi-optical, planar double-diode GaAs chip
[92]	60.4	-0.8	4	15.8	monolithic GaAs circuit, PHEMT diodes and amplifiers
[95]	600–800	45–50	8, 9	10	quasi-optical, whisker-contacted diode
[P6]	650	27	5	10	waveguide, integrated antenna, two planar single-diode chips

6 Schottky Diode Characterisation Through Measurements

The construction of high-efficiency millimetre- and submillimetre-wave devices requires that the circuit design relies on valid component models. The parameters of the models are extracted experimentally by measurements or theoretically by calculations and simulations. One of the best examples of a thorough characterisation of a planar Schottky diode chip is based on dc measurements and extensive electromagnetic structure simulations [96]. In this characterisation procedure, values for dc parameters are extracted from measurements and values for parasitic elements of the diode equivalent circuit are extracted from the results of electromagnetic structure simulations at 555–615 GHz with a circuit simulator.

Some measurement-based characterisations in a planar transmission line environment are presented in [97]–[101]. Coaxial line and waveguide measurement structures have also been used. However, the simplicity of planar transmission line test mounts makes them more desirable. In [97], a two-port microstrip line test mount is used with a TRL (thru-reflect-line) calibration method. Both small and large signal measurements are carried out. The small signal measurements at 0.6–4.2 GHz are done at eight bias points. After this, a circuit simulator is used to extract diode parameter values from the measurement results. On the contrary in [98], a one-port microstrip line mount has been applied. The reflection coefficient has been measured at six bias points (from –5 to 0 V) at 0.5–1.5 GHz. In [99], a large signal model for a Schottky varactor, accurately valid up to 40 GHz, is created through CPW on-wafer measurements. Varactors are measured at 40 different bias points from –8 V to 0.5 V in a two-port microstrip line mount with CPW probe pads. Values of the model are found by fitting the response of the model to the measured data. Reference [100] presents a method for measuring the resistance and capacitance of a diode in a two-port microstrip line test fixture. A vacuum pump holds down the diode providing a non-destructive measurement. Measurements have been carried out up to 20 GHz. Dc and single-frequency *S*-parameter measurements are applied in [101] for extraction of the diode parameter values. In [102], diodes are characterised with a one-port *S*-parameter measurement up to 62.5 GHz. The measurement set-up comprises just a network analyzer with CPW probes and a specific diode test structure suitable for contacting with the CPW probe.

This thesis presents a characterisation procedure based on dc, capacitance, and wide-band *S*-parameter measurements [P7]. The idea is to provide designers with wide-band (up to 220 GHz) *S*-parameter data in addition to parameter values of a simple equivalent circuit, which also includes the junction capacitance and parasitic capacitance. The *S*-parameter data can then be used to extract a more detailed equivalent circuit. This data is measured for diode chips mounted in a CPW environment. In the first measurements single-diode chips, varactors and mixer diodes, of TUD have been used. The measurements have been carried out at different frequency ranges on the chips, which are flip-soldered on quartz CPW test beds. Calibration is done with TRL elements fabricated on the same substrate. First characterisation measurements have been successfully carried out. The results have been analysed to make conclusions of the diode quality.

7 Summary of Publications

Publication [P1] describes the design, construction, and measurement of a new wide-band coplanar waveguide-to-rectangular waveguide transition. It uses a rectangular probe to couple the energy from the waveguide TE_{10} mode to the CPW mode. Due to a unilateral and an in-line structure, the fabrication of the CPW circuit and waveguide block is easy. The transition has been validated through designing and testing of an X-band (8.2−12.4 GHz) model. For a back-to-back double transition, the measured return loss is more than 17 dB and the insertion loss is less than 0.5 dB over the whole X-band. The measured insertion loss value indicates a loss of less than 0.15 dB for a single transition. The results agree well with the simulated ones obtained with an electromagnetic structure simulator. A transition based on this design is used for a local oscillator feed of a submillimetre-wave harmonic mixer [P6]. This transition has been designed for the F-band (90−140 GHz) with a full frequency band coverage. Having a simple structure to design and fabricate, the new transition is a good, practical, and reliable alternative for millimetre-wave applications. It is characterised by wide-band operation, low insertion loss, and high return loss. Both low- and high-permittivity materials can be applied with this transition.

Publication [P2] introduces a new CPW-to-rectangular waveguide transition using a fin-line taper. Like the transition above this has a uniplanar and an in-line structure enabling easy construction. It applies a fin-line taper to convert the waveguide TE_{10} mode to the fin-line mode and a slotline radial stub in the fin-line-to-CPW transition. X-band transitions with a low- and high-permittivity material (ε_r = 2.33 and 10.8, respectively) have been designed, constructed, and measured. Both transitions work well. The measured insertion losses for back-to-back transitions are less than 0.4 dB and 1.0 dB, respectively. These indicate losses of less than 0.14 dB and 0.36 dB, respectively, at the centre frequency for a single transition. This design provides another good alternative for millimetre-wave applications. The applications might differ between this one and the previous since in this transition the CPW centre conductor is grounded.

Publication [P3] presents the design, construction, and measurement of a new tunable millimetre- and submillimetre-wave waveguide backshort. It is a noncontacting type of backshort based on a fixed waveguide short and movable dielectric slab. The slab is moved in the waveguide, through a hole in the fixed waveguide end, in order to tune the effective phase constant of a wave in a waveguide section. The designed backshort has following advantages: simple design, simple fabrication, easy and accurate tuning, low losses, insensitivity for alignment errors, high reliability, and readiness for scaling. These properties make it suitable for millimetre- and submillimetre-wave applications. The validity of the novel design was demonstrated by designing and testing a W-band (75−110 GHz) backshort. The measured return loss is less than 0.21 dB (i.e., VSWR > 82) over the whole waveguide frequency band. The correspondence between simulated and measured results is excellent.

Publication [P4] shows a low-loss wideband microwave coaxial bias T designed for millimetre- and submillimetre-wave mixers to cover a wide range of IF frequencies. The design presented in [73] was applied here as a coaxial line structure. Low losses, proper matching, and simplicity were the objectives for the design. The implemented structure is easy to design and fabricate and has an excellent performance making it competitive with commercial ones. The bias T was tested in the 1−17 GHz frequency range. The insertion loss is less than 0.5 dB at 3−16 GHz (fractional bandwidth of 137 %). The return loss is more than 20 dB from 4.2 to 15 GHz and more than 30 dB from 5.2 to 14.1 GHz. In the latter range, the insertion loss is about 0.1 dB. The measured RF isolation is more than 30 dB over the whole measurement frequency band.

Publication [P5] describes the design, construction, and measurement of a millimetre-wave subharmonic waveguide mixer based on European quasi-vertical Schottky diodes. The aim of this study was to test the suitability of these diodes for millimetre-wave receivers as a part of an international receiver development program (see Chapter 5). This mixer worked as one qualification step in this development. It applied a diode chip with a pair of anti-parallel quasi-vertical diodes. To obtain a proper matching for the diodes, the mixer has a four-tuner structure [78]. The diode was flip-chip soldered on a quartz circuit. Mixer circuits were designed using a circuit and structure simulator. The mixer was tested at an RF frequency of 215 GHz using a WR-10 InP Gunn oscillator as the LO source. Satisfactory results, an SSB noise temperature of 3500 K and conversion loss of 9.2 dB with an LO power of 3.5 mW, were obtained. The performance of the mixer and experiences attained in the study showed that the quasi-vertical diodes are suitable for sensitive millimetre- and submillimetre-wave applications with further improvement of the diode structure and fabrication process.

Publication [P6] shows the design of a wide-band submillimetre-wave fifth-harmonic waveguide mixer. It was designed for the instrumentation of a submillimetre-wave hologram-based compact antenna test range. The mixer uses two planar single-diode chips in a balanced type of configuration. It applies an InP Gunn oscillator as the LO source at a frequency range of 100−140 GHz enabling an operational RF frequency of 500−700 GHz. The novel waveguide-to-CPW transition [P1] is used in the LO waveguide. The mixer block is compact comprising an integrated diagonal horn antenna and no waveguide tuners. This enables good usability in the antenna test system. The measurements have shown a conversion loss of less than 30 dB at 650 GHz with an LO power of about 10 mW. The loss value is low for a fifth-harmonic mixer operating at these frequencies. The mixer is currently being used for phase locking of a submillimetre-wave backward wave oscillator.

Publication [P7] describes a procedure for characterisation of millimetre-wave planar diode chips. The procedure is based on dc-characteristics, capacitance, and wide-band (up to 220 GHz) *S*-parameter measurements and parameter extraction from the measurement results. Diode chips, varactors and mixer diodes, fabricated by TUD have been used as the first test items. The *S*-parameter measurements are carried out on the diode chips mounted on quartz CPW test beds. Calibration of the test system is done with TRL calibration elements made on the same substrate. The *S*-parameter data are obtained for several different bias voltages and currents. As the output of the procedure, parameters of a simple diode equivalent circuit and results of extensive measurements are available for designers for further use. The results have been analysed to make qualitative conclusions of the fabrication process of the tested diodes.

8 Conclusions

This thesis work is based on seven scientific articles covering different millimetre- and submillimetre-wave receiver front-end circuits and components. As a result of the work, new circuit structures are available for millimetre wave designers to construct devices with improved performance, easier designing, easier fabrication, better reliability, or with possibility for new circuit implementations.

Two new waveguide-to-CPW transitions have been invented and developed. By designing and testing X-band transitions, these structures have shown to have a low-loss performance over a wide band (fractional bandwidth of > 40 %). These are new practical alternatives for future applications. Due to their simple design, unilateral circuit, and easy fabrication, they are suitable for integration of MMIC designs with waveguide systems. One transition has already been applied in a submillimetre-wave mixer.

For tuning of waveguide devices, a new adjustable waveguide backshort has been invented and developed. It has a performance competitive with the best ones published. It offers several advantages for millimetre- and submillimetre-wave devices: low losses, large bandwidth, most accurate adjustment, easy design and fabrication, reliability, and insensitivity for alignment errors.

A stand-alone, coaxial line-based bias T has been designed, constructed, and measured. It provides a wide-band, low-loss performance suitable for, e.g., mixer IF outputs.

Two Schottky diode waveguide mixers have been designed, constructed, and measured. One of these is a subharmonic millimetre-wave mixer at 215 GHz utilising European quasi-vertical Schottky diodes. During the project this work has produced valuable information for the diode manufacturer and has shown the ability of these diodes for high-frequency applications. The other mixer is a fifth-harmonic submillimetre-wave mixer, which has been designed for instrumentation used in antenna test facilities at 500–700 GHz. It was designed to offer an easy usage in the antenna test system. It has been in use in a phase locking system of a backward wave oscillator.

A planar diode chip characterisation procedure has been developed and tested. The procedure provides millimetre wave designers with the results of a variety of extensive measurements together with the parameter values of a simple equivalent diode circuit.

Future research topics are the development of a stand-alone waveguide impedance tuner based on two backshorts invented here, testing of the 5th-harmonic mixer at a wider frequency range, extraction of a more specific diode model from the characterisation results, characterisation of low-series resistance varactors, application of the backshort and characterised diodes in construction of a submillimetre-wave mixer, and construction of a microelectromechanical systems-based waveguide impedance tuner with the invented transitions.

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