UNIVERSIDAD POLITÉCNICA DE MADRID
ESCUELA TÉCNICA SUPERIOR DE INGENIEROS DE TELECOMUNICACIÓN

TECHNOLOGICAL DEVELOPMENTS FOR
MILLIMETER-WAVE RADARS
DESARROLLOS TECNOLÓGICOS PARA RADARES EN
BANDAS DE MILIMÉTRICAS
TESIS DOCTORAL

Gorka Rubio Cidre
Ingeniero de Telecomunicación

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DEPARTAMENTO DE SISTEMAS, SEÑALES Y RADIOCOMUNICACIONES
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TECHNOLOGICAL DEVELOPMENTS FOR MILLIMETER-WAVE
RADARS

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El tribunal nombrado por el Mgfco. y Excmo. Sr. Rector de la Universidad Politécnica de Madrid, el día __ de junio de 2017.

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Calificación:

EL PRESIDENTE

LOS VOCALES

EL VOCAL SECRETARIO

A mis padres.

A Julia.

Hay genios sin estudios e idiotas con doctorados. La verdadera sabiduría no te la otorga un título sino lo que haces con lo que has aprendido a lo largo de tu vida y la manera en cómo tratas a los demás.

Joseph Kapone.

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Steve Jobs.

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Resumen

Los desarrollos tecnológicos en la banda de milimétricas han progresado recientemente de forma significativa gracias a la creciente disponibilidad de fuentes de generación de señal y detectores a estas frecuencias, que ofrecen elevados anchos de banda de trabajo, por lo que resultan de interés en aplicaciones tales como espectroscopía, seguridad o radio astronomía. Además, el desarrollo de tecnología en estas bandas se espera que desempeñe un papel importante en la creación de nuevos campos industriales que satisfagan determinadas necesidades de la sociedad. La banda de milimétricas comprende la región del espectro electromagnético entre 30 GHz y 300 GHz, que corresponde a una longitud de onda entre 1 y 10 mm. Esta región constituye la frontera entre la electrónica y la fotónica, que requiere el desarrollo de nueva circuitería que sea capaz de manejar mayores niveles de potencia y anchos de banda que cubran el vacío tecnológico existente.

Mientras que el desarrollo de tecnología es un tema de interés por sí mismo, el desarrollo de tecnología es más efectivo en un contexto donde tenga importancia para una aplicación determinada. Las aplicaciones de seguridad, biomedicina y observación de la tierra y del espacio imponen unos requisitos que demandan soluciones innovadoras donde la tecnología radar en banda de milimétricas aporta considerables beneficios. Por tanto, la presente tesis doctoral tiene como objetivo principal el desarrollo de tecnología para radares en banda de milimétricas, concretamente para dos radares a 94 GHz y 300 GHz, para aplicaciones de meteorología y seguridad, respectivamente. Ambas aplicaciones implican requisitos completamente distintos con respecto al subsistema de antenas, los tiempos de adquisición, la potencia transmitida y la generación de señal en banda estrecha y banda ancha.

La primera parte de la tesis se centra en el desarrollo de tecnología para un radar a 300 GHz diseñado en el Grupo de Microondas y Radar de la Universidad Politécnica de Madrid en colaboración con la Universidad de Vigo. Los sensores radar activos en la banda de milimétricas son una tecnología competitiva para la detección a distancia de objetos ocultos bajo la ropa sin que la persona que está siendo escaneada sea consciente de ello, a diferencia de los sistemas tipo portal, lo que aumenta la vulnerabilidad de dichos sistemas. Además, al contrario que la tecnología de rayos X, la radiación es no ionizante, por lo que no existe riesgo alguno para la salud. El trabajo realizado en esta primera parte consiste en el desarrollo de tres modificaciones hardware que permiten mejorar el rendimiento del radar, con distintas alternativas en la generación de la señal radar, el uso de la información contenida en la fase y la evaluación de técnicas de polarimetría.

La segunda parte de la tesis se centra en el diseño, implementación y caracterización de un radar perfilador de nubes a 94 GHz. La dinámica de las nubes sigue siendo una fuente importante de incertidumbre respecto al equilibrio energético de la tierra y al calentamiento global, ya que estas contribuyen al ciclo hidrológico y la redistribución del calor y la humedad. Este tipo de sistemas tienen gran interés en meteorología, sobre todo en el análisis de nubes sin precipitación, tanto desde un punto de vista comercial como de investigación.

Abstract

The technological developments in the millimeter-wave band have seen a rapid progress recently, owing to the increasing availability of signal-generation sources and detectors at this band, which offer high bandwidths that are of interest for applications such as spectroscopy, security and radioastronomy. Furthermore, the development of technology at this band is expected to play an important role in the creation of new industrial fields that meet society needs. The millimeter-wave band covers the electromagnetic spectrum from 30 GHz to 300 GHz, which corresponds to the wavelength region from 10 mm to 1 mm. This region is a part of a frontier area between electronics and photonics which demands the development of new circuitry that will be able to handle larger power levels and bandwidths in order to cover this technological gap.

While the development of technology is an interesting topic itself, the development of technology is more effective when it is considered within the context of a certain application. The requirements of security, biomedical and earth-and-space remote sensing applications demand innovative solutions for which millimeter-wave band radar technology can yield considerable benefits compared to other technologies. To this end, the main goal of this Ph.D. thesis is the development of technology for millimeter-wave radars, particularly for two radars at 94 and 300 GHz for weather monitoring and security applications, respectively. Both applications imply completely different system requirements in terms of antenna sub-system, acquisition times, transmitted power and narrowband/broadband signal generation.

The first part of this Ph.D. thesis is focused on the development of technology for a 300-GHz imaging radar designed at the Microwaves and Radar group of the Universidad Politécnica de Madrid in collaboration with the Universidad de Vigo. Active radar sensors at millimeter-wave band are a competing technology for the detection of concealed objects under clothes without the targets being aware of their scanning in risk public areas in contrast to portal systems which increase the system vulnerability. Furthermore, as opposed to X-ray technology, the radiation at millimeter-wave band is non-ionizing and therefore, raise safety concerns. The work carried out in this first part consists of three hardware modifications that improve the radar performance with different alternatives of radar-signal generation, the use of the phase information and the evaluation of polarimetry techniques.

The second part presents the design, implementation and characterization of a cloud profiling radar at 94 GHz. Clouds dynamics remain a major source of uncertainty regarding the earth energy balance and global warming, since they contribute to the hydrological cycle and heat and moisture redistribution. This kind of system has a strong interest in meteorology, particularly in the non-precipitating cloud analysis from a commercial and research perspective.

Contents

1	Introduction	1
1.1	Millimeter-wave spectrum and applications	1
1.2	Radar-based applications	4
1.2.1	Standoff imaging systems	4
1.2.2	Cloud profiling radars	7
1.3	Millimeter-wave spectrum properties	10
1.3.1	Atmospheric propagation	10
1.3.2	Material properties	12
1.4	State-of-the-art of millimeter-wave sources	12
1.5	Thesis motivation and goals	17
1.6	Thesis organization	18
1.7	Publications	18
1.8	Research projects related to this thesis	20

I Development of technology for millimeter-wave radars at UPM

2	Hardware-based modifications of millimeter-wave CW-LFM radars	23
2.1	Operating principle of CW-LFM radars	23
2.2	CW-LFM radars developed in ETSIT-UPM	24
2.2.1	Standoff imaging radar at 300 GHz	25
2.2.2	Multi-purpose radar sensor at 100 GHz	27
2.3	Broadband chirp source based on a DDS/PLL structure	28
2.3.1	DDS-based signal generation	30
2.3.2	Chirp sources characterization	36
2.3.3	Evaluation criteria for radar detection performance	41
2.3.4	Radar performance characterization	42
2.3.5	Imaging performance	51
2.3.6	Chirp sources design optimization	55
2.4	Phase stability analysis	59
2.4.1	DDS phase stability	63

2.4.2	Microwave-subsystem phase stability	64
2.4.3	Imaging-radar phase stability	66
2.4.4	Micrometer displacements by means of a Z-axis micropositioner	69
2.5	Impact of frequency reduction and polarimetry techniques	71
2.5.1	Radar configuration description	71
2.5.2	Radar performance	76
2.5.3	Radar images	81

II Design and development of a cloud profiling radar at 94 GHz 83

3	Preliminary design of a cloud profiling radar at 94 GHz	85
3.1	Physical properties of clouds	85
3.1.1	Drop size distribution	86
3.1.2	Physical moments of the drop size distribution	87
3.1.3	Fall velocity of cloud particles	87
3.2	Interaction between radar signals and clouds	89
3.2.1	Radar cross section	89
3.2.2	Radar reflectivity	91
3.2.3	Doppler velocity	94
3.2.4	Liquid water content	95
3.2.5	Atmospheric attenuation	96
3.2.6	Radar polarimetry	97
3.3	Meteorological instrumentation	99
3.3.1	Weather stations	99
3.3.2	Radiosondes	100
3.3.3	Weather satellites	100
3.3.4	Lidars, ceilometers and disdrometers	100
3.3.5	Passive radiometers	100
3.3.6	Scintillometers	101
3.3.7	Weather radars	101
3.3.8	Cloud radars	102
3.4	State-of-the-art of millimeter-wave cloud radars	103
3.4.1	Research radars	103
3.4.2	Commercial radars	107
3.5	Radar equation for a cloud radar at 94 GHz	109
3.5.1	Received power	109
3.5.2	Sensitivity and dynamic range	116
3.6	Parallax error	119
3.7	Cloud radars calibration	123

4	Design and implementation of a cloud profiling radar at 94 GHz	125
4.1	Introduction	125
4.2	Design of the 94 GHz cloud profiling radar	125
4.2.1	Market survey of millimeter-wave radar components	126
4.2.2	Radar operational parameters definition by means of the radar equation	129
4.3	System architecture	131
4.4	Hardware characterization	133
4.4.1	Digital electronics	134
4.4.2	Microwave subsystem	134
4.4.3	Millimeter-wave subsystem	138
4.5	System characterization	140
4.5.1	Spectral purity of the microwave subsystem	140
4.5.2	Receiver noise floor, sensitivity and dynamic range	140
4.5.3	Point-target detection	144
4.6	Signal processing	148
4.6.1	Radar reflectivity	148
4.6.2	Doppler velocity	156
4.7	Radar data exportation and dimensioning	162
4.8	Analysis and interpretation of cloud observations	166
4.8.1	January 29th, 2016	166
4.8.2	February 5th, 2016	173
4.8.3	February 18th, 2016	175
4.8.4	May 12th, 2016	177
	Conclusions and future work	181
	Conclusions	181
	Guidelines for future work	182
III	Appendices	185
A	Millimeter-wave component datasheets	187
B	Characterization of millimeter-wave components of the cloud profiling radar at 94 GHz	225
B.1	W3 tripler from Pacific Millimeter	225
B.2	QMB balanced mixer from QuinStar Technology, Inc.	227
B.3	LNA from Radiometer Physics GmbH	227
B.4	MPA from Farran Technology, Ltd.	227
B.5	x6 frequency multiplier from Radiometer Physics GmbH	230
B.6	HPA from Farran Technology, Ltd.	231

C	Active filters design	233
C.1	Introduction	233
C.2	Filter I: Active filter for the imaging radar at 300 GHz	234
C.2.1	Simulation results	234
C.2.2	Characterization results	235
C.3	Filter II: Active filter for the cloud profiling radar at 94 GHz	240
C.3.1	Simulation results	240
C.3.2	Characterization results	240
D	Receiver noise power level calculations of the cloud profiling radar at 94 GHz	245
E	Automatic biasing sequencer design	249
E.1	Introduction	249
E.2	Drain power supply	250
E.3	Gate power supply	252
E.4	Power-on sequencer	252
E.5	Measurements	255
F	Weathering protection for a Cassegrain antenna	257
F.1	Introduction	257
F.2	Feeder measurements	260
F.3	Cassegrain antenna measurements	264
G	Photographs of the sky condition at the beginning of the measurements with the cloud profiling radar at 94 GHz	269
G.0.1	January 29th, 2016	269
G.0.2	February 5th, 2016	270
G.0.3	February 18th, 2016	270
G.0.4	May 12th, 2016	271

List of Figures

1.1	The electromagnetic spectrum and the millimeter-wave sub-bands designation.	1
1.2	Overview of millimeter-wave band applications.	2
1.3	Portal-based millimeter-wave imaging systems from L-3 Communications Security and Detection Systems (a) and Smiths Detection (b).	6
1.4	Three proposed solutions from Rohde & Schwarz for faster performance and to increase detection capability.	7
1.5	Atmospheric Regions. Graph extracted from [Ric10].	8
1.6	Photograph of two commercial cloud profiling radar from Prosensing [Wid12] (a) and Radiometer Physics GmbH [Rad14a].	10
1.7	Atmospheric phenomenology at millimeter-wave band.	11
1.8	Attenuation (one way) in dB/km due to rain [ITU 03] (a) and clouds [ITU 09] (b) as a function of the rain rate and water vapor concentration, respectively.	11
1.9	Output-power performance of different sources around the <i>THz gap</i> . Graph extracted from [Cha11]. RTD: Resonant tunnel diode. IMPATT: Impact ionization avalanche transit-time diode. Gunn: Gunn laser. QCL: Quantum-cascade laser. III-V: Denotes groups III and V in the periodic table of elements. UTC-PD: Uni-traveling-carrier photodiode. GaN: Gallium nitride. InP: Indium phosphide. TUNNET: Tunnel injection transit time diode.	13
1.10	Photograph of a THz Quantum Cascade Laser (a), an 1 kW EIK at 94 GHz (a) and a 37 W solid-state GaN amplifier at W band (c).	14
1.11	Saturated output power of different signal sources, active and non-linear components: RPG and VDI transmitter chains, RPG and VDI multipliers, commercial and research MMIC amplifiers, commercial and research tubes and transmitter sources developed in this thesis.	15
1.12	Fractional bandwidth of different signal sources, active and non-linear components: RPG and VDI transmitter chains, RPG and VDI multipliers, commercial and research MMIC amplifiers, commercial and research tubes and transmitter sources developed in this thesis.	16
2.1	Block diagram of a CW-LFM radar with two different antenna configurations: single antenna and two antennas.	23
2.2	Transmitted and received chirp signals of a CW-LFM radar.	24
2.3	Block diagram of the imaging radar architecture. The radar structure is divided into four different parts, corresponding to the technologies involved in the design: Digital electronics, microwave subsystem, millimeter-wave subsystem and scanning antenna subsystem.	25

2.4	Photograph of the imaging radar showing the key components and the transmitting and receiving signal paths.	26
2.5	Photograph of the millimeter-wave subsystem showing the key components. .	26
2.6	Simplified block diagram of the 100-GHz radar system.	27
2.7	Photograph of the 100-GHz radar system.	27
2.8	Simplified block diagram of the chirp sources. Test points are defined with capital letters inside squares in the block diagrams.	32
2.9	Photograph of the DDS-based signal generation module of chirp source 1. . .	33
2.10	Photograph of the microwave subsystem of chirp source 1.	33
2.11	Photograph of the DDS-based signal generation module of chirp source 2. . .	34
2.12	Photograph of the microwave subsystem of chirp source 2.	34
2.13	Power ripple and SFDR at the DDS output (test point C) of both chirp sources. The output power is -3 and 3 dBm, respectively.	37
2.14	Power ripple and SFDR measured at the chirp sources output (test point E). The SFDR is -41 and -61 dBc respectively. The output power is 13 dBm in both cases.	38
2.15	Phase noise measurements at the test points of the developed chirp sources: 10 MHz master oscillator output (A), PLO output (B), DDS output (C), coupler output (D) and chirp sources output (E). The VCO phase noise of the narrowband DDS/PLL is plotted in dashed line.	39
2.16	RF non-linearities measurement at the chirp source outputs (test point E) and of the Agilent E8257D analog signal generator. Inset: Block diagram of the delay-line test system.	40
2.17	Radar performance limitations: RF non-linearities and decorrelation effect of the phase noise around the target response. Spurious signals are caused by the DDS-based signal generation.	42
2.18	Figures of merit for the radar detection performance evaluation: SNR is measured with a single target; ΔR and IDR are measured with two targets. . . .	43
2.19	Simplified set-ups for radar performance evaluation.	44
2.20	Received power from an aluminium plate as a function of the angle of incidence.	44
2.21	Phase noise estimation at 300 GHz based on the measured phase noise at the chirp sources output (Fig. 2.15).	45
2.22	Sphere noise floor measurement (black line) and noise floor estimation (blue line) based on the phase noise measurement of Fig. 2.21.	46
2.23	Range resolution measurement using the developed chirp sources.	48
2.24	IDR measurement by adjusting the angle of incidence on the second target to four different angles: 1° , 1.5° , 2° and 2.5° . The second target is separated 1.6 cm (11.5 kHz of beat frequency variation) from the first target. The IDR is higher than 30 dB with both chirp sources.	49
2.25	Mannequin with threats under clothing.	51
2.26	Radar image results of the external and internal layers of a mannequin with explosives hidden under a T-shirt using the developed chirp sources. The colorbar represents the range axis in meters.	53

2.27	SNR results as a function of the maximum reflected power per pixel for all pixels obtained from a mannequin with hidden threats of Fig. 2.26 using the developed chirp sources.	54
2.28	Phase noise estimates at 300 GHz based on the phase noise measurement at the chirp sources output (test point E, 16 GHz) by replacing the master oscillator before and after decorrelation effect in the mixer.	56
2.29	Radar image results of the internal layer of a mannequin with explosives hidden under a T-shirt using the developed chirp sources with the Vectron oscillator.	57
2.30	Phase noise estimates at 300 GHz based on the phase noise simulation at the chirp sources output (test point E, 16 GHz) of both chirp sources before and after decorrelation effect in the mixer.	58
2.31	Relative displacements precision as a function of the phase drift and the operating center frequency.	59
2.32	Phase drift $\Delta\phi$ as a function of relative humidity and temperature by using the ITU model [ITU 15], considering a Δd of 8 meters and an operating frequency of 94 GHz (a) and 300 GHz (b).	61
2.33	Phase stability over time at 1 GHz.	63
2.34	Phase stability over time at 16 GHz without temperature control.	64
2.35	Phase stability over time at 16 GHz with temperature control.	65
2.36	Phase stability over time at 300 GHz without temperature control.	66
2.37	Phase stability over time at 300 GHz with temperature control.	67
2.38	Power stability over time at 300 GHz.	67
2.39	Temperature, pressure and relative humidity measurements carried out with a weather station during the 300 GHz phase stability measurements.	68
2.40	Relative phase variation $\Delta\phi$ for each displacement measurement.	69
2.41	Comparison among the measured and theoretical range displacement Δd for each range displacement measurement.	70
2.42	Photograph of the circular-polarization radar configuration at 100 GHz on the scanning antenna subsystem.	72
2.43	Linear-polarization radar configuration at 100 GHz.	72
2.44	Circular-polarization radar configuration at 100 GHz.	72
2.45	Block diagram of a directional coupler used as a circulator in a radar system.	73
2.46	Directional coupler characterization.	73
2.47	Principle of operation of circular-polarization duplexer. Graph extracted from [Lea14b].	74
2.48	OMT performance results: Isolation and insertion losses.	75
2.49	SNR results of the linear-and-circular polarization radar configurations at 100 GHz compared to the 300 GHz radar. The measured SNR is 65 dB and 55 dB, respectively.	76
2.50	RF nonlinearities results of the linear-and-circular polarization radar configurations at 100 GHz compared to the 300 GHz radar. The measured SLL are 12 dB, 11 dB and 6 dB, respectively.	77
2.51	Spot size measurement at 100 GHz with both polarizations by using a 2.5-mm radius metallic sphere. The colorbar represents the received power in dBm.	78

2.52	Spot size measurement with the circular polarization. Two 2.5-mm radius metallic beads are spaced 4.5, 5, and 5.5 cm apart in the horizontal and vertical axes. The cross-range resolution is 5 cm in both axes.	78
2.53	Range resolution measurement when the antenna is pointing to two nylon threads with a measured spacing of ~ 3 cm.	79
2.54	Received power from an aluminium plate of 20×20 cm ² as a function of the angle of incidence at 100 GHz with both polarizations compared to the measurements at 300 GHz.	80
2.55	SNR results as a function of the maximum reflected power per pixel for all pixels obtained from a mannequin with hidden threats (see Figs. 2.56-2.57) with both radar configurations.	81
2.56	Radar image results of the internal and external layer surface reconstructions of a mannequin with explosives hidden under a T-shirt with the linear-polarization radar configuration at 100 GHz.	82
2.57	Radar image results of the internal and external layer surface reconstructions of a mannequin with explosives hidden under a T-shirt with the circular-polarization radar configuration at 100 GHz.	82
3.1	DSD distribution results of stratocumulus observations from [Mil00] modeled by the log-normal distribution of Eq. (3.1).	86
3.2	Different velocity vectors that affect cloud particles velocity: Falling velocity due to gravity action (red arrows), up- and downdrafts (green arrows) and turbulences (blue arrows).	88
3.3	Sphere scattering from different regimes: Rayleigh, resonance and optics regions, where a is the radius of the sphere and the sphere radar cross section σ has been normalized to the projected area of the sphere, πa^2 , plotted as a function of sphere circumference normalized to wavelength, $ka = 2\pi a/\lambda$. Graph extracted from [Ric10].	90
3.4	Radar cross section as a function of the particle diameter. Graph extracted from [Kol07].	90
3.5	Diagram of the radar cross section of a volume V that is illuminated by a radar.	91
3.6	Photograph of the different cloud types defined in Table 3.1.	93
3.7	Scatter diagram of radar reflectivity factor z versus LWC for measured spectrum obtained during the experiment ASTEX [Ovt99].	96
3.8	Atmospheric attenuation at standard and dry conditions. The water vapor concentration is 7.5 g/m ³ and 0 g/m ³ , respectively. The total pressure is 1013 hPa and the temperature is 20°C.	97
3.9	Block diagram of a full polarimetric radar system (top) and switch control signal for the synchronization of the chirp signal (bottom).	98
3.10	Meteorological instrumentation: Radiosonde, weather station, weather satellite, lidar, ceilometer, weather radar, passive radiometer, scintillometer and cloud radar.	99
3.11	Scanning strategies of weather radars (a) and cloud radars (b).	101
3.12	Rutherford Appleton Laboratory (RAL) cloud profiling radar.	103
3.13	Chiba University cloud profiling radar.	105
3.14	Miami University cloud profiling radar.	105
3.15	CloudSat cloud profiling radar.	106

3.16	MeTeK cloud profiling radar.	107
3.17	Prosensing cloud profiling radar.	108
3.18	Radiometer Physics cloud profiling radar.	108
3.19	Gain (a) and beamwidth (b) of Cassegrain and lens antennas as a function of frequency and antenna diameter. $\eta_{a_1} = 0.65$ and $\eta_{a_2} = 75^0$ have been considered.	110
3.20	Received power as a function of the radar reflectivity factor, transmitted power and diameter of transmit and receive antennas, D_{tx} and D_{rx} , respectively.	111
3.21	Received power as a function of the transmitted bandwidth, considering a radar reflectivity factor of -20 dBZ, a transmitted power of 31 dBm and antenna diameters of 1.1 m.	112
3.22	Atmospheric attenuation at three different temperatures: -20°C, 0°C and 20°C and three different pressures: 1013 hPa, 800 hPa and 600 hPa. The water vapor concentration is 7.5 g/m^3	113
3.23	Empirical relationship between saturation vapor concentration (SVC) and temperature (T).	114
3.24	Water vapor concentration (WVC) as a function of the temperature (T) and relative humidity (RH).	114
3.25	Atmospheric attenuation as a function of the relative humidity (RH), total pressure (Ph) and temperature (T) for the radar operating frequency (94 GHz).	115
3.26	SNR estimates as a function of the reflectivity factor and the PRF. The transmitted power is 31 dBm, the transmitted bandwidth is 15 MHz and the antenna diameter is 1.1 m.	117
3.27	SNR improvement as a function of integrated chirp periods (M).	118
3.28	SNR estimates as a function of the reflectivity factor and the PRF, considering a non-coherent integration of 35.000 chirp signals. The transmitted power is 31 dBm, the transmitted bandwidth is 15 MHz and the antenna diameter is 1.1 m.	118
3.29	Diagram of parallax error for two-antenna radar configuration with parallel-beams (a) and non-parallel-beams (b).	119
3.30	Correction factor as a function of the antenna diameter at 94 GHz. The legend shows the corresponding Fresnel and far field regions of each antenna diameter.	121
3.31	Power losses as a function of the range for different alignment error angles θ_s and ϕ_s , considering a separation between antennas Δ of 1 m and an antenna beamwidth θ_0 of 0.2° taking into account both parallax error and correction factor for Fresnel and far field region of antennas.	121
3.32	Power losses as a function of the separation between antennas Δ for different ranges. The alignment error angles θ_s and ϕ_s are 0.2° and 0° , respectively. The antenna beamwidth θ_0 is 0.2°	122
3.33	Sensitivity improvement with two-different antennas with beamwidths θ_{0_1} and θ_{0_2} as a function of the range for different alignment error angles θ_s and ϕ_s , considering a separation between antennas Δ of 1 m.	122
4.1	Photograph of the selected antennas on the radar mechanical structure.	127
4.2	SNR as a function of the reflectivity factor using the operational parameters of the UPM cloud profiling radar.	129

4.3	Block diagram of the radar system at 94 GHz. It is made up of four subsystems: the DDS-based signal generation source, the microwave subsystem, the millimeter-wave subsystem, and the data acquisition subsystem. Test points are defined with circle markers and the swept bandwidth of the chirp signals at each test point is indicated in parenthesis. The measured output power at the operating frequency of the radar system is shown in red.	132
4.4	Photograph of the microwave subsystem enclosure (a), the millimeter-wave transmitter (b) and the millimeter-wave receiver (c).	133
4.5	Power spectrum measurement at the DDS output (test point A).	134
4.6	Power spectrum measurement at the microwave subsystem output (test point B).	135
4.7	Photograph of the doubler-MPA MMICs (a) and the doubler module housing with the automatic biasing sequencer (b). This circuit is highlighted with a dashed line box in Fig. 4.3.	136
4.8	Doubler-MPA module: Output power measurement as a function of its input power.	136
4.9	Doubler-MPA module: S_{11} -and- S_{22} parameters measurement.	137
4.10	Farran HPA amplifier: Output power measurement.	138
4.11	Power measurements at the output of different components of the radar architecture.	139
4.12	Phase noise measurement at the output of different components of the radar architecture.	139
4.13	RF non-linearities measurement at the microwave subsystem output with the cloud-profiling-radar operational parameters.	140
4.14	Noise power floor of the receiver chain after non-coherent integration of 2.5 seconds of IF signal with a PRF of 6 kHz. The ADC and the receiver chain noise power floor are measured with a 50- Ω load at their inputs in both cases. Thirdly, the noise power floor is measured in clear-sky conditions, with receiver and transmitter, including both antennas, operating simultaneously. . .	141
4.15	Noise distribution histogram of the noise power floor in clear-sky conditions. .	142
4.16	Receiver sensitivity in dBZ considering SNR=0 dB of the developed cloud radar as a function of the transmitted bandwidth after non-coherent integration of 2.5 seconds of IF signal with a PRF of 6 kHz.	143
4.17	Sensitivity reduction at lower ranges due to transmitter leakage.	143
4.18	Photograph of the drone <i>DJI Phantom III Professional</i> used for checking the radar operation.	144
4.19	Range measurement comparison with the radar and the drone GPS measurement.	145
4.20	Four range profiles when drone is detected at a distance from the radar of 26, 83, 124 and 180 meters, and the receiver noise floor in order to compare the increase of the noise floor due to the transmitted phase noise.	145
4.21	Noise floor measurement as a function of the drone received power.	146
4.22	Noise floor comparison between the phase noise estimate and the radar measurement of the detected drone.	146
4.23	Doppler velocity measurement comparison with the radar and the drone GPS measurement.	147

4.24	Doppler-range matrix of the drone.	147
4.25	Signal processing procedure in order to obtain the 3D radar reflectivity in dBZ of the cloud profiling radar.	149
4.26	Block diagram of the double-threshold detection scheme.	149
4.27	Block diagram of the binary hypothesis test of the detection scheme.	150
4.28	Probability of false alarm as a function of detection threshold T for the radar-reflectivity signal processing for different values of integrated chirp periods: 1000, 2000, 3000, 5000 and 15000 chirps. The probability of false alarm obtained from the radar data (solid black line) and an estimate (dashed black line) obtained from the mean and variance of the analyzed radar data without spurious signals are also shown.	151
4.29	Histogram of the noise distribution of the radar data with and without spurious after non-coherent integration.	151
4.30	Noise floor comparison between the phase noise estimate and the radar measurement of the detected cloud when there was drizzle.	152
4.31	Measured MD on range profiles of cloud measurement results that were carried out on January 29th, 2016. A cubic polynomial fitting of measured MD is also plotted in order to define a MD as a function of the maximum received power of the range profile.	153
4.32	Three range profiles of detected clouds. Black and blue curves are not affected by phase noise (there was no drizzle). The noise floor of the range profile represented in red is affected by phase noise (there was drizzle), and a dynamic range of 18 dB is applied that is shown with green crosses.	154
4.33	3D reflectivity profile with (top) and without (bottom) second threshold of cloud measurements obtained on January 29th, 2016. The range profiles where the second threshold is applied (and, therefore, there was drizzle) are highlighted in grey.	154
4.34	dBm-to-dBZ conversion of range profiles from detected clouds on January 29th, 2016 after 5 minutes from the initial time.	155
4.35	3D reflectivity profile in dBm (a) and in dBZ (b).	155
4.36	Signal processing procedure in order to obtain the Doppler profile of the cloud profiling radar.	156
4.37	Doppler spectrum as a function of different shifts of the STFT.	157
4.38	Doppler spectrum as a function of different overlaps of the STFT.	158
4.39	Doppler spectrum as a function of different sliding-window lengths of the FFT.	158
4.40	Probability of false alarm as a function of scaling factor K for the Doppler-velocity signal processing.	160
4.41	Parameter extraction from the Doppler spectrum of a detected cloud at a given range of 2 km. Green crosses denote the detections that pass the detection threshold (red line).	161
4.42	3D mean Doppler velocity profile.	161
4.43	Developed cloud radar for continuous measurements in outdoor conditions.	162
4.44	Required FFT points as a function of the transmitted bandwidth.	164
4.45	Captured data size as a function of the acquisition time and the sampling frequency considering a radar PRF of 6 kHz.	165
4.46	Sky condition at the initial time on January 29th, 2016.	166

4.47	Mixed-phase cloud measurement carried out on January 29th, 2016.	167
4.48	Doppler spectrum for four different heights: 540, 1450, 2915 and 3050 meters after 70 minutes from the initial time (there was no drizzle) on January 29th, 2016.	168
4.49	Doppler spectrum for four different heights: 540, 1450, 2915 and 3050 meters after 30 minutes from the initial time (there was drizzle) on January 29th, 2016.	169
4.50	Doppler spectrograms at different moments of the measurement: 20 and 70 minutes from the initial time.	170
4.51	Examples of multimodal Doppler spectra at six different heights: 1187, 1337, 1472, 1863, 2118 and 2404 meters after 80 minutes from the initial time.	171
4.52	Peak-picking results on multimodal Doppler spectra at six different heights: 1187, 1337, 1472, 1863, 2118 and 2404 meters after 80 minutes from the initial time.	172
4.53	Fog measurement carried out on February 5th, 2016.	174
4.54	Range profiles after 19 minutes and 36 minutes from the initial time of the cloud measurement carried out on February 18th, 2016.	175
4.55	Mixed-phase cloud measurement carried out on February 18th, 2016.	176
4.56	Detected cumulus and stratus on May 12th, 2016.	177
4.57	Stratus and cumulus clouds measurement carried out on May 12th, 2016.	178
4.58	Doppler spectrum of cumulus 80 minutes (no drizzle) after the initial time for four different heights: 5105, 5330, 5808 and 6012 meters on May 12th, 2016. The mean Doppler velocities correspond to 1.29, 1.29, 1.2 and 0.9 m/s, respectively.	179
4.59	Doppler spectrum of stratus 4 hours (no drizzle) after the initial time for three different heights: 2133, 3426 and 5470 meters on May 12th, 2016. The mean Doppler velocities correspond to -0.84, +0.74 and +1.15 m/s, respectively.	179
B.1	x3 frequency multiplier: S_{11} -and- S_{22} parameters measurement.	226
B.2	x3 frequency multiplier: Output power measurement as a function of the doubler input power.	226
B.3	Quinstar balanced mixer measurements: RF-IF conversion losses, return losses at RF and LO ports and the RF-LO isolation.	227
B.4	RPG low-noise amplifier: S-parameters measurement.	228
B.5	Farran MPA amplifier: S-parameters measurement.	228
B.6	Farran MPA amplifier: Output power measurement as a function of the doubler input power.	229
B.7	RPG x6 frequency multiplier: Output power measurement as a function of its input power.	230
B.8	RPG x6 frequency multiplier: S_{11} -parameter measurement.	230
B.9	Farran HPA amplifier: S-parameters measurement.	231
B.10	Farran HPA amplifier: Output power, current consumption and PAE measurement results as a function of the HPA input power.	232
C.1	Frequency range for some filter technologies as a function of the fractional bandwidth and center frequency. Graph extracted from [Mur11].	234

C.2	Schematic of the active bandpass filter of the imaging radar at 300 GHz (filter I).	236
C.3	Schematic of the active bandpass filter of the imaging radar at 300 GHz (filter I) taking into account the S-parameters of the inductors used.	237
C.4	S-parameter simulation results of filter I taking into account the ideal (dashed line) and real (solid line) behavior of Murata inductors.	238
C.5	S-parameter measurement results of filter I.	238
C.6	Power compression measurement results of filter I.	239
C.7	Photograph of the active bandpass filter of the imaging radar at 300 GHz (filter I).	239
C.8	Schematic of the active bandpass filter of the cloud profiling radar at 94 GHz (filter II).	241
C.9	Schematic of the active bandpass filter of the cloud profiling radar at 94 GHz (filter II) taking into account the S-parameters of the used inductors.	242
C.10	S-parameter simulation results of filter II taking into account the ideal (dashed line) and real (solid line) behavior of Murata inductors.	243
C.11	S-parameter measurement results of filter II.	243
C.12	Power compression measurement results of filter II.	244
C.13	Photograph of the active bandpass filter of the cloud profiling radar at 94 GHz (filter II).	244
D.1	Block diagram of the 100-GHz receiver.	245
D.2	Schematic of the receiver chain for ADS budget analysis of noise power.	247
D.3	Results obtained by an ADS noise budget analysis of the receiver noise power levels at the output of each component.	247
E.1	Photographs of the automatic biasing sequencer.	249
E.2	Schematic of the automatic biasing sequencer.	251
E.3	Schematic of a voltage sequencer using a RC-circuit and two power MOSFETs.	252
E.4	Output voltage at the RC-circuit and gate outputs of the automatic biasing sequencer.	253
E.5	Output voltages of the automatic biasing sequencer.	253
E.6	Noise voltage at the drain and gate outputs of the automatic biasing sequencer.	254
E.7	Output voltage measurements at the drain outputs of the automatic biasing sequencer.	255
F.1	Details of the drainage systems and fixing screws for the radome on the feeder structure.	258
F.2	Photograph of the flat radome.	259
F.3	Photograph of the inclined radome.	259
F.4	Photograph of the antenna feeder under test in the anechoic chamber.	260
F.5	Measured co-polar and cross-polar radiation pattern of the feeder without radome.	261
F.6	Measured co-polar and cross-polar radiation pattern of the feeder with flat radome.	261

F.7	Measured co-polar and cross-polar radiation pattern of the feeder with inclined radome.	262
F.8	Power losses of the developed radomes measured in the co-polar radiation pattern of the feeder.	262
F.9	S_{11} results comparison of the Cassegrain antenna feeder with the developed radomes.	263
F.10	Photograph of the Cassegrain antenna under test in the anechoic chamber. . .	264
F.11	Measured co-polar and cross-polar radiation pattern of the Cassegrain antenna without radome.	265
F.12	Measured co-polar and cross-polar radiation pattern of the Cassegrain antenna with flat radome.	265
F.13	Measured co-polar and cross-polar radiation pattern of the Cassegrain antenna with inclined radome.	266
F.14	Power losses of the developed radomes measured in the co-polar radiation pattern of the Cassegrain antenna.	266
F.15	S_{11} results comparison of the Cassegrain antenna with the developed radomes.	267
G.1	Sky condition on January 29th, 2016 at 10:21 AM.	269
G.2	Sky condition on February 5th, 2016 at 15:19 PM.	270
G.3	Sky condition on February 18th, 2016 at 10:39 AM.	270
G.4	Sky condition on May 12th, 2016 at 12:34 PM.	271

List of Tables

1.1	Cloud-base height type above ground level in temperate regions. Data extracted from [Guid08].	9
1.2	Optical properties of 5-mm thick materials. Data extracted from [App07]. . .	12
2.1	Overview of imaging radars above 100 GHz. The main parameters are the center frequency (f_c), the transmitted bandwidth (BW), the chirp period (T_m), the fractional bandwidth (FBW) and the chirp rate (BW/T_m).	29
2.2	Overview of commercial broadband and narrowband DDS. This table contains the DDS bandwidth (BW), the number of bits of DAC (b), the number of bits of phase accumulator (n), the frequency resolution (Δf), the minimum frequency update time (Δt), the spurious-free dynamic range (SFDR), the power consumption and the evaluation board cost.	31
2.3	Summary of the PLL parameters of the narrowband DDS/PLL architecture.	35
2.4	Chirp source cost and power consumption.	35
2.5	Summary of the results for the developed chirp sources.	50
2.6	ITU-R P.453 [ITU 15] coefficients for liquid water particles.	61
2.7	Summary of phase drift results after warm-up at the different points of the radar electronics.	63
2.8	Variance results of the measured phase in the short term.	63
2.9	Comparison results of the radar performance at 100 GHz with both polarizations and 300 GHz.	79
3.1	Interpretation of the dBZ value of cloud types extracted from [Kol07].	92
3.2	Comparison of the operational parameters of remarkable cloud profiling radars.	104
4.1	Commercial antennas at W-band for cloud profiling radars.	126
4.2	Research and commercial power amplifiers at millimeter-wave bands for the UPM cloud profiling radar at 94 GHz. The amplifiers parameters such as central frequency, bandwidth, output power and gain are shown in conjunction with the employed technology and the number of levels of power combination (if applicable).	128
4.3	Operational parameters of the UPM cloud profiling radar at 94 GHz.	130
4.4	Radar sensitivity of the UPM cloud profiling radar at 94 GHz for different reflectivity factors considering SNR=0 dB.	130
4.5	Power consumption of the cloud-profiling radar system.	133

4.6	Spurious signals of the receiver noise power floor and its power level difference regarding the noise floor (-94.2 dBm).	141
4.7	Noise distribution parameters of the noise power floor in clear-sky conditions.	142
4.8	Results of comparison between STFT overlap and shift and FFT sliding window.	159
4.9	Statistical parameters of Doppler spectra of mixed-phase cloud observation on January 29th, 2016.	169
4.10	Statistical parameters of Doppler spectrums of cumuliform-and-stratiform clouds measurement carried out on May 12th, 2016.	177
C.1	Comparison of simulation and measurement results of the active filters developed for the imaging radar at 300 GHz (I) and the cloud profiling radar at 94 GHz (II).	240
D.1	Receiver parameters at 94 GHz.	246
F.1	Oratex 600 specifications.	258

Acronyms

3D	Three Dimensional
ADC	Analog-to-Digital Converter
ADS	Advanced Design System software
AEMET	Agencia Española de Meteorología (<i>Spanish National Weather Agency</i>)
ALMA	Atacama Large Millimeter/submillimeter Array
ARM	Atmospheric Radiation Measurement Program
ATD	Automatic Target Detection
ATR	Automatic Target Recognition
BiCMOS	Bipolar Complementary Metal-Oxide-Semiconductor
BJT	Bipolar Junction Transistor
BW	Bandwidth
CA-CFAR	Cell Averaging Constant False Alarm Rate
CARMA	Combined Array for Research in Millimeter-wave Astronomy
CATR	Compact Antenna Test Range facilities
CES	Committee on Earth Sciences of the US Global Change Research Program
COTS	Commercial-Off-The-Shelf Components
COST	European Cooperation in Science and Technology
CPI	Communications & Power Industries, Inc.
CPR	Cloud Profiling Radar
CSA	Canadian Space Agency
CW-LFM	Continuous Wave Linear Frequency Modulation
DAC	Digital-to-Analog Converter
DARPA	Defense Advanced Research Projects Agency
DDS	Direct Digital Synthesis
DHBT	Double Heterojunction Bipolar Transistor
DHEMT	Dual-channel High-Electron-Mobility Transistor
DJI	Dà-Jiāng Innovations Science and Technology Co., Ltd.
DoE	The United States Department of Energy
DRIE	Deep Active Reactive Ion Etching Process
DSD	Drop Size Distribution

EarthCARE	Earth Clouds, Aerosols and Radiation Explorer mission
EIK	Extended Interaction Klystron
ESA-JAXA	European Space Agency-Japan Aerospace Exploration Agency
ETSIT	Escuela Técnica Superior de Ingenieros de Telecomunicación (<i>Telecommunications Engineering School</i>)
EUMETSAT	European Organization for the Exploitation of Meteorological Satellites
EW	Electronic Warfare
FCC	US Federal Communications Commission
FFT	Fast Fourier Transform
FHR	Fraunhofer Institute for High Frequency Physics and Radar Techniques
FOD	Foreign Object Debris
GaAs	Gallium Arsenide
GaN	Gallium Nitride
GMR	Microwaves and Radar Group
GPS	Global Positioning System
HDMI	High-Definition Multimedia Interface
HEMT	High-Electron-Mobility Transistor
HEXFET	Hexagonal Field Effect Transistor
HFET	Heterostructure Field Effect Transistor
HPA	High-Power Amplifier
HPBW	Half-Power Beamwidth
IDR	Instantaneous Dynamic Range
IATA	International Air Transport Association
IED	Improvised Explosive Device
IF	Intermediate Frequency
IMPATT	Impact ionization Transit-Time
InP	Indium Phosphide
ISAR	Inverse Synthetic Aperture Radar
ITU-R	International Telecommunication Union - Radiocommunication sector
JPL	Jet Propulsion Laboratory
LDR	Linear Depolarization Ratio
LEHA	Laboratorio de Ensayos y Homologación de Antenas (<i>UPM Antenna Test Facilities</i>)
LHCP	Left-Hand Circular Polarization
LNA	Low-Noise Amplifier
LO	Local Oscillator
LWC	Liquid Water Content
MetOp-SG	Meteorological Operational - Second Generation
mHEMT	Metamorphic High-Electron-Mobility Transistor
MIMO	Multiple-Input Multiple-Output

MMIC	Monolithic Microwave Integrated Circuit
MOSFET	Metal-Oxide-Semiconductor Field-Effect Transistor
MPA	Medium-Power Amplifier
NASA	National Aeronautics and Space Administration
NDI	Nondestructive Inspection
NDT	Nondestructive Testing
NOAA	National Oceanic and Atmospheric Administration
NWC SAF	EUMETSAT Nowcasting Satellite Application Facility
OCXO	Oven Controlled Crystal Oscillator
OMT	Orthomode Transducer
PBIED	Person-Borne Improvised Explosive Device
PAE	Power-Added Efficiency
PC	Personal Computer
PLL	Phase-Locked Loop
PLO	Phase-Locked Oscillator
PO	Physics Optics
PRF	Pulse Repetition Frequency
QCL	Quantum Cascade Laser
RAL	Rutherford Appleton Laboratory
RAM	Random Access Memory
RH	Relative Humidity
RHCP	Right-Hand Circular Polarization
RPG	Radiometer Physics GmbH
RTD	Resonant Tunnel Diode
SAR	Synthetic Aperture Radar
SFDR	Spurious Free Dynamic Range
SHM	Subharmonic Mixer
SiGe	Silicon Germanium
SLL	Sidelobe Level
SMT	Surface-Mount Technology
SNR	Signal to Noise Ratio
STFT	Short-Time Fourier Transform
SVC	Saturation Vapor Concentration
TUNNET	Tunnel Injection Transit Time Diode
TWT	Traveling-Wave Tube
UPM	Universidad Politécnica de Madrid (<i>Technical University of Madrid</i>)
USB	Universal Serial Bus
UTC-PD	Uni-Traveling-Carrier PhotoDiode
UWB	Ultra Wideband

VCO	Voltage-Controlled Oscillator
VDI	Virginia Diodes, Inc.
WMO	World Meteorological Organization
WPAN	Wireless Personal Area Network
WVC	Water Vapor Concentration
XPD	Cross-Polar Discrimination
YIG	Yttrium Iron Garnet

Chapter 1

Introduction

1.1 Millimeter-wave spectrum and applications

Millimeter waves occupy the frequency spectrum from 30 GHz to 300 GHz. They are found in the electromagnetic spectrum between microwaves (1 to 30 GHz) and submillimeter waves (300 to 3.000 GHz), as shown in Fig. 1.1, which includes the millimeter-wave sub-bands designation. The wavelength is in the 1 to 10 mm range, hence their name millimeter waves*. A general overview of the different applications, technologies and physical phenomena related to the millimeter-wave band will be presented below.

Figure 1.1: The electromagnetic spectrum and the millimeter-wave sub-bands designation.

Millimeter-wave applications can be broadly divided into five main categories, as shown in Fig. 1.2. The interest in the millimeter-wave band has grown out of the radioastronomy community [Cro05, Sie07, Cha11, Dhi17]. Although currently there are many sectors involved in the millimeter-wave band research for biological issues, remote sensing or security applications, the main drivers are the automotive industry and the telecom sector, which is strongly devoted to test 5G wireless broadband communications technology at this band [Sie02, Rap13, Co014]. The rapid growth in the interest of millimeter-wave systems in security applications is due to the advantage of the detection of concealed objects under clothes or inside luggage. Besides millimeter waves are non ionizing and therefore they do not raise safety concerns [She09].

*In order to be rigorous, as the terahertz band is loosely defined between 100 GHz to 10 THz [Sie02], and the radar systems presented in this thesis work at an operating frequency lower than 300 GHz, this Introduction will be focused on the millimeter-wave band properties and related applications.

Astronomy	Remote sensing	Security	Biological issues	Communications
<ul style="list-style-type: none"> - Interstellar medium. - Spectral line emissions from CO in galaxies. - Gas kinematics in stars and planetary disks. - Comets observation. - Extraterrestrial life. 	<ul style="list-style-type: none"> - Phenomenology of propagation through foliage and forest. - Cloud physics investigation. - Atmospheric pollution. - Icing hazards. - Terrestrial gases. 	<ul style="list-style-type: none"> - Standoff detection of concealed threats. - Assisted landing of aircrafts. - Foreign object debris. - Civil structures deformation. - Automotive (Collision avoidance) 	<ul style="list-style-type: none"> - Tissue identification. - Biomolecular sensing. - Dosimetry and spectroscopy through cultured cells. - Non-contact food monitoring for industrial production quality. - Vital-sign detection. 	<ul style="list-style-type: none"> - Backhaul point-to-point communications. - WirelessHD. - ECM387. - 802.15.3C.

Figure 1.2: Overview of millimeter-wave band applications.

The first use of millimeter-wave signals was carried out in 1895 by J.C. Bose, who achieved communications signal transmission at 60 GHz to a distance of a mile [Eme97]. The early applications of the millimeter-wave technology were related to radioastronomy in 1960, when the first systems were developed [Wal99]. Later, in the 80's, the development of millimeter-wave integrated circuits by the United States Government research programs created opportunities for mass manufacturing of millimeter wave products for commercial applications, like automotive, imaging and weather radars, radiometers, or satellite communications [Adh08]. In 1995, the FCC (US Federal Communications Commission) opened the spectrum between 59 GHz and 64 GHz for unlicensed wireless communications. Broadband communications are interesting in applications such as high speed data transmission and video distribution [Tes01].

Currently, the technology in millimeter-wave band is considerable mature for exploring different applications across industries. It is becoming increasingly important both in civil and military applications. The millimeter-wave technology market was valued at \$346.8 Millions in 2015 and is expected to grow at a rate of 43% in 2022 [Mar15a], mainly in Europe and North America. Its growth is expected to be driven by applications in the telecom sector. Regarding military applications, the industry is looking to millimeter-wave technology to increase connectivity and sensing for the next generation of tactical networks, security, electronic warfare (EW), and even active denial systems, besides missile guidance, navigation guidance, surveillance systems for air defense, and location tracking for snipers [Bai93, Woo07, Zhu09]. A summary of the current and emerging applications in the civil and military fields is given below.

- **Astronomy.** The millimeter-wave band contains spectral and spatial information on the cosmic background radiation, on very distant newly formed galaxies, and on the early stages of star formation in gas clouds within our own galaxy [Cha99]. Two examples of current instruments are the Combined Array for Research in Millimeter-wave Astronomy (CARMA) and the Atacama Large Millimeter/submillimeter Array (ALMA). CARMA is a university-based interferometer encompassing 23 antennas used in combination to image the universe at millimeter wavelengths. ALMA is an astronomical interferometer of radio telescopes in the Atacama desert of northern Chile. The CARMA and ALMA receivers operate at these frequency bands: 27 GHz-270 GHz and 31 GHz-1565 GHz, respectively.
- **Weather monitoring,** such as cloud physics investigation, icing hazards and atmospheric pollution [Kro96, Ton07, Kol07, Bat14], mainly at 35 GHz, 94 GHz, 140 GHz and 220 GHz. Currently, the Meteorological Operational Second Generation (MetOp

SG) satellites are under development, and will begin to replace the current MetOp in 2019-2020 [Tho14, Tho16, Rsc16]. These satellites integrate different instruments such as microwave imagers, ice cloud imagers and microwave sounders. They cover the frequency bands of 23-230 GHz, 18-183 GHz and 183-664 GHz, respectively.

- Dosimetry and spectroscopy issues through cultured cells and isolated organs to animals and humans for medical and biological studies [Pak98, Sas14, Tay15] at 25 GHz, 60 GHz, 75 GHz, 94 GHz and 100 GHz.
- Non-contact food monitoring for industrial product and quality [Ess11a]. A remarkable system at millimeter waves for this application has been developed by Fraunhofer Institute for High Frequency Physics and Radar Techniques (FHR) at 78 GHz. The scanned area is 30x30 cm² and the acquisition time is 1 minute.
- Non-destructive testing (NDT) of civil structures with micrometer precision of deformation [Oka08, Abr11], in the whole W-band (75-110 GHz). These systems require a high phase stability of the transmitter in order to offer an acceptable precision.
- Vital-sign detection, such as heart-beat monitoring and cardio-respiratory control [Bak11, Die12, Kao13, Lin14], mainly at 60 GHz and 94 GHz.
- Standoff detection of concealed threats under clothing in civil and military security points based on active and passive systems [She09, Mar12, Rob12, Cool14, Cool15, Gra15, Rob16b, Gra17] at 60 GHz, 300 GHz, 340 GHz and 670 GHz.
- Foreign object debris (FOD) of airport runways [Ess11b, Meh12] or space manmade objects in orbit around Earth [Bat03, Cze14] at 36 GHz, 94 GHz and 220 GHz.
- Short distance wireless communications, such as back-haul for wireless base stations and wireless personal area network (WPAN) at 60 GHz, at gigabit rates, allowing the wireless transmission of popular video interfaces such as high-definition multimedia interface (HDMI) 1.3 or DisplayPort 1.2 [Tes01].
- Characterization and modeling of phenomenology of wave propagation through foliage and forest ground reflectivity for vegetation patterns studies [Ham96, Nas04, Rap15] at 35 GHz and 73 GHz.
- Automotive applications, such as collision-avoidance systems and intelligent-vehicle highway systems. Designers are also working with greater resolution at 77 GHz, 79 GHz and 94 GHz for cars that will be equipped with radar-based systems which leads to high volume production with low cost potential [Wal03, Hoa06, Has12].
- Assisted landing of aircrafts, such as helicopters [Ran07, Hag08, Pag09], mainly at 94 GHz and 95 GHz.
- Remote chemical detection of terrestrial gases [Gop01, Gop08] and nuclear radiation [Gop09] of industrial emissions, mainly at 146 GHz, 154 GHz, 225 GHz and 315 GHz.

1.2 Radar-based applications

This Ph.D. thesis is focused on radar-based applications such as security and weather monitoring. The advantages of active-radar-based systems instead of other active or passive systems are well known by the scientific community [Sko08, Ric10]. Generally speaking, active radar systems provide superior penetration capability through any type of weather condition compared to lasers, and can be used in the day or night time, unlike passive cameras. Moreover, active radar systems are capable of affording the main parameters of detected targets: reflectivity, range and velocity. Regarding radar disadvantages, it is also well known that they are unable to detect objects under metallic enclosures and under water.

Two applications have been considered in this thesis where the use of millimeter-wave radar technology has been considered appropriate:

1. Standoff detection imaging. These systems establish strict criteria in terms of fast and broadband signal generation, since it requires high transmitted bandwidths (around 27-30 GHz [Coo11, Men13b]) to achieve a good range resolution (around 1 cm), in short periods of time in order to reduce the image acquisition time of the pixel-by-pixel scanning. The scanned human is located a few meters away, so high transmitted power is not required (powers below 1 milliwatts are enough [Coo12, Gra15, Gra17]).
2. Weather applications, particularly cloud monitoring. In this application, the main requirement is the transmitted power in order to maximize the detection range and observe weak meteorological phenomena at long range. The required transmitted power is around one watt to detect clouds at 15-20 km [Kol07]. Narrowband signal generation (in the order of a 3-15 MHz) is enough since the required range resolution is between 10-50 meters.

A comprehensive introduction and current trends of these two applications are described below.

1.2.1 Standoff imaging systems

Securing air traffic has become very challenging nowadays due to unpredictable threats, such as Person-Borne Improvised Explosive Devices (PBIED) [Int14, IAT15]. Additionally, recent incidents of terrorist activity have led to increased demand for standoff imaging technologies capable to detect threats [Yin11] in critical infrastructures (airports, secure building entrances, meeting venues...).

Currently deployed security technologies in airports include passive and active millimeter-wave systems, X-ray back scatter and X-ray transmission [Sin03, Roe15]. Efforts are already under way on the integration of the available imaging technologies and automatic target recognition (ATR) software algorithms to screen for a wider variety of threats in the form of liquids, gels, plastics, ..., as proposed by the International Air Transport Association (IATA) in their executive summary *Checkpoint of the future* [Inte11]. In addition, the operational requirements of standoff imaging are aimed at continuing to improve the operational efficiency in terms of increase throughput and optimize the asset utilization.

Body scanners were implanted in airports some years ago to achieve detection of body-borne threats composed of metallic and nonmetallic objects, because plastic explosives and ceramic weapons are not detected by traditional metal detectors. X-ray security scanners are still used in airports. However, X-rays are an ionizing radiation, and, therefore, they raise safety concerns [Com96]. Two types of X-ray security scanners have currently been developed for airport security use: X-ray units using backscattered X-rays and transmission X-rays [Eur12]. Backscatter radiation is the radiation reflected (scattered) from a material

back towards the X-ray radiation source. Transmission radiation is the radiation that passes directly through the person being examined. Furthermore, X-ray systems are portal-based system, therefore they require to be in close proximity to the target, which increases the system vulnerability.

On the other hand, the specific properties of electromagnetic waves at millimeter-wave band allow the detection of concealed objects under clothes or inside luggage [App07], with non-ionizing radiation and maintaining safety standards [Mou12]. As indicated before, there are two inherent imaging limitations:

- It is not possible to detect objects under a metallic container.
- Objects inside the body are not detected due to the strong attenuation of water at the millimeter-wave band.

Two kinds of millimeter-wave imaging systems have been proposed: Active and passive systems [Yuj03, Pet08]. Passive systems take advantage of the contrast in temperature between the background, e.g. the human skin, and the target (explosives) [Gro10], produced by their emissivity and reflectivity differences. Active systems include an illumination source to improve the contrast of the image [She09]. The image quality of these non-radar systems is highly degraded by the specular reflection effect, and, therefore, the current trend is towards substituting them for radar systems to take advantage of their high-range resolution capability to form high-contrast 3-D images [Coo08a, Rob12, Gra15]. These systems are non-portal based, i.e. standoff detection without the targets being aware of their scanning. These systems are commonly based on monostatic radar front-ends in conjunction with a specific antenna subsystem that allow the standoff scanning of targets of interest while keeping safety standards. High-speed mechanical scanning is the selected strategy for systems such as [Coo08a, She09, Gra15], or based on galvanometers [Rob12] and two-axis gimbals [Rob16a, Rob16b]. Most of these systems are continuous-wave linear-frequency modulation (CW-LFM) radars with high transmitted bandwidth [Coo11, Rob12, Gra15, Rob16b].

Current efforts of the millimeter-wave imaging research community are being put into two directions, both maintaining the key goal of faster scanning for a safe real-time high-throughput screening:

- Development of compact transceiver technology [Abb11, Bry13, Cha13, Coo14, Cha14, Rec15] to build multi-pixel active imaging radars, where multiple focused beams would simultaneously scan over a fixed field of view. To this end, Bryllert *et al.* [Bry13] have developed a transceiver module for a 3D imaging radar at 220 GHz that consists of a frequency doubler that also acts as a subharmonic mixer (SHM) based on 3 μm -thick Gallium arsenide (GaAs) semiconductor membrane technology. In addition, in some cases, the horn antenna is included in the transceiver module to achieve a higher integration level. For instance, Reck *et al.* [Rec15] have developed an eight-pixel transceiver array for a 340 GHz imaging radar [Coo14] with an average transmitter power of 0.3-0.6 mW per pixel, based on silicon micro-machined technology by utilizing deep reactive ion etching (DRIE) process. Chattopadhyay *et al.* [Cha13, Cha14] are also developing multi-pixel instruments with integrated microlens antennas at terahertz frequencies using a DRIE process. Lastly, Abbasi *et al.* [Abb11] have developed an heterodyne transceiver at 220 GHz based on monolithic microwave integrated circuits (MMICs) with on-chip integrated lens antenna fabricated in 0.1 μm GaAs metamorphic high electron-mobility transistor technology (mHEMT).
- Development of multistatic imaging systems with portal-based configuration. The current trend of the state-of-the-art active radar millimeter-wave scanners is based on a portal-based configuration and monostatic or multistatic radar technology. For

instance, as proposed in [She01, Gon16a], two rotating sets of transmitters/receivers produce and receive the millimeter wave signals as the person being screened stands still inside the portal. In [Ahm13], an advanced multistatic millimeter-wave imaging configuration is presented. Multiple panels are used to create an image of the whole body in a portal configuration. A compromise between coverage aperture, antenna elements and computation cost should be made, besides keeping in mind the elevated costs of millimeter-wave components and its integration. Novel architectures have recently proposed in [Gon16a, Gon16b]. They do not include any mechanical movement, and images are obtained taking advantage of the movement of the person under test when passing through the system on a moving walkway.

Currently employed commercial millimeter-wave systems in airports work in a multistatic configuration. Automatic target detection (ATD) technology ensures passenger privacy by highlighting threats and anomalies using a generic mannequin that resembles a human outline. The millimeter-wave imaging systems of the three remarkable manufacturers are presented below: L-3 Communications Security and Detection Systems, Smiths Detection and Rohde&Schwarz.

1. L-3 Communications Security and Detection Systems. The most mature system is the Provision ATD system (see Fig. 1.3a), that is currently deployed in american and european airports. It uses two columns with linear arrays of millimeter-wave CW-LFM transmitters and receivers operating at 33 GHz [Pop11, L3 15]. The scan procedure requires only a single stationary position during a 1.5-second scan. The system does all the work for the scanned human and requires no special movements or motions and it processes up to 300 people per hour depending on the application.
2. Smiths Detection. The Ego radar-based system works at 24 GHz. The system has two primary components, a transceiver and a reflector panel that redirects and focuses the signal to the scanned subject, i.e. as a walk-through portal [Pop11, Roe15], as shown in Fig. 1.3b. The subject is scanned by programming the reflector to focus the beam and then the focused beam is scanned throughout the volume surrounding the subject. While this process is occurring, the subject is making a 360 degrees turn in the scan volume.

Figure 1.3: Portal-based millimeter-wave imaging systems from L-3 Communications Security and Detection Systems (a) and Smiths Detection (b).

3. Rohde & Schwarz. The QPS 100 system consists of a single panel with an aperture of approximately $2 \times 1 \text{ m}^2$ that operates between 70 to 80 GHz [Ahm15], as shown in Fig. 1.4a. It presents 32 clusters, where each cluster integrates 94 Tx and 94 Rx antennas. The automatic threat detection is in around 3 seconds. For the applications where an even faster performance is required, a triple panel arrangement can be utilized. Fig. 1.4b illustrates this solution, where two passenger lines can be served by one screening system. This arrangement would ensure a higher level of illumination on the screened person, thus increasing detection capability or alternatively helping reduce false alarm rates. The scanning hardware architecture is actually not restricted to be confined in panels. It is also conceivable to extend the raster arrangement to cover any needed flat geometry, e.g., walls or corridors. Fig. 1.4c illustrates a future vision to utilize this technology for a scan-while-walking tunnel. This would deliver the highest throughput possible with the best comfort for the scanned humans. The realization of such a fast and comprehensive system is still, however, constrained by the speed and complexity of the digital processing units to deal with the huge data amounts produced.

Figure 1.4: Three proposed solutions from Rohde & Schwarz for faster performance and to increase detection capability.

1.2.2 Cloud profiling radars

The contribution of clouds on the earth energy balance, and how it may change as a result of man-made perturbations, remains one of the greatest sources of uncertainties in climate simulations and predictions [Kol07, Fab15]. Furthermore, studies of the formation of clouds and how precipitation initiates remain an active research area. Clouds are immensely important factors in climate change research [Ste13, Bon15] by means of their effects on the transfer of radiant energy through the atmosphere, in addition to their importance as a vital link in the hydrological cycle, as recognized by the Committee on Earth Sciences (CES) of the United States Global Change Research Program [Bru10]. On the other hand, from a commercial perspective, two current needs are observed after consultation with weather industry and academia:

- Fog and clouds nowcasting in airports and sport race circuits. Conventional weather radars are not able to detect fog and non-precipitating clouds.
- Improvement of solar energy produced by terrestrial solar power stations, in order to better estimate the solar energy production and to plan what type of energy should be used in the short-term.

Clouds are an accumulation of very small water droplets, ice crystals, or a mixture of both, in the order of $200\text{ }\mu\text{m}$ of diameter [Guid08]. They are formed mainly as the result of the vertical motion of air, in convection, in forced ascent over high ground, or in the large-scale vertical motion associated with depressions and fronts. At temperatures below 0°C , cloud particles frequently consist entirely of water droplets supercooled down to about -10°C in the case of layer clouds and to about -25°C in the case of convective clouds. At temperatures below these limits and above about -40°C , many clouds are mixed, with ice crystals predominating in the lower part of the temperature range.

The atmosphere can be characterized by a number of key radial layers, as shown in Fig. 1.5. These layers can be characterized by the vertical distribution of temperature fluctuation, as well as the water vapor in the layer [Ric10].

Figure 1.5: Atmospheric Regions. Graph extracted from [Ric10].

A cloud classification was proposed in [Guid08] in terms of the level-high, middle or low - at which the various cloud types are usually encountered, as shown in Table 1.1. In close regions, the approximate limits are: High, 6-12 km; middle, surface-6 km; and low, surface-1.5 km. The high clouds are cirrus, cirrocumulus and cirrostratus; the middle clouds are altocumulus and altostratus (the latter often extending higher) and nimbostratus (usually extending both higher and lower); and the low clouds are stratocumulus, stratus, cumulus and cumulonimbus.

Cloud dynamics study has become a topic of great interest in the last 20-30 years and cloud radars at millimeter band have emerged as an important tool in characterizing their properties [Lhe87, Clo95, Mor98, Clo99, Kol07]. Cloud radars are now being used in conjunction with lidar systems, one of the principal sensors from the 1980s in the study of clouds. Whereas the horizontal distribution of clouds over the earth is characterized currently by passive radiometry from satellites [Gra02, Gas13, Pal14], mapping the vertical distribution of clouds is achieved best using lidars and radars. Lidars are able to detect all types of cloud particles up to 10-15 km in circumstances in which the lidar beam is able to penetrate to the location of the cloud particles. However, complete extinction of a lidar beam during heavy overcast, or a period with multiple cloud layers, is common [Ste03, Kra05].

On the other hand, millimeter-wave radar observations are a natural complement to the popular and widely used centimeter-wavelength radars, because the radar cross section is proportional to the fourth power of the operating frequency, therefore improving

Table 1.1: Cloud-base height type above ground level in temperate regions. Data extracted from [Guid08].

Cloud region	Cloud type	Height from the surface (m)
Low	Stratus	0- 600
	Stratocumulus	300- 1.350
	Cumulus	300- 1.500
	Cumulonimbus	600- 1.500
Middle	Nimbostratus	0- 3.000
	Altostratus	0- 600
	Alto cumulus	0- 600
High	Cirrus	6.000- 12.000
	Cirrostratus	6.000- 12.000
	Cirrocumulus	6.000- 12.000

their sensitivity in the millimeter-wave region [Kol07]. Millimeter-wave radar beam generally penetrates all cloud types except those that occur during periods of heavy rain. In precipitation-free conditions, a range of tens of kilometers is possible with millimeter-wave cloud radars, but with the onset of precipitation the useful observation range decreases [Gos97, Kol07]. Radars operating at 35 GHz and 94 GHz are used the most since the attenuation of the beam as a result of absorption by oxygen and water vapor is at a local minimum near these frequencies. Currently, some cloud radars at G band (140 GHz to 220 GHz) in combination with lower frequencies are under development to further improve the retrievals of microphysical properties [Bat14]. Millimeter-wave cloud profiling radars operate in a vertically pointing mode, have narrow beams ($< 1^\circ$) and low range resolution (50-100 m). Some comparison between ground-based and satellite-based cloud radars have been carried out [Kok09, Mer15]. It should be noted that ground-based radar provides an inexpensive alternative to in situ cloud measurements, enabling the collection of large datasets at high temporal and spatial resolutions instead of airborne or satellite-based radars [Im90, Li94, Ste03].

Regarding commercial interest, many cloud radars have appeared on the market in recent years, both in Europe and the United States. Two main commercial cloud radars can be highlighted from Radiometer Physics GmbH (RPG) [Rad14a] and Prosensing [Wid12], respectively. A photograph of both systems is shown in Fig. 1.6.

1. Prosensing dual-frequency (35 and 94 GHz) dual-polarized cloud radar [Wid12]. Besides the dual-frequency and polarimetry techniques, an elevation-over-azimuth positioner provides scanning capabilities in order to increase the scanned area. Five units of this system are used by the Atmospheric Radiation Measurement (ARM) Climate Research Facility of the United States Department of Energy (DoE).
2. Radiometer Physics cloud radar at 94 GHz [Rad14a]. It has an extra option to scan in elevation and azimuth. Moreover, it includes an 89 GHz passive radiometer to measure the liquid water profile of detected particles and an automatic weather station to measure the temperature, barometric pressure, rain rate,...

(a) Prosensing

(b) RPG

Figure 1.6: Photograph of two commercial cloud profiling radar from Prosensing [Wid12] (a) and Radiometer Physics GmbH [Rad14a].

1.3 Millimeter-wave spectrum properties

Two main characteristics of millimeter waves are crucial in the design and development of systems at this band, particularly to the radar-based applications related to this Ph.D. thesis: Atmospheric propagation and material properties. A brief description is given below.

1.3.1 Atmospheric propagation

Atmospheric attenuation due to the presence of oxygen and water vapor is the limiting factor of millimeter-wave systems for remote sensing and communications [Bat73, Mill97, ITU 13]. Much research work has been developed in the last decades in order to improve the attenuation-estimation techniques [Bru95, Ayd00, Gra16]. As shown in Fig. 1.7a, there are regions of the spectrum in these bands that are attenuated by as much as 70 dB/km, which makes operation at ranges of even a few kilometers challenging. The minima in Fig.1.7a show the atmospheric windows that are used to perform some interesting applications with an acceptable atmospheric attenuation. In the millimeter-wave region they are typically 26 to 40 GHz, 70 to 110 GHz, 130 to 150 GHz, and 200 to 340 GHz. Furthermore, two weather phenomena can dominate the power transmission losses: Rain and clouds, as shown in Fig. 1.7b as a function of the operating frequency.

In the case of rain, the power loss occurs both by absorption and by scattering, and it depends on the raindrop size [ITU 03]. Commonly, the rain attenuation is represented as a function of the rain rate (in mm/h), as shown in Fig. 1.8a. For clouds, where the droplets are smaller than the wavelength, and, therefore, the requirements of the Rayleigh scattering theory are satisfied [ITU 09, Ric10], the attenuation can be expressed as a function of the water content per unit volume, as shown in Fig. 1.8b.

(a) Atmospheric windows at millimeter-wave band. Graph extracted from [Ric10].

(b) Zenith transmissivity of the clear atmosphere and of rain and clouds as a function of the operating frequency. Graph extracted from [Fab15].

Figure 1.7: Atmospheric phenomenology at millimeter-wave band.

(a) Rain.

(b) Clouds.

Figure 1.8: Attenuation (one way) in dB/km due to rain [ITU 03] (a) and clouds [ITU 09] (b) as a function of the rain rate and water vapor concentration, respectively.

1.3.2 Material properties

The interest of the use of millimeter-wave band in imaging applications for enabling detection of weapons and improvised explosive devices (IEDs) is related to its capability to penetrate non-metallic materials and the non-ionizing characteristic of its radiation. Each material presents three properties that characterize its return signal: Emissivity e , reflectivity r , and transmittance t . The emissivity e defines the material by its thermal radiation capabilities, the reflectivity r is defined by the ability of a material to reflect a part of the incident energy, and the transmittance t is the portion of the incident power that is transmitted through the objects. These parameters vary from 0 to 1 and depend on many factors (angle of illumination, polarization, shape, physical properties,...) [Kap14]. Furthermore, the law of conservation of energy requires that

$$e + r + t = 1 \quad (1.1)$$

At thermal equilibrium, the emissivity e of a body equals its absorptivity α , as states the Kirchhoff law of thermal radiation. Security systems at millimeter-wave band are designed to detect objects and have a limited capability to identify materials, as happens in spectroscopy with explosives, which present a unique signature. In [App07], measured values of emissivity, reflectivity, and transmittance for certain materials such as metal, skin or denim are listed at 100-1000 GHz, as shown in Table 1.2.

Table 1.2: Optical properties of 5-mm thick materials. Data extracted from [App07].

Frequency (GHz)	Emissivity (e)			Reflectivity (r)			Transmission (t)		
	100	500	1000	100	500	1000	100	500	1000
Explosive on skin	0.76	0.95	0.94	0.24	0.05	0.06	0	0	0
Skin	0.65	0.91	0.93	0.35	0.09	0.07	0	0	0
Metal	0	0	0	1	1	1	0	0	0
Denim	0.09	0.49	0.85	0.01	0.01	0.05	0.9	0.5	0.1
T-shirt	0.04	0.2	0.3	0	0	0.05	0.96	0.8	0.65

1.4 State-of-the-art of millimeter-wave sources

One of the most important elements in any radar system is the power source. The millimeter-wave and terahertz bands have historically lacked convenient power sources, due to the limited cut-off frequency and breakdown voltage of active devices as well as the lower quality factor of passive components caused by ohmic and substrate losses [Sie02, Dav04, Sie10]. This issue is known by the scientific community as *THz gap* due to the lack of sources capable of producing adequate output power levels (see Fig. 1.1). Fig. 1.9 shows the output power performance of different types of sources around the *THz gap*. Three types of sources are currently available:

- Photonics. Quantum cascade lasers (QCL) are competitive in the terahertz band (1-10 THz), with output power levels of more than 1 W [Wil07, Li14a]. QCLs cannot

Figure 1.9: Output-power performance of different sources around the *THz gap*. Graph extracted from [Cha11]. RTD: Resonant tunnel diode. IMPATT: Impact ionization avalanche transit-time diode. Gunn: Gunn laser. QCL: Quantum-cascade laser. III-V: Denotes groups III and V in the periodic table of elements. UTC-PD: Uni-traveling-carrier photodiode. GaN: Gallium nitride. InP: Indium phosphide. TUNNET: Tunnel injection transit time diode.

operate at room temperature, so they still require cryogenically cooling for high-power operation [Bel09, Cha15].

- Vacuum tubes. Gyrotrons, extended interaction klystrons (EIKs) and traveling wave tubes (TWTs) are examples of millimeter-wave vacuum tubes [Gil11, Tru12, Joy14, Cai15]. The main advantage of EIKs is their wide bandwidth and the high power capability in the order of kilowatts [Hor07, Ste09, Ber14]. Furthermore, modern millimeter-wave TWTs include multistage depressed collectors which sort electrons by energy, slow the electrons down in retarding fields, and thus return energy to the power supply. This means overall efficiencies around 70% [Tru12].
- Solid-state devices. A variety of solid-state sources operate in the millimeter-wave band with wide outputs power levels and bandwidths. Currently, the signal generation is based on cascaded frequency-multiplication and amplification stages, starting from the microwaves band with an oscillator [Sie03, Cro05]. This signal-generation architecture offers better performance than millimeter-wave IMPATT oscillators, that are very noisy due to the avalanche process [Tes01]. The current trend in solid-state frequency multiplier and amplifier technology is of special interest for millimeter-wave radar systems in order to increase the transmitted bandwidth and the transmitted power, respectively [Sam06, Sam11]. Regarding millimeter-wave amplifiers, Gallium Nitride (GaN) is increasingly attracting interest for high power requirements, instead of Indium phosphide (InP) or GaAs, which are taken into account for the development of low-noise and medium-power amplifiers, respectively [Sam08, Mic05, Abb11]. On the other hand, frequency multipliers based on planar GaAs Schottky technology play a crucial role in this field, with significant improvements in efficiency and operating frequencies [Cha11, Hro13, Sil15].

A photograph of these three types of sources is shown in Fig. 1.10: A THz Quantum Cascade Laser with a peltier cooling device for temperature stabilization from *Hamamatsu* (a), an EIK at 94 GHz with a 1 kW of output power and 1 GHz of bandwidth from [CPI16]

(b), and a solid-state 37 W GaN amplifier with power-combination of 12 levels from 75 GHz to 110 GHz [Sch16] (c).

Figure 1.10: Photograph of a THz Quantum Cascade Laser (a), an 1 kW EIK at 94 GHz (a) and a 37 W solid-state GaN amplifier at W band (c).

In view of the above general outline, this thesis is based on solid-state devices due to their affordability, efficiency, reliability and by their commercial availability. To this end, a detailed analysis of the output power and the fractional bandwidth of the state-of-the-art of millimeter-wave multipliers and amplifiers, both commercial and research, is presented in Figs. 1.11-1.12. This analysis has been used for the millimeter-wave components selection of this thesis. Four types of devices are presented: Multipliers, amplifiers (solid-state MMICs and tubes) and commercial transmitter chains. The commercial components are selected from the two main manufacturers in Europe and the United States: Radiometer Physics GmbH and Virginia Diodes (VDI), respectively. For comparison purposes, the output power over the swept bandwidth of the transmitter chains of the proposed radar systems in this thesis are also shown.

As observed in Fig. 1.11, the maximum output power is achieved with commercial EIKs, with a bandwidth of 2 GHz. These devices are offered by Communications & Power Industries, Inc. (CPI) [CPI16], that were developed under a Defense Advanced Research Projects Agency (DARPA) program called *High Frequency Integrated Vacuum Electronics (HiFIVE)*. The current cost of these devices (~ 250 k€) is too high for the budget of the research projects related to this thesis. An acceptable and affordable high power is achieved by GaN MMICs of HRL [HRL13], around 27 dBm. Some research projects have developed power combined amplifiers based on corporate networks of 4-8 levels to increase the output power up to 37 dBm, as measured in [Sch10] for a bandwidth of 5 GHz, that are suitable for high-range radar systems. Recently, a new outstanding power-combined amplifier with 12 levels was presented in [Sch16], offering 45 dBm over 25 GHz of bandwidth at W band. Lastly, commercial chains based on frequency multipliers achieve larger bandwidths up to 100 GHz with an output power between 0-7 dBm, which is interesting for instrumentation equipments or high-resolution low-range radar systems.

Figure 1.11: Saturated output power of different signal sources, active and non-linear components: RPG and VDI transmitter chains, RPG and VDI multipliers, commercial and research MMIC amplifiers, commercial and research tubes and transmitter sources developed in this thesis.

Figure 1.12: Fractional bandwidth of different signal sources, active and non-linear components: RPG and VDI transmitter chains, RPG and VDI multipliers, commercial and research MMIC amplifiers, commercial and research tubes and transmitter sources developed in this thesis.

1.5 Thesis motivation and goals

This thesis is mainly concerned with two millimeter-wave radars at 94 GHz and 300 GHz for weather applications and security tasks, respectively. Both applications imply completely different system requirements in terms of antenna subsystem, acquisition times, transmitted power and narrowband/broadband signal generation [Kol07, Coo09]. The operating frequencies of both systems has been allocated within windows of atmospheric attenuation. Low cost and hardware versatility have been kept in mind along the design and development cycles.

In particular, this work contributes to the development of technology and experimental work in the fields of millimeter-wave imaging and cloud monitoring as summarized below:

1. Technological developments for CW-LFM radars at 100 GHz and 300 GHz developed at the Escuela Técnica Superior de Ingenieros de Telecomunicación (ETSIT) of the Universidad Politécnica de Madrid (UPM). Three hardware modifications have been carried out with the aim of cost-and-power-consumption reduction in order to develop a pre-industrial prototype:
 - (a) Design and development of a direct digital synthesizer (DDS)/phase-locked loop (PLL) signal generation architecture at 16 GHz.
 - (b) Phase stability analysis of the transmitted signal in order to enable new capabilities of the radars, with the aim of obtaining micrometer precision for small-displacements measurements.
 - (c) Evaluation of polarimetry techniques at 100 GHz in order to characterize the scattering issues in millimeter-wave imaging radars, in terms of specular reflection effects and its possible mitigation.
2. Development of a 94 GHz cloud profiling radar system. The main subjects addressed in this work are:
 - (a) Description of physical properties of clouds and its interaction with radar signals.
 - (b) Radar design and characterization of different subsystems.
 - (c) Development and evaluation of the signal processing techniques for this application.
 - (d) Analysis and interpretation of cloud observations. This system is currently under a continuous measurement campaign in collaboration with the Spanish National Weather Agency (AEMET) in order to validate this radar prototype.

1.6 Thesis organization

The remainder of this dissertation is organized in two parts:

- Technological developments for millimeter-wave radars developed at ETSIT-UPM.
 - Chapter 1 addresses the three hardware modifications indicated previously for the 100-GHz and 300-GHz radars. A comparison of these hardware modifications in terms of RF and radar performance with the radar early state is also carried out to determine the advantages of the new hardware modifications.
- Design and development of a cloud profiling radar at 94 GHz.
 - Chapter 3 links atmospheric science with radar engineering, by describing the interaction between cloud particles and radar signals. Furthermore, it addresses the preliminary radar design by means of an analysis of the state-of-the-art cloud radars at millimeter-wave and the radar equation.
 - Chapter 4 presents the implementation, characterization and evaluation performance of the cloud profiling radar at 94 GHz. To this end, the system architecture is shown and the RF measurements of each subsystem are presented. Furthermore, the signal processing is introduced in order to obtain the reflectivity and Doppler-velocity 3D profiles. Lastly, the analysis and interpretation of experimental measurement results of non-precipitating clouds are presented.
- The last chapter sums up the conclusions and the planned future work.
- Appendices A through G complement the information provided in these chapters.

1.7 Publications

The results obtained in this thesis have been published in the following journals, international conferences and workshops:

- Grajal, J.; Mencia-Oliva, B.; Yeste-Ojeda, O. A.; Garcia-Fernandez, A. F. and Rubio-Cidre, G. “A prototype of High Resolution ISAR imaging system at millimetre-wave band”. In *IEEE CIE International Conference on Radar*, oct. 2011.
- Mencia-Oliva, B.; Grajal, J.; Yeste-Ojeda, O. A.; Rubio-Cidre, G. and Badolato, A. “Multiple purpose hardware for sensors at 100 GHz”. In *IET International Radar Conference*, 2012.
- Mencia-Oliva, B.; Grajal, J.; Yeste-Ojeda, O. A.; Rubio-Cidre, G. and Badolato, A. “Low-Cost CW-LFM Radar Sensor at 100 GHz”. In *IEEE Transactions on Microwave Theory and Techniques*, Vol. 61, No. 2, pp. 986 - 998, feb. 2013.
- Rubio-Cidre, G.; Mencia-Oliva, B.; Badolato, A.; Úbeda-Medina, L.; Grajal, J.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “Characterization of a 300 GHz imaging radar for standoff detection”. In *International Radar Conference*, oct. 2014.
- Grajal, J.; Badolato, A.; Rubio-Cidre, G.; Úbeda-Medina, L.; Mencia-Oliva, B.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “3D High Resolution Imaging Radar at 300 GHz with Enhanced FoV”. In *IEEE Transactions on Microwave Theory and Techniques*, Vol. 63, No. 3, pp. 1097 - 1107, jan. 2015.

- Badolato, A.; Rubio-Cidre, G.; Úbeda-Medina, L.; Grajal, J.; Mencia-Oliva, B.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “A 300 GHz Imaging Radar for Standoff Anomaly Detection”. In *9th European Conference on Antennas and Propagation*, apr. 2015.
- Rubio-Cidre, G.; Badolato, A.; Úbeda-Medina, L.; Grajal, J.; Mencia-Oliva, B. and Dorta-Naranjo, B. P. “DDS-based Signal-Generation Architecture Comparison for an Imaging Radar at 300 GHz”. In *IEEE Transactions on Instrumentation and Measurement*, Vol. 64, No. 11, pp. 3085 - 3098, jun. 2015.
- Rubio-Cidre, G.; Grajal, J.; “THz Radar Technologies”. In *Spain-Japan Joint Workshop on Millimeter-wave and Terahertz issues*, mar. 2016.
- Rubio-Cidre, G.; Grajal, J.; Garcia-Pino, A. and Rubiños, O.; “Design and Preliminary Results of a Ground-Based Cloud Profiling Radar at 94 GHz”. In *IEEE Radar Conference (RadarConf)*, may 2016.
- García, F.; Úbeda-Medina, L.; Badolato, A.; Grajal, J.; Rubio-Cidre, G.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “Optimization of a Compact THz Imaging Radar for Real-Time Operation”. In *9th Global Symposium on Millimeter-Waves (GSMM)*, jun. 2016.
- Grajal, J.; Rubio-Cidre, G.; Badolato, A.; Úbeda-Medina, L.; García, F.; Garcia-Pino, A. and Rubiños, O.; “Compact Radar Front-End for an Imaging Radar at 300 GHz”. In *IEEE Transactions on Terahertz Science and Technology*, Vol. 7, No. 3, pp. 268 - 273, may 2017.
- Rubio-Cidre, G.; Rodríguez, C.; Ramírez-Torres, M.; Grajal, J. and Rubiños, O.; “Understanding cloud dynamics using a ground-based radar at 94 GHz”. In *IEEE Radar Conference (RadarConf)*, may 2017.
- Grajal, J.; Rubio-Cidre, G.; García, F.; Ramírez-Torres, M.; “Terahertz Radar Developments at UPM”. In *Sub-millimetre wave and THz imaging for security workshop at European Microwave Week (EuMW)*, oct. 2017 (invited).

The thesis results also have been published in the following national conferences:

- Grajal, J.; Badolato, A.; Mencia-Oliva, B.; Rubio-Cidre, G.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “Sensor radar a 300 GHz para la generación de imágenes tridimensionales”. In *I Congreso Nacional de I+D en Defensa y Seguridad (DESEi+d)*, nov. 2013.
- Rubio-Cidre, G.; Mencia-Oliva, B.; Badolato, A.; Grajal, J.; Úbeda-Medina, L.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “Characterization of a 300 GHz imaging radar for standoff detection”. In *XXIX Simposium Nacional de la Unión Científica de Radio (URSI)*, sep. 2014.
- Grajal, J.; Rubio-Cidre, G.; Badolato, A.; Úbeda-Medina, L.; Garcia-Pino, A.; Gonzalez-Valdés, B. and Rubiños, O.; “Tecnología de milimétricas para radares de imágenes a 300 GHz”. In *II Congreso Nacional de I+D en Defensa y Seguridad (DESEi+d)*, nov. 2014.
- Rubio-Cidre, G.; Grajal, J.; Garcia-Pino, A. and Rubiños, O.; “Design and Preliminary Results of a Ground-Based Cloud Profiling Radar at 94 GHz”. In *XXXI Simposium Nacional de la Unión Científica de Radio (URSI)*, sep. 2016.

It is worth mentioning that the paper “Characterization of a 300 GHz imaging radar for standoff detection” was awarded as the “Accésit Premio Jóvenes Científicos URSI 2014” during the “XXIX Symposium Nacional de la Unión Científica Internacional de Radio (URSI)”, which took place in Valencia, Spain, in September of 2014.

1.8 Research projects related to this thesis

The work described in next chapters has been funded by several research projects listed below:

- Proyecto CENIT “Sistemas para la detección de explosivos en centros e infraestructuras públicas (SEDUCE)”. Centro para el desarrollo Tecnológico Industrial (CDTI). Ministerio de Ciencia e Innovación. 2008-2011.
- Terahertz Technology for Electromagnetic Sensing Applications (TERASENSE). CSD2008-00068. Consolider Ingenio - Ministerio de Ciencia e Innovación. 2009-2013.
- Sensores heterodinos activos en la banda de terahercios para detección remota. TEC2011-28683-C02-01. Comisión Interministerial de Ciencia y Tecnología. 2012-2015.
- Space Debris Radar - Radar para detección de basura espacial (SPADERadar-CM). P2013/ICE 3000. Comunidad de Madrid. 2014-2016.

Part I

Development of technology for millimeter-wave radars at UPM

Chapter 2

Hardware-based modifications of millimeter-wave CW-LFM radars

2.1 Operating principle of CW-LFM radars

A CW-LFM radar consists essentially of a transceiver and an analog-to-digital converter (ADC), as shown in Fig. 2.1. A two-antenna configuration is usually implemented in order to improve the Tx-Rx isolation, instead of a single antenna with a circulator, which offers an isolation lower than 30 dB at millimeter-wave band. The transmitted signal is generated by a DDS, and then upconverted to the selected operating frequency by using frequency multiplication or upconversion stages. A sample of the transmitted signal feeds the local-oscillator (LO) port of the mixer, which downconverts the received signal -previously amplified by a low-noise amplifier (LNA)- to intermediate frequency (IF). If high-power transmission is required by the radar application, a high-power amplifier (HPA) should be placed in the transmitter chain. After IF signal digitization, IF signal conditioning is mandatory to reduce the influence of receiver and ADC noises on the detected targets. In order to guarantee system coherence and synchronization, a master oscillator is required, both for the ADC and DDS clock inputs.

Figure 2.1: Block diagram of a CW-LFM radar with two different antenna configurations: single antenna and two antennas.

The stretch-processing technique is adopted in order to deal with CW-LFM signals [Jan07]. As shown in Fig. 2.2, this technique basically consists in mixing the local os-

cillator signal, that is a sample of the transmitted signal, and its delayed version due to the round-trip propagation time of the target reflection. Thus, after mixing, a sinusoid of frequency f_b is obtained for a point-like target located at a range R :

$$f_b = \frac{2 \cdot R \cdot BW}{c \cdot T_m} \quad (2.1)$$

where c is the free-space propagation velocity, BW is the transmitted bandwidth and T_m is the chirp period, which is the inverse of the pulse repetition frequency (PRF).

Figure 2.2: Transmitted and received chirp signals of a CW-LFM radar.

After the digitization of the IF signal, a Fast Fourier Transform (FFT) per chirp period is applied. It consists of a bank of matched filters that are separated by the inverse of the chirp period, allowing the detection of different beat frequencies that correspond to the location of different targets. Each frequency point is called *bin*. Taking into account the demanding requirement for high instantaneous dynamic range in these radars, the IF signal is windowed, in spite of the degradation of the theoretical range resolution by a factor α due to the mainlobe widening of the filters [Har78]. A Hamming window is adopted in this work due to the compromise between the high sidelobe performance (-43 dB), the mainlobe width ($\alpha=1.81$ bins measured at -6 dB point, see Fig. 2.18b) and the equivalent noise bandwidth ($\alpha_{noise}=1.36$ bins). Therefore, considering that the frequency resolution due to the FFT and windowing is $\Delta f = \frac{\alpha}{T_m}$ and substituting it into Eq. (2.1), the theoretical range resolution of a CW-LFM radar only depends on the transmitted bandwidth:

$$\Delta R \approx \frac{c}{2 \cdot BW} \cdot \alpha \quad (2.2)$$

Consequently, the transmitted bandwidth maximization is crucial in order to achieve high-range-resolution and to resolve closely spaced targets [Jan07], particularly in imaging applications.

2.2 CW-LFM radars developed in ETSIT-UPM

The Grupo de Microondas y Radar (GMR) of the UPM is a research group with more than 20 years of experience in simulation and development of microwave and millimeter-wave CW-LFM radar systems. Since 1996, with the participation in the ARIES project (a high-resolution radar developed by Indra Sistemas S.A. in co-operation with the UPM and the Spanish Ministry of Defence [Per02]), GMR has developed several millimeter-wave CW-LFM radar systems [Alm07, Gon07, Alm09, Gon09, Fel11, Men13b, Men13a, Gra15] for multiple

applications such as synthetic aperture radar (SAR), inverse SAR (ISAR), interferometry, through-the-wall surveillance, standoff imaging,...

The last two millimeter-wave radars have been selected in order to improve their performance by means of three hardware modifications that will be shown after a brief description of both radar systems:

2.2.1 Standoff imaging radar at 300 GHz

The block diagram of the CW-LFM imaging radar at 300 GHz is presented in Fig. 2.3. It is divided into the following subsystems: the microwave signal generation source, the millimeter-wave subsystem, the scanning antenna subsystem and the data acquisition subsystem.

Figure 2.3: Block diagram of the imaging radar architecture. The radar structure is divided into four different parts, corresponding to the technologies involved in the design: Digital electronics, microwave subsystem, millimeter-wave subsystem and scanning antenna subsystem.

The scanning antenna subsystem consists of a bifocal ellipsoidal Gregorian reflector [Gar13, Gar14]. A photograph is shown in Fig. 2.4. This configuration ensures a spot size about 1.6 cm at 300 GHz on the field of view of $50 \times 90 \text{ cm}^2$ at a standoff distance of 8 m. The main reflector aperture is 60 cm. The beam scanning is based on the rotation and vertical tilting of a flat mirror, which depicts an elliptical scanning. The rotation motion of the scanning mirror is provided by a EC-4 Pole Maxon motor which turns the mirror at 2,000 rpm. The tilting movement is obtained with a 35H4C Haydon Kerk motor by displacing its stroke 30.4 mm at $60.9 \frac{\text{mm}}{\text{s}}$ speed. These values allow a complete scan of the mannequin image in a total acquisition time of 0.5 seconds.

The millimeter-wave subsystem shown in Fig. 2.5 is based on the traditional homodyne architecture [Rub14]. This architecture is selected due to its low cost and compactness compared to the heterodyne architecture, in which extra oscillators are usually needed [Sch93]. The reduction of sensitivity in zero-IF homodyne receivers does not affect our imaging radar because the IF of interest is 8 MHz. The microwave chirp signal with a bandwidth of 1.5 GHz is input to an active $\times 6$ multiplier followed by a passive $\times 3$ multiplier. Therefore, the system sweeps 27 GHz centered at 300 GHz. The LO signal for the receiver chain is achieved with two triplers to pump a SHM at 143.1-156.6 GHz. This subsystem is presented in detail in [Men13b].

The radar system coherence is guaranteed by a master oscillator of 10 MHz. The radar system is controlled by a personal computer (PC) that manages the DDS chirp configuration and the ADC. An active bandpass filter is included before the digitization of the IF signal. It amplifies and filters out only the frequencies corresponding to a suitable range around the standoff distance of 8 m.

Figure 2.4: Photograph of the imaging radar showing the key components and the transmitting and receiving signal paths.

Figure 2.5: Photograph of the millimeter-wave subsystem showing the key components.

2.3 Broadband chirp source based on a DDS/PLL structure

The first hardware modification consists in the development of a new microwave signal generator at 16 GHz, also called *chirp source* due to the generation of CW-LFM signals, based on a different architecture than the previous microwave subsystem. In this Section, a complete characterization of the RF electronics and radar performance and its comparison with the earlier configuration [Men14] are carried out.

As indicated in Introduction, these high-resolution radar systems require broadband and fast-switch signal generation architectures in order to achieve a high-resolution and near-video-rate imaging system. Broadband CW-LFM signal generation architectures with emphasis on fast switch features have been considered in the research and industrial communities in the last years. It is due to the need for near real-time images obtained with imaging radars based on scanning antenna subsystems [She09, Co09, Qua09, Co11, Fri11, Cha12, Rec12, Co13, Bla14]. On the other hand, the industrial market demands the measurement time reduction in automated RF test systems [Oze02, Joa08, Che09]. In addition, real-time chemical sensing systems for terahertz frequency-domain spectroscopy applications are incorporating the electronic signal generation for replacing pulsed laser generation, primarily due to the higher spectral resolution [Kar05, Liu11, Liu13].

CW-LFM signals are commonly used for low-probability-of-intercept and high-range-resolution radars [Weh95, Pac09]. The use of this kind of signals is advantageous because of the development of simple and cheap low-power solid-state transmitters compared with high-power pulsed sources [Sko08, Ric10]. Secondly, the range measurement can be effectively performed by an FFT processor. Taking advantage of these features, CW-LFM signals are usually employed for imaging radars above 100 GHz, as summarized in Table 2.1. All these systems are designed for standoff sensing of person-borne concealed threats with the exception of the system described in [Loe09] that is used for material inspection. Although there are competing technologies like yttrium iron garnet (YIG) tuned oscillators or free-run broadband voltage-controlled oscillators (VCO), the most recent radars are based on direct digital synthesizers (DDS). YIG oscillators are not suitable for fast-switching generation, in spite of their excellent phase noise performance, and free-run VCO generation is limited by the non-linearity of the VCO tuning characteristic [Bre11]. The advantages of DDS in comparison with all-analog chirp sources for radar systems are discussed extensively in the literature [Ana99, Jan07, Van10]. It is important to mention the high chirp linearity, the fine frequency resolution and the high-frequency update rate.

A reduction of cost and power consumption is mandatory to offer an affordable and energy-efficient imaging system solution for security applications [Tan13]. Furthermore, an increase in the transmitted power of the imaging radar is not helpful for the improvement of the radar performance if the signal-to-noise ratio (SNR) of the detected targets is limited by the transmitted phase noise, which is originated in the signal-generation architecture [Coo12, Rob13]. Taking all these factors into account, the new implemented generator is based on a narrowband DDS within a PLL. This architecture is commonly used owing to the linearity of the frequency sweep provided by the PLL circuit [Mus97, Wu09, Dao13]. The first signal generator was based on a broadband DDS with high clock speed combined with an up-conversion stage [Men13b]. It is employed because of its wide chirp period configuration (milli- to nanosecond sweeps). Consequently, the key goal is the selection of the signal-generation architecture that ensures the proper operation of the imaging radar with a cost-saving objective based on three figures of merit: radar image quality, power consumption and cost.

Table 2.1: Overview of imaging radars above 100 GHz. The main parameters are the center frequency (f_c), the transmitted bandwidth (BW), the chirp period (T_m), the fractional bandwidth (FBW) and the chirp rate (BW/T_m).

	f_c	BW	T_m	FBW	BW/T_m	Architecture
Loeffler [Loe09]	275 GHz	90 GHz	240 μ s	32.72 %	375 MHz/ μ s	Free-run VCO
Dengler <i>et al.</i> [Den07]	604 GHz	88 GHz	1 s	14.61 %	88 kHz/ μ s	YIG
Cooper <i>et al.</i> [Coo08a]	582.5 GHz	13 GHz	50 ms	2.17 %	260 kHz/ μ s	YIG
Cooper <i>et al.</i> [Coo08b]	590.5 GHz	29 GHz	12.5 ms	4.87 %	2.32 MHz/ μ s	Narrowband DDS/PLL
Dengler <i>et al.</i> [Den09]	670 GHz	28.8 GHz	1 ms	4.29 %	28.8 MHz/ μ s	Narrowband DDS/PLL
Cooper <i>et al.</i> [Coo11]	676.7 GHz	28.8 GHz	100 μ s	4.26 %	288 MHz/ μ s	Narrowband DDS/PLL
Bryllert <i>et al.</i> [Bry13]	220 GHz	10.8 GHz	30 μ s	4.91 %	360 MHz/ μ s	Narrowband DDS/PLL
Robertson <i>et al.</i> [Rob12]	340 GHz	3.6 GHz	51.2 μ s	1.05 %	70.31 MHz/ μ s	Broadband DDS
Robertson <i>et al.</i> [Rob13]	340 GHz	3.6 GHz	76 μ s	1.05 %	47.36 MHz/ μ s	Narrowband DDS/PLL
Robertson <i>et al.</i> [Rob16a]	340 GHz	30 GHz	100 μ s	8.82 %	300 MHz/ μ s	Broadband DDS
This work	299.7 GHz	27 GHz	54 ns-1.5 ms	9 %	18-5000 MHz/ μ s	Broadband DDS
This work	299.7 GHz	27 GHz	$\geq 30 \mu$ s	9 %	≤ 900 MHz/ μ s	Narrowband DDS/PLL

As indicated in Section 2.1, the core of CW-LFM radars is a DDS-based signal generation. In order to guarantee a high performance of the chirp sources, special attention should be paid to the specifications of the DDS and its reference clock. Firstly, a low-phase-noise reference clock is desirable to guarantee the DDS phase noise indicated by the manufacturer. On the other hand, the spurious signals content at the DDS output depends on the internal digital-to-analog converter (DAC) quantization errors and phase truncation and the reference-clock spurious signals [Ana99].

For the purpose of comparing the commercial DDS available in the market, Table 2.2 shows the broadband and narrowband DDS for chirp signal generation of the two main DDS manufacturers, Analog Devices and Euvis. The distinction between narrowband and broadband DDS is defined arbitrarily by a bandwidth limit of 400 MHz. The low-pass reconstruction filter, that is needed for removing the aliased image components of the DAC *sinc* envelope response [Ana99], is not included in the evaluation board of the broadband DDS in Table 2.2. The DDS used in the imaging radars of Table 2.1 are also included.

After the description and characterization of previous and new chirp sources in Sections 2.3.1 and 2.3.2, respectively, the evaluation criteria for the radar performance is defined in Section 2.3.3. Subsequently, the radar performance characterization and the radar images of threats hidden under clothing are carried out in Sections 2.3.4 and 2.3.5, respectively, in order to compare the radar image quality with both chirp sources. Lastly, an additional optimization of the chirp sources is proposed in Section 2.3.6.

2.3.1 DDS-based signal generation

The two DDS-based signal generation sources shown in Fig. 2.8 are described below.

2.3.1.1 Chirp source 1.- Broadband DDS

The chirp source 1 block diagram is shown in Fig. 2.8a. It was developed and characterized in [Men14], It is based on a chirp signal with a bandwidth of 500 MHz generated by a broadband DDS from EUVIS and then upconverted to 5.3-5.8 GHz and multiplied by 3. This DDS offers a wide chirp time configuration between 54 ns and 1.5 ms. The LO signal of the up-conversion stage and the DDS reference clock are obtained with a phase-locked oscillator (PLO) from Synergy at 2.4 GHz that is locked to a commercial 10-MHz oven controlled crystal oscillator (OCXO) from Oscilloquartz. A photograph of these components is shown in Fig. 2.9.

Fig. 2.10a and 2.10b present two photographs of the microwave subsystem of chirp source 1. Three coupled-line filters have been implemented in the different frequency bands in order to remove the aliased image components of the DDS, the undesired mixing products and the harmonic bands resulting from frequency multiplication. In addition, it should be pointed out that an equalization filter has been designed to flatten the amplitude envelope of the chirp signal generated by the DDS (Fig. 2.10b), which mitigates the DAC *sinc* envelope response, and the mixer response.

Table 2.2: Overview of commercial broadband and narrowband DDS. This table contains the DDS bandwidth (BW), the number of bits of DAC (b), the number of bits of phase accumulator (n), the frequency resolution (Δf), the minimum frequency update time (Δt), the spurious-free dynamic range (SFDR), the power consumption and the evaluation board cost.

Manufacturer	Model	BW	b	n	Δf	Δt	SFDR	Consumption	Cost
Euvis	DSM309	DC - 2 GHz	11	32	0.93 Hz	2 ns	-45 dBc	14.6 W	3,400 €
Euvis (CS1) ¹	DSM303	DC - 1.5 GHz	11	32	0.69 Hz	2.6 ns	-40 dBc	9 W	1,700 €
Euvis	DSM202	DC - 1 GHz	11	32	0.46 Hz	16 ns	-45 dBc	5.8 W	1,500 €
Analog [Rob16a]	AD9914	DC - 1.75 GHz	12	32	190 pHz ²	6.8 ns	-50 dBc	3 W	600 €
Analog	AD9915	DC - 1.25 GHz	12	32	135 pHz ²	9.6 ns	-57 dBc	2.9 W	600 €
Analog [Rob12]	AD9910	DC - 0.4 GHz	14	32	0.23 Hz	4 ns	-55 dBc	0.95 W	500 €
Analog (CS2) ¹ , [Rob13]	AD9958	DC - 0.2 GHz	10	32	0.12 Hz	8 ns	-64 dBc	0.4 W	400 €
Analog [Den09]	AD9956	DC - 0.2 GHz	14	48	1.4 μ Hz	10 ns	-55 dBc	0.4 W	400 €
Analog [Coo11]	AD9954	DC - 0.2 GHz	14	32	0.01 Hz	10 ns	-52 dBc	0.22 W	400 €
Analog	AD9913	DC - 0.125 GHz	14	32	0.058 Hz	4 ns	-55 dBc	0.1 W	400 €

¹ CS1 and CS2 are the two chirp sources presented in this work.

² These DDS have an auxiliary accumulator to modify the DDS output frequency. It allows to improve n up to 64 bits.

(a) Chirp source 1 (CS1): Broadband DDS architecture.

(b) Chirp source 2 (CS2): Narrowband DDS/PLL architecture.

Figure 2.8: Simplified block diagram of the chirp sources. Test points are defined with capital letters inside squares in the block diagrams.

Figure 2.9: Photograph of the DDS-based signal generation module of chirp source 1.

Figure 2.10: Photograph of the microwave subsystem of chirp source 1.

2.3.1.2 Chirp source 2.- Narrowband DDS/PLL

The chirp source 2 is made up of a narrowband AD9958 DDS from Analog Devices that generates a chirp signal with a bandwidth of 9.375 MHz and a $\times 80$ PLL to bring up the chirp signal to 7.95-8.7 GHz. A photograph of both subsystems is presented in Fig. 2.11 and Fig. 2.12, respectively. This signal is doubled to achieve a bandwidth of 1.5 GHz centered at 16.65 GHz. The block diagram is shown in Fig. 2.8b. The DDS reference clock is generated by a PLO from Synergy at 500 MHz that is locked to the 10 MHz master oscillator from Oscilloquartz. It should be noted that this system has been developed in collaboration with Universidad de Las Palmas de Gran Canaria [Rub15].

Figure 2.11: Photograph of the DDS-based signal generation module of chirp source 2.

Figure 2.12: Photograph of the microwave subsystem of chirp source 2.

The PLL loop filter is based on an active lead-lag filter topology using the differential phase detector outputs. According to the PLL design theory [Ban06], the PLL 3 dB bandwidth (2.3) and the PLL noise bandwidth (2.4) follow these expressions:

$$BW_{PLL-3dB} = w_n \cdot \sqrt{1 + 2\zeta^2 + \sqrt{(1 + 2\zeta^2)^2 + 1}} \quad (2.3)$$

$$BW_{noise} = \frac{w_n}{2} \cdot \left(\zeta + \frac{1}{4 \cdot \zeta} \right) \quad (2.4)$$

where w_n is the natural frequency and ζ is the damping factor. In order to maintain the PLL locked handling a CW-LFM signal with a bandwidth BW_{out_chirp} , the maximum chirp

rate (BW_{out_chirp}/T_{m_min}) at the PLL output is limited by the lock frequency w_L (2.5) and the loop filter zero frequency w_2 (2.6) of the PLL [Wol91], as shown in (2.7). Table 2.3 summarizes the aforementioned PLL parameters of the developed narrowband DDS/PLL architecture. The first six values are obtained from the PLL parameters: VCO and phase detector gains, PLL multiplication factor and the capacitors and resistors values of the loop filter.

$$w_L = 4\pi\zeta w_n \quad (2.5)$$

$$w_2 = \frac{w_n}{2\zeta} \quad (2.6)$$

$$\frac{2 \cdot \pi \cdot (BW_{out_chirp}/N_{PLL})}{T_{m_min}} \leq w_L \cdot w_2 \quad (2.7)$$

Table 2.3: Summary of the PLL parameters of the narrowband DDS/PLL architecture.

Parameter	Value
Natural frequency (w_n)	560 krad/s
Damping factor (ζ)	1.1
PLL 3-dB bandwidth ($BW_{PLL-3dB}$)	225 kHz
PLL noise bandwidth (BW_{noise})	374 kHz
Lock frequency (w_L)	7.8 Mrad/s
Loop filter zero frequency (w_2)	252 krad/s
PLL output bandwidth (BW_{out_chirp})	750 MHz
PLL multiplication factor (N_{PLL})	80
Minimum chirp period (T_{m_min})	30 μ s
Maximum chirp rate at the PLL output (BW_{out_PLL}/T_{m_min})	25 MHz/ μ s

As regards the phase noise performance in PLL circuits [Kro03], the phase noise of the PLL reference signal is multiplied by the PLL multiplication factor and it dominates the output phase noise at offset frequencies inside the PLL 3-dB bandwidth while the VCO phase noise dominates the output phase noise at higher offset frequencies.

Table 2.4 shows the market cost in Euros and the power consumption in Watts for the components of the chirp sources. The narrowband DDS/PLL architecture reduces the cost and the power consumption compared to the broadband DDS architecture due to the DDS features.

Table 2.4: Chirp source cost and power consumption.

	Broadband DDS (CS1)		Narrowband DDS/PLL (CS2)	
	Component	Power	Component	Power
	Cost	Consumption	Cost	Consumption
DDS	1,700 €	9 W	400 €	0.4 W
Oscilloquartz oscillator	1,300 €	8.7 W	1,300 €	8.7 W
Microwave subsystem	250 €	7.3 W	250 €	6 W
Total	3,250 €	25 W	1,950 €	15.1 W

2.3.2 Chirp sources characterization

This section shows the RF characterization of the developed chirp sources described previously. Firstly, the power spectrum, the SFDR and the power ripple results of both chirp sources are presented in Section 2.3.2.1. Secondly, the phase noise measurements at different test points of the chirp sources are shown in Section 2.3.2.2. Lastly, the RF non-linearities of both chirp sources are presented in Section 2.3.2.3. The chirp period is 250 μ s for all of the measurements presented in this section.

2.3.2.1 Power Spectrum, SFDR and power ripple

The power ripple and the SFDR measurement at the DDS output (test point C) of both chip sources are represented in Fig. 2.13. It is observed that the narrowband DDS/PLL architecture has a SFDR of -57 dBc and a power ripple (peak-to-peak variation) of 0.1 dB. However, the broadband DDS architecture has -41 dBc of SFDR and 2.2 dB of power ripple due to the *sinc* roll-off response of the broadband DDS. As indicated before, RF non-linearities appear close to the beat frequency due to the non-flat amplitude response throughout the swept bandwidth of the RF circuitry. Therefore, an important RF design strategy is to flatten the amplitude response of the chirp signal. In this scheme, an equalization filter between the broadband DDS and the mixer (see Fig. 2.8a) is used to compensate this degradation.

The output chirp signal (test point E) of both chirp sources are plotted in Fig. 2.14. The SFDR of the narrowband DDS/PLL is -61 dBc. This represents an improvement of 4 dB with respect to the DDS output in this particular case. It is due to the PLL bandwidth that reduces the amplitude of out-of-band spurious signals. The SFDR at the output of the broadband DDS architecture is -41 dBc, that is equal to the SFDR at the broadband DDS output. Therefore, the SFDR of the narrowband DDS/PLL architecture is 19 dB better with respect to the broadband DDS architecture. Both chirp sources present a good power ripple lower than 1.5 dB throughout a bandwidth of 1.5 GHz.

2.3.2.2 Phase noise

The phase noise measurements at different test points of the chirp sources are plotted in Fig. 2.15. For the broadband DDS architecture, it can be noted that the phase noise after the mixer (test point D) is imposed by the PLO phase noise (test point B) that provides the LO signal of the mixer. Then, the tripler adds a multiplication factor of $20 \cdot \log_{10}(3) = 9.5$ dB.

The output phase noise of the narrowband DDS/PLL architecture is limited by the DDS phase noise (test point C) at offset frequencies inside the PLL 3 dB bandwidth (225 kHz). The DDS phase noise reported by the manufacturer is deteriorated by the PLO phase noise (test point B). According to PLL theory for offset frequencies inside the PLL 3 dB bandwidth [Ban06], the PLL multiplication factor ($20 \cdot \log(80) = 38$ dB) is added to the phase noise of the narrowband DDS (test point C), obtaining the phase noise at the test point D.

For offset frequencies outside the PLL 3 dB bandwidth, the phase noise is limited by the noise of the PLL frequency divider, which in this case is higher than the VCO phase noise at 8 GHz supplied by the manufacturer (dashed line in Fig. 2.15b). Lastly, the doubler adds a multiplication factor of $20 \cdot \log_{10}(2) = 6$ dB.

Thus, the output phase noise of the chirp sources gets better with the broadband DDS architecture at all offset frequencies between 100 Hz-10 MHz. In both chirp sources, the output phase noise at 16.65 GHz is limited by the PLOs at low offset frequencies. This is not a significant limitation on the radar performance because the decorrelation effect at the mixer compensates its impact on the noise floor, as will be shown in Section 2.3.3. Overall,

(a) Broadband DDS architecture

(b) Narrowband DDS/PLL architecture

Figure 2.13: Power ripple and SFDR at the DDS output (test point C) of both chirp sources. The output power is -3 and 3 dBm, respectively.

Figure 2.14: Power ripple and SFDR measured at the chirp sources output (test point E). The SFDR is -41 and -61 dBc respectively. The output power is 13 dBm in both cases.

it would not be appropriate to generalize that the phase noise results are always better with broadband DDS architectures. However, it is worth mentioning that narrowband DDS/PLL architectures always present a lower power consumption and cost as regards broadband DDS architectures with up-conversion stages primarily due to DDS and PLL features [Wu09].

2.3.2.3 RF non-linearities

A delay-line test system is used to monitor the signal spectral purity at the chirp sources output (test point E). The block diagram of the test system is shown in the inset of Fig. 2.16. A delay line based on a coaxial cable of 10.7 m in length is used with a simple transceiver made up of general purpose surface-mount technology components [Men13b]. The transceiver output signal is digitized and an FFT is applied in the same manner as described in Section 2.1, considering that the transmitted bandwidth is 1.5 GHz.

The RF non-linearities of both chirp sources are shown in Fig. 2.16. The Hamming windowing function is plotted as a reference to observe easily the spectral purity. In addition, for comparison purposes, the RF non-linearities of the Agilent E8257D analog signal generator measured with the delay-line test system are also plotted in Fig. 2.16. It generates a chirp signal of 1.5 GHz of bandwidth centered at 16.65 GHz (as the developed chirp sources) and the measured power ripple throughout the swept bandwidth is 0.5 dB. The received power and the target location are normalized in order to ease the interpretation.

From these results, the sidelobe level (SLL) is -33 dBc with both chirp sources. Additionally, the lab instrumentation (Agilent E8257D) offers a similar SLL, demonstrating the high spectral purity of the developed DDS-based signal generation architectures.

(a) Broadband DDS architecture

(b) Narrowband DDS/PLL architecture

Figure 2.15: Phase noise measurements at the test points of the developed chirp sources: 10 MHz master oscillator output (A), PLO output (B), DDS output (C), coupler output (D) and chirp sources output (E). The VCO phase noise of the narrowband DDS/PLL is plotted in dashed line.

Figure 2.16: RF non-linearities measurement at the chirp source outputs (test point E) and of the Agilent E8257D analog signal generator. Inset: Block diagram of the delay-line test system.

2.3.3 Evaluation criteria for radar detection performance

The radar detection performance in imaging radars is commonly limited by three effects: the RF non-linearities [Gri91], the transmitted phase noise [Bro05] and DDS spurious signals. These effects are illustrated in Fig. 2.17 and their impact on the radar detection performance is described below.

- The RF non-linearities generate SLL which are higher than the SLL defined by the windowing function (Fig. 2.17a) at low offset frequencies from the beat frequency of detected targets, degrading the instantaneous dynamic range. This limitation is generated by the non-flat amplitude and phase response throughout the swept bandwidth of the RF circuitry [Men13b]. It can be partially compensated by the calibration procedure explained in [Kno93] because of the deterministic behavior of RF non-linearities.
- The transmitted phase noise generates a noise floor (Fig. 2.17b) that is commonly higher than the receiver thermal noise floor far from the beat frequency, reducing the SNR of detected targets*. This degradation is diminished at low offset frequencies due to the noise decorrelation effect [Bud93, Bur93, Sch93], which acts as a high-pass filter on the phase noise of the signal. The decorrelation factor follows the next equation:

$$C = 10 \cdot \log \left(4 \cdot \sin^2 \left(\pi \cdot \frac{2R}{c} \cdot \Delta f_{beat} \right) \right) \quad (2.8)$$

where Δf_{beat} is the offset frequency from the beat frequency, that corresponds to the standoff distance R .

Taking advantage of the decorrelation effect, the phase noise requirements of the signal-generation architectures can be reduced at low offset frequencies (see Fig. 2.17b). This might result in a cost reduction of master oscillators and PLOs.

- DDS spurious signals are caused by the inherent digital nature of the DDS-based signal generation. Usually they appear far from the beat frequency of the detected targets (see Fig. 2.17b). Thus, this limitation imposes that a range gating at the standoff distance should be used in the radar image formation algorithm. To overcome this limitation without considering DDS spurious-reduction techniques [Wil91], closed-loop architectures such as PLLs perform better than open-loop architectures due to the PLL bandwidth reduces the power level of out-of-band spurious signals [Che07, Van10].

Three figures of merit are specified in order to evaluate the impact of the signal-generation architectures on the radar detection performance: SNR, range resolution (ΔR) and instantaneous dynamic range (IDR), see Fig 2.18. The SNR is measured with a single target located at a distance corresponding to f_{beat} , whereas the ΔR and the IDR are measured with two targets located at distances corresponding to f_{b1} and f_{b2} , respectively. In general, the SNR measurement is limited by the transmitted phase noise and the two-targets measurements are limited by the RF non-linearities.

- SNR. A single target presents an SNR that can be found from its power level and the noise power [Tsu04]. The noise power can be imposed by the phase noise, or by the receiver thermal noise, as indicated before (Fig. 2.18a). The set-up for the SNR measurement also allows to obtain the relative power level of spurious signals with respect to the reflected power of a single target.

*If the thermal noise is the SNR limiting factor, this can be improved by increasing the transmitted power or by the reduction of the receiver noise figure. However, when phase noise is the SNR limiting factor, these improvements are not useful because the relation between the noise floor and the reflected power of the target is constant and it does not depend on the transmitted power, as shown in [Bro05].

Figure 2.17: Radar performance limitations: RF non-linearities and decorrelation effect of the phase noise around the target response. Spurious signals are caused by the DDS-based signal generation.

- ΔR . It is the radar ability to resolve two close targets in the range dimension. As illustrated in Fig. 2.18b, two targets with the same power level move away until they are distinguished with a local minimum at -6 dB [Ric10].
- IDR. To determine the radar ability to resolve multiple targets with different power levels separated Δf_{IDR} (see Fig. 2.18c), the IDR concept is used [Tsu04].

2.3.4 Radar performance characterization

The figures of merit of the radar specified in Section 2.3.3 have been assessed by means of experiments based on the set-ups of Fig. 2.19. The antenna subsystem is not scanning during the measurements and, consequently, it points to the target location. The SNR is measured with a sphere while the two target measurements (ΔR and IDR) are carried out with a cardboard and an aluminium plate. The set-up of Fig. 2.19b allows to separate two targets at a fixed distance using a linear positioner and to adjust the angle of incidence of the second target using an azimuth positioner to modify the target reflected power.

Due to specular reflection, the received power is highly dependent on the angle of incidence and thus the material recognition of detected objects is impossible based on the radar cross section measurement [Coo11, Nas16]. For instance, an incidence variation of 2° with respect to the maximum reflected power of the aluminium plate produces a reflected power reduction of 35 dB in our system, as shown in Fig. 2.20. The problems of non-radar active imaging systems associated to the specular reflection effect [She09] are currently solved with radar systems taking advantage of their high-range resolution capability. These systems represent the 3D images as a function of the range dimension in order to exhibit strong contrast images [Coo08a].

Figure 2.18: Figures of merit for the radar detection performance evaluation: SNR is measured with a single target; ΔR and IDR are measured with two targets.

Figure 2.19: Simplified set-ups for radar performance evaluation.

Figure 2.20: Received power from an aluminium plate as a function of the angle of incidence.

2.3.4.1 Single-target measurement

The single target measurement has been carried out using a 40 mm diameter sphere. This sphere is used due to its measured noise floor being higher than the receiver thermal noise (-84 dBm, see [Men14]) and, therefore, it serves to ensure that the measured noise floor is imposed by the transmitted phase noise. The single-target set-up allows to characterize the SNR, the nearest spurious signals to the target and the RF non-linearities.

2.3.4.1.1 SNR-limited by phase noise

The transmitted phase noise at 300 GHz using both signal-generation architectures is plotted in Fig. 2.21. These results are estimations based on the measured phase noise at the chirp sources output (Fig. 2.15) and then translated to 300 GHz by adding the multiplication factor of the millimeter-wave subsystem ($20 \cdot \log_{10}(18) = 25$ dB). The decorrelation factor (see Eq. 2.8) for a target located at the standoff distance of 8 m is presented with the decorrelated phase noise of both chirp sources.

Fig. 2.22 shows the sphere noise floor measurement (black line) with the developed chirp sources and the noise floor estimation (blue line). This noise floor estimation is obtained by considering the phase noise results at 300 GHz of Fig. 2.21 integrated in the detection bandwidth of FFT processing ($10 \cdot \log_{10}(\frac{\alpha_{noise}}{T_m}) = 10 \cdot \log_{10}(\frac{1.36}{250\mu s}) = 37.4$ dB) and 3 dB due to the phase noise addition of the two input signals of the SHM [Coo11]. Fig. 2.22 shows a good agreement between the measured noise floor (black line) and the phase noise floor estimation (blue line) far from the beat frequency, demonstrating that the SNR limitation is due to the transmitted phase noise.

From the sphere reflected power and its noise floor measurement (black line), the SNR results are 55 dB and 41 dB for the broadband DDS and narrowband DDS/PLL architectures, respectively. The noise power is estimated with the noise bins that correspond to the noise flat region without considering the spurious signals.

Figure 2.21: Phase noise estimation at 300 GHz based on the measured phase noise at the chirp sources output (Fig. 2.15).

Figure 2.22: Sphere noise floor measurement (black line) and noise floor estimation (blue line) based on the phase noise measurement of Fig. 2.21.

2.3.4.1.2 Spurious signals

The nearest spurious signals to the target are clearly identified in Fig. 2.22a for the broadband DDS architecture. They are caused by the spurious-signal content of the transmitted signal (see Fig. 2.14). The spurious-signal level with respect to the reflected power of the target is -35 dBc, that is 6 dB worse than the SFDR of Fig. 2.13a for the broadband DDS. As regards the narrowband DDS/PLL architecture where the SFDR obtained in the previous section was -61 dBc (see Fig. 2.13b), the spurious signals are masked by the decorrelated phase noise and they cannot be observed in the radar measurement[†].

2.3.4.1.3 RF non-linearities

The measurement results of Fig. 2.22 demonstrate that the target response close to the beat frequency is dominated by the RF non-linearities, which cause the degradation of SLL to -7 dBc in both cases. The main contribution to this degradation is the non-flat efficiency of the last tripler of the transmit chain and mismatching between components [Men13b]. This effect can be partially compensated by the aforementioned calibration procedure described in [Kno93].

2.3.4.2 Two-targets measurements

Whereas the SNR measurement allows to observe the influence of the spurious signals and the phase noise limitation of the signal generation architectures, the two-targets measurements are not influenced by the transmitted phase noise due to the decorrelation effect. Furthermore, the RF non-linearities whose effect appears close to the beat frequency are partially compensated by the calibration procedure. Thus, the two-targets measurements are only carried out to verify the proper operation of the imaging radar with both architectures.

2.3.4.2.1 ΔR

The range resolution of the radar is measured by adjusting the distance between the two targets to the theoretical range resolution considering the Hamming windowing function (using (2.2) is $\frac{3 \times 10^8}{2 \cdot 27 \times 10^9} \times 1.81 = 1.01$ cm). The angle of incidence of the second target is selected to achieve the same received power for the two targets (see Fig. 2.18). The results of Fig. 2.23 confirm the expected range resolution of 1.01 cm due to the use of the calibration procedure. Noncoherent averaging has been performed so as to reduce the fluctuations of the local minimum at -6 dB point.

2.3.4.2.2 IDR

The instantaneous dynamic range is obtained by modifying the angle of incidence on the second target separated 1.6 cm from the first target (see experimental set-up of Fig. 2.19). Fig. 2.24 shows the power variation of the aluminium plate for several angles of incidence (1° , 1.5° , 2° and 2.5°). The results show that the second target can be distinguished with an IDR higher than 30 dB in both cases.

[†]If the detection bandwidth of FFT processing is reduced by increasing the chirp period beyond 20 ms, some spuri appear at a relative power level of -60 dBc.

Figure 2.23: Range resolution measurement using the developed chirp sources.

(a) Broadband DDS architecture

(b) Narrowband DDS/PLL architecture

Figure 2.24: IDR measurement by adjusting the angle of incidence on the second target to four different angles: 1° , 1.5° , 2° and 2.5° . The second target is separated 1.6 cm (11.5 kHz of beat frequency variation) from the first target. The IDR is higher than 30 dB with both chirp sources.

A summary of the results of Sections 2.3.2 and 2.3.4 is presented in Table 2.5. Some conclusions can be extracted from the chirp sources and the radar performance characterization:

- The SNR results are higher than those obtained with similar imaging radars: 40 dB in [Coo11] and 45 dB in [Rob12], mainly with the broadband DDS architecture, that is 14 dB higher than the measured SNR using the narrowband DDS/PLL architecture.
- The nearest spurious signals to the target are visible at -35 dBc using the broadband DDS architecture due to the high spurious-signal content of the transmitted signal. The spurious signals generated by the narrowband DDS/PLL architecture are masked by the transmitted phase noise and, therefore, are lower than -41 dBc.
- The RF non-linearities are dominated by the millimeter-wave subsystem and not by the chirp sources. The calibration procedure described in [Kno93] allows to get a range resolution of 1.01 cm and an IDR better than 30 dB.

Table 2.5: Summary of the results for the developed chirp sources.

		Broadband DDS	Narrowband DDS/PLL
Chirp sources characterization @ 16.65 GHz	Cost	3,250 €	1,950 €
	Power Consumption	25 W	15.1 W
	Center frequency (f_0)		16.65 GHz
	Bandwidth		1.5 GHz
	Output power		13 dBm
	Power ripple		1.5 dB
	SFDR	-41 dBc	-61 dBc
	1 kHz	-67 dBc/Hz	-59 dBc/Hz
	Phase noise at f_0		
	10 kHz	-72 dBc/Hz	-64 dBc/Hz
	100 kHz	-93 dBc/Hz	-85 dBc/Hz
	1 MHz	-122 dBc/Hz	-105 dBc/Hz
Radar performance characterization @ 300 GHz	10 MHz	-135 dBc/Hz	-124 dBc/Hz
	SLL		-33 dBc
	Center frequency		299.7 GHz
	Bandwidth		27 GHz
	SLL		-7 dBc
	Spurious level	-35 dBc	$\leq$ -41 dBc
	SNR	55 dB	41 dB
	IDR		$\geq$ 30 dB
	ΔR		1.01 cm

2.3.5 Imaging performance

This section deals with the analysis of the impact of the signal generation architectures on the final application: 3D imaging radar. It is based on the radar images of a mannequin with concealed threats under clothing. The mannequin is presented in Fig. 2.25 and it is located at the standoff distance of 8 m. After the IF signal processing explained in Section 2.1, the next steps are carried out to form the data set of the radar images [Gra15]:

- Each FFT of a chirp period is associated to its (x_i, y_i) coordinates in the target plane.
- A range gating is established around the standoff distance.
- A noise power threshold is defined for fixing the false alarm rate.

After this pre-processing, a peak finder algorithm is applied to every FFT to obtain two sets of 3D points corresponding to the external and internal layer of detections. It allows to respectively identify both the visible and hidden objects at the scene. The external layer is generated by taking into account only the first detection (the one with the lowest beat frequency), while the internal layer represents only the last detection (the one with the highest beat frequency). Lastly, the alpha-hulls algorithm [Ede94] is used to reconstruct the target surfaces.

Figure 2.25: Mannequin with threats under clothing.

Fig. 2.26 shows the radar image results of the external and internal layers of a mannequin with threats hidden under a T-shirt. As regards the radar image quality, it is appropriate to set out the following considerations:

- The range resolution determines our capability to separate the different layers of detected targets.
- Radar images are a collection of points in the 3D space (x,y,z) with enough reflected power to be above the noise power threshold. The color map of the images is not related to the reflected power due to the specular reflection effect. As indicated before, the radar images are represented as a function of the range dimension in order to enhance the contrast of the radar image [Coo08b].
- It can be noted that the radar is capable of detecting anomalies under the T-shirt with both chirp sources in a similar manner, and therefore an SNR of 40 dB is enough to obtain good-quality radar images.

Given that both signal generation architectures achieve this SNR and the radar image quality is similar, the selection of the appropriate architecture for CW-LFM imaging radars is guided by a cost effective criterion. Thus, the narrowband DDS/PLL is the optimum scheme for this specific application based on the three figures of merit evaluated in this section: radar image quality, power consumption and cost. A fair comparison among both chirp sources would require the selection of a metric regarding image quality. Due to the lack of this requirement, a subjective quality has to be used.

Fig. 2.27 shows the measured SNR as a function of the maximum reflected power per pixel in a collection of pixels obtained from the image results of Fig. 2.26 using the developed chirp sources. These results prove that the noise power floor with the narrowband DDS/PLL architecture is dominated by the transmitted phase noise when the maximum reflected power is higher than roughly -56 dBm, whilst in the broadband DDS architecture it is limited for reflected powers higher than -45 dBm. These SNR results are consistent with the SNR measurement of both architectures presented in Section 2.3.4, where a SNR increase (ΔSNR) of 14 dB is obtained with the broadband DDS architecture.

Broadband DDS

(a) External layer.

(b) Internal layer.

Narrowband DDS/PLL

(c) External layer.

(d) Internal layer.

Figure 2.26: Radar image results of the external and internal layers of a mannequin with explosives hidden under a T-shirt using the developed chirp sources. The colorbar represents the range axis in meters.

Figure 2.27: SNR results as a function of the maximum reflected power per pixel for all pixels obtained from a mannequin with hidden threats of Fig. 2.26 using the developed chirp sources.

2.3.6 Chirp sources design optimization

Having established the performance impact of the developed chirp sources on the radar image quality, two further steps towards the radar optimization are proposed in this section:

1. Power consumption and cost optimization of the chirp sources without affecting the radar performance.
2. Improvement of the radar performance by reducing the output phase noise of both chirps sources.

2.3.6.1 Power consumption and cost optimization

The 10 MHz master oscillator from Oscilloquartz can be replaced by one from Vectron (C2310 model) that presents a phase noise that is 10-15 dB worse in the range of 100 Hz to 10 MHz. The output phase noise results (test point E) at the chirp sources output with the Vectron oscillator becomes worse up to 30 kHz, as shown in Fig. 2.28, and, therefore, it does not affect the SNR result due to decorrelation effect. Additionally, the measurement results of Fig. 2.22 have not changed in terms of spurious signals, noise floor and RF non-linearities.

The radar image results of the internal layer using the Vectron oscillator with both chirp sources are presented in Fig. 2.29 to corroborate that the radar image quality is similar to that obtained with the Oscilloquartz oscillator. These results indicate that the radar will present the same performance with a large reduction of power consumption and cost as regards the Oscilloquartz oscillator (see Table 2.4), taking into account that the power consumption and cost of the Vectron oscillator are 0.01 W and 130 €, respectively. Therefore, a reduction of 75% in power consumption and cost is achieved with the narrowband DDS/PLL architecture using the Vectron oscillator as regards the initial configuration of the imaging radar based on the broadband DDS/PLL architecture with the Oscilloquartz oscillator.

2.3.6.2 Radar performance optimization

From the point of view of the improvement of the radar performance, the RF components that limit the output phase noise of the chirp sources should be replaced by others with better performance. In the case of the broadband DDS architecture, the 2.4 GHz PLO dominates the output phase noise, as indicated in Section 2.3.2. It should be replaced by a PLO from Nexyn with better phase noise performance allowing to improve the output phase noise, in spite of the increase of the cost (~ 2500 €) and the power consumption (0.4 W). On the other hand, the phase noise improvement of the narrowband DDS/PLL architecture should be carried out by replacing the frequency divider with a Hittite programmable divider and the same PLO from Nexyn, both with better performance. Fig. 2.30 shows the simulated phase noise at 300 GHz using Advanced Design System (ADS) from Keysight Technologies (formerly Agilent Technologies).

(a) Broadband DDS architecture

(b) Narrowband DDS/PLL architecture

Figure 2.28: Phase noise estimates at 300 GHz based on the phase noise measurement at the chirp sources output (test point E, 16 GHz) by replacing the master oscillator before and after decorrelation effect in the mixer.

Figure 2.29: Radar image results of the internal layer of a mannequin with explosives hidden under a T-shirt using the developed chirp sources with the Vectron oscillator.

(a) Broadband DDS architecture

(b) Narrowband DDS/PLL architecture

Figure 2.30: Phase noise estimates at 300 GHz based on the phase noise simulation at the chirp sources output (test point E, 16 GHz) of both chirp sources before and after decorrelation effect in the mixer.

2.4 Phase stability analysis

As indicated in Section 2.3.5, the information we get after radar signal processing is a complex number for each pixel. For imaging formation purpose, we have used until now only the information in the amplitude. On the other hand, phase information could be useful at the millimeter-wave band to measure relative small range displacements with precision in the order of tenths of microns, which are of special interest to certain applications, such as infrastructure monitoring [Abr11, Igl14, Mar15b], e.g. building or bridge deformations, nondestructive inspection (NDI) to detected surface-breaking fatigue cracks of materials [Zou08, Kha09] and biomedical applications such as contact-less monitoring of vital constants and blood pulse signatures of humans [Mar16]. In contrast to Eq. (2.2), the range displacement measurement is not proportional to the transmitted bandwidth, but only depends on the operating frequency, since it is based on a relative phase drift $\Delta\phi$ following Eq. (2.9):

$$\Delta d = \frac{\Delta\phi \cdot c}{4\pi f_0} \quad (2.9)$$

where Δd is the range displacement, c is the free-space propagation velocity and f_0 is the operating center frequency.

Phase stability must be guaranteed in order to make micrometer precision at the millimeter-wave band achievable. Phase stability can be defined as the phase drift $\Delta\phi$ over time with a certain bounds of the returned phase from a completely static target at a certain range. Fig. 2.31 shows the relative displacements precision as a function of the phase drift and the operating center frequency. For instance, a micrometer precision at 300 GHz can be obtained if the phase stability is between 0.005-0.05 rad.

Figure 2.31: Relative displacements precision as a function of the phase drift and the operating center frequency.

Phase stability is constrained at the millimeter-wave band due to three physical realities:

- Millimeter-wave hardware. Commonly, the transmit and receive chains of millimeter-wave radar systems are not under controlled-temperature conditions. Therefore, their performance, and most specially of their active devices, varies over time. Furthermore, a master reference clock is required to achieve a high degree of phase stability of the

transmitted signal. It should be noted that the phase stability of the master reference clock will be degraded at the millimeter-wave band due to the frequency multiplication stages.

- **Atmospheric conditions.** The environmental variability of the atmosphere, in terms of pressure, relative humidity and temperature, particularly at outdoor conditions, modifies the phase of the radar signal path over time due to air permittivity changes, which have a strong impact although there are only a few meters of distance from the radar to the target. It has been studied for indoor conditions, in particular for compact antenna test range (CATR) facilities at the millimeter-wave band [Muo12], and outdoor conditions, such as in [Mar15b] for W-band high resolution imaging and interferometry.
- **Vibrations.** The signal reflected from a detected target is extremely sensitive to target vibrations of the measurement scenario [Che11, Xia16]. Therefore, the phase stability measurement is affected by the variation in range introduced by these vibrations. In order to properly measure the phase stability, the used targets should be firmly fixed to the floor and the measurement scenario should be located in the ground floor of the building, in order to mitigate the influence of building vibrations.

To overcome these limitations, regarding the millimeter-wave hardware, the components temperature can be stabilized by using a thermal camera, or by using fans that are controlled by a thermometer in a more cost-efficient manner [Mar16]. This reduces notably the phase drift due to the hardware temperature instability. Regarding atmospheric conditions, the International Telecommunication Union - Radiocommunication Sector (ITU-R) recommendation P.453 [ITU 15] is usually applied to compensate the phase drift. The corresponding ITU model is shown below.

The atmospheric refractive index n can be computed as:

$$n = 1 + N \times 10^{-6} \quad (2.10)$$

where N is the radio refractivity, that takes into account the dry and wet terms:

$$N = N_{dry} + N_{wet} = \left(77.6 \frac{P}{T}\right) + \left(-5.6 \frac{e}{T} + 3.75 \times 10^5 \frac{e}{T^2}\right) \quad (2.11)$$

where P is the total atmospheric pressure in hPa, T is the absolute temperature in Kelvin and e is the water vapor pressure in hPa. e can be directly related to the relative humidity RH , since it is the standard parameter extracted from weather stations:

$$e = \frac{RH}{100} EF_{water} \cdot a_x \cdot e^{\frac{t(b_x - t/d_x)}{t + c_x}} \quad (2.12)$$

where EF_{water} is a variable related to the pressure P and temperature t (in Celsius degrees) for liquid water particles valid between -40°C and 50°C , see Eq. (2.13), and a_x - d_x are coefficients that are summarized in Table 2.6.

$$EF_{water} = 1 + 10^{-4}[7.2 + P(0.0032 + 5.9 \cdot 10^{-7} \cdot t^2)] \quad (2.13)$$

The atmospheric radio refractive index n can be computed by using a weather station that provides pressure, temperature and relative humidity measurements. Once computed, the resulting phase drift should be subtracted from the phase radar measurement estimated by using Eq. (2.9). An example of this phase drift as a function of temperature and relative

Table 2.6: ITU-R P.453 [ITU 15] coefficients for liquid water particles.

Coefficient	Value
a_x	6.1121
b_x	18.678
c_x	257.14
d_x	234.5

Figure 2.32: Phase drift $\Delta\phi$ as a function of relative humidity and temperature by using the ITU model [ITU 15], considering a Δd of 8 meters and an operating frequency of 94 GHz (a) and 300 GHz (b).

humidity is shown in Fig. 2.32, considering a pressure (P) of 1030 hPa, a range separation Δd of 8 meters and an operating center frequency (f_0) of 94 GHz and 300 GHz.

Having introduced the phase stability limitations, the phase stability characterization of the imaging radar at 300 GHz with the architecture proposed in [Gra17] is carried out to evaluate its capabilities and limitations for microdeformation monitoring. To this end, the phase stability at different points of the radar electronics is analyzed in order to determine the limiting subsystem of the phase stability. The phase stability is measured both in the short term (~ 5 s) and long term (≥ 1 hour). This is mainly intended to determine the radar ability in the time domain to measure in a stable form. Specifically, the ADC captures the received signal during 5 s once every a few minutes. The chirp period is 150 μ s for all measurements presented in this section[‡]. Zero-padding is used to obtain an accurate estimate of the measured phase with larger FFTs of 65536 points instead of 8192 points, which correspond to the next power of two of the samples of the chirp-signal length, taking into account a sampling frequency of 40 MHz.

Firstly, the phase stability at the DDS output (1 GHz) and at microwave subsystem output (16 GHz) is presented in Sections 2.4.1 and 2.4.2, respectively. In both cases, the phase stability is characterized by using a delay-line test system, in the same way as in Section 2.3.2.3 (see block diagram of Fig. 2.16), based on a coaxial cable of 10.7 m in length. Therefore, the measured phase corresponds to the range bin where the maximum amplitude is observed in the range profile. Secondly, the phase stability results with the imaging radar at 300 GHz is shown in Section 2.4.3. It has been carried out by using an aluminium plate located at 8 m from the radar. After these measurements and its phase stability improvement through a thermal camera at a fixed temperature of 20°C, microdisplacements measurement with a Z-axis micropositioner is carried out to demonstrate the radar capabilities for microdeformation monitoring in Section 2.4.4.

[‡]The phase stability has been evaluated with different values of chirp period, such as 1500 μ s, but the results are similar.

2.4.1 DDS phase stability

First of all, the phase stability measurement at the DDS output (1 GHz) is carried out. As indicated before, the phase measurement corresponds to the phase of a detected target at a certain range, where the maximum power of the range profile is detected, obtained through the FFT. In this case, it matches with the cable length of 10.7 meters. The phase measurement of Fig. 2.33 shows a peak-to-peak phase drift of 0.007 radians after a warming up period of 30 minutes of the master reference clock, as indicated by the manufacturer. In the short term, the phase drift is 0.003 radians. The phase stability results are identical if the DDS is introduced into the thermal camera. Tables 2.7 and 2.8 summarizes the phase drift results after warm-up and its variance at the different points of the radar electronics, respectively.

Figure 2.33: Phase stability over time at 1 GHz.

Table 2.7: Summary of phase drift results after warm-up at the different points of the radar electronics.

Duration	Thermal camera	DDS output	16 GHz	300 GHz
Short-term	No	0.003 rad	0.08 rad	0.14 rad
	Yes		0.015 rad	0.03 rad
Long-term	No	0.007 rad	0.1 rad	2.35 rad
	Yes		0.03 rad	0.03 rad

Table 2.8: Variance results of the measured phase in the short term.

Duration	Thermal camera	DDS output	16 GHz	300 GHz
Short-term	No	$7.02 \cdot 10^{-7} \text{ rad}^2$	$6.6 \cdot 10^{-6} \text{ rad}^2$	$2.63 \cdot 10^{-4} \text{ rad}^2$
	Yes		$2.2 \cdot 10^{-6} \text{ rad}^2$	$1.55 \cdot 10^{-5} \text{ rad}^2$

2.4.2 Microwave-subsystem phase stability

In the same way as in the phase stability measurement at the DDS output, a delay-line test system is used at the microwave subsystem output. As observed in Fig. 2.34, the phase drift is 0.1 radians and 0.08 radians in the long and short term, respectively. These results are more than 10 times higher than those obtained at the DDS output. Although, theoretically, there is a direct relationship between the phase drift degradation and the multiplication factor between the DDS output and the microwave subsystem output, as was the case of the phase noise degradation, it cannot be observed in the phase stability measurements. This is due to the influence of temperature instability of components at higher frequencies.

Figure 2.34: Phase stability over time at 16 GHz without temperature control.

To improve this results, the thermal camera is used to fix the ambient temperature of the microwave subsystem. Fig. 2.35 shows the phase drift results. 0.03 radians and 0.015 radians of phase stability are obtained in the long and short terms, respectively. It is observed that the temperature control improves the phase stability of the microwave subsystem. A detailed study of the main limiting component of the phase stability of the millimeter-wave subsystem was carried out. To this end, each active component of the microwave subsystem was individually tested in the thermal camera during the phase stability measurements. The main limiting component of the phase instability was the 2.4 GHz PLO from Synergy.

Figure 2.35: Phase stability over time at 16 GHz with temperature control.

2.4.3 Imaging-radar phase stability

Lastly, the phase stability measurement at 300 GHz is carried out by using an aluminium plate at 8 m of range from the radar. Without controlling the system temperature, the phase cannot be stabilized. Fig. 2.36 shows a phase drift of 2.35 radians and 0.14 radians in the long and short terms, respectively.

Figure 2.36: Phase stability over time at 300 GHz without temperature control.

By using the thermal camera (including the millimeter-wave subsystem and the power supply of the millimeter-wave components) and a fan for the MPA, the phase drift can be stabilized. As observed in Fig. 2.37, the phase drift is 0.03 radians both in the long and short term. This phase drift corresponds to a precision of $2.4 \mu\text{m}$, by using Eq. (2.9). These results are better than those obtained in [Mar16], since it provides a similar phase stability of 0.025 radians, that corresponds to a precision of $6.35 \mu\text{m}$ due to the operating frequency being three times lower approximately.

While the power drift at the DDS and microwave subsystem outputs is negligible (lower than 0.05 dB), that is not true for the 300 GHz phase stability measurements. As observed in Fig. 2.38, the received power from the aluminium plate is stabilized when temperature control is considered. By contrast, this power cannot be stabilized if the thermal camera is not used, mainly due to the MPA of the local oscillator chain. Such results were sufficient justification to control the temperature of the radar electronics.

Furthermore, there was no need to compensate the atmospheric conditions, since the 300 GHz measurements were carried out in indoor conditions, with a minimum variation of temperature, pressure and relative humidity, as shown in Fig. 2.39. The refractive index, obtained using Eq. (2.10) and the aforementioned measured parameters change slightly during the 300 GHz measurements and its impact on the measured phase is negligible.

Figure 2.37: Phase stability over time at 300 GHz with temperature control.

Figure 2.38: Power stability over time at 300 GHz.

Figure 2.39: Temperature, pressure and relative humidity measurements carried out with a weather station during the 300 GHz phase stability measurements.

2.4.4 Micrometer displacements by means of a Z-axis micropositioner

In order to show the radar capabilities of measuring micrometer displacements, a Z-axis micropositioner from *Edmund Optics* fixed to a wall is used to displace the aluminum plate located at a range of 8 meters. The travel per knob rotation is $10\text{ }\mu\text{m}$ and the maximum travel is 1 cm. Therefore, three types of relative displacements have been carried out: $10\text{ }\mu\text{m}$, $20\text{ }\mu\text{m}$, $50\text{ }\mu\text{m}$ and $100\text{ }\mu\text{m}$, in order to evaluate the capabilities of the radar for microdeformation monitoring. Each displacement has been repeated several times until reaching the maximum travel of the micropositioner. These measurements have been carried out in the corridor of the ground floor of building B of the ETSIT. A comparison of the results considering the temperature control is shown. Atmospheric conditions remain constant in the short period of the measurement.

Fig. 2.40 shows the relative phase variation $\Delta\phi$ for each displacement measurement, compared to the theoretical phase variation that is obtained from its theoretical range displacement Δd by using Eq. (2.9). The corresponding range displacement Δd is observed in Fig. 2.41 compared to the theoretical values.

Figure 2.40: Relative phase variation $\Delta\phi$ for each displacement measurement.

Thus, it might be concluded that the micrometer displacement measurement is improved by considering the temperature control, since the results are very poor without using the thermal camera. The temperature control in a cost-efficient manner of the microwave and millimeter subsystems is mandatory for this particular application. The error on micrometer displacement measurement considering temperature control is lower than $3\text{ }\mu\text{m}$, according to the precision estimate of Section 2.4.3. This should enable this radar for microdeformation monitoring. For applications that require better precision (below micrometer), an optical system is the best choice [Meg15].

Figure 2.41: Comparison among the measured and theoretical range displacement Δd for each range displacement measurement.

After this analysis, the radar is being adapted for a measurement campaign in collaboration with the Universidad Politcnica de Catalua. They are working on the microdeformation measurement of bridges and some materials such as concretes after compression strength with a 94-GHz radar. Our system provides two main advantages: the increase of the operating frequency up to 300 GHz that improves the micrometer precision of their 94-GHz radar, as indicated before, and the focusing capabilities of the scanning antenna subsystem, which allow to measure the microdeformation in a specific spot.

2.5 Impact of frequency reduction and polarimetry techniques

This section consists of the evaluation of the imaging radar performance at lower operating frequency, 100 GHz, based on two polarization radar configurations with a single antenna, based on the radar electronics of Section 2.2.2. Polarimetry can play a critical role in some radar applications like geoscience remote sensing and imaging [Ula90, Lee09, Dai09]. Therefore, linear and circular polarizations have been tested in the framework of high-resolution millimeter-wave imagery. To this end, two radar configurations have been implemented based on circular and linear polarizations through the use of an orthomode transducer (OMT) and a directional coupler, respectively, in order to replace the two horn antennas of the previous configuration see Section 2.2.

As will be shown, the proposed approach with a single antenna has a significant drawback in terms of SNR in the case of receiver thermal noise predominance, due to the low isolation between transmitter and receiver. This leakage produces a powerful short-range target which leads to the receiver saturation and, therefore, the masking of weak targets. As the interest of this radar is for short-range applications, the reduction of the transmitted power is a feasible solution to overcome this limitation. Furthermore, this approach results in an SNR compensation of 6 dB because of the removal of the beamsplitter [Lea14a]. On the other hand, the SNR limitation due to transmitted phase noise is also partially compensated due to the reduction of the operating frequency.

In order to maintain the focusing capabilities at 100 GHz of the scanning antenna subsystem, the 100 GHz antenna should be properly located as occurs with the transmit horn antenna at 300 GHz (see Fig. 2.5). To this end, the phase center[§] of the 100 GHz horn antenna is calculated in order to position the radar electronics in its accurate place to avoid the blurring of images. As will be demonstrated below, the reduction by three of the operating frequency causes the spot size enlargement in the same proportion at the same standoff range of 8 m.

2.5.1 Radar configuration description

The block diagram and the corresponding photograph of the two radar configurations are shown in Figs. 2.43-2.44. It should be noted that the millimeter-wave components are reused from the 100 GHz radar described in Section 2.2.2. Some differences are introduced in order to transform this system into a single antenna configuration, as indicated below. Datasheets of these components are presented in Appendix A.

In both cases, a variable attenuator from Quinstar is used to adjust the transmitted power to overcome LNA saturation due to the finite isolation between transmitter and receiver. The transmitted power is adjusted until -23 dBm are measured using a power meter at the LNA input in both radar configurations. The maximum input power of the LNA is -20 dBm, as indicated by the manufacturer, so a 3 dB security margin is fixed to prevent the LNA damage.

[§]The phase center of an antenna is a reference point where the radiated fields are spherical waves with ideal spherical wave fronts. These radiated fields measured on the surface of a sphere whose center coincides with the phase center have the same phase [Bal05].

Figure 2.42: Photograph of the circular-polarization radar configuration at 100 GHz on the scanning antenna subsystem.

Figure 2.43: Linear-polarization radar configuration at 100 GHz.

Figure 2.44: Circular-polarization radar configuration at 100 GHz.

2.5.1.1 Linear polarization

A directional coupler can be used as a circulator in a radar system, taking advantage of the isolation between the through and coupled ports, as shown in Fig. 2.45. In general, this is not a typical configuration because of its low isolation, but due to the lack of broadband circulators at W band, it is a viable choice [Loe09]. In this case, a 10 dB directional coupler from Radiometer Physics GmbH has been used. The directional coupler characterization is shown in Fig. 2.46. The insertion losses are lower than 1 dB, the coupling factor is 10 dB as indicated by the manufacturer, and the isolation between the input and coupled ports is 18 dB at 100 GHz.

Figure 2.45: Block diagram of a directional coupler used as a circulator in a radar system.

Figure 2.46: Directional coupler characterization.

On the other hand, due to the lack of the required waveguide straights and bends, a 1 mm coaxial cable from Astrolab and two waveguide-to-coaxial transitions from *Flann Microwave* are used to connect the coupled output of the directional coupler to the radar receiver. The insertion losses are ~ 7 dB.

2.5.1.2 Circular polarization

The circular-polarization radar configuration is based on a septum OMT duplexer designed by the *Electromagnetismo y teoría de circuitos* research group of the Technical University of Madrid [Lea14b]. It is based on the principle of circular polarized plane waves on isotropic targets, as shown in Fig. 2.47. It consists of transmitting, for instance, a right-hand circular polarization (RHCP) field, and, therefore, the reflected wave results in a left-hand circular polarization (LHCP) field, since the direction of propagation changes while maintaining the original sense of electric field rotation. In that way, the OMT duplexer is in charge of separating both orthogonal circular polarizations at its output ports. Circular polarization offers some advantages for standoff imaging by allowing differentiation of smooth features, such as the body, and sharper features present in many concealed threats, as presented in [She05]. It will be analyzed in Section 2.5.3.

As indicated in the OMT design [Lea14b], the return losses at the OMT common port strongly impact on its isolation. Therefore, a joint design with a horn antenna in the same structure is mandatory, as shown in Fig. 2.44. The horn antenna performance in terms of beamwidth is similar to the used antennas in the linear-polarization radar configuration.

Figure 2.47: Principle of operation of circular-polarization duplexer. Graph extracted from [Lea14b].

The OMT performance is shown in Fig. 2.48. At 100 GHz, the OMT presents an isolation of 30 dB. It is measured as the transmission response between the two OMT ports. The insertion loss estimate is obtained by measuring with a short circuit at the horn aperture. Therefore, the measurement results correspond to the double path insertion losses, being between 1 dB and 2 dB (see Fig. 2.48). Thereby, the insertion loss are 0.5-1 dB. On the other hand, the beamwidth is 12° and the axial ratio is below 0.8 dB.

Taking into account the LNA maximum input power (-23 dBm) and the isolation with both radar configurations, the transmitted power is 4 dBm and 8 dBm, respectively. More power is transmitted with the circular polarization due to the OMT isolation being higher than the directional coupler isolation. In any case, as will be shown below, the transmitted power with the linear polarization is enough for short-range applications, as was the case with the 300 GHz radar configuration (~ 0 dBm [Gra15]).

Figure 2.48: OMT performance results: Isolation and insertion losses.

2.5.2 Radar performance

The purpose of this section is the comparison of the two radar configurations at 100 GHz with regard to the 300 GHz imaging radar. To this end, the performance of the imaging radar at 100 GHz with both polarization configurations has been assessed by means of six parameters [Men14, Rub15]: SNR, RF nonlinearities, spot size, cross-range resolution, range resolution and specular reflection effect.

2.5.2.1 SNR

The SNR measurement at the receiver output is presented in Fig. 2.49 considering both polarizations at 100 GHz. It has been carried out using a 40-mm-diameter sphere located at 8 m. The measured SNR value is limited by the transmitted phase noise, which is clearly higher (more than 20 dB) than the receiver thermal noise. The noise power is estimated with the noise bins that correspond to the noise flat region without considering the spurious signals. The measured SNR is 65 dB and the spurious-signal level with respect to the reflected power of the target is -45 dBc. Therefore, the SNR value is ~ 10 dB better than the SNR value obtained with the 300 GHz imaging radar (see Section 2.3.4.1.1), as expected, due to the reduction of the transmitted phase noise through the operating frequency decrease by three. The reduction of the multiplication factor corresponds to $20 \log(3) = 9.5$ dB.

Figure 2.49: SNR results of the linear-and-circular polarization radar configurations at 100 GHz compared to the 300 GHz radar. The measured SNR is 65 dB and 55 dB, respectively.

2.5.2.2 RF nonlinearities

The RF nonlinearities comparison due to the non-flat amplitude response throughout the swept bandwidth of the RF circuitry is shown in Fig. 2.50. Regarding the differences between linear and circular polarizations at 100 GHz, the RF nonlinearities are similar (12 dB and 11 dB, respectively). As expected, the RF nonlinearities are lower at 100 GHz compared to 300 GHz due to the reduction of components and the transmitted bandwidth. A SLL improvement of 5-6 dB is obtained, therefore the beat signals are less distorted and the calibration procedure may not be necessary.

Figure 2.50: RF nonlinearities results of the linear-and-circular polarization radar configurations at 100 GHz compared to the 300 GHz radar. The measured SLL are 12 dB, 11 dB and 6 dB, respectively.

2.5.2.3 Spot size

Measurements have been carried out to determine the half-power beamwidth (HPBW) at 6 dB of the spot at 100 GHz with both polarizations. The experimental setup was made up of a 2.5-mm-radius metallic sphere, that has been glued to a toothpick, 1 mm in radius, and 50 mm in length, and then, stuck in an expanded polystyrene plate. Fig. 2.51 presents the results obtained. The colorbar represents the received power in dBm. Range gating has been applied to the data in order to eliminate the returns from the expanded polystyrene plate.

The measured spot size at 6 dB is 4.6 cm of width and 4.4 cm of height with the linear polarization and 4.5 cm in both dimensions with the circular polarization. As expected, the cross-range results are roughly three times higher than the measured spot size at 300 GHz in [Men14] (1.56 cm of width and 1.84 cm of height). As occurs at 300 GHz, the spots become elliptical due to the introduced astigmatic phase error when scanning far from the circular focused ring of the bifocal configuration [Gar14].

2.5.2.4 Cross-range resolution

An experiment with two metallic beads getting closer in the horizontal and vertical dimensions is presented. The 2.5-mm- radius metallic beads are spaced 4.5, 5, and 5.5 cm apart in both axes. Fig. 2.52 presents the results of this experiment with the circular polarization, since the results obtained with the linear polarization are very similar. The two beads are readily identified up to a distance of 5 cm in both axes. However, for a separation of 4.5 cm, the two beads are not properly resolved. Therefore, cross-range resolution is 5 cm in both axes, which is in good agreement with the previously obtained spot dimensions and the measured cross-range resolution at 300 GHz in [Men14] (1.5 cm).

Figure 2.51: Spot size measurement at 100 GHz with both polarizations by using a 2.5-mm radius metallic sphere. The colorbar represents the received power in dBm.

Figure 2.52: Spot size measurement with the circular polarization. Two 2.5-mm radius metallic beads are spaced 4.5, 5, and 5.5 cm apart in the horizontal and vertical axes. The cross-range resolution is 5 cm in both axes.

2.5.2.5 Range resolution

The range-resolution measurement has been assessed by means of a simple experiment. It is based on a grid of two parallel 400- μm -diameter nylon threads. The thread spacing is 3 cm approximately, that is adjusted to the theoretical range resolution considering the Hamming windowing function (using (2.2) is $\frac{3 \times 10^8}{2.9 \times 10^9} \times 1.81 = 3.03$ cm). Fig. 2.53 shows the range profile when the antenna is pointing to these threads. This experiment shows that it is possible to resolve the two threads separated 3 cm with a local minimum at -6 dB point. It should be noted that the calibration procedure is used to compensate the nonlinearities. Noncoherent averaging has been performed so as to reduce the fluctuations of the local minimum at -6 dB point.

Figure 2.53: Range resolution measurement when the antenna is pointing to two nylon threads with a measured spacing of ~ 3 cm.

A summary of the results of Section 2.5.2 and its comparison with the radar performance at 300 GHz of Section 2.3.4 and [Men14] is presented in Table 2.9.

Table 2.9: Comparison results of the radar performance at 100 GHz with both polarizations and 300 GHz.

	300 GHz	100 GHz (Linear)	100 GHz (Circular)
Bandwidth	27 GHz		9 GHz
SNR	55 dB		65 dB
SLL	6 dB	12 dB	11 dB
Spot size (width/height)	1.56/1.84 cm		4.6/4.4 cm
Cross-range resolution	1.5 cm		5 cm
ΔR	1 cm	3 cm	3 cm

2.5.2.6 Specular reflection effect

As indicated in Section 2.3.4, the received power is highly dependent on the angle of incidence due to the specular reflection effect. Fig. 2.54 shows the comparison of the received power from an aluminium plate as a function of the angle of incidence at 100 GHz compared to 300 GHz. It is observed that the received power is 10 dB higher at 100 GHz as of 2.5° than the received power at 300 GHz. Furthermore, these results are compared to those obtained with a Physics Optics (PO) simulator[¶] developed by the Universidad de Vigo for that particular scanning antenna subsystem. As observed in Fig. 2.54, the simulation results are consistent with the specular reflection measurements. In this measurement, a flat plate of $20 \times 20 \text{ cm}^2$ is used.

Figure 2.54: Received power from an aluminium plate of $20 \times 20 \text{ cm}^2$ as a function of the angle of incidence at 100 GHz with both polarizations compared to the measurements at 300 GHz.

[¶]Closed expressions of the radar cross section of a flat metallic plate as a function of the angle of incidence [Kno93, Nas16] cannot be used due to the fact that they assume that the system is working in the far field regime and they do not take into account the focusing capabilities of the scanning antenna subsystem.

2.5.3 Radar images

This section analyzes the image results at 100 GHz with both polarizations of a mannequin with hidden threats in the same way as presented in Fig. 2.25, also located at a standoff distance of 8 m. The image results with linear and circular polarizations are shown in Fig. 2.56 and Fig. 2.57, respectively.

The spot-size and range-resolution degradations due to the reduction of the operating frequency shows that the imaging radar is still capable of separating both external and internal layers but it cannot distinguish the set of three individual capacitors, as occurs at 300 GHz (see Section 2.3.5). The image results are similar with both polarizations, no major differences are observed although some mannequin parts have rounded shape with the circular polarization, as chests and explosives. This is consistent with the similar imaging comparison carried out in [She01].

Thus, for this particular application, the use of polarization techniques does not add anything to the threat detection and the loss of image quality is clear compared to the 300-GHz system.

As in Section 2.3.5 (see Fig. 2.27), Fig. 2.55 shows the measured SNR as a function of the maximum reflected power per pixel in a collection of pixels obtained from the image results of Figs. 2.56-2.57 using both radar configurations. These results prove that the noise power floor is dominated by the transmitted phase noise when the maximum reflected power is higher than roughly -20 dBm with both configurations. Therefore, it is also consistent with the SNR results of both radar configurations presented in Section 2.5.2. Furthermore, the transmitted power increase (ΔP_{tx}) of ~ 4 dB with the circular polarization owing to its high isolation is also observed in Fig. 2.55, as indicated in Section 2.5.1.

Figure 2.55: SNR results as a function of the maximum reflected power per pixel for all pixels obtained from a mannequin with hidden threats (see Figs. 2.56-2.57) with both radar configurations.

Figure 2.56: Radar image results of the internal and external layer surface reconstructions of a mannequin with explosives hidden under a T-shirt with the linear-polarization radar configuration at 100 GHz.

Figure 2.57: Radar image results of the internal and external layer surface reconstructions of a mannequin with explosives hidden under a T-shirt with the circular-polarization radar configuration at 100 GHz.

Part II

Design and development of a cloud profiling radar at 94 GHz

Chapter 3

Preliminary design of a cloud profiling radar at 94 GHz

The preliminary design of a cloud profiling radar at 94 GHz is addressed in this chapter. The main goal of this chapter is to link atmospheric science with radar engineering and on how the driving science creates technical challenges at the system level, since weather phenomena present certain features that are not common in radar detection due to their volumetric nature. First of all, the physical properties of clouds and their interaction with radar signals are described. After that, a study of the meteorological instrumentation and a survey of commercial and research cloud profiling radars at millimeter-wave band is presented. Subsequently, a preliminary design of the cloud profiling radar is carried out based on the well-known radar range equation in order to estimate the received power and the signal-to-noise ratio of detected clouds. It directly determines the system performance in terms of sensitivity and dynamic range. Therefore, a set of requirements for the design and implementation has been defined.

The technological contribution of the development of a CW-LFM cloud profiling radar at W-band is aimed at improving the radar sensitivity for longer range regarding the state-of-the-art cloud profiling radars, on the basis of two premises:

- Cost-effective radar solution to offer an affordable prototype to the research community.
- Versatility in terms of radar signal generation to enable different operating modes in order to detect several types of weather phenomena at different ranges. Furthermore, the radar prototype is designed to facilitate in the future the inclusion of polarimetric techniques to enrich the extracted information of the aforementioned meteorological processes.

3.1 Physical properties of clouds

Clouds are composed of liquid water drops, ice crystals or a mixture of both floating in the atmosphere by condensation or water-vapor deposition processes [Sau92]. Commonly, liquid clouds are located in low regions, while ice clouds are observed in medium and high layers [Fab15]. The corresponding cloud types to each region are summarized in the Introduction (see Table 1.1).

Liquid water drops can be easily approximated by spheres, but the ice crystals cannot. Liquid clouds are made of droplets up to $25\text{ }\mu\text{m}$ of average diameter. It has been reported

[Wei53, Dow90, Pru10] that the water content of clouds ranges from 1 to a $2.5 \text{ g}\cdot\text{m}^{-3}$. In the case of drizzle formation, the droplets can measure up to $300 \text{ }\mu\text{m}$ of diameter. Regarding ice clouds, crystals generally measure hundreds of microns and the water content is often less than $0.1 \text{ g}\cdot\text{m}^{-3}$. Lastly, raindrops vary in size from about 0.5 mm to as much as 8 mm in thunderstorms [Sko08].

3.1.1 Drop size distribution

The distribution of particles inside clouds is represented by the function $N(D)$, also called Drop Size Distribution (DSD), which represents the numerical concentration of particles per volume unit with a D diameter [Sau92]. This function has been modeled by log-normal, modified gamma and exponential distributions [Mil00]. For instance, the general form of the log-normal distribution is:

$$N(D) = \frac{N_T}{\sqrt{2\pi} (\ln \sigma_N) D} e^{-\frac{\ln^2(D/D_g)}{2 \ln^2(\sigma_N)}} \quad (3.1)$$

where N_T is the total number of drops per volume unit, D_g is the mean geometrical diameter and σ_N is the standard deviation of the distribution. Fig. 3.1 shows the DSD distribution results of stratocumulus observations obtained in [Mil00].

Figure 3.1: DSD distribution results of stratocumulus observations from [Mil00] modeled by the log-normal distribution of Eq. (3.1).

3.1.2 Physical moments of the drop size distribution

The DSD moment on the order n is defined by:

$$m_n = \int_0^{\infty} D^n N(D) dD \quad (3.2)$$

On this basis, the cloud parameters extraction can be obtained for the different DSD moments [Sau92]. The moment of the zero order corresponds to the total number of particles per volume unit m_0 :

$$N_T = \int_{D_{min}}^{D_{max}} N(D) dD = m_0 \quad (3.3)$$

The mean diameter can be obtained with the two first moments as follows:

$$D_g = \frac{1}{N_T} \int_{D_{min}}^{D_{max}} D \cdot N(D) dD = \frac{m_1}{m_0} \quad (3.4)$$

The liquid water content (LWC) is proportional to the third moment:

$$LWC = \frac{\pi}{6} \rho \int_{D_{min}}^{D_{max}} D^3 N(D) dD = \frac{\pi}{6} \rho \cdot m_3 \quad (3.5)$$

where ρ is the particle density. For pure water, $\rho=10^6 \text{ g}\cdot\text{m}^{-3}$ [Pru10].

The radar reflectivity factor z depends on the sixth moment:

$$z = \int_{D_{min}}^{D_{max}} D^6 N(D) dD = m_6 \quad (3.6)$$

3.1.3 Fall velocity of cloud particles

Another parameter of interest for meteorologists is the fall velocity of cloud particles, i.e. the vertical velocity to or from the ground, which is a valuable information for the DSD measurement [Kol11]. Each cloud particle has an associated velocity vector, that is affected by the action of gravity, up- and downdrafts, and turbulences, as shown in Fig. 3.2. The latter makes it difficult to measure the inherent fall velocity of particles. An explicit formulation to model this complex issue is suggested in [Pru10], where the particle velocity depends on the inertial forces that proceed from the difference between the weight of the particle and the weight of the displaced air, known as buoyancy force [Sau92]. These forces are difficult to define for all the particle dimensions that exist, particularly if the particles are not perfect spheres.

Figure 3.2: Different velocity vectors that affect cloud particles velocity: Falling velocity due to gravity action (red arrows), up- and downdrafts (green arrows) and turbulences (blue arrows).

3.2 Interaction between radar signals and clouds

In this section the different parameters that can be related to the information obtained by a cloud radar from weather targets will be analyzed.

3.2.1 Radar cross section

When a radar signal from the radar transmitter reaches a cloud, a part of the energy contained in the radiation is reflected, a part is transmitted through, a part is scattered in all directions and a part is absorbed by the cloud. From the point of view of radar engineering, the interest is in receiving the reflected and scattered radiation. This effect is quantified by the radar cross section, that is proportional to the fourth power of the operating frequency [Sau92], and can also be related to the cloud type.

The above consideration applies if the particle size is smaller than the operating wavelength ($D_g \ll \lambda$), i.e., Rayleigh region. In simplified terms, it can be assumed that the radar scanned volume is made up of spherical particles and, therefore, there are three scattering regimes of the radar cross section [Ric10], as shown in Fig. 3.3:

- Rayleigh region. It occurs when the operating wavelength is much greater than the particle size, and the radar cross section increases as the fourth power of frequency and sixth power of radius sphere [Kno93].
- Resonance or Mie region, that is originated when the operating wavelength is on the order of the particle size. In this case, the radar cross section oscillates due to surface wave effects [Kno93].
- Optics region, which is caused when the operating wavelength becomes much smaller than the particle size. In this region, the radar cross section of a sphere is independent of frequency [Kno93].

Therefore, the use of millimeter-wave band in the Rayleigh region improves radar sensitivity of conventional weather radars due to the radar cross section being proportional to the fourth power of the operating frequency, that is of interest for non-precipitating clouds detection owing to their low reflectivity. For example, the radar cross section increases 60 dB if the operating frequency is 94 GHz, compared to a conventional weather radar at 3 GHz [Kol07]. Thus, millimeter cloud radars are commonly used to complement the sensitivity performance of conventional weather radars.

On the other hand, in case of drizzle or rain, the millimeter-wave system is working in the Mie region, instead of Rayleigh region, where radar cross section of the particles is not proportional to the fourth power of the frequency. Therefore, the radar system does not provide useful information about the weather phenomenon which is detecting. An example of this situation is shown in Fig. 3.4, where the radar cross section depends on the diameter of a spherical particle. It can be observed from less than 1 mm of particle diameter, a 94 GHz radar does not work in the Rayleigh region, while a 3 GHz radar extends its Rayleigh region up to a few centimeters.

Figure 3.3: Sphere scattering from different regimes: Rayleigh, resonance and optics regions, where a is the radius of the sphere and the sphere radar cross section σ has been normalized to the projected area of the sphere, πa^2 , plotted as a function of sphere circumference normalized to wavelength, $ka = 2\pi a/\lambda$. Graph extracted from [Ric10].

Figure 3.4: Radar cross section as a function of the particle diameter. Graph extracted from [Kol07].

3.2.2 Radar reflectivity

As indicated before, the radar cross section in the Rayleigh region is proportional to the fourth power of the operating frequency. The radar cross section depends directly on the radar reflectivity (η , in m^{-1}), of the scanned volume [Sko08]:

$$\eta = \frac{\pi^5}{\lambda^4} |K|^2 \sum_{i=1}^N D_i^6 = \frac{\pi^5}{\lambda^4} |K|^2 z \quad (3.7)$$

where D_i is the diameter of i th droplet, N is the number of droplets per unit volume, λ is the operating wavelength, $|K|^2$ is a physical constant related to the properties of water and z is the radar reflectivity factor, presented in Section 3.1.2. The conventional unit of z is mm^6/m^3 since it is common to use volume density expressions in radar meteorology. At 94 GHz, $|K|^2 = 0.813$ for the water phase and $|K|^2 = 0.173$ for the ice phase [Lhe87].

For targets distributed in a volume, as is the case for cloud particles, the radar cross section σ and the volume V are related as follows

$$\sigma = \eta \cdot V \quad (3.8)$$

$$V = \frac{\pi \phi \theta r^2 \Delta R}{4} \quad (3.9)$$

where r is the range to the target, ϕ and θ are the azimuth and elevation antenna beamwidths and ΔR is the range resolution. The volume shape is a truncated cone (see Fig. 3.5).

Figure 3.5: Diagram of the radar cross section of a volume V that is illuminated by a radar.

In Section 3.5, the radar equation is presented in order to relate the radar cross section σ and radar reflectivity η with the received power by the radar. This radar reflectivity factor z can be directly related to the cloud type, and it is considered more appropriate to represent the radar reflectivity factor in the logarithmic form dBZ [Sko08]. Table 3.1 shows the correspondence between the dBZ value and the cloud type. Fig. 3.6 shows a photograph of each cloud type in order to facilitate the Table 3.1 interpretation. At 94 GHz, the particles of the two first cloud types (up to -20 dBZ) are in the Rayleigh region. For higher reflectivity factors, the measured reflectivity factor value at the millimeter-wave band is commonly denoted by dBZe , and it is called effective reflectivity factor [Sko08], since the radar reflectivity expression of Eq. (3.7) is not valid.

Table 3.1: Interpretation of the dBZ value of cloud types extracted from [Kol07].

Reflectivity factor	Interpretation
< -50 dBZ	Cumulus ¹ , alto-cumulus and thin cirrus ³
(-40,-20) dBZ	Stratus ² , alto-stratus and thick cirrus
(-20,0) dBZe	Drizzle and precipitating cirrus
(0,10) dBZe	Heavy drizzle
(10,30) dBZe	Stratiform precipitation
(30,40) dBZe	Convective precipitation

¹ Cumulus are a class of clouds of low altitude, about 2 km, that are characterized by dense individual elements in the form of puffs, mounds, or towers, with flat bases and tops that often resemble cauliflower: as such clouds develop vertically, they form cumulonimbus.

² Stratus clouds are characterized by a generally uniform gray sheet or layer, of medium altitude, about 2.5-6 km.

³ Cirrus are a class of clouds composed of ice crystals typically found at heights greater than 6 km in the form of white, delicate filaments, mostly white patches or narrow bands.

Figure 3.6: Photograph of the different cloud types defined in Table 3.1.

3.2.3 Doppler velocity

As is well-known, Doppler radars measure the radial velocity of detected targets, which corresponds to the target velocity toward or away from the radar [Sau92]. This velocity is of interest for meteorologists, as indicated in Section 3.1.3, because it can be related to the cloud dynamics and its particles size [Kol11].

Firstly, in a simplified manner, the sole presence of liquid particles in an ideal cloud can be considered in order to model the Doppler spectrum. In this case, an unimodal Doppler spectrum is obtained from cloud particles falling at a falling velocity v_f [Luk13]. Thus, an ideal quiet air Doppler spectrum $S_Q(v_f)$ in $\text{mm}^6/\text{m}^3\text{s}^{-1}$ can be modeled using the following formula proposed in [Kol11]:

$$S_Q(v_f) = \frac{\lambda^4}{\pi^5 |K|^2} N(D) \sigma \frac{\partial D}{\partial v_f} \quad (3.10)$$

If the air motion, i.e. up- and downdrafts, and turbulences are considered, the Doppler spectrum can be rewritten as proposed again in [Kol11], obtaining a multimodal Doppler spectrum. The air motion causes a simple translation of the Doppler spectrum w_{air} in the velocity domain, while turbulences broaden the Doppler spectrum [Gos97]. Turbulences are parameterized by the convolution of a Gaussian function $g(V_f)$ having a width σ_t . The convolution operation on the quiet air Doppler spectrum and turbulence function is represented by the following expression:

$$S_t(v_f) = [S_Q(v_f) * g(v_f)] = \frac{1}{\sqrt{2\pi}\sigma_t} \int_{-v_N}^{v_N} S_Q(v) \cdot e^{-\frac{(v_f-v)^2}{2\sigma_t^2}} dv \quad (3.11)$$

$$S_t(v_f) = S_w(v_f + w_{air}) \quad (3.12)$$

where $S_t(v_f)$ is the radar Doppler spectrum resulting from the convolution of the turbulence function $g(v_f)$ with the quiet air Doppler spectrum $S_Q(v_f)$, with the integral covering the velocities v_N of the Doppler spectrum that the radar is able to measure, and $S_w(v_f)$ is the radar Doppler spectrum after the vertical air motion w_{air} velocity shift.

In the same way as in Section 3.1.2, different moments of the Doppler spectrum obtained by a cloud radar can be defined [Kol11]. The first three Doppler moments (e.g., the spectral reflectivity Z_e , the mean Doppler velocity V_D and the spectral width σ_D) are related to dynamical and microphysical parameters, as proposed in [Fri95, Gos97]. As indicated before, the mean Doppler velocity is the fall velocity of cloud particles plus the air motion contribution. These three Doppler moments are expanded to include the skewness s_D and kurtosis k_D , which correspond to the third and fourth moments of the Doppler spectrum, and are, respectively, a measure of the degree of asymmetry and of steepness exhibited by the spectrum with respect to a symmetric Gaussian distribution. Skewness and kurtosis are a sensitive indicator of early drizzle onset rather than radar reflectivity or Doppler velocity, which are sensitive only when the drizzle contribution to the total observed reflectivity is significantly larger than the cloud contribution [Kol11].

Regarding radar Doppler processing, the radar is in charge of obtaining the Doppler frequency f_d that can be related to the falling velocity v_f as follows:

$$f_d = \pm 2 \frac{v_f}{\lambda} \quad (3.13)$$

By convention, this falling velocity v_f is negative for targets moving away from the radar and positive for targets approaching the radar [Sko08]. The Doppler frequency f_d is

measured by performing a spectral analysis of the received signal. The spectral analysis is usually performed by transmitting a sequence of M successive chirp signals, performing an FFT for each range bin. In the absence of windowing, the Doppler resolution is PRF/M [Ric10].

The range and velocity ambiguities, also called *Doppler Dilemma*, should be taken into account in the radar design [Ric10]. Regarding the range ambiguity, if the PRF is too high, the received signal from a far target may return after the radar transmitter has emitted another chirp signal, thereby appearing as a close target. Thus, to prevent range ambiguities, the following condition of maximum unambiguous range R_{mur} must be satisfied:

$$\frac{1}{PRF} \geq \frac{2 \cdot R_{mur}}{c} \Rightarrow R_{mur} \leq \frac{c}{2 \cdot PRF} \quad (3.14)$$

From Eq. (3.14), lowering the PRF will increase the maximum unambiguous range. However, lowering the PRF also has a negative consequence on the Doppler measurement. In a CW-LFM radar, the Doppler frequency is sampled at the PRF [Ric10]. Therefore, the maximum unambiguous Doppler frequency $f_{d_{muv}}$ and velocity $v_{f_{muv}}$ (using Eq. 3.13) that can be unambiguously measured is:

$$f_{d_{muv}} = \pm \frac{PRF}{2} \Rightarrow v_{f_{muv}} = \pm \frac{\lambda \cdot PRF}{4} \quad (3.15)$$

Maximizing unambiguous range leads to lower PRFs. However, maximizing unambiguous Doppler frequency leads to higher PRFs. Furthermore, the radar ability to measure an unambiguous range and Doppler frequency only depends on the operating wavelength:

$$v_{f_{muv}} \cdot R_{mur} = \frac{c\lambda}{8} \quad (3.16)$$

A solution to mitigate this limitation is to operate at multiple PRFs, collecting reflectivity information at low PRFs and velocity information at high PRFs [Sko08].

3.2.4 Liquid water content

The liquid water content of clouds is an important parameter for the research community [Kol11, Luk13, Kol14] because it can be related to the third moment of the DSD.

Early studies deduced theoretical relationships between radar reflectivity and LWC derived from the detected clouds. These empirically derived relationships are obtained during aircraft-mounted in-situ probe observations. These z-LWC relations have been calculated as regressions:

$$z = a \cdot LWC^b \quad (3.17)$$

where a and b are the fitting coefficients. The radar reflectivity factor z is in mm^6/m^3 and the LWC is in g/m^3 . A few z-LWC relations have been published in literature, but all of them are useful only in absence of drizzle (below -20 dBZ, as indicated in Table 3.1):

1. The first z-LWC theoretical relationship was published by [Atl54], proposing $a = 0.048$ and $b = 2$.
2. [Sau87, Fox97] propose two alternative measurements based on a the COCAGNE 3 campaign measurement in France: $a = 0.03$ and $b = 1.31$, and $a = 0.012$ and $b = 1.16$, respectively.

Recent works regarding z-LWC relationships have determined that there is a direct but nonlinear relationship between LWC and radar reflectivity. In particular, there are specific ranges for the different life cycle stages of clouds:

1. [Hag03, Kra05]. They propose three different relationships for three cloud states using a synergistic radar-lidar technique: No drizzle, light drizzle and heavy drizzle.
2. [Mai12]. The z-LWC relationship could be specified for all four fog development stages: Ground fog, formation stage, mature fog and dissipation stage. For a more precise relationship between these two key figures, a fog differentiation by type and evolutionary stage is mandatory.

Neither of these relationships are capable of obtaining the LWC with precision, as shown in the scatter diagram obtained during the experiment ASTEX [Ovt99] of Fig. 3.7. Therefore, there is no one-to-one relationship between both magnitudes. They are indirectly related to the third and six moments of the DSD, as indicated in Section 3.1.2. Slight variations in the DSD cause very large variations on the z-LWC relationship [Mai12, Kol14]. To overcome this limitation, the combination of different instruments such as cloud radars with lidars or passive radiometers improves the precision of the LWC estimation [Lhn01].

Figure 3.7: Scatter diagram of radar reflectivity factor z versus LWC for measured spectrum obtained during the experiment ASTEX [Ovt99].

3.2.5 Atmospheric attenuation

As indicated in the Introduction, the atmospheric attenuation due to the presence of oxygen and water vapor is the limiting factor of remote-sensing millimeter-wave systems. It is, therefore, essential to take into account the attenuation effect on the radar signal in order to obtain the accurate value of reflectivity of detected clouds.

The ITU-R P.676-10 recommendation [ITU 13] proposes an approximate estimate of specific attenuation at frequencies up to 350 GHz. Fig. 3.8 shows the atmospheric attenuation

at standard and dry conditions. The water vapor concentration is 7.5 g/m^3 and 0 g/m^3 , respectively. The total pressure is 1013 hPa and the temperature is 20°C .

Figure 3.8: Atmospheric attenuation at standard and dry conditions. The water vapor concentration is 7.5 g/m^3 and 0 g/m^3 , respectively. The total pressure is 1013 hPa and the temperature is 20°C .

From the ideal point of view of the radar detection, it is indispensable to obtain the atmospheric attenuation at each range measured by the radar, in order to take into account the total attenuation in the radar signal path. To this end, the pressure, temperature and water vapor concentration at each range should be measured periodically. In Section 3.5.1.1, the procedure to compute the atmospheric attenuation in the radar measurements will be outlined.

3.2.6 Radar polarimetry

Finally, it can be noted that polarimetry techniques add the ability to estimate additional information of meteorological targets. The sequential transmission of the two linear polarizations (horizontal and vertical), and the use of a dual receiver allows to extract up to six meteorological variables related to a scanned volume, which are listed below [Cao12].

- Reflectivity factor of the horizontal polarization z_{HH} .
- Differential reflectivity z_{DR} , which is the ratio of received powers of each polarization ($P_{rx_{HH}}/P_{rx_{VV}}$).
- Correlation coefficient $\rho_{HV} \cdot e^{j\delta}$ between the echoes of the co-polar signals of horizontal and vertical polarizations. The phase of correlation coefficient is called differential phase δ .
- Linear depolarization ratio LDR , defined by the ratio of the cross-polar and co-polar received powers ($P_{rx_{VH}}/P_{rx_{VV}}$).
- Complex correlations between co-polarized and cross-polarized echoes, $E(V_{HH}^* V_{HV})$ and $E(V_{VV}^* V_{VH})$, where V is the measure of the received signal for each polarization defined in the subscript.

Each polarimetric variable has a certain importance in terms of information extracted from the radar system [Mot07]. It significantly improves the physical characterization and classification of detected weather events both with conventional weather radars [Zrn99, Cao12] and cloud radars [Loh97]. The first four listed variables (z_{HH} , z_{DR} , ρ_{HV} and δ) extract the main properties of detected cloud particles: density, dielectric constant, shape, size and orientation [Zrn99].

From the point of view of the radar system design, the inclusion of polarimetry techniques in the radar system increases the complexity and, thus, the overall cost of the system [Sko08, Zrn99, Bri01]. A switch should be used for alternating the transmitted polarization every certain number of chirp periods through the transmitter OMT, while the dual receiver captures the two signals extracted at each receiver orthomode outputs simultaneously. An outline of this technique is shown in Fig. 3.9.

Figure 3.9: Block diagram of a full polarimetric radar system (top) and switch control signal for the synchronization of the chirp signal (bottom).

On the other hand, the current developments of duplexers based on orthomode transducers at millimeter-wave band [Lea13, Lea14a] enable cost reduction for the inclusion of polarimetry techniques in millimeter-wave radars, as indicated in Section 2.5. It allows to work in a single-antenna radar configuration and is based on transmitting and receiving in orthogonal circular polarizations, respectively, with the consequent cost saving through the hardware reduction instead of the full polarimetric scheme. This single-antenna configuration limits the detection sensitivity due to transmitter noise leaking into the receiver, since the OMT isolation is 25-30 dB [Hid06]. A recent work of Galletti *et al.* [Gal14] suggests that circular polarization offer some advantages over linear polarization, in particular in the presence of anisotropic (aligned) weather returns, that is, when azimuthal symmetry is broken, and, therefore, the dynamic range with circular polarization can be significantly larger than the dynamic range obtained with linear polarization.

3.3 Meteorological instrumentation

Multiple remote-sensing equipment is used to obtain information about the state of the atmosphere in order to provide weather forecasts and to study the climate. Each system measures one or more specific atmospheric variables, that can be merged with the results provided by other systems from different technologies, as will be shown below.

The evolution of meteorological instrumentation has a long history spanning centuries [Hoc00]. Summarizing, in the 15th century the rain gauge, the anemometer and the hygrometer were developed. The 17th century brought along the development of the barometer while the 18th century saw the development of the thermometer with the Fahrenheit and Celsius scales. Radiosondes were developed in the 19th century based on kites and meteographs, which is a recording device that measures pressure and temperature. The 20th century resulted in new remote sensing tools, such as weather radars, lidars and satellites. A photograph of existing meteorological instrumentation is shown in Fig. 3.10. A brief description of each system in terms of technology, measured variables, range and operating frequency is shown below, giving emphasis on the radar-based systems.

Figure 3.10: Meteorological instrumentation: Radiosonde, weather station, weather satellite, lidar, ceilometer, weather radar, passive radiometer, scintillometer and cloud radar.

3.3.1 Weather stations

A weather station is a ground-or-sea-based facility with different instruments for measuring atmospheric conditions, such as temperature, pressure, humidity, wind speed, precipitation and solar radiation [Oreg17]. To this end, typical weather stations have the corresponding instruments: Thermometer, barometer, hygrometer, anemometer, rain gauge and pyranometer.

3.3.2 Radiosondes

A radiosonde is a small weather station coupled with a radio transmitter at 403 MHz or 1680 MHz bands [Vai17]. The radiosonde is attached to an helium or hydrogen-filled balloon. It lifts the radiosonde to altitudes up to 35 km and drifts more than 200 km from the release point. As the radiosonde rises, sensors on the radiosonde transmit pressure, temperature and relative humidity measurements. Wind speed and direction aloft are currently obtained by tracking the position of the radiosonde in flight using the Global Positioning System (GPS).

3.3.3 Weather satellites

In general, weather satellites fall into two categories defined by their orbit class, namely, polar and geostationary [Tan14]. Weather satellites are able to collect global data for a variety of environmental applications, including weather forecasting, climate research, global sea surface temperature measurements, atmospheric soundings of temperature and humidity, ocean dynamics research, volcanic eruption and forest fire monitoring, global vegetation analysis and atmospheric pollution.

Several portions of the spectrum are commonly used to obtain weather images: microwaves, millimeter waves, visible and infrared. These satellites incorporate mainly imagers and radiometers. It should be noted that the CloudSat system incorporates a cloud profiling radar at 94 GHz [Ste03], that was launched in 2006. Lastly, the joint European Space Agency-Japan Aerospace Exploration Agency (ESA-JAXA) Earth Clouds, Aerosols and Radiation Explorer (EarthCARE) mission is scheduled for launch in 2018 and features the first atmospheric cloud profiling radar with Doppler capability in space [Kol14].

3.3.4 Lidars, ceilometers and disdrometers

Several remote systems are based on laser signals, such as lidars, ceilometers and disdrometers. The two first systems present minor differences and are often confused with each other. A ceilometer is a device that uses a single-wavelength laser to determine the height of a cloud base. As radar systems, they can measure the time between the transmit and receive signals in order to obtain the range from detected clouds. Additionally, it can also be used to measure the aerosol concentration. Lidars - and more specifically multi-wavelength Raman lidars - can provide continuous measurements of the aerosol backscatter vertical profiles up to 15.000 km [Rei14]. The laser light source used in ceilometers is less powerful and spectrally broader compared to that of a lidar system, therefore, the ability of ceilometers to detect aerosols is limited up to around 3 km [Tsa11].

Lastly, a disdrometer is a ground-based instrument used to measure the drop size distribution and velocity of falling particles at ground level. Even though there are microwave and optical disdrometers, laser technology is the most employed technology in these systems [Bal16]. They are composed of a laser source of light and a light detector facing each other, which are used to measure the signals that are generated by precipitation particles passing through the measuring area.

Their performance is strongly degraded with multiple cloud layers or overcast days due to the high attenuation of the laser signal compared to a cloud radar [Clo99].

3.3.5 Passive radiometers

As is known, passive radiometry is the measurement of the electromagnetic radiation which is emanated from all objects and surfaces, like the atmosphere. In the absence of scatter-

ing, the atmospheric attenuation can be estimated by radiometric measurements of the sky brightness temperature [Sko89]. Besides, dual-frequency radiometer measurements provide meteorological parameters like liquid and ice water content and temperature and moisture vertical profiles [Wu05]. For weather monitoring, radiometers are operated from space [Bue07] as well as from the ground [Sil13]. The main advantages of space monitoring is its high field of view compared to ground-based systems. Radiometers have receivers in frequency bands from microwaves up to 325 GHz [Sko89, Eva05, Bue07, Nor08].

3.3.6 Scintillometers

A scintillometer is a device used to measure small fluctuations of the refractive index of air caused by variations in temperature, humidity, and pressure up to 10 km, that is of interest in hydrology and radiation budget studies. Both optical and millimeter-wave scintillometers have been proposed in the literature [Bru02, Mei02, Rad14b]. A scintillometer consists of a transmitter and a receiver unit and measures intensity fluctuations. By emitting a signal through the atmosphere over a horizontal path, intensity fluctuations can be observed at the receiver. The intensity fluctuations are caused by turbulent temperature and humidity fluctuations and can be expressed as the structure parameter of the refractive index of air. It is known that in the visible to near-infrared wavelength region scintillations are primarily caused by temperature fluctuations. Humidity fluctuations also contribute at millimeter-wave band.

3.3.7 Weather radars

Weather radars are the most important instrument for monitoring the occurrence and movement of precipitation patterns and the presence of many hazardous weather conditions such as thunderstorms or tornadoes [Fab15]. The maximum range of these radars is a few of hundreds kilometers and they work on C-and-S bands.

Weather surveillance requires that the radar uses scanning capabilities to sample the atmosphere as thoroughly as possible, since hazard weather phenomena can move and evolve rapidly. To this end, a typical scanning strategy of weather radar consists of a scan in elevation and azimuth for a certain amount of time to complete a 3D volume, as shown in Fig. 3.11a. Since 2003, the United States National Oceanic and Atmospheric Administration (NOAA) has been experimenting with phased-array antennas as a replacement for conventional mechanically-scanned parabolic antennas [Hei08].

Figure 3.11: Scanning strategies of weather radars (a) and cloud radars (b).

Furthermore, polarimetry techniques are currently added to weather radars, also referred to as dual-polarization radars, to improve estimates of rainfall and to distinguish between rain, snow or hail [Zrn99].

3.3.8 Cloud radars

As indicated in Section 3.2, millimeter-wave cloud profiling radars (CPRs), also called cloud radars, seek to address the conventional-weather-radars problem of small particles detection [Kol07], such as fog or non-precipitating clouds. Given that the radar cross section is proportional to the fourth power of the operating frequency, and assuming the same transmitted power and scanned volume, the millimeter-wave radar is capable of detecting smaller particles contained in this volume, complementing the sensitivity performance of conventional weather radars.

Although some cloud profiling radars scan in azimuth and elevation [Lhe87, Wid12, MeT17], the antenna pointing CPR is usually vertical, toward the zenith, as shown in Fig. 3.11b. In order to analyze a radar-section volume as homogeneous as possible regarding the detected particles, the beamwidth of the used antennas is around $0.1\text{-}0.2^\circ$, in order to scan an area of 0.5 km^2 at 10 km of height. In addition, to compensate the atmospheric attenuation and to maintain a considerable range of 10-15 km, the gain of the antennas is considerably high, about 50-60 dB. This antenna performance is obtained mainly with Cassegrain antennas, which have a diameter of 0.8-1.2 m, and, to a lesser extent, with lens antennas at least 0.3 m in diameter. Another antenna parameter of interest in this system is the antenna sidelobe level in order to minimize the influence of ground clutter [Sko08, Kol07]. This consideration is specific to conventional weather radars, which perform scans in azimuth and elevation, and are more influenced by ground clutter when performing scans with a low elevation angles (see Fig. 3.11). Commercial weather radars have a sidelobe level of around 27-30 dB. Focusing on commercial cloud profiling radars, the sidelobe level is around 18-20 dB, making it less restrictive due to the above reasoning.

Regarding the radar electronics, due to the requirement of dynamic range of weather signals, about 70-80 dB*, the radar electronics architecture of most cloud profiling radars is heterodyne, in order to extend the receiver dynamic range [Sko08]. It increases the overall cost of the system, due to the need of extra oscillators, amplifiers and filters. Currently, cloud radar systems are implemented with an homodyne structure, with a receiver dynamic range up to 80 dB [Men13a, Rad14a]. On the other hand, a two antenna configuration is also common in these systems to improve the isolation up to 70-80 dB between the transmitter and the receiver, since ferrite-based circulators at W-band present an isolation lower than 30 dB. Another alternative is the use of a reflected power canceller to improve the low isolation of W-band circulators, allowing the design of a radar system with a single antenna, as proposed in [Hid06]. It limits the radar sensitivity [Sch93], due to the leakage power from the transmitter being stronger than the reflected power from clouds, particularly at lower ranges. Regarding the chirp period, most cloud profiling radars use a chirp period of 0.1-1 ms, which is enough for short-range cloud measurements, compared to the long-range measurements of conventional weather radars [Sko08], whose chirp period is slightly higher (around 6 ms), in order to increase the maximum unambiguous range up to 150 km. Due to the received signals from non-precipitating clouds being very weak, non-coherent integration is applied in order to improve the SNR of detected targets [Ric10]. The integration time is in the range of 2-60 s.

*Dynamic range is defined as the the power ratio between the weakest and strongest received signals that the radar receiver can handle, from close range to the maximum range of the radar. Weather signals span a dynamic range of about 70-80 dB due to the r^2 dependence and the variation of the reflectivity factor from non-precipitating clouds to drizzle (from -50 to 0 dBZ) [Sko08].

3.4 State-of-the-art of millimeter-wave cloud radars

This section deals with a performance analysis of remarkable cloud radars at millimeter-wave band. A summary of their operational parameters is shown in Table 3.2.

3.4.1 Research radars

3.4.1.1 Rutherford Appleton Laboratory (RAL)

The RAL cloud profiling radar [Hug08] is centered at a frequency of 94.02 GHz, sweeping a bandwidth of 15 MHz. It is a CW-LFM radar, the transmitted power is 23 dBm and the dynamic range is 60 dB. This radar is based on an heterodyne structure and uses two antennas to increase the transmitter/receiver isolation. The offset parabolic reflector antennas with 0.7 m diameter are pointing towards the zenith, as shown in Fig. 3.12. The gain is 55 dBi and the beamwidth is 0.15° . Only one linear polarization is transmitted.

Figure 3.12: Rutherford Appleton Laboratory (RAL) cloud profiling radar.

3.4.1.2 Chiba University (Japan)

The Chiba University cloud profiling radar [Yam06] is centered at 94.79 GHz and the transmitted bandwidth is 10 MHz. It is also a CW-LFM radar. The transmitted power is 27 dBm. Like the previous radar system, it presents a heterodyne structure and uses two antennas, as shown in Fig. 3.13. The antennas are in a conventional Cassegrain configuration of 1 m diameter pointing to the zenith. The gain is 57 dBi and the beamwidth is 0.18° . Only transmits a linear polarization.

3.4.1.3 Miami University

The first cloud profiling radar was proposed by R. Lhermitte at the University of Miami in 1987 [Lhe87]. It is a pulsed radar at 94 GHz with a peak power of 1.2 kW and a heterodyne structure with two Cassegrain antennas of 0.9 m diameter and 0.27° of beamwidth. The structure allows scanning in azimuth and elevation, as shown in Fig. 3.14.

Table 3.2: Comparison of the operational parameters of remarkable cloud profiling radars.

Parameter	MeTeK	Miami	Chiba	CloudSat	Prosensing	RAL	RPG	This work
	[MeT17]	[Lhe87]	[Yam06]	[Spi01]	[Wid12]	[Hug08]	[Rad14a]	
Frequency	35.2 GHz	94 GHz	94.7 GHz	94 GHz	94 GHz	94 GHz	94 GHz	94 GHz
Radar signal	Pulsed	Pulsed	CW-LFM	Pulsed	Pulsed	CW-LFM	CW-LFM	CW-LFM
Radar type	Ground-based	Ground-based	Ground-based	Satellite-based	Ground-based	Ground-based	Ground-based	Ground-based
Radar architecture	Heterodyne	-	Heterodyne	-	-	Heterodyne	Homodyne	Homodyne
Bandwidth/Pulse width	100-400 ns	0.1-2 μ s	10 MHz	3.3 μ s	2 μ s	15 MHz	5-150 MHz	5-2000 MHz
Range resolution	30 m	60 m	15 m	500 m	300 m	10 m	1-30 m	0.075-30 m
Transmitted power	30 kW(pk)	1 kW (pk)	27 dBm	1.8 kW (pk)	1.7 kW (pk)	23 dBm	31.8 dBm	30.2 dBm ¹⁰⁴
PRF	2.5-7 kHz	2.5-20 kHz	1 kHz	3.7-4.3 kHz	20 kHz	1.3 kHz	-	0.03-33 kHz
Noise figure	3.6 dB	6.5 dB	5.5 dB	5 dB	6 dB	-	4 dB	4.6 dB
Antenna type	Cassegrain	Cassegrain	Cassegrain	Offset	Off-axis	Off-axis	Cassegrain	Lens (Tx)
				parabolic	parabolic	parabolic		Cassegrain
				reflector	reflector	reflector		(Rx)
Antenna gain	50.4 dB	57 dB	57 dB	63.1 dB	53.5 dB	55 dB	52.3 dB	48/59 dB
Antenna beamwidth	0.6 ⁰	0.27 ⁰	0.27 ⁰	0.1 ⁰	0.33 ⁰	0.15 ⁰	0.45 ⁰	0.8/0.2 ⁰
Radar weather constant C	165 dB	172 dB	122 dB	180 dB	165 dB	111 dB	123.8 dB	124 dB
Receiver sensitivity	-100 dBm	-93 dBm	-140 dBm	-114 dBm	-110 dBm	-	-	-131.6 dBm

Figure 3.13: Chiba University cloud profiling radar.

Figure 3.14: Miami University cloud profiling radar.

3.4.1.4 CloudSat satellite

The CloudSat system [Spi01, Ste03] is a 94-GHz nadir-looking radar which was developed jointly by the National Aeronautics and Space Administration (NASA) Jet Propulsion Laboratory (JPL) and the Canadian Space Agency (CSA) under the CloudSat mission. It consists in the study of the role that clouds and aerosols play in regulating Earth weather, climate and air quality. The vertical resolution is 500 m and the along-track sampling[†] is 2 km. It is a pulsed radar with a peak power of 1.5 kW and the dynamic range is 70 dB, with a maximum range of 25 km. The antenna subsystem consists of the collimating antenna of 1.85 m diameter and 63.1 dB of gain. CloudSat orbits in formation as part of the A-Train constellation of satellites (Aqua, CloudSat, CALIPSO, PARASOL and Aura) in sun-synchronous orbit at an altitude of 705 km above the Earth.

Figure 3.15: CloudSat cloud profiling radar.

[†]Along-track sampling consists of the horizontal distance between two adjacent scan satellite measurements.

3.4.2 Commercial radars

3.4.2.1 MeTeK

One of the commercial cloud profiling radar is the MIRA-36, offered by the German company MeTeK [MeT17]. It is a pulsed radar at 36.5 GHz with a peak power of 30 kW. Working at this frequency, it slightly extends the radar operation in the Rayleigh regime of cloud particles, as indicated in Section 3.2. It presents a homodyne structure with a Cassegrain antenna of 1 m diameter, as shown in Fig. 3.16. The antenna gain is 50.4 dBi and the beamwidth is 0.6° . The sidelobe level is -20 dB. The structure allows scanning in azimuth and elevation. The antenna transmits only a linear polarization and receives two linear polarizations simultaneously.

Figure 3.16: MeTeK cloud profiling radar.

3.4.2.2 Prosensing

This cloud profiling radar is offered by Prosensing [Wid12]. This system is used by the ARM Climate Research Facility of the United States DoE. The system works at dual frequency (35 GHz and 94 GHz). The antennas are parabolic reflectors, one for each frequency, with diameters of 1.8 m and 0.9 m, respectively, with a beamwidth of 0.33° and 0.29° . The gain is 53.5 dB and 54.5 dB, respectively. Sidelobe level is -18 dB in both frequency bands. This system has the ability to scan in elevation and azimuth, as shown in Fig. 3.17. It also transmits a linear polarization and receive both linear polarizations simultaneously.

3.4.2.3 Radiometer Physics

A recent cloud profiling radar is presented by Radiometer Physics [Rad14a]. The system works at 94 GHz and presents a wide selection of the transmitted chirp signal in terms of bandwidth (5 to 150 MHz). The transmitted power is 1.5 W. It is based on a heterodyne structure with two Cassegrain antennas, with a beamwidth of 0.45° and a gain of 52.3 dB. It has an extra option to scan in elevation and azimuth.

Figure 3.17: Prosensing cloud profiling radar.

Figure 3.18: Radiometer Physics cloud profiling radar.

3.5 Radar equation for a cloud radar at 94 GHz

The basic equations required for computing the radar reflectivity factor are presented in this section. This is the essential parameter for meteorologists because it is directly related to the cloud type [Sko08].

3.5.1 Received power

Taking into account Eqs. (3.7)-(3.9), the received power at the receive antenna output of the radar from detected clouds in the Rayleigh region is obtained by means of the radar equation [Sko08] as follows:

$$P_{rx} = \frac{\beta \cdot \sigma}{r^4} \quad (3.18)$$

where r is the range to the detected cloud, σ is the radar cross section (defined in Section 3.2.2) and β is a constant that depends on the operational parameters of the radar:

$$\beta = \frac{P_t G_{tx} G_{rx} \lambda^2}{(4\pi)^3 \cdot 2 \ln 2 \cdot L_{atm}(r)} \quad (3.19)$$

where G_{tx} and G_{rx} are the transmit and receive antenna gain, respectively, and P_t is the transmitted power. $L_{atm}(r)$ is the correction for power losses from the ground radar to the cloud by specific attenuation due to atmospheric gases (mainly dry air and water vapor). It is mandatory for cloud radars at 94 GHz, since it is appreciable at these wavelengths [ITU 13] and it is range dependant. Because the antenna beam is Gaussian, a correction factor of $2 \ln(2)$ is applied [Pro62].

Taking into account Eqs. (3.7),(3.9) and (3.19), and substituting into Eq. (3.18) gives

$$P_{rx} = \frac{\pi^3 |K|^2}{512(\ln 2)} \left[\frac{P_t G_{tx} G_{rx} \phi \theta \Delta R}{\lambda^2} \right] \cdot \frac{z}{r^2 \cdot L_{atm}(r)} \quad (3.20)$$

where the factor inside brackets corresponds to C , which is known as the weather radar constant. Table 3.2 summarizes the weather radar constant of state-of-the-art cloud radars at millimeter-wave band, where the pulsed radars clearly present better performance due to their high transmitted power. The reflectivity factor z is usually expressed in logarithmic scale as dBZ :

$$dBZ = P_{rx}(dBm) + 20 \log_{10} r - C(dB) - 10 \log_{10} \left(\frac{\pi^3 |K|^2}{512(\ln 2) \cdot L_{atm}(r)} \right) \quad (3.21)$$

Equations (3.20)-(3.21) are valid if it is assured that the cloud particles are in the Rayleigh region (see Section 3.2), and the scanned volume is filled completely with these particles. Furthermore, the cloud particles should be in the far field of the antennas. As indicated in Section 3.2.2, if these considerations cannot be ensured, the measured reflectivity factor value is denoted by $dBZe$, and it is called effective reflectivity factor [Sko08].

As indicated in Section 3.3.8, most of the millimeter-wave cloud radar antennas are Cassegrain reflectors or lens. The gain and HPBW of these antennas are directly related to the antenna diameter and the operating wavelength, as indicated in [Kay66]:

$$G(dB) = 10 \cdot \log_{10} \left(\eta_{a1} \left(\frac{\pi \cdot D_a}{\lambda} \right)^2 \right) \quad (3.22)$$

$$HPBW(^{\circ}) = \eta_{a2} \frac{\lambda}{D_a} \quad (3.23)$$

where D_a is the antenna diameter, η_{a1} is the antenna normal-aperture efficiency and η_{a2} is the antenna beam efficiency. Fig 3.19 shows the gain and beamwidth of these antennas as a function of frequency and antenna diameter. In this case, typical values at millimeter-wave band of $\eta_{a1} = 0.65$ and $\eta_{a2} = 75^0$ have been considered [Kay66]. The maximum diameters of commercially available Cassegrain and lens antennas at millimeter-wave band are 1.2 m and 0.3 m, respectively [Mil16].

(a) Gain.

(b) HPBW.

Figure 3.19: Gain (a) and beamwidth (b) of Cassegrain and lens antennas as a function of frequency and antenna diameter. $\eta_{a1} = 0.65$ and $\eta_{a2} = 75^0$ have been considered.

Following Eq. (3.21), the received power as a function of the radar reflectivity factor, the transmitted power and diameters of transmit and receive antennas, D_{tx} and D_{rx} , respectively, is shown in Fig. 3.20. Three different values of radar reflectivity factor are selected: -40 dBZ, -20 dBZ and 0 dBZ. The two first values (see Table 3.1) are in the Rayleigh region for a 94 GHz cloud radar. The signal bandwidth is 15 MHz and the particles are in liquid water phase ($|K|^2 = 0.813$). Atmospheric losses are not considered. The transmitted power and antenna diameters are defined arbitrarily taking into account the typical values of remarkable cloud radars.

(a) Fixed P_{tx} , variation of D_{tx} and D_{rx} .

(b) Fixed D_{tx} and D_{rx} , variation of P_{tx} .

Figure 3.20: Received power as a function of the radar reflectivity factor, transmitted power and diameter of transmit and receive antennas, D_{tx} and D_{rx} , respectively.

Regarding the transmitted bandwidth, which is directly related to the range resolution ($\Delta R = c/(2 \cdot BW)$), the received power is reduced if the transmitted bandwidth is increased, according to Eq. (3.21). This makes sense since the scanned volume becomes smaller and therefore the number of detected particles is lower. Fig. 3.21 shows the received power variation as a function of the transmitted bandwidth, considering a transmitted power of 31 dBm and antenna diameters of 1.1 m. For example, an increment of 3 dB on the received power is obtained by modifying the transmitted bandwidth from 10 MHz to 5 MHz. Current versions of commercial cloud radars offer different operational modes as a function of the weather phenomena of interest [Rad14a]. Radar PRF and transmitted bandwidth are parameters selected by the user to change the operational modes.

Figure 3.21: Received power as a function of the transmitted bandwidth, considering a radar reflectivity factor of -20 dBZ, a transmitted power of 31 dBm and antenna diameters of 1.1 m.

3.5.1.1 Atmospheric losses

As indicated in Section 3.2.5, a correction for power losses by specific attenuation due to atmospheric gases (mainly dry air and water vapor) is mandatory in the radar equation for cloud radars at 94 GHz.

For this purpose, the ITU-R P.676-10 recommendation [ITU 13] is taken into account. This recommendation considers the individual resonance lines from oxygen and water vapor, together with small additional factors for the non-resonant Debye spectrum of oxygen, pressure-induced nitrogen attenuation above 100 GHz and a wet continuum to account for the excess water vapor absorption. For quick and approximate estimates of specific attenuation at frequencies up to 350 GHz, the simplified algorithms given in Annex 2 of the ITU-R P.676-10 recommendation are used. Fig. 3.22 shows the atmospheric attenuation at three different temperatures: -20°C, 0°C and 20°C and three different pressures: 1013 hPa, 800 hPa and 600 hPa.

In order to compute the atmospheric attenuation, three parameters, besides the operating frequency, are required by the algorithms of the ITU-R recommendation:

- Total air pressure (Ph) in *hectopascals* (hPa).
- Temperature (T) in *Kelvin*.

Figure 3.22: Atmospheric attenuation at three different temperatures: -20°C , 0°C and 20°C and three different pressures: 1013 hPa, 800 hPa and 600 hPa. The water vapor concentration is 7.5 g/m^3 .

- Water vapor concentration (WVC) in g/m^3 .

Two out of the three parameters provided by a radiosonde are directly used by the algorithms of the ITU-R recommendation to obtain the atmospheric losses: Total air pressure (Ph) and temperature (T). However, the water vapor concentration should be obtained from the relative humidity measurement that is provided by the radiosonde. To this end, a simple equation relates both parameters:

$$RH = \frac{WVC}{SVC} \cdot 100\% \quad (3.24)$$

where SVC is the saturation vapor concentration. [Ahr09] proposes an empirical relationship between saturation vapor concentration and temperature measured by the radiosonde, as shown in Fig. 3.23.

Therefore, the water vapor concentration depends on the temperature and relative humidity, both measured by the radiosonde, as shown in Fig. 3.24.

Lastly, the atmospheric attenuation is computed using the aforementioned parameters as shown in Fig. 3.25.

Due to the unavailability of radiosondes at the cloud radar location in order to ensure reliable results of atmospheric attenuation for each radar measurement, these parameters are obtained everyday from a radiosonde located in the Adolfo Suarez Madrid-Barajas Airport of Madrid (Spain), whose measurements are available on the website of the Department of Atmospheric Science of the University of Wyoming [Univ16]. This radiosonde is $\sim 13 \text{ km}$ away from the cloud radar location. As all of these devices, the range interval between measurements is not uniform, it varies from 100 m to 500 m. Therefore, several ranges of the radar measurement correspond to the same range of the radiosonde due to the different granularity. As indicated in Section 3.2.5, from the ideal point of view of the radar detection, it is indispensable to obtain the atmospheric attenuation at each range measured by the radar, which is a source of error in the radar equation. However, this approximation implies an improvement compared to assuming a constant value.

Figure 3.23: Empirical relationship between saturation vapor concentration (SVC) and temperature (T).

Figure 3.24: Water vapor concentration (WVC) as a function of the temperature (T) and relative humidity (RH).

Figure 3.25: Atmospheric attenuation as a function of the relative humidity (RH), total pressure (Ph) and temperature (T) for the radar operating frequency (94 GHz).

3.5.2 Sensitivity and dynamic range

The sensitivity of a cloud profiling radar is usually defined in the literature as the minimum reflectivity factor z detected within a specified range. For example, the Chiba University radar [Yam06] indicates a minimum reflectivity factor of -32 dBZ detected at 5 km. From the point of view of a radar engineer, this sensitivity is the minimum signal-to-noise ratio (SNR) required to get a selected detection probability once a probability of false alarm is fixed. The SNR is obtained by dividing the received power by the receiver noise power (P_n) at the receive antenna output, as follows:

$$SNR = \frac{P_{rx}}{P_n} = \frac{P_{rx}}{k \cdot T_s \cdot BW_{rx}} \quad (3.25)$$

where k is the Boltzmann constant, T_s is the equivalent noise temperature at the receiver input and BW_{rx} is the receiver bandwidth. In CW-LFM radars, BW_{rx} coincides with the PRF[‡]. Assuming no loss in the connection between antenna and receiver[§], the equivalent noise temperature T_s at the receiver input is calculated as:

$$T_s = T_a + T_0 (F - 1) \quad (3.26)$$

where F is the noise factor of the receiver, T_0 is the reference temperature (290 K) and T_a is the antenna temperature. As indicated in [ITU 01, Mar07, Pre16], due to sky noise from main individual sources such as oxygen and water vapor, the value of T_a is ~ 60 K for 94 GHz receivers with very narrow beam antennas at zenith pointing.

Some cloud profiling radars of Table 3.2 indicate their receiver noise power (P_n) or the parameters necessary to calculate it, such as the noise figure, allowing us then to analyze the signal to noise ratio of these systems. This happens for example for the University of Miami radar [Lhe87], which has a noise power of -93 dBm, or Chiba University radar [Yam06], with a noise power of -140 dBm (see Table 3.2).

In radar detection theory, at least 18 dB of SNR is required to detect the particles of the scanned volume, which is resembled to a Swerling 1 target (it consists of a number of equally large isotropic reflectors which are distributed on a volume) [Sko08]. This SNR ensures a detection probability of 0.9 and a probability of false alarm of 10^{-3} [Ric10]. This consideration will be analyzed in the sensitivity estimation of our developed cloud profiling radar in Chapter 4, in order to determine the clouds detection threshold. On the other hand, in radar meteorology, the sensitivity of a cloud radar is also defined in several articles [Lhe87, Kno07] as the reflectivity factor for an SNR of 0 dB at a determined range. Therefore, the calculation of sensitivity is direct using Eq.3.21:

$$dBZ_{min} = dBZ|_{P_{rx}=P_n} = P_n(dBm) + 20\log_{10}r - C(dB) - 10\log_{10}\left(\frac{\pi^3|K|^2}{512(\ln 2) \cdot L_{atm}(r)}\right) \quad (3.27)$$

Considering an arbitrary receiver noise figure of 4 dB, the SNR can be obtained with this value and the received power results of Fig. 3.20, following Eq. (3.25). T_0 is 290 K and T_a is 60 K. The SNR results as a function of the reflectivity factor and PRF are shown in Fig. 3.26. For this SNR estimate, the transmitted power is 31 dBm, the transmitted bandwidth is 15 MHz and the antenna diameter is 1.1 m.

[‡]It corresponds to the radar PRF in the absence of IF signal windowing. On the other hand, as indicated in Section 2.1, the equivalent noise bandwidth α of the windowing function is multiplied by the PRF value to obtain the receiver bandwidth.

[§]In the case of a waveguide section being used to connect the antenna to the receiver [Bar13], the equivalent noise temperature is $T_s = T_a + T_r + L_r \cdot T_0 (F - 1)$, where L_r and T_r are the losses and noise temperature of the waveguide section, respectively.

Figure 3.26: SNR estimates as a function of the reflectivity factor and the PRF. The transmitted power is 31 dBm, the transmitted bandwidth is 15 MHz and the antenna diameter is 1.1 m.

As shown in Fig. 3.26, as the PRF increases, the SNR gets worse, since the receiver noise power is increased. Furthermore, it limits the Doppler ambiguity measurement, see Eq. (3.15). Therefore, a suitable balance between the PRF and the maximum unambiguous Doppler velocity should be achieved. The maximum unambiguous Doppler velocity $v_{f_{muv}}$ for the three studied PRFs (1 kHz, 5 kHz and 10 kHz) at 94 GHz is ± 0.8 m/s, ± 4 m/s and ± 8 m/s, respectively.

On the other hand, it can be observed in Fig. 3.26 that the SNR results at low radar reflectivity factors present a value below 0 dB, i.e. the received power from detected clouds is lower than the receiver power noise P_n . In order to detect these particles, since each received chirp presents different phase at the cloud range due to cloud return samples being quickly decorrelated, non-coherent integration is applied to improve the SNR [Lhe87]. The SNR improvement in the worst case [Ric10] depends on the integrated chirp periods (M) as follows:

$$SNR_{out}(dB) \simeq SNR_{in}(dB) + 10 * \log_{10} \sqrt{M} \quad (3.28)$$

Fig. 3.27 shows the SNR improvement as a function of integrated chirp periods. The analyzed cloud profiling radars consider an integration of 32.000 [Lhe87] or 40.000 [Hug08] chirp periods. Therefore, the SNR improvement is 22.5 dB and 23 dB, respectively. The corresponding integration time is 5 s and 60 s, respectively. Fig. 3.28 shows the SNR results of Fig. 3.26 considering a non-coherent integration of 35.000 chirp signals.

Regarding the dynamic range, it was defined in Section 3.3.8 as the the power ratio between the weakest and strongest received signals that the radar receiver can handle, from close range to the maximum range of the radar. It is shown in Fig. 3.28 with a maximum range of 15 km, and considering a minimum reflectivity factor of -40 dBZ, the required dynamic range is 75 dB, taking into account the maximum received power at 0.15 km with a reflectivity factor of 0 dBZ. Lastly, the sensitivity is -40 dBZ at 5.1 km.

Figure 3.27: SNR improvement as a function of integrated chirp periods (M).

Figure 3.28: SNR estimates as a function of the reflectivity factor and the PRF, considering a non-coherent integration of 35,000 chirp signals. The transmitted power is 31 dBm, the transmitted bandwidth is 15 MHz and the antenna diameter is 1.1 m.

3.6 Parallax error

The difficulty in the precise alignment of the cloud radar antennas at millimeter wave band produces a high reduction of the received power from detected targets [Sek02], mainly due to non-parallel beams of the antennas, which present large dimensions and a narrow beamwidth, as shown in Fig. 3.29. This sensitivity reduction mainly affects low-range clouds that are of special interest in fog detection.

Figure 3.29: Diagram of parallax error for two-antenna radar configuration with parallel-beams (a) and non-parallel-beams (b).

Assuming Gaussian beams in the antennas, as indicated in Section 3.5, the power losses can be estimated by comparing the scanned volume with a single antenna V_{single} and a two-antenna configuration V_{dual} , known as the parallax factor $f(r)$, as proposed by [Sek02]. V_{single} models the scanned volume by a single transmit-receive antenna and V_{dual} models the overlap between the transmit and receive antenna beams. In this estimation, the influence of antenna side lobes is not considered.

$$f(r) = \frac{V_{dual}(r)}{V_{single}(r)} \quad (3.29)$$

$$V_{single}(r) = \Delta R \int_{y=-\infty}^{\infty} \int_{x=-\infty}^{\infty} \left(\exp\left(-\left(\frac{x}{\sigma_x}\right)^2\right) \exp\left(-\left(\frac{y}{\sigma_y}\right)^2\right) \right)^2 dx dy \quad (3.30)$$

$$V_{dual}(r) = \Delta R \int_{y=-\infty}^{\infty} \int_{x=-\infty}^{\infty} \left(\exp\left(-\left(\frac{x}{\sigma_x}\right)^2\right) \exp\left(-\left(\frac{y}{\sigma_y}\right)^2\right) \right) \left(\exp\left(-\left(\frac{x+r \cdot \sin(\phi_s)}{\sigma_x}\right)^2\right) \exp\left(-\left(\frac{y+r \cdot \sin(\theta_s)+\Delta}{\sigma_y}\right)^2\right) \right) dx dy \quad (3.31)$$

where r is the range from the radar to the scanned volume, ΔR is the range resolution, Δ is the separation between the antennas center, θ_s and ϕ_s represents the alignment error in the yz and xz planes, θ_0 is the antenna beamwidth, and the parameters σ_x and σ_y are obtained from the antenna beamwidths for xz and yz planes, since the antenna beamwidth may not be identical in both planes and depends on the range r as follows: $\sigma_i = r \cdot \theta_{oi}/2 \cdot \sqrt{\ln 2}$, as proposed in [Pro62], considering a Gaussian antenna pattern. To simplify the analysis in this work, we have considered that the antenna beamwidth is identical in both axes.

This analysis assumes that the antennas are in the far field, as was also assumed in the radar equation, see Eq. (3.20). It gives underestimated radar reflectivity values, since some low-range clouds lie in the Fresnel region of the antennas, where the antenna gain and pattern shape vary significantly. To improve this estimation, [Sek02] proposes a correction factor to these clouds:

$$F_{cor}(r/r_{ff}) = \frac{5.26 \cdot 10^{-5} + (r/r_{ff})^{2.5}}{0.01117 + (r/r_{ff})^{2.5}} \quad (3.32)$$

where r_{ff} is the far-field range, that corresponds to $2 \cdot D^2/\lambda$. The Fresnel region starts at a range $r_{fresnel} = 0.62\sqrt{D^3/\lambda}$ and it ends at r_{ff} [Bal05]. This relationship is an analytic approximation to the results derived from numerical antenna pattern simulations. It is independent of radar PRF, transmitted bandwidth and receiver bandwidth. Fig. 3.30 shows the correction factor F_{cor} at 94 GHz for different antenna diameters and its corresponding starting range of the far field r_{ff} . As expected, the correction factor goes to 0 dB at the r_{ff} value of the corresponding antenna diameter.

Firstly, Fig. 3.31 evaluates the power losses represented by the parallax factor $f(r)$ and the correction factor F_{cor} as a function of the range for different alignment error angles θ_s and ϕ_s , considering a separation between antennas Δ of 1 m and an antenna beamwidth θ_0 of 0.2° . As observed, the power losses are extremely sensitive to alignment error higher than 0.3° (19 dB of power losses are obtained at 1 km of range). Furthermore, the power losses dependence on θ_s is not symmetric as negative values of θ_s are less harmful to the sensitivity reduction.

On the other hand, the influence of the separation between antennas on the power losses is significant at lower ranges. However, it decreases rapidly with range as shown in Fig. 3.32.

Lastly, the analysis of a two-antenna configuration with two antennas of different diameters is presented. A smaller antenna with higher beamwidth is selected in order to overcome the sensitivity reduction at lower ranges, although the smaller antenna presents a lower gain due to the antenna diameter reduction.

Fig. 3.33 evaluates the power losses as a function of the range for two different alignment error angles θ_s and ϕ_s , considering a separation between antennas Δ of 1 m. The first antenna presents a beamwidth θ_{01} of 0.2° and three antenna beamwidths θ_{02} have been considered for the second antenna: 0.2° , 0.4° and 0.8° . As observed, 46.5 dB of sensitivity improvement is obtained with a second antenna of 0.8° of beamwidth at 100 m of range, considering the parallax factor and the correction factor. To this value, the gain reduction of the smaller antenna should be subtracted. Using Eq. (3.22), the gain reduction is ~ 12 dB,

Figure 3.30: Correction factor as a function of the antenna diameter at 94 GHz. The legend shows the corresponding Fresnel and far field regions of each antenna diameter.

Figure 3.31: Power losses as a function of the range for different alignment error angles θ_s and ϕ_s , considering a separation between antennas Δ of 1 m and an antenna beamwidth θ_0 of 0.2° taking into account both parallax error and correction factor for Fresnel and far field region of antennas.

therefore, the sensitivity improvement is 34.5 dB at 100 m of range. This improvement is enhanced if alignment errors are considered, as shown in Fig. 3.33.

Figure 3.32: Power losses as a function of the separation between antennas Δ for different ranges. The alignment error angles θ_s and ϕ_s are 0.2° and 0° , respectively. The antenna beamwidth θ_0 is 0.2° .

Figure 3.33: Sensitivity improvement with two-different antennas with beamwidths θ_{0_1} and θ_{0_2} as a function of the range for different alignment error angles θ_s and ϕ_s , considering a separation between antennas Δ of 1 m.

3.7 Cloud radars calibration

The calibration procedure is mandatory in this kind of systems in order to know the precise values of radar equation constants. In general, in radar meteorology, there are four types of calibration that are currently in use [Sau92]:

1. Measurement of the different radar components in order to calculate the radar transmitted power and the receiver gain, and then introduce it with the radar equation. This calibration presents the disadvantage that component responses (in particular, active devices) change over time.
2. Use of an isolated target with known radar cross section. Additionally, the target should be located in the far field of the radar antennas, where antenna gain is well defined. This is the most common and oldest method in radar meteorology, although it is logistically difficult. The target is often a metal sphere or a trihedral reflector. The calibration precision with this method is about 1 dB [Sau92].
3. Dicke-based receiver calibration and transmitted power monitoring [Rad14a]. Firstly, the receiver can be periodically calibrated as occurs in radiometry, by using a Dicke switch and a high precision load [Sko89]. On the other hand, the transmitted power is also periodically monitored to improve the radar calibration precision.
4. Rain-based calibration of millimeter-wave systems [Hog03]. The radar reflectivity factor of rain at 94 GHz is practically stable, since the rain particles are in the optics region (see Section 3.2). To this end, by the use of a rain gauge close to the radar site, the relationship between rainfall rate and the radar reflectivity factor can be empirically obtained, and therefore, the calibration can be carried out. This technique has been demonstrated in [Hog03] with a precision better than 1 dB for rainfall rates between 3 and 10 mm⁻¹, which corresponds to a radar reflectivity factor of ~19 dBZ at 94 GHz.

There are some peculiarities of millimeter-wave cloud radars that should be taken into account that make the absolute radar calibration difficult:

1. Regarding the second type of calibration, the cloud-radar antenna pointing is toward the zenith, so the known target should be risen up to 1 km approximately, since the far field of a Cassegrain antenna with 1.2 m of diameter begins at 900 m following the equation $2D^2/\lambda$. The use of small drones to this end is not allowed at this range for air security concerns. A sphere hanging from the drone using a thread of 1.5 m of length at a range of 150 m has been evaluated without success. The received power from the sphere fluctuates and the drone masks the received power from the sphere. The known target should be stable at this range to avoid measurement imprecisions. An expensive alternative is the use of a balloon although it is sometimes difficult to keep it inside the antenna beamwidth. We are considering this measurement for the future in the framework of our collaboration with AEMET, which has a balloon in use in combination with radiosondes. This method is currently used in the UK Met Office measurement campaigns [Ben09].
2. As indicated in Section 3.2.5, the temperature at each range should be periodically measured in order to obtain the radar signal path attenuation, particularly within the clouds. For instance, $|K|^2$, that is related to the radar reflectivity (see Section 3.2), varies as a function of the temperature at the millimeter-wave band [Hog03]. It increases from 0.67 at 0°C to 0.813 at 20°C [Lie89], and therefore, it limits the calibration precision.

3. Radome losses change if drizzle or rain occur during the measurements. The power losses are considerable, in some cases up to 14 dB [Sek98, Hog03]. To overcome this limitation, new versions of commercial cloud radars incorporate a dew blower and a heater for avoiding liquid drops and ice crystals on the antenna radome [Rad14a].

Considering these difficulties, the calibration of the cloud radar developed in this thesis is currently carried out by calculating the overall Tx-Rx performance as indicated in the first type of calibration, based on the RF measurements of Section 4.4. The measurement error is lower than 1 dB considering the lab instrumentation precision, besides, matching between components is taken into account. Furthermore, the transmitted power monitoring is carried out by using a digital amperimeter connected to the HPA. The relationship between the output power and the HPA current consumption is previously obtained in order to define a calibration function. The amperimeter is connected via universal serial bus (USB) to the PC that manages the radar system and the current consumption is updated and stored each second.

Chapter 4

Design and implementation of a cloud profiling radar at 94 GHz

4.1 Introduction

This chapter addresses the design and implementation of a CW-LFM radar working at 94 GHz for cloud monitoring. As indicated in the previous chapter, the main requirements of the radar architecture for this application are: high transmitted power, high-gain antennas and low transmitted bandwidth (thus, low range resolution). Both the transmission of high-power and high gain antennas are desired to compensate the atmospheric attenuation in order to achieve a maximum range of 10-15 km, where some interesting cloud types are located. On the other hand, low transmitted bandwidth (in comparison with other radar systems for different applications, as occurs with the imaging radar system of Chapter 2) up to 15 MHz is enough for this application, since the desired range resolution is 10 m at most.

The chapter structure consists of the following parts. First of all, Section 4.2 shows the design of the cloud radar by means of a review of commercially available components and the radar equation presented in the previous chapter in order to set the radar operational parameters. Secondly, the system architecture is presented in Section 4.3, and the RF characterization of the transmitter and receiver chains and components is shown in Section 4.4. Section 4.5 is devoted to the characterization of sensitivity and dynamic range of the system. Data acquisition and signal processing of the radar reflectivity and the Doppler velocity, both introduced in the previous chapter, are described in Section 4.6. The radar data exportation and dimensioning is shown in Section 4.7. Lastly, the analysis and interpretation of some results of cloud observations are presented in Section 4.8.

4.2 Design of the 94 GHz cloud profiling radar

The analysis of the millimeter-wave components market is divided into: antennas, high-power amplifiers and radar receivers. The transmitter and receiver chains are based on new and reused components of a previous millimeter-wave radar system (see Section 2.2.2) that worked at a different operating frequency (100 GHz) [Men13a]. As indicated in Chapter 2, the microwave signal generation is based on the DDS/PLL chirp source, and a 30 GHz MMIC-based doubler-MPA module that has been developed to achieve a high degree of integration in the microwave subsystem with cost-saving purposes. On the other hand, the data acquisition subsystem is based on the same ADC that the imaging radar uses with a

redesigned active filter to the beat frequencies of interest for the cloud radar.

The radar operational parameters are determined in collaboration with AEMET meteorologists. They are interested in non-precipitating clouds of ~ -40 dBZ up to 10-15 km. The falling velocity of these clouds is lower than 5 m/s. It corresponds to a radar PRF of 6 kHz, according to Eq. (3.15). Furthermore, the required range resolution is 50 m, so the transmitted bandwidth should be 3 MHz. The area of interest should be as small as possible, in order to detect sufficiently homogeneous volumes of clouds. Therefore, the antenna should have the largest diameter available. Lastly, in order to validate the developed cloud radar with AEMET weather radars and satellites, one minute of time interval between consecutive measurements is enough.

4.2.1 Market survey of millimeter-wave radar components

4.2.1.1 Radar antennas

Taking into account the strict requirements of this application in terms of antenna beamwidth and gain (see Sections 3.5 and 3.6), apart from the cloud profiling radars analysis presented in Section 3.4, a market analysis of commercial antennas offered by some manufacturers has been conducted. Table 4.1 shows the main features of the selected antennas. In this analysis, the main parameters, such as gain, antenna diameter, operating bandwidth (B), HPBW and SLL are shown, including the antenna cost. As indicated in Section 3.5, the maximum diameter of commercially available Cassegrain and lens antennas at the millimeter-wave band is 1.2 m and 0.3 m, respectively [Mil16]. Additionally, they offer customized designs for higher diameters.

Table 4.1: Commercial antennas at W-band for cloud profiling radars.

Manufacturer	Antenna	Diameter	B	Gain	HPBW	SLL	Cost
Quinstar	Cassegrain	1.2 m	Fullband	59 dB	0.2°	16 dB	15 k€
Quinstar	Lens	0.3 m	Fullband	47 dB	0.72°	25 dB	8.3 k€
Millitech	Cassegrain	0.45 m	Fullband	50 dB	0.49°	16 dB	24.5 k€
Millitech	Lens	0.3 m	Fullband	47 dB	0.73°	23 dB	7.2 k€
Mi-Wave	Cassegrain	1.2 m	2 GHz	59 dB	0.2°	18 dB	9.8 k€
Sage	Lens	0.3 m	4 GHz	48 dB	0.8°	23 dB	6.3 k€

In view of the cost, the available budget and performance of commercially available antennas and the study of the parallax error (see Section 3.6), two different antennas (marked in bold in Table 4.1) have been selected for the developed cloud radar. For the receiver, the Cassegrain antenna of 1.2 m diameter from Mi-wave with a beamwidth of 0.2° , and for the transmitter, a lens antenna of 0.3 m diameter from Sage Millimeter with 0.8° of beamwidth have been selected. This beamwidth compensates the parallax error, particularly at lower ranges, although the antenna gain is reduced 11 dB (see Fig. 3.33). The selected antennas on the radar mechanical structure are shown in Fig. 4.1.

Due to the large dimensions of both antennas, particularly the Cassegrain antenna, and its lower beamwidth, the minimum range of detection is limited by the separation between both antennas. In the developed radar, a minimum separation of 90 cm is established between antennas center (5 cm between antenna edges, see Fig. 4.1). Additionally, as indi-

Figure 4.1: Photograph of the selected antennas on the radar mechanical structure.

cated before, the metallic structure is mechanized taking into account a small displacement of ± 5 cm of the lens antenna. Furthermore, in order to determine the exact value of isolation between transmit and receive antennas, the Erickson PM4 Power meter has been used, obtaining that the isolation between antennas is higher than ~ 70 dB*.

4.2.1.2 Radar high-power amplifier

In order to analyze the state-of-the-art of power amplifiers at millimeter-wave band, Table 4.2 summarizes the main parameters of research and commercial power amplifiers for CW-LFM radars at W-band. As indicated in the Introduction, most of them are based on GaN HEMT technology, although early efforts were devoted to InP or heterostructure field effect transistor (HFET) technologies with worse results [Che98, Mic06]. Furthermore, corporate networks for power combination are usually implemented in order to increase the output power, as indicated in Section 1.4.

First of all, the HRL high-power amplifier was selected as the radar high power amplifier. A housing with two microstrip-to-waveguide transitions was designed and manufactured, but the output results were very poor, mainly due to assembly capabilities of our laboratory at this band. Therefore, the Farran power amplifier[†] was finally selected due to its affordability (~ 15 k€) and performance regarding other commercial amplifiers.

4.2.1.3 Radar receiver

The radar receiver is based on the reuse of previous components of a 100-GHz radar [Men13a]. These components are characterized at 94 GHz in Appendix B. It is composed of a mixer, an isolator and a LNA from RPG and an active filter that is presented in Ap-

*This measurement is limited by the sensitivity of the power meter. Therefore, the isolation between antennas should be higher than ~ 70 dB.

[†] Some reliability issues of the Farran power amplifier have appeared during the radar measurement campaigns, which have been repaired twice. In this context, Quinstar informs us during the HPA selection process that the lifetime of their commercial W-band HPA amplifiers is not expected to exceed 1,000 hour.

Table 4.2: Research and commercial power amplifiers at millimeter-wave bands for the UPM cloud profiling radar at 94 GHz. The amplifiers parameters such as central frequency, bandwidth, output power and gain are shown in conjunction with the employed technology and the number of levels of power combination (if applicable).

	Frequency	Bandwidth	Output power	Gain	Technology	Comb.
<i><u>Research HPAs</u></i>						
[Mas09]	76.5 GHz	10 GHz	25.4 dBm	11.6 dB	GaN HEMT	-
[Mic06]	84 GHz	9 GHz	26.3 dBm	17 dB	GaN HFET	-
[Che98]	95 GHz	1 GHz	26.4 dBm	8.8 dB	InP HEMT	-
[Sch10]	96 GHz	4.5 GHz	37.2 dBm	13 dB	GaN HEMT	12
[Che13]	94 GHz	4 GHz	20 dBm	23 dB	InP HEMT	2
[Mic12]	94 GHz	4 GHz	31.7 dBm	10 dB	GaN HEMT	4 & 8
[Sch13]	90 GHz	20 GHz	31 dBm	12 dB	GaN DHEMT	-
[Sch16]	88 GHz	25 GHz	45 dBm	25 dB	GaN HEMT	12
[Bro11]	91 GHz	2 GHz	32.3 dBm	15 dB	GaN HEMT	No
[Fun11]	89 GHz	8 GHz	34.7 dBm	6.6 dB	GaN HEMT	2 & 4
<i><u>Commercial HPA MMICs</u></i>						
HRL	92 GHz	5 GHz	27 dBm	14 dB	GaN HEMT	No
Northrop	94 GHz	4 GHz	25 dBm	7.5 dB	GaN HEMT	No
<i><u>Commercial HPA MMICs with housing</u></i>						
Farran	92 GHz	4 GHz	31 dBm	25 dB	GaN HEMT	2
Quinstar	92 GHz	5 GHz	30 dBm	30 dB	GaN HEMT	-
Millitech	93 GHz	6 GHz	34.3 dBm	27 dB	GaN HEMT	-

pendix C. This active filter of 20 dB of gain[‡] overcomes the ADC noise-power limitation that affects the radar sensitivity. The receiver noise power calculations are presented in Appendix D.

[‡]This gain value is justified in Appendix D taking into account the ADC noise power and the receiver noise power previous to this active filter.

4.2.2 Radar operational parameters definition by means of the radar equation

According to the previous analysis, the radar parameters suggested by AEMET meteorologists and the selection of the power amplifier, the receiver and antennas, the operational parameters of the developed cloud profiling radar at 94 GHz are shown in Table 4.3. These parameters have been obtained through the radar equation stated in Eq. (3.20) (see Section 3.5).

As occurs with the imaging radar of Chapter 2, a baseband sampling is considered. The required sampling frequency (f_s) taking into account the sampling theorem by Nyquist as a function of the maximum range:

$$f_s \geq 2 \cdot f_{b_{max}} = 2 \cdot \frac{2 \cdot R_{max} \cdot BW}{c \cdot T_m} \quad (4.1)$$

Considering a maximum range of interest of 20 km, a radar PRF of 6 kHz and a transmitted bandwidth of 3 MHz as suggested by AEMET meteorologists, the sampling frequency should be 4.8 MHz.

The SNR estimation as a function of range is shown in Fig. 4.2. In this estimation, the sensitivity is -50 dBZ at 11.4 km. As indicated in Section 3.5.2, the sensitivity of a cloud radar is defined in radar meteorology as the reflectivity factor for an SNR of 0 dB. The dynamic range is 90 dB if the reflectivity factor margin is between -50 dBZ and 0 dBZ at a maximum range of 11.4 km.

Table 4.4 shows the sensitivity for different reflectivity factors. Therefore, the requirements suggested by AEMET meteorologists (-40 dBZ at 10-15 km) are fulfilled. Regarding the number of integrated chirp signals (M), sampling frequency and the acquisition time, a compromise among them is achieved to reduce the captured data size and the FFT points. This will be shown in Section 4.7.

Figure 4.2: SNR as a function of the reflectivity factor using the operational parameters of the UPM cloud profiling radar.

Table 4.3: Operational parameters of the UPM cloud profiling radar at 94 GHz.

Parameter	Value
Center frequency	94 GHz
Transmitted bandwidth	3 MHz
Transmitted power	30.2 dBm
Noise figure	4.6 dB
Sampling frequency	4.8 MHz
Tx antenna gain	48 dB
Tx antenna beamwidth	0.8°
Rx antenna gain	59 dB
Rx antenna beamwidth	0.2°
PRF	6 kHz
Maximum unambiguous Doppler velocity	± 4.8 m/s
Acquisition time	2.5 s
Integration gain ($M = 15000$)	20.9 dB
Number of FFT points	1024
Radar sensitivity ($SNR = 0$ dB)	-50 dBZ @11.4 km
Dynamic range	90 dB @0-11.4 km, (-50,0) dBZ

Table 4.4: Radar sensitivity of the UPM cloud profiling radar at 94 GHz for different reflectivity factors considering SNR=0 dB.

Reflectivity factor	Range
-70 dBZ	1.1 km
-60 dBZ	3.6 km
-50 dBZ	11.4 km
-40 dBZ	35.9 km
-30 dBZ	113.7 km
-20 dBZ	359.2 km

4.3 System architecture

Fig. 4.3 shows a simplified version of the block diagram of the radar architecture. It is divided into four subsystems as a function of the chosen technology: The digital electronics subsystem, the microwave subsystem, the millimeter-wave subsystem and the data acquisition.

The digital electronics subsystem is in charge of the chirp signal generation based on a narrowband DDS, that is managed from a PC. The advantages of DDS signal generation for radar systems were discussed in Section 2.3, providing a high chirp linearity and fine frequency resolution. CW-LFM cloud radars of Table 3.2 also generate the chirp signal with a DDS. Focusing on this application, special attention should be paid to the spurious signals content at the DDS output. Low level of spurious signals is required to avoid uncertainties involved in radar reflectivity factor estimates. Its high performance will be demonstrated by means of the SFDR measurement in Section 4.4.1.

The millimeter-wave subsystem is divided into two chains: the transmitter and receiver chains which generates the LO signal. Additionally, we have considered a radar design alternative by dividing the transmitter and receiver chains at 94 GHz (test point E) instead of at 15.6 GHz (test point B) in order to reduce the number of components of the millimeter-wave subsystem. It implies the use of a long waveguide section to connect both chains since the antennas separation is around 0.9 m. The insertion losses are around 3 dB. This alternative was discarded due to the high cost of the custom design of the waveguide section in comparison with the selected architecture, since some millimeter-wave components were reused. Furthermore, it limits the separation between antennas, which cannot be adjusted such as in the proposed architecture.

The operating frequency is 94 GHz and the maximum transmitted bandwidth is 2 GHz, limited by the Cassegrain antenna bandwidth (see Appendix A). As indicated before, a transmitted bandwidth of 3 MHz is enough for this application. However, this versatility in the bandwidth selection allow us to use this radar for different applications, such as small objects detection and identification [Wav06] or perimeter security surveillance [Nav16].

As indicated in Chapter 3, a two-antenna configuration is used to increase the isolation between the transmitter and the receiver and two different antennas are used to overcome the parallax error degradation. The homodyne configuration has been chosen in order to reduce cost and compactness regarding remarkable heterodyne radars presented in Chapter 3. This selection produces a sensitivity reduction [Sch93] due to the targets received signal being affected by the low frequency noise of the receiver noise floor due to the transmitter leakage. It will be demonstrated in Section 4.5.2 that it minimally affects the radar sensitivity in our configuration.

Regarding the data acquisition subsystem, the sampling frequency is adjustable up to 40 MHz, driven by the second channel of the DDS generator (see Fig. 4.3). This versatility of the sampling frequency allows to modify the chirp signal parameters, taking into account the range of distances and its beat frequencies of interest for the cloud radar operation. In the same way as in Chapter 3, the IF signal is previously amplified and filtered out by an active bandpass filter, and then is digitized by a 16-bit digitizer from ADLINK Technology assembled on a conventional PC.

Figure 4.3: Block diagram of the radar system at 94 GHz. It is made up of four subsystems: the DDS-based signal generation source, the microwave subsystem, the millimeter-wave subsystem, and the data acquisition subsystem. Test points are defined with circle markers and the swept bandwidth of the chirp signals at each test point is indicated in parenthesis. The measured output power at the operating frequency of the radar system is shown in red.

4.4 Hardware characterization

In this section, the characterization of the radar architecture is presented. The output power and the phase noise at each point of the radar architecture will be shown after the individual characterization of each component. Fig. 4.4 shows the photographs of the subsystems that compose the radar architecture. Fig. 4.4a shows the microwave subsystem enclosure, where the PLL frequency multiplier, the DDS generator and the master oscillator are also included. Fig. 4.4b-c show the transmitter and receiver chains. The data acquisition subsystem is assembled on a conventional PC.

Figure 4.4: Photograph of the microwave subsystem enclosure (a), the millimeter-wave transmitter (b) and the millimeter-wave receiver (c).

The DC power consumption of the radar electronics is 30 W, which is shown in Table 4.5 in conjunction with the current consumption for each voltage. 26 W corresponds to the millimeter-wave subsystem (the HPA consumption is 15 W) and 4 W corresponds to the microwave subsystem. The PC power consumption, including the ADC, is not contemplated in this estimate.

Table 4.5: Power consumption of the cloud-profiling radar system.

1.8 V	150 mA	6 V	1.2 A	15 V	1.05 A
3.3 V	220 mA	7 V	350 mA	-5 V	50 mA
5 V	665 mA	12 V	5 mA	Total	30 W

4.4.1 Digital electronics

The chirp source is based on the AD9958 narrowband DDS from Analog Devices, which is the same as the one presented in Section 2.3.1.2, although the operating frequency and the bandwidth are modified to adapt it to this system at 94 GHz. The selected DDS chirp signal presents a PRF of 6 kHz and a bandwidth of 2.08 MHz, which corresponds to a maximum transmitted bandwidth of the radar system[§] (2 GHz), since the multiplication factor of the transmitter chain is $\times 960$ ($\times 80 \times 2 \times 6$, see Fig. 4.3). The power ripple and the SFDR measurement results at the DDS output (test point A, see Fig. 4.3) are represented in Fig. 4.5. It is observed that the SFDR is -65 dBc and the power ripple (peak-to-peak variation) is 0.05 dB, see zoom of Fig. 4.5.

Figure 4.5: Power spectrum measurement at the DDS output (test point A).

4.4.2 Microwave subsystem

As indicated before, the DDS-based signal generation subsystem is made up of a narrowband DDS that provides a chirp signal to a $\times 80$ PLL to bring up the chirp signal to 7.83 GHz. This signal is doubled to achieve a chirp signal centered at 15.66 GHz (test point B). Then, the output chirp signal is divided with a 10-dB directional coupler. The through signal of the directional coupler is introduced in the LO chain of the millimeter wave subsystem, and the coupled signal is introduced into a doubler module (test point C) that provides the input signal to the W-band tripler of the transmitter chain (test point D).

The output power at the microwave subsystem (test point B) is shown in Fig. 4.6. The chirp signal is ramping from 15.5 to 15.83 GHz to obtain the maximum transmitted bandwidth of the radar system. The power ripple is 1.5 dB and the measurement of single tones show an SFDR of roughly 55 dBc.

[§]Although 3 MHz of transmitted bandwidth is enough for this application, the measurements that are shown in this characterization correspond to the maximum transmitted bandwidth of 2 GHz in order to determine exactly the center frequency around 94 GHz where the components offer the best performance in terms of output power.

Figure 4.6: Power spectrum measurement at the microwave subsystem output (test point B).

4.4.2.1 Doubler-MPA module

This circuit is based on two *Hittite* (currently *Analog Devices*) MMICs: HMC598 doubler and HMC499 MPA, as shown in Fig. 4.7a. It is highlighted with a dashed line box in Fig. 4.3. It has been designed by using a commercial CuClad substrate ($\epsilon_r = 2.17$, $h = 254 \mu\text{m}$). The output power results as a function of its input power (test point C) are presented in Fig. 4.8. The grey shaded area corresponds to the maximum transmitted bandwidth of the cloud profiling radar (2000 MHz centered at 94 GHz), as will be the case of the rest of figures. Considering that the W-band tripler (test point D) requires an input power of 20 dBm to obtain its optimal output power (around 5 dBm), which allows to generate the maximum power at the HPA output (1 W, test point G), the input power of the doubler should be higher than 2 dBm, as can be seen in Fig. 4.8. It implies that the coupling factor of the directional coupler should be 10 dB, since the output power at the test point B is 12 dBm. On the other hand, an automatic biasing sequencer based on power metal-oxide-semiconductor field-effect transistors (MOSFETs) is designed and fabricated to adequately polarize both MMICs [Tor04]. This sequencer is placed over the doubler housing as shown in Fig. 4.7b and it generates a regulated drain voltage and automatically adjusts the gate voltage, as described in further detail in Appendix E.

The return losses at the input and output ports of the doubler are presented in Fig. 4.9. For both measurements, the return losses are better than 15 dB at the operating frequency of the radar system.

Figure 4.7: Photograph of the doubler-MPA MMICs (a) and the doubler module housing with the automatic biasing sequencer (b). This circuit is highlighted with a dashed line box in Fig. 4.3.

Figure 4.8: Doubler-MPA module: Output power measurement as a function of its input power.

(a) S_{11}

(b) S_{22}

Figure 4.9: Doubler-MPA module: S_{11} -and- S_{22} parameters measurement.

4.4.3 Millimeter-wave subsystem

This subsystem is made up of two chains: The transmitter chain and the LO chain, which generates the transmitted power and the LO power to the mixer, respectively. The components used in the LO chain are: an active $\times 6$ frequency multiplier, an LNA and a fundamental mixer. The components are commercial items from RPG except the fundamental mixer, which has been supplied by Quinstar. The transmitter chain is composed by a tripler from Pacific Millimeter, an MPA and a HPA from Farran Technology, Ltd. that provides more than 1 W of output power at 94 GHz. The characterization of each component is presented in Appendix B.

The HPA output power (test point G) is presented in Fig. 4.10. The transmitted power of the radar system is 30.2 dBm at 94 GHz. At this output power, the power-added efficiency (PAE) at 94 GHz is 6.9%, since the current consumption is 1.22 A at 15 V. This result is practically constant in the radar bandwidth. As can be observed, the maximum power over the 2-GHz bandwidth occurs both at 93 GHz and 94 GHz of center frequency. Therefore, the selected center frequency for this radar is 94 GHz.

Figure 4.10: Farran HPA amplifier: Output power measurement.

Lastly, Fig. 4.11 shows the power measurement at different test points defined in Fig. 4.3. The DDS output power (test point A) is -0.1 dBm. The output power of the microwave subsystem (test point B) is 12 dBm and the OL-signal power (test point D) is 6 dBm, both at the operating frequency of the radar system. The doubler module (test point C) generates an output power of 21 dBm, which is transferred to the MPA, which produces an output power of 16 dBm (test point F). The transmitted power (test point G) is 30.2 dBm at 94 GHz. The power ripple is 0.1 dB and 0.4 dB considering that the transmitted bandwidth is 3 MHz and 2000 MHz centered at 94 GHz, respectively. This great power flatness will produce a reduction of the RF-nonlinearities, in contrast to the RF-nonlinearities of the imaging radar system of Chapter 2 due to its bandwidth being more than 9 times the bandwidth of the cloud profiling radar.

In the same way, the phase noise measurements at the different test points defined in Fig. 4.3 are presented in Fig. 4.12. The phase noise at 94 GHz is estimated by adding the multiplication factor of $20 \cdot \log_{10}(3)$ to the measurement at 31 GHz.

Figure 4.11: Power measurements at the output of different components of the radar architecture.

Figure 4.12: Phase noise measurement at the output of different components of the radar architecture.

4.5 System characterization

4.5.1 Spectral purity of the microwave subsystem

A delay-line test system is used to monitor the signal spectral purity at the microwave subsystem output (test point B, see Fig. 4.3) in the same way as in Chapter 2, considering that the transmitted radar bandwidth is 2000 MHz and the PRF is 6 kHz. Therefore, the beat frequency corresponds to 15 kHz, following Eq. (4.2). From this measurement, the SLL is 35 dBc. The RF non-linearities are lower with respect to the imaging radar system due to the reduction of the transmitted bandwidth. A Hamming windowing function is used, which offers a sidelobe level of -43 dB.

Figure 4.13: RF non-linearities measurement at the microwave subsystem output with the cloud-profiling-radar operational parameters.

4.5.2 Receiver noise floor, sensitivity and dynamic range

Fig. 4.14 shows the noise power floor at three different points of the receiver chain: ADC, receiver chain, both with a 50-Ω load at their inputs, and the noise power floor in clear-sky conditions (receiver and transmitter including both antennas are operating simultaneously). All of these measurements are obtained after non-coherent integration of 2.5 seconds of IF signal with a PRF of 6 kHz, as defined in Table 4.3. In the latter case, the noise power floor is -94.2 dBm and the coupled power noise of the transmitter chain is observed. This agrees well with the noise power simulation of Appendix D. The noise power difference between the ADC noise floor and the receiver noise floor of 7 dB also agrees with the noise calculations of Appendix D and the active filter gain. It is crucial for this application to identify and locate each spurious signal on the receiver noise power floor in order to avoid interpretation errors of the non-precipitating cloud measurements. To this end, the use of linear power supplies properly filtered, the ADC housing inside the PC and an USB opto-isolator to separate the electrical connection between the PC and the DDS are mandatory. Table 4.6 shows the detected spurious and its power level difference regarding the noise floor (-94.2 dBm). Fig. 4.14 shows that the maximum difference of the noise power level (black curve) and the spurious signal level is lower than 1 dB. In Section 4.6.1.1, spurious impact on the detection threshold definition will be further discussed.

Figure 4.14: Noise power floor of the receiver chain after non-coherent integration of 2.5 seconds of IF signal with a PRF of 6 kHz. The ADC and the receiver chain noise power floor are measured with a $50\text{-}\Omega$ load at their inputs in both cases. Thirdly, the noise power floor is measured in clear-sky conditions, with receiver and transmitter, including both antennas, operating simultaneously.

Table 4.6: Spurious signals of the receiver noise power floor and its power level difference regarding the noise floor (-94.2 dBm).

(kHz)	(dB)	(kHz)	(dB)	(kHz)	(dB)	(kHz)	(dB)	(kHz)	(dB)
11	0.9	103	0.7	306	0.2	830	0.9	1246	0.6
34	0.3	135	0.9	415	0.9	919	0.4	1454	0.5
57	0.2	207	0.2	623	0.3	1038	0.7	1532	0.5

Fig. 4.15 shows the noise distribution histogram of the noise power floor (after windowing and FFT) without considering spurious signals of Table 4.6 in clear-sky conditions of 10 data captures of 2.5 s, obtained every ten seconds. A Gaussian distribution is observed in the real and imaginary parts. Assuming that both distributions are independent, zero-mean and with variance equal to σ^2 , the distribution of the modulus corresponds to a Rayleigh distribution, as observed in Fig. 4.15. Table 4.7 shows the mean and the variance results of the noise distributions of Fig. 4.15. Furthermore, the stationarity of the receiver noise has been evaluated. The mean noise variation between consecutive noise captures is a few hundredths of a dB.

Figure 4.15: Noise distribution histogram of the noise power floor in clear-sky conditions.

Table 4.7: Noise distribution parameters of the noise power floor in clear-sky conditions.

Noise Distribution	Mean (V)	Variance (V ²)
Real part	$-4.37 \cdot 10^{-7}$	$3.35 \cdot 10^{-11}$
Imaginary part	$1.3 \cdot 10^{-6}$	$3.36 \cdot 10^{-11}$
Modulus	$3.08 \cdot 10^{-6}$	$-1.49 \cdot 10^{-11}$

As indicated in Section 3.5.2, regarding the radar sensitivity in radar meteorology, it is not clearly defined in the literature by means of a minimum required SNR that is commonly used in radar detection theory. [Lhe87] considers a minimum SNR of 0 dB and other systems [Ste03, Rad14a, MeT17] indicate the sensitivity in dBZ as a function of range without indicating the minimum required SNR. Fig. 4.16 shows the radar sensitivity of our cloud radar as a function of different transmitted bandwidths, considering the noise power floor of Fig. 4.14 (black curve), since it corresponds to the sensitivity when SNR is 0 dB and converting it to dBZ using Eq. (3.27) and radar operational parameters of Table 4.4. The radar sensitivity is degraded if the range resolution is improved by the increase of the transmitted bandwidth, since the sampled volume by the radar gets smaller. Additionally, the radar sensitivity in dBZ of the most radars of Table 3.2 extracted from their published information is shown for comparison purposes. The radar sensitivity of the cloud radar offers better performance than those of the aforementioned systems, excepting of MeTeK cloud radar that works at a lower operating frequency (36.5 GHz).

Fig. 4.17 shows the sensitivity reduction at lower ranges due to the transmitter leakage as a function of different transmitted bandwidths. As observed, when the radar is transmitting, a low frequency noise is coupled and therefore it limits the sensitivity of the radar. The

Figure 4.16: Receiver sensitivity in dBZ considering SNR=0 dB of the developed cloud radar as a function of the transmitted bandwidth after non-coherent integration of 2.5 seconds of IF signal with a PRF of 6 kHz.

minimum range R_{min} where the sensitivity limitation does not affect to the radar detection is indicated in Fig. 4.17, using Eq. (4.2) for different bandwidths: 3.6, 5, 10 and 15 MHz. The R_{min} value is 190, 115, 57 and 38 m, respectively.

Figure 4.17: Sensitivity reduction at lower ranges due to transmitter leakage.

4.5.3 Point-target detection

In order to check the radar operation, a quadcopter drone (see Fig. 4.18) from Dà-Jiāng Innovations Science and Technology Co., Ltd. (DJI) is used as point target. The drone dimensions are 29x29x18 cm³. In this case, the drone measurement allows us to compare the height and velocity measurements between the cloud radar and the GPS of the drone. The transmitted bandwidth is 150 MHz centered at 94 GHz, the chirp period is 6 kHz and the integration time is 1 second.

Additionally, since the drone is a point target for the cloud radar, it also enables us to characterize the phase noise influence on the target response. The radar cross section is high enough, and, therefore, the target noise floor is higher than the thermal noise floor of the receiver. When the drone is not well aligned with the antennas, the received power will be lower and allows us to observe the receiver noise floor. This phenomenon will also occur with cloud detection, particularly if drizzle is detected, since the radar reflectivity factor is high. It will be analyzed in Section 4.6.1.1.

Figure 4.18: Photograph of the drone *DJI Phantom III Professional* used for checking the radar operation.

4.5.3.1 Range detection

Fig. 4.19 shows the range measurement comparison between the cloud radar and the GPS of the drone. A good agreement is obtained between both measurements of the drone trajectory. The colorbar of the radar measurement represents the received power in dBm.

As indicated before, the range profile of the drone is affected by the transmitted phase noise, as shown in Fig. 4.20. The receiver noise floor is shown in red, and four range profiles are plotted when the drone is detected at 16, 83, 124 and 180 meters, respectively.

Up to a certain value of the received power from the drone, P_{rxlim} , the noise floor is imposed by the receiver, and for higher powers, the noise floor is increased in the same amount as the received power, as shown in Fig. 4.21.

Figure 4.19: Range measurement comparison with the radar and the drone GPS measurement.

Figure 4.20: Four range profiles when drone is detected at a distance from the radar of 26, 83, 124 and 180 meters, and the receiver noise floor in order to compare the increase of the noise floor due to the transmitted phase noise.

Fig. 4.22 shows the comparison of the noise floor estimate by using the phase noise measurement at 16 GHz of a single tone (see Fig. 4.12) and the drone measurement by the radar. The decorrelation factor, the detection bandwidth of FFT processing (6 kHz) and 3 dB due to the phase noise addition of the two input signals of the mixer [Coo11] should be taken into account. As was shown in Section 2.3.4.1.1, the estimate is consistent with the radar measurement. The reason behind the differences between the noise floor of the four range profiles of Fig. 4.20 correspond to the decorrelation factor, which depends on the beat frequency of detected target (see Eq. 2.8) and, therefore, it will be different as a function of the drone height.

Figure 4.21: Noise floor measurement as a function of the drone received power.

Figure 4.22: Noise floor comparison between the phase noise estimate and the radar measurement of the detected drone.

4.5.3.2 Doppler velocity

The Doppler velocity comparison of the drone measurement is shown in Fig. 4.23. The maximum ascent/descent velocity of the drone is 3 m/s. Therefore, the Doppler measurement velocity is unambiguous, since the unambiguous velocity of the radar is 4.8 m/s, taking into account that the PRF is 6 kHz and the operating frequency is 94 GHz. The Doppler velocity at each measurement interval corresponds to the maximum received power of the Doppler-range matrix.

Fig. 4.24 shows the Doppler-range matrix for a certain measurement interval of the drone. The effect of range migration can be observed, which consists of the drone not being located during the integration time in the same range bin [Ric10], since during 1 second of integration the drone has moved a few meters. Additionally, the broadening of the drone response in the frequency domain is due to the propellers rotation.

Figure 4.23: Doppler velocity measurement comparison with the radar and the drone GPS measurement.

Figure 4.24: Doppler-range matrix of the drone.

4.6 Signal processing

This section describes the signal processing algorithm in the cloud profiling radar with the aim of obtaining the 3D reflectivity profile in dBZ and the 3D Doppler velocity profile. The signal procedure will be described by using the cloud measurement results that were carried out on January 29th, 2016 in order to facilitate the reader comprehension.

4.6.1 Radar reflectivity

With the aim of obtaining the 3D radar reflectivity in dBZ, the signal processing procedure is depicted in Fig. 4.25. The following steps are necessary in order to obtain the 3D radar reflectivity profile:

1. First of all, the IF signal of M chirps is obtained in every measurement interval time ($T_m \cdot M$ seconds) of the total acquisition time ($N \cdot T_m \cdot M$ seconds). Commonly, a standby time between each measurement interval time is defined arbitrarily. After each interval time, the column-wise FFT is carried out to obtain the spectral analysis of the IF signal and then transform the frequency domain into the range domain following Eq. (4.2), where c is the free-space propagation velocity, BW is the transmitted bandwidth and T_m is the chirp period. Previously, the IF signal is windowed to reduce the sidelobe levels due to the required dynamic range of this application.

$$f_b = \frac{2 \cdot r \cdot BW}{c \cdot T_m} \quad (4.2)$$

2. Subsequently, since cloud return samples are quickly decorrelated, non-coherent integration of the M chirps of the measurement interval time is carried out to average down the noise and obtain higher SNR than the individual chirp signals [Ric10].
3. Lastly, the radar reflectivity profile is formed by the integrated IF signal of each measurement interval time. After that, a double-threshold detection scheme is applied to each profile, which will be detailed in Section 4.6.1.1. Finally, the integrated range profile is represented in 3D in dBZ, as indicated in Section 3.5. It is obtained by transforming the reflected power of each target of the range profile following Eq. (3.21).

4.6.1.1 Double-threshold detection scheme

As presented in Section 4.5.3, a double-detection threshold scheme is required to mitigate the influence of the transmitted phase noise when the received power from a cloud is considerably high[¶]. This is often the case in noisy systems [Igl15], such as millimeter radars [Coo12]. This phenomenon was already analyzed in Section 2.3.3 (see Fig. 2.17b). The block diagram of the double-threshold detection scheme is presented in Fig. 4.26.

4.6.1.1.1 First threshold: Detection of clouds in receiver noise

Firstly, in order to determine the detected weather phenomena over the receiver noise, a classic detection threshold scheme related to the probability of false alarm P_{FA} is applied to the received power at each integrated range profile^{||} following a binary hypothesis test, as

[¶]It is worth noting that the second threshold related to the transmitted phase noise is only used in case of drizzle or rain, and this system is of interest for non-precipitating clouds, see Section 3.2.

^{||}In [Clo95], an alternative detection threshold was proposed, but it cannot be related to the probability of false alarm.

Figure 4.25: Signal processing procedure in order to obtain the 3D radar reflectivity in dBZ of the cloud profiling radar.

Figure 4.26: Block diagram of the double-threshold detection scheme.

shown in Fig. 4.27. The first hypothesis is denoted as H_0 , which corresponds to a integrated range profile without clouds, and the second as H_1 , where a target is presented in the integrated range profile.

Figure 4.27: Block diagram of the binary hypothesis test of the detection scheme.

The probability of false alarm P_{FA} is defined as the probability that the incorrect hypothesis is chosen: an interference or noise can cross the detection threshold, leading to a false alarm [Ric10]. It can be expressed as the integral of the probability density function of the input signal X , which corresponds in our case to the received signal by the cloud radar:

$$P_{FA} = \int_T^{\infty} f_{dp}(X)_{H_0} dX \quad (4.3)$$

Given that non-coherent integration presents a Gaussian noise distribution [Ric10], then:

$$P_{FA} = \int_T^{\infty} \frac{1}{\sqrt{2\pi\frac{\sigma_n^2}{M}}} \exp\left(-\frac{(X-\eta)^2}{2\frac{\sigma_n^2}{M}}\right) dX = Q\left(\frac{T-\eta}{\sqrt{\frac{\sigma_n^2}{M}}}\right) = \frac{1}{2} \operatorname{erfc}\left(\frac{T-\eta}{\sqrt{\frac{\sigma_n^2}{M}}}\right) \quad (4.4)$$

where Q is the Q function and erfc is the complementary error function and η and σ_n are the mean and the variance of the receiver noise floor, respectively, obtained previously (see Table 4.7). Using the exponential approximation of Q function [Bor79] gives:

$$P_{FA} = \frac{1}{2} \frac{\exp\left(\frac{-(T-\eta)^2}{4\frac{\sigma_n^2}{M}}\right)}{\sqrt{2\pi} \sqrt{1 + \frac{(T-\eta)^2}{2\frac{\sigma_n^2}{M}}}} \quad (4.5)$$

Fig. 4.28 shows the probability of false alarm as a function of detection threshold T for different values of integrated chirp periods: 1000, 2000, 3000, 5000 and 15000 chirps.

From this theoretical basis, it is intended to obtain experimentally the curve of the probability of false alarm from the cloud radar by using conventional Monte Carlo methods and, therefore, determine the radar detection threshold from experimental data. Firstly, an analysis of the receiver noise distribution between 0.9-2.7 MHz** considering $M=15000$ is shown in Fig. 4.29, with and without spurious signals. In both cases, the corresponding Gaussian probability density function is presented, obtained from the mean and variance of the analyzed data. As is clear from the results of Fig. 4.29, the data without spurious signals

**A reduced bandwidth in comparison with histogram of Fig. 4.15 is considered to avoid the influence of low-frequency noises due to the transmitter leakage.

Figure 4.28: Probability of false alarm as a function of detection threshold T for the radar-reflectivity signal processing for different values of integrated chirp periods: 1000, 2000, 3000, 5000 and 15000 chirps. The probability of false alarm obtained from the radar data (solid black line) and an estimate (dashed black line) obtained from the mean and variance of the analyzed radar data without spurious signals are also shown.

seem to follow a Gaussian distribution. The same cannot be said of the data with spurious signals. Therefore, their corresponding bins will not be considered in the probability of false alarm estimate.

Figure 4.29: Histogram of the noise distribution of the radar data with and without spurious after non-coherent integration.

Fig. 4.28 shows the results of the probability of false alarm as a function of detection threshold T obtained from the radar data (solid black line) and an estimate (dashed black line) obtained from the mean and variance of the analyzed data without spurious signals.

The threshold is $9.18 \cdot 10^{-6}$ V for a probability of false alarm of 10^{-3} . It can be noted that the obtained curves are shifted towards higher thresholds compared to theoretical results for $M=15000$. This may be due the noise samples not being completely uncorrelated and the effect of variation over time of receiver noise temperature. The precision of the Monte Carlo method is 6.63 % in the obtained results for a P_{FA} of 10^{-3} . Lastly, for P_{FA} values higher than 10^{-5} , the measured curve is distorted due to the amount of data being relatively low to obtain a proper precision.

4.6.1.1.2 Second threshold: Removal of transmitted phase noise on detected clouds

As shown in Section 4.5.3 for a point target such as drone, the influence of the transmitted phase noise on the range profile noise can be estimated. Despite clouds are distributed targets, the phase noise influence on the range profile can be estimated by considering the detected cloud as a summation of point targets. Fig. 4.30 shows a good agreement between the range profile measurement and the phase noise estimate. In this estimate, the selected point targets correspond to the received powers that pass the double-threshold detection scheme. The individual contribution of each point target at each range of the range profile is added in natural units and then converted to decibel scale. As indicated in Section 4.5.3.1, the detection bandwidth of FFT processing and 3 dB due to the phase noise addition of the two input signals of the mixer [Coo11] should be taken into account.

Figure 4.30: Noise floor comparison between the phase noise estimate and the radar measurement of the detected cloud when there was drizzle.

Therefore, fixed a certain received power (Pr_{lim} , see Fig. 4.21) from the range profile, a maximum dynamic range (MD) is defined, and the received power for all range bins that are below this dynamic range are not represented. Fig. 4.31 shows the measured MD as a function of the maximum received power on range profiles of cloud measurement results that were carried out on January 29th, 2016. The noise power is measured at the range bin which corresponds to 5000 m. The obtained curve is fitted by a cubic function, as is shown in Fig. 4.31. This curve allows to define the MD function as a function of the maximum received power of the range profile. Furthermore, a linear function is fitted based on the lower received powers which are not affected by the transmitted phase noise. Thus, the crossing point between the linear and cubic functions defines the value of $Pr_{lim} = -63$ dBm,

which agrees with the result obtained in the drone measurement (see Fig. 4.21).

Figure 4.31: Measured MD on range profiles of cloud measurement results that were carried out on January 29th, 2016. A cubic polynomial fitting of measured MD is also plotted in order to define a MD as a function of the maximum received power of the range profile.

An example of the use of the second threshold is shown in Fig. 4.32. Three range profiles are presented, where two of them are not affected by the transmitted phase noise (black and blue curves). The red curve shows a typical range profile affected by the transmitted phase. In this case, a dynamic range of 18 dB is applied and therefore only the range bins represented with green crosses overcome the second threshold. Fig. 4.33 shows 3D reflectivity profile with and without second threshold of cloud measurement results that were carried out on January 29th, 2016. The range profiles where the second threshold is applied are highlighted in grey.

Figure 4.32: Three range profiles of detected clouds. Black and blue curves are not affected by phase noise (there was no drizzle). The noise floor of the range profile represented in red is affected by phase noise (there was drizzle), and a dynamic range of 18 dB is applied that is shown with green crosses.

Figure 4.33: 3D reflectivity profile with (top) and without (bottom) second threshold of cloud measurements obtained on January 29th, 2016. The range profiles where the second threshold is applied (and, therefore, there was drizzle) are highlighted in grey.

4.6.1.2 dBm-to-dBZ conversion

Following the Eq. (3.21) of Section 3.5, the conversion of range profiles in dBm to dBZ is straightforward, taking into account the radar parameters, as shown for a certain range profile from a detected cloud on January 29th, 2016 in Fig. 4.34:

$$dBZ = P_{rx}(dBm) + 20\log_{10}r - C(dB) - 10\log_{10}\left(\frac{\pi^3|K|^2}{512(\ln 2)}\right) - L_{atm}(r) \quad (4.6)$$

Figure 4.34: dBm-to-dBZ conversion of range profiles from detected clouds on January 29th, 2016 after 5 minutes from the initial time.

This dBm-to-dBZ conversion compensates the dependence on the squared range in Eq. (3.21). Thus, only the radar cross section of detected clouds is taken into account in the dBZ-reflectivity representation, as can be shown in the 3D reflectivity profile both in dBm and dBZ of Fig. 4.35.

Figure 4.35: 3D reflectivity profile in dBm (a) and in dBZ (b).

4.6.2 Doppler velocity

The second parameter of interest is the Doppler velocity. As indicated in Section 3.2.3, the Doppler profile is broadened because air motion and turbulences cause different velocities to the single components within the scanned volume. This means that different moments should be extracted to interpret the Doppler profile and to understand the physical properties that it represents: Mean frequency, width, skewness and kurtosis [Kol11]. The mean velocity and width velocity are obtained taking into account the spectral density of the Doppler spectrum [Coh95]:

$$V_D = \int v_f |S_t(v_f)|^2 dv_f \quad (4.7)$$

$$\sigma_D^2 = \int (v_f - V_D)^2 |S_t(v_f)|^2 dv_f \quad (4.8)$$

where $|S_t(v_f)|^2$ is the normalized spectral density of the Doppler spectrum, v_f is the falling velocity of cloud particles, V_D is the mean velocity and σ_D^2 is the width velocity of the Doppler spectrum, following the nomenclature adopted in Section 3.2.3.

Starting from the IF signal-slow time matrix of the previous procedure, the next steps are required to obtain the Doppler measurement following the scheme of Fig. 4.36:

Figure 4.36: Signal processing procedure in order to obtain the Doppler profile of the cloud profiling radar.

1. As occurs in the first step of reflectivity processing (see Section 4.6.1), the column-wise FFT is carried out to obtain the spectrum of the IF signal of each interval time ($T_m \cdot M$ seconds) and then it is transformed into the range domain following Eq. (4.2).

2. After that, two alternatives are considered to allow non-coherent integration of each range bin in the same way as in the reflectivity processing:
 - The row-wise short-time Fourier transform (STFT) for each range bin can be carried out, as proposed by [Shu08]. An analysis with different shift and overlap of samples is carried out to determine the execution time of each STFT per range bin, as shown in Figs. 4.37 and 4.38, respectively, considering a 256-point FFT. Table 4.8 summarizes the results of mean velocity, width velocity, kurtosis, skewness and execution time. Considering the deviations of the results of the Doppler-spectrum parameters and the execution time, an overlap of 128 samples and a shift of 32 samples are used in order to speed up the real-time processing.

Figure 4.37: Doppler spectrum as a function of different shifts of the STFT.

- Alternatively, the use of a single FFT per each range bin and a sliding window to smooth the resulting Doppler profile can be used. It was also proposed by [Shu04]. Fig. 4.39 shows the Doppler profile as a function of different sliding-window lengths.

Furthermore, as in the previous case, Table 4.8 summarizes the results of mean velocity, width velocity, kurtosis, skewness and execution time. It can be noted that the kurtosis, and to a lesser extent the skewness, are highly variable for the evaluated window sizes, although the execution times are lower than those obtained in the STFT processing. It was also stated in [Shu04], therefore this alternative is discarded for the cloud radar signal processing.

Figure 4.38: Doppler spectrum as a function of different overlaps of the STFT.

Figure 4.39: Doppler spectrum as a function of different sliding-window lengths of the FFT.

Table 4.8: Results of comparison between STFT overlap and shift and FFT sliding window.

Samples	Mean velocity	Width velocity	Kurtosis	Skewness	STFT execution time
STFT - Shift					
1	-0.646	0.364	0.008	2.449	0.148
16	-0.646	0.364	0.008	2.449	0.013
32	-0.646	0.364	0.008	2.449	0.003
64	-0.646	0.364	0.009	2.447	0.002
128	-0.646	0.364	0.007	2.45	0.003
255	-0.622	0.347	-0.027	2.463	0.001
STFT - Overlap					
1	-0.622	0.347	-0.027	2.463	0.002
16	-0.637	0.36	-0.03	2.446	0.002
32	-0.648	0.364	0.003	2.437	0.002
64	-0.629	0.357	-0.019	2.409	0.002
128	-0.646	0.364	0.007	2.45	0.003
255	-0.646	0.364	0.008	2.449	0.137
FFT - Sliding window					
1	-0.415	1.56	-0.71	4.532	0.001
16	-0.6	0.778	-1.737	17.6	0.001
32	-0.636	0.49	-2.117	27.56	0.001
64	-0.656	0.359	0.01	2.386	0.001
128	-0.657	0.361	0.007	2.381	0.001
256	-0.648	0.363	-0.049	2.386	0.001

3. Lastly, a cell averaging constant false alarm rate (CA-CFAR) detection scheme [Ric10] is applied to each profile, after a Doppler frequency-to-velocity conversion following Eq. (3.13). The detection threshold T_D is obtained for each Doppler profile by considering the summation of the first and last noise samples of the Doppler spectrum (5% of the total range bins R of the Doppler profile) and then multiplied by a scaling factor K , as shown in Eq. 4.9. The means of determining which threshold is the most adequate is through the curves of probability of false alarm, as shown in Fig. 4.40 for the two integration alternatives of the previous step. A scaling factor K of 1.19 for the STFT and 1.22 for the FFT with sliding window were determined for a probability of false alarm of 10^{-3} .

$$T_D = K \cdot \sum_{|i| < 0.025 \cdot R} S_t(v_f) \quad (4.9)$$

Figure 4.40: Probability of false alarm as a function of scaling factor K for the Doppler-velocity signal processing.

Fig. 4.41 shows a Doppler profile of a detected cloud at a given range of 2 km. Once the power spectrum is computed, the different Doppler moments are calculated. Furthermore, the detection threshold for this Doppler profile is represented in red. Finally, a 3D mean Doppler velocity profile is usually represented, as shown in Fig. 4.42.

Figure 4.41: Parameter extraction from the Doppler spectrum of a detected cloud at a given range of 2 km. Green crosses denote the detections that pass the detection threshold (red line).

Figure 4.42: 3D mean Doppler velocity profile.

4.7 Radar data exportation and dimensioning

At the time of this writing, a four-month campaign is being carried out in collaboration with AEMET. To this end, the cloud radar has been prepared for autonomous operation and outdoor conditions, as shown in Fig. 4.43, through waterproof metallic enclosing of the radar structure and a radome for the Cassegrain antenna to prevent the intrusion of water to the radar electronics, as described in detail in Appendix F. Furthermore, a proper cooling of radar electronics has been performed in order to prevent the deterioration of the active components. The cloud radar is managed from a remote desktop connection via an Ethernet cable. Data acquisition and processing are carried out with Matlab, which alerts the user of errors and warnings in radar operation via email. Additionally, a weather station from Oregon Scientific [Oreg17] is used to measure the rain rate, temperature and pressure close to the radar. The campaign purpose is the validation of the developed cloud radar, by using the satellite-and-radar weather information provided by AEMET. These systems provide the type and range of detected clouds every 15 minutes, therefore, the first goal of this validation will be the correlation between the reflectivity profiles of the cloud radar with the cloud information provided from the AEMET systems.

Figure 4.43: Developed cloud radar for continuous measurements in outdoor conditions.

To this end, the cloud radar data is exported in *netCDF*, a widely used format in meteorology. This format was initially developed by meteorologists studying climate models to ensure that data from different systems were interoperable and intercomparable. The netCDF header is split into a number of components, namely; dimensions, variables and global attributes, as shown below in an example file:

```
>> ncdisp('ExampleFileCloudRadar_UPM.nc')
Source:
C:\Cloud Profiling Radar\netcdf\ExampleFileCloudRadar_UPM.nc
Format:
classic
Dimensions:
time = 5
range = 800
Variables:
system (A string identifying the instrument: 'UPM 94-GHz Cloud radar')
Size: 1x1
Dimensions:
Datatype: string
day (The day of the month)
Size: 1x1
Dimensions:
Datatype: integer
month (The month)
Size: 1x1
Dimensions:
Datatype: integer
year (The year)
Size: 1x1
Dimensions:
Datatype: integer
range (A vector containing the range from the antenna, in km.)
Size: 800x1
Dimensions: range
Datatype: double
time (A vector containing the time of each captured data,
in decimal hours UTC)
Size: 5x1
Dimensions: time
Datatype: double
reflectivity (A matrix containing the radar reflectivity factor,
in dBZ.)
Size: 800x5
Dimensions: range,time
Datatype: double
doppler (A matrix containing the mean Doppler velocity, in m/s.)
Size: 800x5
Dimensions: range,time
Datatype: double
longitude (The longitude of the UPM cloud radar: 3°43'36.25"0)
Size: 1x1
Dimensions:
Datatype: double
latitude (The latitude of the UPM cloud radar: 40°27'6.98"N)
Size: 1x1
Dimensions:
Datatype: double
altitude (The altitude of the UPM cloud radar: 681 m)
Size: 1x1
Dimensions:
Datatype: double
frequency (The frequency of the UPM cloud radar in GHz: 94 GHz)
Size: 1x1
Dimensions:
Datatype: double
prf (Pulse repetition frequency: 6 kHz)
Size: 1x1
Dimensions:
Datatype: double
BW (Transmitted bandwidth: 3 MHz)
Size: 1x1
Dimensions:
Datatype: double
```

Regarding the dimensioning of the digital data processing subsystem, it is required to estimate the data storage and the required digital processing capacity.

Although the transmitted bandwidth of 3 MHz was defined in Section 4.2.2, the interest of different operational modes of the cloud radar for the analysis of different weather phenomena implies the use of different bandwidths, as proposed in [Rad14a]. To this end, the cloud radar is also configured by default to operate at three additional transmitted bandwidths (BW): 5 MHz, 10 MHz and 15 MHz, although it can be increased up to 2000 MHz, as indicated before. Therefore, with these bandwidths, the sampling frequency (f_s) is 8 MHz, 16 MHz and 32 MHz, respectively, by using Eq. (4.1) and considering a maximum range of interest of 20 km and a PRF of 6 kHz (± 4.8 m/s of maximum unambiguous Doppler velocity).

On the other hand, the number of FFT points is the product of the sampling frequency by the inverse of the PRF, that corresponds to the chirp period T_m . Because the FFT works with a numbers of points that are powers of two^{††}, the number of points N_c is the nearest power-of-two, as indicated as follows:

$$\tilde{N}_c = f_s \cdot T_m = \frac{f_s}{PRF} = \frac{4 \cdot R_{max} \cdot BW}{c} \Rightarrow N_c = 2^{\lceil \log_2(\tilde{N}_c) \rceil} \quad (4.10)$$

Fig. 4.44 shows the number of required FFTs as a function of the maximum range for different transmitted bandwidths. These results try to be realistic indicators of the required system digital processing power, since an FFT per chirp signal is required by the radar processing, as was presented in Section 4.6.

Figure 4.44: Required FFT points as a function of the transmitted bandwidth.

^{††}Due to the long period of the validation measurement campaign in collaboration with AEMET, and bearing in mind an interval time between measurements of one minute that does not imply hard-real time constraints, the use of non-power-of-two FFT has been considered for the reduction of the stored data, despite the slight increase of the processing time.

The acquisition time is directly related to the non-coherent integration process that it is mandatory in order to improve the SNR, as indicated in Section 3.5.2. Therefore, the acquisition time T_{acq} is:

$$T_{acq} = M \cdot T_m \quad (4.11)$$

The acquisition time also determines the required size of the ADC captured data. Fig 4.45 shows the captured data size in MB as a function of the acquisition time and the sampling frequency (4.8, 8, 16 and 32 MHz) considering a PRF of 6 kHz and a 16-bit ADC. Taking into account these results and that these kinds of systems are used in a four-month continuous campaigns of measurements, the data dimensioning plays a key role in the definition of operational radar parameters.

Figure 4.45: Captured data size as a function of the acquisition time and the sampling frequency considering a radar PRF of 6 kHz.

For the validation measurement campaign, considering an acquisition time of 2.5 s, i.e. 15,000 chirp signals at a PRF of 6 kHz, the data size is 22.9 MB per measurement interval. Therefore, in four months and a standby time between each measurement interval time of one minute, the total data size is ~ 9.5 TB. On the other hand, 15,000 FFTs of 1024 points should be carried in less than a minute. It has been tested with a conventional PC with 16 GB of RAM memory. The reflectivity-and-Doppler processing time of each measurement in Matlab is less than 12 seconds, that is enough for this measurement campaign, since the interval between measurements is 1 minute.

4.8 Analysis and interpretation of cloud observations

In this section some occasional observations that were carried out during the first six months of operation of the developed radar before launching the long-term validation campaign will be analyzed. The 3D radar reflectivity in dBZ and the Doppler velocity are shown. Additionally, the range-time matrix with and without the threshold is presented in order to better understand the reflectivity processing procedure. The photographs of the sky condition of each measurement are depicted in Appendix G.

The 94-GHz cloud radar observations were measured with a range resolution of 30 m and 10 s of standby time between each measurement interval time. The radar measurements are based on a non-coherent average over the measurement interval time of 2.5 s ($T_m \cdot M = 166\mu s \cdot 15000$). Therefore, the unambiguous Doppler spectrum is obtained between ± 4.8 m/s in increments of 0.0375 m/s using a 256-point FFT. Additional operational parameters pertaining to the radar measurements are summarized in Table 4.3.

4.8.1 January 29th, 2016

The measurements of January 29th, 2016 correspond to multilayered mixed-phase clouds up to a height of 3 km. Mixed-phase clouds play an important role in global climate and aircraft safety, but they are not yet completely understood in terms of multimodal Doppler modeling [Shu16]. Regarding the sky conditions, it was cloudy, as observed in Fig. 4.46. After 15 minutes, there was a slight drizzle during 20 minutes.

Figure 4.46: Sky condition at the initial time on January 29th, 2016.

Figure 4.47: Mixed-phase cloud measurement carried out on January 29th, 2016.

Firstly, the melting layer can be clearly recognized in Fig. 4.47 at a height of 1 km, both on the radar reflectivity and the Doppler 3D profiles. The melting layer is defined as the layer below the 0°C isotherm where ice particles melt and turn into liquid water particles. It can be located up to a height of 5 km [Bal06]. It appears as a band of enhanced reflectivity due to an increase of the dielectric constant of melting ice particles. Thus, it is commonly known as the *radar bright band*, as indicated by the ARM Climate Research Facility.

The Doppler spectrum also provides extra information to estimate the physical composition of mixed-phase clouds [Luk10, Kol11]. The detection of up- and downdrafts of ice particles can be easily observed in the Doppler profile because vertical air motions play an important role for the formation of mixed-phase clouds [Shu08]. At the cloud top, the ice particle formation is initiated and their movement is dominated by the air motion due to their small size. This means that positive and negative falling velocities at the cloud top should be observed in the 3D Doppler profile, as occurs in Fig. 4.47. Four Doppler spectra at different heights are presented in Fig. 4.48, after 70 minutes from the initial time: 540, 1450, 2915 and 3050 meters. The mean Doppler velocities correspond to +2.13, +1.43, -0.38 and -0.1 m/s, respectively. The negative Doppler velocities in Fig. 4.48 indicates the presence of an updraft [Shu08]. On the other hand, descending through the cloud structure, the Doppler spectra show stronger downward falling velocities that indicate larger liquid water particles. The broadening of the Doppler spectrum, clearly observed at 540 m, is due to the turbulences, as indicated in Section 3.2.3. The results of mean velocity and width velocity of the four Doppler spectra are summarized in Table 4.9.

Figure 4.48: Doppler spectrum for four different heights: 540, 1450, 2915 and 3050 meters after 70 minutes from the initial time (there was no drizzle) on January 29th, 2016.

As indicated in [Kol07], in case of drizzle, the cloud radar is working in the Mie region, instead of Rayleigh region, where η is not proportional to the fourth power of the frequency. Therefore, the radar system does not provide the right dBZ value through the reflectivity profile although some useful information is contained in the Doppler spectra. Additionally, a sharp change occurs on the falling velocity between ice and liquid water particles, that is the other main factor causing the high reflectivity in the melting layer [Bal06]. It can be discerned in the Doppler spectrum of Fig. 4.49, for four different heights (540, 1450, 2915 and 3050 meters), below and above the melting layer, after 30 minutes from the initial time, that the mean falling velocity is +2.49, +1.51, +0.23 and +1.63 m/s, respectively. The

width velocity is also summarized in Table 4.9. Furthermore, it is observed the Doppler spectrum at 540 m is wider than the 2915-m Doppler spectrum. On the other hand, two peaks are clearly observed in the Doppler spectrum at 1450 m, known as bimodal spectrum [Shu08]. It is due to the large differences in size distributions and fall regimes between the liquid droplets and ice crystals in the mixed-phase cloud, causing these two phases at different falling velocities.

Figure 4.49: Doppler spectrum for four different heights: 540, 1450, 2915 and 3050 meters after 30 minutes from the initial time (there was drizzle) on January 29th, 2016.

Table 4.9: Statistical parameters of Doppler spectra of mixed-phase cloud observation on January 29th, 2016.

Height	Mean Velocity	Width Velocity
Fig. 4.48 - No drizzle		
540 m	+2.13 m/s	0.63 m/s
1450 m	+1.43 m/s	0.41 m/s
2915 m	-0.38 m/s	0.45 m/s
3050 m	-0.1 m/s	0.41 m/s
Fig. 4.49 - Drizzle		
540 m	+2.49 m/s	0.46 m/s
1450 m	+1.51 m/s	0.32 m/s
2915 m	+0.23 m/s	0.52 m/s
3050 m	+1.63 m/s	0.36 m/s

A different way to examine the Doppler spectrum is proposed by [Shu04]. The Doppler spectrogram reveals the location and evolution with height of the detected mixed-phase cloud, as well as the detection of supercooled liquid water particles or new ice particle formation [Luk10]. The Doppler spectrogram is represented as a function of height and SNR for a certain measurement interval. The SNR is obtained as indicated in Section 3.5.2,

by subtracting the noise floor power to the received power of each range bin. Fig. 4.50 shows two Doppler spectrograms after 30 and 70 minutes. In Fig. 4.50a, the supercooled liquid water particles are detected close to the cloud top. They are small in size and, therefore, fall slower than large ice particles. The drizzle is observed up to 800 m, where the Doppler spectrum is broadened due to the large size of rain particles and turbulences. On the other hand, Fig. 4.50b shows the Doppler spectrogram without drizzle. Besides the detection of supercooled liquid water particles close to the cloud top, different sizes of liquid water particles can be differentiated. It is called multimodal Doppler spectrum, and it is caused by different particle size distributions and by variability in the air motion due to turbulences [Kol11].

(a) After 30 minutes (drizzle).

(b) After 70 minutes (no drizzle).

Figure 4.50: Doppler spectrograms at different moments of the measurement: 20 and 70 minutes from the initial time.

Fig. 4.51 shows six multimodal Doppler spectra at different heights: 1187, 1337, 1472, 1863, 2118 and 2404 meters after 80 minutes from the initial time. Excepting the Doppler spectrum at 1472 m, where three peaks are clearly observed, the remainder Doppler spectra are bimodal. It is due to the large differences in size distributions and fall regimes between the liquid droplets and ice crystals in the mixed-phase cloud, causing these two phases at different falling velocities. Modeling of these Doppler spectra evolution allows to retrieve the drop size distributions and air motion parameters [Tri15].

Figure 4.51: Examples of multimodal Doppler spectra at six different heights: 1187, 1337, 1472, 1863, 2118 and 2404 meters after 80 minutes from the initial time.

A spectral peak-picking algorithm is proposed in [Shu04] to identify the significant Doppler spectral moments of multimodal spectra by using a 35 GHz cloud radar. This algorithm determines the peak-picking criteria empirically by manual inspection, adjusting to maximize the algorithm ability to correctly identify the major spectral features. It is due to criteria strongly depending on the cloud formation stage, as indicated in [Shu04]. The following criteria are followed to identify the peaks, with the parameters adjusted to the spectra of Fig. 4.51. The criteria have been modified from [Shu04] since they do not apply a detection threshold to Doppler spectra.

- Any spectral mode must have a width of at least $0.2 \text{ m}\cdot\text{s}^{-1}$.
- For two continuous modes above the noise to be considered distinct modes, the saddle point between the peaks must be higher than 1 dB of the lowest of the two peaks.

The results of the modified spectral peak-picking algorithm are shown in Fig. 4.52, including the mean velocity of each detected peak.

Figure 4.52: Peak-picking results on multimodal Doppler spectra at six different heights: 1187, 1337, 1472, 1863, 2118 and 2404 meters after 80 minutes from the initial time.

4.8.2 February 5th, 2016

The developed cloud radar is capable of fog detection, as shown in the measurements on February 5th, 2016 (see Fig. 4.53). Fogs are a type of clouds that are composed of very small spherical water droplets less than $20\text{ }\mu\text{m}$ of diameter suspended near the surface with the top no higher than 300 m. Fog detection is currently of great interest [Li14b, Del16] because it is a significant factor that affects all forms of traffic: roads, sailing and aviation. The current efforts of the research community are devoted to retrieving the fog microphysical properties through the fog drop size distributions. To this end, millimeter-wave cloud radars in conjunction with optical sensor are required to obtain reliable fog models [Li15].

The results of Fig. 4.53 show the detection of fogs up to 250 m. The radar reflectivity factor is ~ 50 dBZ and the mean falling Doppler velocity is 0.5 m/s.

Figure 4.53: Fog measurement carried out on February 5th, 2016.

4.8.3 February 18th, 2016

The results of radar measurement carried out on February 18th, 2016 correspond to mixed-phase clouds, as shown in Fig. 4.55, in the same way as in Section 4.8.1. No significant differences are observed in terms of Doppler spectrograms or radar reflectivity factor measurements. In this case, after 40 minutes from the initial time, it began to drizzle. Furthermore, the melting layer is not so easily identified, but in the 3D Doppler profile it can be slightly appreciate around 600 m. It is worth noting that plankton (non-hydrometeoric targets such as insects or birds, as stated in radar meteorology [Lhe66]) is detected after 18 and 36 minutes from the initial time. The range profiles at these minutes are shown in Fig. 4.54. Plankton can be easily identified due it occupies a single time interval, as shown Fig. 4.55. Furthermore, as occurs with the drone measurements of Section 4.5.3, plankton is often a point target and the received power is too high and, therefore, the noise floor is dominated by the transmitted phase noise.

Figure 4.54: Range profiles after 19 minutes and 36 minutes from the initial time of the cloud measurement carried out on February 18th, 2016.

Figure 4.55: Mixed-phase cloud measurement carried out on February 18th, 2016.

4.8.4 May 12th, 2016

Fig. 4.57 shows the detection of stratiform and cumuliform clouds corresponding to the May 12th, 2016 measurement. A photograph of both clouds are shown in Fig. 4.56. Stratiform clouds are important in boundary layer dynamics and global climate and most measurements have been made with aircrafts [Fri95, Fri97]. These clouds are characterized by horizontal layering with a uniform base, as opposed to cumuliform clouds, that sometimes look like pieces of floating cotton, as observed in the reflectivity profile of Fig. 4.57. 6-7 layers of clouds such as fogs, stratus and cumulus are also detected. After 4 and 6 hours from the initial time, it began to drizzle during short periods of time (~ 10 minutes).

Figure 4.56: Detected cumulus and stratus on May 12th, 2016.

Figs. 4.58 and 4.59 show several Doppler spectra of cumulus and stratus clouds after 80 minutes and 4 hours, respectively. Table 4.10 summarizes the mean and width Doppler velocity of Doppler spectra of Figs. 4.58 and 4.59. The mean Doppler velocities results are consistent with the typical values of cumulus and stratus clouds [Pru10]. Furthermore, up- and downdrafts are observed in the Doppler profiles of the stratus clouds of Fig. 4.59, therefore it determines the presence of ice particles.

Table 4.10: Statistical parameters of Doppler spectrums of cumuliform-and-stratiform clouds measurement carried out on May 12th, 2016.

Height	Mean Doppler velocity	Width Doppler velocity
Fig. 4.58 - Cumuliform clouds		
5105 m	+1.29 m/s	0.19 m/s
5330 m	+1.29 m/s	0.18 m/s
5808 m	+1.2 m/s	0.2 m/s
6012 m	+0.9 m/s	0.16 m/s
Fig. 4.59 - Stratiform clouds		
2133 m	-0.84 m/s	0.89 m/s
3426 m	+0.74 m/s	0.93 m/s
5470 m	+1.15 m/s	0.17 m/s

Figure 4.57: Stratus and cumulus clouds measurement carried out on May 12th, 2016.

Figure 4.58: Doppler spectrum of cumulus 80 minutes (no drizzle) after the initial time for four different heights: 5105, 5330, 5808 and 6012 meters on May 12th, 2016. The mean Doppler velocities correspond to 1.29, 1.29, 1.2 and 0.9 m/s, respectively.

Figure 4.59: Doppler spectrum of stratus 4 hours (no drizzle) after the initial time for three different heights: 2133, 3426 and 5470 meters on May 12th, 2016. The mean Doppler velocities correspond to -0.84, +0.74 and +1.15 m/s, respectively.

As a summary of this section, some interesting weather phenomena are recognized in these observations through the reflectivity and Doppler profiles, such as the melting layer, up- and downdrafts, fog and supercooled liquid water particle formation. Additional information from European Organization for the Exploitation of Meteorological Satellites (EUMETSAT) Nowcasting Satellite Application Facility (NWC SAF) has facilitated the verification of the detected cloud types.

The maximum height of detected clouds is 6 km, with a radar sensitivity of -50 dBZ. This result is in good agreement with the simulation taking into account the antenna gain reduction of 7 dB (see Appendix F) and radar operational parameters of Section 4.2.2. This antenna gain variation produces a reduction of the maximum height to 10.1 km for a -40 dBZ of reflectivity. In any case, this maximum range is enough for detection of most clouds of interest for meteorologists in this Earth region, which are below the tropopause (see Fig. 1.5). On the other hand, the developed cloud radar has demonstrated its ability to penetrate several cloud layers, as indicated by [Kol07] because of the superior sensitivity to non-precipitating cloud layers of these radar systems instead of lidar systems.

Conclusions and future work

Conclusions

The overall motivation of this Ph.D. thesis has been to contribute to the technological developments at the millimeter-wave band from an experimental perspective, spanning from circuit level to system approach. This thesis has been split into two main parts, one per radar system, at 94 GHz and 300 GHz for weather applications and security tasks, respectively.

The first part is concerned with three hardware modifications for a 300-GHz imaging radar developed at the Grupo de Microondas y Radar of the Universidad Politécnica de Madrid in collaboration with the Universidad de Vigo (Spain), with the aim of developing a pre-industrial prototype for standoff sensing of person-borne concealed threats at 8 m.

The first modification has involved the development of a new microwave signal generator at 16 GHz based on a narrowband DDS/PLL architecture, that provides a radar image quality similar to the previous subsystem with a considerable reduction on cost and power consumption.

The second modification has involved the analysis and improvement of the phase stability over time in order to make the 300-GHz imaging radar ready for phase measurements that allows measuring range micrometer displacements, which is of interest for different applications from the original one. The results show that this system is capable of achieving a precision lower than $3\text{ }\mu\text{m}$ with temperature control, improving therefore the precision of other radar systems used in nondestructive inspection and biomedical applications.

Lastly, the third hardware modification has consisted in the evaluation of the imaging radar performance at a lower operating frequency, 100 GHz, based on two polarization configurations, linear and circular, with a single horn antenna. For standoff detection, it has been proven that the use of polarization techniques does not improve threat detection and the loss of image quality is clear compared to the 300-GHz system, mainly due to the frequency reduction which increases the spot size by a factor of three.

Regarding the second part of this Ph.D. thesis, it consists of the design, implementation and characterization of a ground-based cloud radar at 94 GHz. The main advantage of this system is its high sensitivity regarding other cloud radars and the radar electronics, where low cost and hardware versatility have been kept in mind.

On the other hand, the analysis of cloud observations based on 3D reflectivity profiles and Doppler spectra has shown some interesting phenomena, such as the melting layer, up- and downdrafts and supercooled liquid water particle formation on detected mixed-phase, stratiform and cumuliform clouds and fog up to 8 km of range.

This cloud radar could be an interesting tool for understanding cloud dynamics. The data results of the developed cloud radar about the dynamics of different cloud layers are complementary to data from others instruments, and it is a new instrument for study of physics and dynamics of meteorological phenomena. At this time, the developed cloud radar takes part in a four-month campaign of measurements for its validation, in collaboration with AEMET, which provides their satellite-and-radar weather information. The provided information consists of the cloud type and its height. It will be compared to the reflectivity, Doppler velocity and height data of detected clouds by the developed radar.

Guidelines for future work

There are some interesting issues for the immediate future that should be carried out in order to continue the work presented in this Ph.D. thesis:

1. ***Circuit integration.*** As indicated in the Introduction, the compactness of radar electronics is the current key goal of the imaging research community in order to develop compact transceivers. Current developments with new packaging concepts have been recently proposed [Sch15, Hit16, Fra16]. It is worth mentioning that signal generation sources including a Silicon germanium (SiGe) bipolar complementary metal-oxide-semiconductor (BiCMOS) VCO and InP mixing and power amplification stages operating at up to 330 GHz have been demonstrated [Hos15]. This approach combines the advantages of both technologies: highly complex BiCMOS analog and digital circuits are augmented with the high bandwidth and output power of InP double heterojunction bipolar transistor (DHBT) amplification and mixing stages.
2. ***Multistatic radar networks.*** There is an increasing demand to improve concealed object detection that requires advanced multistatic radar configurations. In a real-time dynamic scenario, a millimeter-wave radar network based on a multistatic radar configuration with multiple panels is aimed at creating an image of the whole body and overcoming the specular reflection limitation, as proposed in [Gon16a].
3. ***Integration of different imaging technologies.*** Active imaging systems should be combined with different kinds of sensors to fuse their information in order to improve the detection, classification, and identification of concealed threats. To this end, the integration of the imaging radar at 300 GHz with a multistatic configuration, a passive imager [Luu12, Zie13] and an ultra wideband (UWB) multiple-input multiple-output (MIMO) imaging radar system [Her16, Now16, Ble17] is feasible. In fact, several project proposals have been presented in the Horizon 2020 work programme “Secure societies - Protecting freedom and security of Europe and its citizens” [Hori14].
4. ***Integration of the cloud radar on a ground-based remote-sensing station.*** Regarding the following steps of the cloud radar, combined observations with Doppler lidar, radiometers and wind profilers should be carried out at the same location to provide synergistic results to mitigate errors in the prediction of each instrument. After the current cloud radar validation, the participation on a new measurement campaign in the European Cooperation in Science and Technology (COST) framework, such as in [Ruf06, Hae16], or in the Cloudnet project [Ill07], is desirable in order to develop an integrated ground-based remote-sensing station for atmospheric profiling.

5. ***Increase of cloud detection area.*** Cloud nowcasting in airports requires cloud detection in an area of 8-10 km², instead of a small volume obtained from a single cloud radar. To this end, two approaches can be envisaged:

- *Cloud radar network.* The easiest solution is the deployment of a cloud radar network around the area of interest. This idea has been implemented in Tokyo (Japan), where five cloud radars at 34 GHz were installed in the metropolitan area, covering ~14.000 km² [Mae15]. The main goal was to observe the cloud development from cumulus into several cumulonimbus.
- *Electronical beam steering.* Since hard real-time constraints are not imposed by this application, the area of interest can be covered by a single cloud radar with beam steering capabilities. An attractive option is based on liquid crystal reflectarrays, which allow covering the required area of interest with a beam steering delay time of a few seconds. Some developments at W-band have been carried out with excellent performance, as well as low fabrication cost and ease of integration [Bil11, Per13, Per15].

6. ***Polarimetry techniques.*** As indicated in Chapter 3, polarimetry techniques provide additional information of meteorological targets. The sensitivity of polarimetric radar parameters to size, shape, orientation, and dielectric constant of cloud particles make them suitable for use in classification algorithms. For instance, a 95-GHz polarimetric cloud airborne radar has been developed [Ayd04], allowing the cloud ice crystal classification based on a fuzzy logic scheme taking into account the linear depolarization ratio and the differential reflectivity parameters.

Part III

Appendices

Appendix A

Millimeter-wave component datasheets

The datasheets of the millimeter-wave components used in the radar systems of this thesis have been collected in this appendix. They have been sorted by manufacturer as follows:

1. RPG AFM6-110 $\times 6$ frequency multiplier.
2. RPG W-MPA-90-110 medium power amplifier.
3. RPG WPD-110 hybrid power divider.
4. RPG WFI-170 isolator.
5. RPG SHM-300 subharmonic mixer.
6. RPG FM3-300-6 frequency tripler.
7. RPG W-LNA-75-110 low-noise amplifier.
8. RPG WSC-110 waveguide single coupler.
9. Quinstar QMB-9999WS balanced mixer.
10. Quinstar QMM-4910103W frequency tripler.
11. Quinstar QAL-W00000 level set attenuator.
12. Virginia Diodes VDI6.5X3HP-1 frequency tripler.
13. Pacific Millimeter W3 tripler.
14. Farran FPA-10-0005 high-power amplifier (FTL 10627).
15. Farran FPA-10-0002 power amplifier (FTL 8089).
16. Sage 12" lens antenna.
17. Mi-Wave 48" Cassegrain antenna.
18. Astrolab 32061E 1 mm cable.
19. Flann 27373-WF60 W-band end launch adaptor.

Product Data Sheet:

Active Frequency Multiplier
Order Code: AFM6-110
Serial - No.: 205

Specifications:

AFM6-110	WR10
Frequency Range	75 - 110 GHz
RF - Interface	WR10 with Flange UG387
Multiplication Factor	6
DC-Requirement	450 mA max. @ +7 VDC
RF - IN	12,5 - 18,33 GHz
RF - Input Power	+7 dBm opt. (+10 dBm max)

Output Power [dBm] @ +7 dBm

Product Data Sheet:

W-MPA 90-110
Order Code: W-MPA-90-110
Serial - No.: 003

Specifications:

W-MPA 90-110	WR10
RF - Range	90 - 110 GHz
RF - Interface ¹⁾	WR10 / UG#387
max. RF input power	+15 dBm
DC Requirements	+5,0 VDC @ 500 mA
Gain	See Plot

Note

¹⁾ HP/A flange for 0,635" pin

Outline:

Gain and Return Loss Performance

AFM6-110 & W-MPA Output Power

Product Data Sheet:

Waveguide Power Divider
Order Code: WPD-110
Serial - No.: 101
Project - No.: 2010 271

Specifications:

WPD-110	WR10
RF - Range	90 -110 GHz
RF - Interface	WR10 with Flange UG387
max. Input Power	+20dBm

Product Data Sheet:

Waveguide Full Band Isolator
Order Code: WFI-170
Serial - No.: 10130
Project - No.: 2010 271

Specifications:

WFI-170	WR6,5
RF - Range	110 -170 GHz
RF - Interface	WR6,5 with Flange UG387
Insertion Loss	typ. 1.5dB
Isolation	> 20dB
VSWR	typ. 1.5:1
max. Input Power	+20dBm

Product Data Sheet:

Subharmonic Mixer

Order Code: SHM-300

Serial No: 103

Specifications:

SHM300	WR3.4
RF - Range	275 – 330 GHz
RF – Interface	WR3.4 with Flange UG387
LO - Interface	WR6.5 with Flange UG387
LO – Power ¹⁾	+4...+6 dBm @ RF/2
IF – Frequency ²⁾	0.01 – 8 GHz
Conversion Loss (DSB)	Typ. -8 dB

Note:

¹⁾ Maximum LO – Power +10dBm

²⁾ IF – Connector SMA-F 2.92mm

Product Data Sheet:

Type: FM3-300-6
Serial Number: 10002

Specifications:

RF Output Range	285 – 315 GHz
RF Output Interface	WR3.4 / UG-387/U-M
Multiplication Factor	3
RF Input Range	95 – 105 GHz
RF Input Interface	WR10 / UG-387/U-M
Bias DC	-5 to -7 V
Bias DC Interface	SMA-f
RF Output Power	See Plot

Absolute Maximum Ratings:

RF Input Power	+18 dBm
Case Temperature	+85 °C
Bias DC	-4 to -8 V
Bias Current	5 mA

Do not apply any RF Power if Bias DC is not applied!

Product Data Sheet:

Type: Bias DC Voltage Control
Serial Number: 2013-2

Specifications:	
DC Input Voltage	-12 V
Startup Time	500 ms
Bias DC	-2 to -11 V
Bias DC Output Interface	SMA-f
Monitor DC Output Interface	SMA-f

Absolute Maximum Ratings:	
DC Input Voltage	-15 V
Case Temperature	+85 °C
Startup Time is 500 ms!	

Product Data Sheet:

W-LNA 75-110
Order Code: W-LNA75110-20-4
Serial - No.: 2009 183-1

Specifications:

W-LNA75110-20-4	WR10
RF - Range	75 - 110 GHz
RF - Interface ¹⁾	WR10 / UG-387
max. RF input power	-20 dBm
DC Requirements	+5,0 VDC / 50mA
Noise Figure	3,8 dB
Gain	See Plot

Note

¹⁾ HP/A flange for 0,635" pin

Outline:

ALL DIMENSIONS ARE IN MILLIMETERS

Gain and Return Loss Performance

Product Data Sheet:

Waveguide Single Coupler
Type: WSC-110
Serial – Number: 20004

Specifications:

WSC-110	WR10
Frequency Range	75 – 110 GHz
Interface	WR10 / UG387/U-M
Couple Factor	-10 dB
Insertion Loss	typ. 0.9 dB

BALANCED MIXER DATA SHEET

CUSTOMER: Sematron Espana PO# 500/1111

MODEL # QMB-9999WS S/O# 12659-301S S/N 12659001001

SPECIFICATONS				
LO INPUT FREQUENCY	<u>93 to 105</u>	GHz @	<u>13</u>	dBm
RF INPUT FREQUENCY	<u>93 to 105</u>	GHz @	<u>-10</u>	dBm
IF OUTPUT FREQUENCY	<u>DC to 5</u>	GHz @	CONNECTOR	<u>SMA (F)</u>
CONVERSION LOSS	<u>8.5</u>	dB typ.		

LO INPUT TEST DATA									
FREQUENCY (GHz)	93	99	105						
POWER (dBm)	13	13	13						

RF INPUT TEST DATA	
FREQUENCY (GHz)	93 to 105
POWER (dBm)	-10

IF OUTPUT TEST DATA									
RF FREQUENCY (GHz)	93	94	95	96	97	98			
IF FREQUENCY (GHz)	DC	1	2	3	4	5			
C.L. (dB) @ 93 GHz LO	5.3	5.8	6.1	6.6	6.5	6.3			
RF FREQUENCY (GHz)	94	95	97	99	101	103	104		
IF FREQUENCY (GHz)	5	4	2	DC	2	4	5		
C.L. (dB) @ 99 GHz LO	6.8	7.0	6.3	5.6	6.5	6.8	7.0		
RF FREQUENCY (GHz)	100	101	102	103	104	105			
IF FREQUENCY (GHz)	5	4	3	2	1	DC			
C.L. (dB) @ 105 GHz LO	7.0	6.8	7.0	6.1	5.8	5.3			
RF FREQUENCY (GHz)									
IF FREQUENCY (GHz)									
C.L. (dB) @ GHz LO									

NOTE: LO Input Power +15 dBm Max. IF Input Power 0 dBm Max.

TEST TECH: B. Chen

DATE: 11/16/11

QUALITY: [Signature]

DATE: 11/17/11

S/O 13369-001

S/N 13369001001

[illegible]

NOTES:

- 1) ALL DATA TAKEN @ **25°C** UNLESS OTHERWISE SPECIFIED
- 2) DO NOT EXCEED +65.0 °C
- 3) TEST @ SMALL SIGNAL UNLESS OTHERWISE SPECIFIED
- 4) CONNECTORS **NOT** TO EXCEED 8lbs. OF TORQUE PER INCH

DATE: 8/04/2010

DATE: 8/5/10

QAL Micrometer Type Level Set Attenuators

- For level set attenuators 00 micrometer reading gives the maximum attenuation.
- To increase the attenuation turn the micrometer screw in counter clockwise direction **but do not turn further in counter clockwise when you have reached the 00 micrometer reading** as shown in picture below because it may damage the mechanical assembly or attenuation vane of the attenuator.

Attenuator set to maximum attenuation reading

- You can turn micrometer screw back in clockwise direction to decrease the attenuation.

VDI USER GUIDE

Product: WR6.5x3 Frequency Triplers

Serial Number(s): VDI-WR6.5x3R2 2-06

Virginia Diodes Inc., (434) 297-3257

082410-10180-Grupo de Microondas y Radar-VDI6.5X3HP-1

10/08/2010 JD

Section 1 User Guide Overview

The customer must read the entire User Guide prior to handling, testing, biasing, or using this product. This document gives a general description of the product, several important warnings to be considered when using the product, and basic operating instructions.

This VDI product is one WR6.5x3 frequency tripler.

Warning: Read the entire VDI User Guide for this product prior to operation.

Warning: This product can be permanently damaged by ElectroStatic Discharge (ESD). It is recommended that engineers and technicians wear a special grounded wrist strap when handling this component. In addition, the work environment around the component should be well grounded.

Warning: Opening the blocks, parts, or components will damage the internal components. VDI is not responsible for the warranty or guaranty of products damaged as a result of improper handling, testing, biasing, or use by customer.

Caution: VDI assumes the customer is familiar with microwave, millimeter wave, and VDI products. The user and customer are expected to understand all safety guidelines, health hazards, and general advisories that may exist and are associated with the use of this device. VDI is not responsible for any human hazards that may exist or may occur while using this device.

Figure 1: Photograph of similar product is shown.

Figure 2: Diagram of the product is shown.

Section 2 General Operating Procedures and Guidelines

VDI assumes the customer is familiar with VDI products, and VDI is not responsible for the warranty or guaranty of products damaged as a result of improper handling, testing, biasing, or use by customer. VDI offers the following general guidelines for using these products and recommends the customer contact VDI at (434) 297-3257 for assistance if needed. The following procedures are a quick guide for turning on and off the product. In each case the individual steps must be followed in the proper sequence to avoid damaging critical components.

Warning: RF drive limitations, voltage bias limitations, and current limitations may exist for this device and are described below. Exceeding these limitations and guidelines may cause permanent damage to the device.

The drive and bias guidelines, limitations, and recommendations for this product are:

- RF input power to the WR6.5x3 must be <180mW for safe operation (40-180 mW expected for normal operation); exceeding this limitation will destroy the device
- The WR6.5x3 does not require bias.
- Remove protective tape from waveguide flanges prior to initial operation.

Contact VDI with questions or concerns regarding the product RF drive limitations, voltage bias limitations, and current limitations.

The following procedure is to be used when **turning on** the device.

Step A: The user and test bench must be properly grounded and protected against ESD.

Step B: Connect the VDI multiplier to the user system.

Step C: Cautiously increase the RF input power (<180mW, 40-180mW expected).

The following procedure is to be used when **turning off** the device.

Step A: The user and test bench must be properly grounded and protected against ESD.

Step B: Turn off the RF input power.

Section 3 Product Performance and Guidelines

The product performance is shown below.

Figure 3: Performance of VDI-WR6.5x3R2 S2-06 (output efficiency vs. output frequency for 50mW input power) is shown.

Section 4 End of Document

VDI is not responsible for the warranty or guaranty of products damaged as a result of improper handling, testing, biasing, or use by customer.

ALL DIMENSIONS IN INCHES

W3.DAT

PACIFIC MILLIMETER PRODUCTS

OUTLINE DRAWING
TRIPLER
MODEL W3

PACIFIC
MILLIMETER
PRODUCTS

FINAL TEST DATA

FREQUENCY TRIPLER
MODEL W3 S/N 439

INPUT POWER + 20 dBm

INPUT FREQUENCY (GHz)	OUTPUT FREQUENCY (GHz)	OUTPUT POWER (dBm)
26.00	76	5.2
26.33	76	5.7
26.67	77	5.8
26.00	78	6.1
26.33	79	6.6
26.67	80	7.3
27.00	81	7.2
27.33	82	7.1
27.67	83	7.2
28.00	84	6.8
28.33	85	7.1
28.67	86	6.9
29.00	87	6.4
29.33	88	6.6
29.67	89	6.6
30.00	90	6.5
30.33	91	6.1
30.67	92	5.5
31.00	93	5.9
31.33	94	6.0
31.67	95	6.5
32.00	96	6.7
32.33	97	6.6
32.67	98	7.0
33.00	99	7.2
33.33	100	7.0
33.67	101	7.0
34.00	102	6.7
34.33	103	6.4
34.67	104	6.4
35.00	105	6.1
35.33	106	5.7
35.67	107	5.3
36.00	108	5.1
36.33	109	5.1
36.67	110	4.7

E.H.
12/18/14

FPA-10-0005

W-Band High Power Amplifier
FTL 10627

User Manual

Customer:	Universidad Politecnica de Madrid	FTL Job Number: 901068 Serial number: FTL 10627
Farran Technology LTD. Unit 1 Airport East Business Park Farmers Cross Co. Cork, Ireland Tel. +353-21-4849170 Fax. +353-21-4849192		Approvals Design: <i>Michael Crowley</i> Date: <i>27/3/15</i> Production: <i>M. J. J. J.</i> Date: <i>27/3/15</i>

Farran Technology	Manual		Date: 27-Mar-15
	FPA-10-0005 W-Band Power Amplifier		Issue: A MC
	Filename: FPA-10-0005_901068.doc		Job no.: 901068
			Page: 2/6

Table of contents

TABLE OF CONTENTS	2
1. DESCRIPTION	3
2. SPECIFICATION	3
3. INSTALLATION.....	4
4. TEST RESULTS	5
4.1 Gain Measurement.....	5
4.2 Output Power Measurement.....	6

Farran Technology	Manual		Date: 27-Mar-15
	FPA-10-0005 W-Band Power Amplifier		Issue: A MC
	Filename: FPA-10-0005_901068.doc		Job no.: 901068
			Page: 3/6

1. Description

The FPA-10-0005 is a high power W-band High Power Amplifier giving excellent output power and gain performance from 88 – 96 GHz.

The module will run from +15V power supply with a typical current consumption of 750mA in standby and 1,200mA in full output power operation.

2. Specification

WPA-10-903020 Specification:	
Frequency (GHz) :	88 - 95
Gain typical (dB) :	25
Output Power (dBm):	+29
DC+15v (mA) :	1,200 typical

Farran Technology	Manual		Date:	27-Mar-15
			Issue:	A MC
	FPA-10-0005 W-Band Power Amplifier		Job no.:	901068
	Filename:	FPA-10-0005_901068.doc	Page:	4/6

3. Installation

- The FPA-10-0005 runs from +15V
- There is no negative voltage requirement as this is provided internally in the module
- **Note output power in excess of 1W – high power mm-wave can damage – never operate without suitable terminations on the waveguide ports**
- Current consumption in Saturation (typical)
 - 1,200mA
- Amplifier is fitted with heatsinks & fans to ensure operation at room temperature – no additional heatsinking is required but mounting holes are provided if customer needs to install in system

4. Test Results

4.1 Gain Measurement

Figure 1 Gain v RF Frequency

4.2

Output Power Measurement

Output Power v RF Frequency

Figure 2 Output Power v RF Frequency 93.5 – 95 GHz

Output Power v RF Frequency

Figure 3 Output Power v RF Frequency Broadband

Output Power v RF Frequency

**Farran
Technology**

FPA-10-0002
W-Band Power Amplifier
FTL 8089

User Manual

Customer: Sample	FTL Job Number: Sample Serial number: FTL 8089
Farran Technology LTD. Unit 1 Airport East Business Park Farmers Cross Co. Cork, Ireland Tel. +353-21-4849170 Fax. +353-21-4849192	Approvals Design: <i>Michael Cowley</i> Date: <i>23/4/15</i> Production: <i>[Signature]</i> Date: <i>23/4/15</i>

Farran Technology	Manual		Date: 23-Apr-15
	FPA-10-0002 W-Band Power Amplifier		Issue: A MC
	Filename: FPA-10-0002_SAMPLE.doc		Job no.: SAMPLE
			Page: 2/6

Table of contents

TABLE OF CONTENTS	2
1. DESCRIPTION	3
2. SPECIFICATION	3
3. INSTALLATION.....	4
4. TEST RESULTS	5
4.1 S-Parameter Measurements.....	5
4.2 Saturated Output Power Measurements.....	6
4.3 DC Bias Measurements	6

Farran Technology	Manual		Date:	23-Apr-15
			Issue:	A MC
	FPA-10-0002 W-Band Power Amplifier		Job no.:	SAMPLE
	Filename:	FPA-10-0002_SAMPLE.doc	Page:	3/6

1. Description

The FPA-10-0002 is a W-band Power Amplifier with good gain and output power performance from 92 – 98 GHz.

The module will run from +4.5V to +8V power supply with a typical current consumption of 400mA.

2. Specification

Product Specification	Minimum	Typical	Maximum	Unit
RF Frequency	92		98	GHz
Linear Gain		19		dB
Psat	13	15		dBm
Input Drive Level			9	dBm
Supply Voltage	4.5	5	8	V
Current		400		mA

Farran Technology	Manual		Date: 23-Apr-15
	FPA-10-0002 W-Band Power Amplifier		Issue: A MC
	Filename: FPA-10-0002_SAMPLE.doc		Job no.: SAMPLE
			Page: 4/6

3. Installation

- The FPA-10-0002 runs from +4.5V to +8V on the input filtercon
- There is no negative voltage requirement as this is provided internally in the module
- Current consumption (typical)
 - 400mA

4. Test Results

4.1 S-Parameter Measurements

4.2 Saturated Output Power Measurements

FPA-10-0002 Saturated output power

4.3 DC Bias Measurements

+5 V @ 441 mA

94 GHz Lens Corrected Antenna Final Test Report

Customer: Giza Technologies
P.O. Number: 1599UVG-1
M/N: SAL-9239634801-094-C1
S/N: 10900-01

Date: 10/02/14
Tested by: RM
Q.A. by:

Tested Data:

Parameters	Specifications	10900-01
Frequency	94.0 GHz	94.0 GHz
Bandwidth	± 2.0 GHz	Comply
Antenna Beamwidth (Typ)	0.8° @ E-Plane, 0.7° at H-Plane	Comply
Antenna Gain (Typ)	48 dBi	48 dBi
Return Loss (Typ)	-20 dB	-30 dB
Antenna Type	12" Lens	Comply
Polarization	Linear	Comply
Sidelobes (Typ)	-20 dB	Comply

Mounting Details:

Millimeter Wave Products Inc.

2100 Tall Pines Drive
Largo, FL. 33771
Tel. (727) 536-0033
Fax. (727) 536-0012

Test Data Sheet
Part no. 223W-48/94/387
Serial No. 576

Specifications

RF Band 93 to 95 GHz

VSWR 1.3:1 Nominal Side lobe level max. -18 dbc

-3 db Bandwidth @ 94 GHz 0.2 degrees (E plane)

-3 db Bandwidth @ 94 GHz 0.2 degrees (H plane)

Gain vs Frequency

59 db @ 94 Ghz

Tested by

Date

12/8/2014

223W-48/387

CONTROL DRAWING

32061E-29840-29840-XX

NOTES:

1. DESCRIPTION
CABLE ASSEMBLY, 1mm PLUG CONNECTOR TO SAME.

2. CABLE,
COAXIAL CABLE ASTROLAB P/N 32061E
MEETS OR EXCEEDS MIL-DTL-17
SEE ASTROLAB CONTROL DRAWING
FOR MATERIALS AND FINISHES.

3. CONNECTORS -A-, 1mm PLUG,
ASTROLAB P/N 29840-32-61E
INTERFACE IS COMPATIBLE TO IEEE P287
SEE ASTROLAB CONTROL DRAWING
FOR MATERIALS AND FINISHES.

4. CONNECTORS -B-, 1mm PLUG,
SAME AS CONNECTOR -A-.

5. MARKING,
MARKER SLEEVE, FLOATING ONTO CABLE BUT
CAPTIVATED FROM FALLING OFF
MATERIAL, HEAT SHRINKABLE SLEEVING PER
SAE-AMS-DTL-23053
MARKING, .060 [1.52] MIN. TALL CHARACTERS IN
CONTRASTING COLOR AND IAW MIL-STD-130,
DATE CODE PER MIL-STD-1285
MARKING PERMANENCE IAW SAE-AMS-81531
CABLE ASSEMBLIES 4 INCHES OR SHORTER WILL HAVE
NO MARKER SLEEVE

6. ELECTRICAL CHARACTERISTICS,
IMPEDANCE,
50.0 Ohms NOMINAL
FREQUENCY, INSERTION LOSS AND VSWR,
SEE CHART
*TESTED UP TO 67.0 GHz *

7. MECHANICAL,
OPERATING TEMPERATURE RANGE,
-55° C TO +125° C
CABLE ASSEMBLY LENGTH "L"
MEASURED IAW MIL-PRF-55427

ASTROLAB PART NUMBER	DIMENSION "L" IN [mm]	18.0 GHz VSWR IL dB	4.0 GHz VSWR IL dB	65.0 GHz VSWR IL dB	*100.0 GHz * VSWR IL dB
32061E-29840-29840-2	2.00±.08 [50.8±2.0]	1.351 0.66	1.451 0.95	1.551 1.42	1.901 1.76
32061E-29840-29840-3	2.50±.08 [63.5±2.0]	1.351 0.76	1.451 1.10	1.551 1.60	1.901 1.99
32061E-29840-29840-3	3.00±.08 [76.2±2.0]	1.351 0.85	1.451 1.24	1.551 1.79	1.901 2.22
32061E-29840-29840-3	3.50±.08 [88.9±2.0]	1.351 0.94	1.451 1.38	1.551 1.97	1.901 2.45
32061E-29840-29840-4	4.00±.08 [101.6±2.0]	1.351 1.03	1.451 1.52	1.551 2.15	1.901 2.68
32061E-29840-29840-4	4.50±.08 [114.3±2.0]	1.351 1.13	1.451 1.66	1.551 2.33	1.901 2.91
32061E-29840-29840-5	5.00±.08 [127.0±2.0]	1.351 1.22	1.451 1.80	1.551 2.51	1.901 3.14
32061E-29840-29840-5	5.50±.08 [139.7±2.0]	1.351 1.31	1.451 1.94	1.551 2.69	1.901 3.37
32061E-29840-29840-6	6.00±.08 [152.4±2.0]	1.351 1.40	1.451 2.08	1.551 2.88	1.901 3.60
32061E-29840-29840-6	6.50±.08 [165.1±2.0]	1.351 1.49	1.451 2.21	1.551 3.06	1.901 3.83
32061E-29840-29840-7	7.00±.13 [177.8±3.3]	1.351 1.58	1.451 2.35	1.551 3.24	1.901 4.06
32061E-29840-29840-7	7.50±.13 [190.5±3.3]	1.351 1.68	1.451 2.49	1.551 3.42	1.901 4.29
32061E-29840-29840-8	8.00±.13 [203.2±3.3]	1.351 1.77	1.451 2.63	1.551 3.60	1.901 4.52
32061E-29840-29840-9	9.00±.13 [228.6±3.3]	1.351 1.95	1.451 2.91	1.551 3.96	1.901 4.98
32061E-29840-29840-10	10.00±.13 [254.0±3.3]	1.351 2.13	1.451 3.19	1.551 4.33	1.901 5.43
32061E-29840-29840-11	11.00±.13 [279.4±3.3]	1.351 2.32	1.451 3.47	1.551 4.69	1.901 5.89
32061E-29840-29840-12	12.00±.13 [304.8±3.3]	1.351 2.50	1.451 3.75	1.551 5.05	1.901 6.35
32061E-29840-29840-					

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<div> <div>PREP.</div> <div>EB</div> </div> <div> <div>ELEC.</div> <div>RF</div> </div> <div> <div>MECH.</div> <div>GSG</div> </div> <div> <div>Q.C.</div> <div></div> </div>	<div>DATE</div> <div>04/21/08</div> <div>05/05/08</div> <div>05/05/08</div>	<div>NAME</div> <div>EB</div> <div>RF</div> <div>GSG</div> <div></div>	<div>CODE IDENT.</div> <div>16301</div>
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32061-29840-29840-6

INSTRUMENT SPECIFICATIONS MANUAL

ISM Instrument, Report

Authorised User

Alan Long 18/07/2013
Sales Manager

Spec # 435

(29 May 2013)

H7626

27373-WF60 UG-387/ U-Mod

WG27 WR10 R900
End Launch Adaptor
Standard Instrument

Series Specific Specifications:

Non-Defined

Instrument Specific Specifications:

Microwave Specifications

Parameter:	Value:
Frequency band	73.8GHz to 112GHz
VSWR	1.40 MAX
Insertion loss	0.55dB MAX

Instrument Interface Specifications

Parameter:	Value:
Flange	UG-387/U-Mod (Provisional value)
Connector	1.0mm Female

Mechanical Specifications

Parameter:	Value:
Base Material	Brass
External Finish	Gold Plated & stainless steel
Interface Finish	Gold Plated

Environmental Specifications

Parameter:	Value:
Operating ambient temperature	-30°C to 70°C
RoHS compliant	Yes

Dimension Specifications

Parameter:	Value:
A	10.1mm
B	4.1mm

Electrical Specifications

Parameter:	Value:
Grounded centre conductor	Yes

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Appendix B

Characterization of millimeter-wave components of the cloud profiling radar at 94 GHz

This appendix presents the S-parameters results and the output power measurements of the millimeter-wave components, whose datasheet are collected in Appendix A. These components are part of the millimeter-wave subsystem of the cloud profiling radar at 94 GHz, whose block diagram is shown in Fig. 4.3. The power measurements at W-band have been obtained by means of an Erickson PM4 Power meter from VDI and the S-parameters results have been obtained with a 8757D scalar network analyzer from Keysight Technologies (formerly Agilent Technologies).

B.1 W3 tripler from Pacific Millimeter

The Pacific Millimeter tripler provides the W-band chirp signal to the MPA (test point D). The return losses at the input and output ports of the tripler are presented in Fig. B.1.

The output power as a function of the doubler input power is presented in Fig. B.2. The tripler input signal of 20 dBm is generated by using the doubler input due to the lack of lab instrumentation to provide this power level at this frequency band. The expected output power is obtained with input powers of 2 dBm and 3 dBm, according to the manufacturer datasheet. However, the measured output power above 95 GHz is lower than the output power indicated by the manufacturer. It is due to the doubler and MPA module performance, whose output power drops at the end of the measured bandwidth (32-33 GHz, see Fig. 4.8).

Figure B.1: x3 frequency multiplier: S_{11} -and- S_{22} parameters measurement.

Figure B.2: x3 frequency multiplier: Output power measurement as a function of the doubler input power.

B.2 QMB balanced mixer from QuinStar Technology, Inc.

The balanced mixer is characterized by using the LO signal of the radar system (test point E). According to the mixer datasheet, the recommended LO input power level is 11-13 dBm, providing an RF-IF conversion loss of about 7 dB. However, the maximum achieved LO power is 6 dBm (see Fig. 4.11). The resulting RF-IF conversion losses are about 7.5 dB with 1.5-dB ripple, as shown in Fig. B.3. The return losses at the RF and LO input ports and the LO-RF isolation are also shown. Return losses are better than 7 dB at both input ports and the LO-RF isolation is higher than 20 dB.

Figure B.3: QuinStar balanced mixer measurements: RF-IF conversion losses, return losses at RF and LO ports and the RF-LO isolation.

B.3 LNA from Radiometer Physics GmbH

The LNA S-parameter results are shown in Fig. B.4. The gain is 25 dB with 2 dB ripple and the return loss are higher than 9 dB, according to the manufacturer datasheet.

B.4 MPA from Farran Technology, Ltd.

The MPA S-parameter results are shown in Fig. B.5. It presents a gain of 15 dB, an input return loss below 15 dB and a poor output return loss of 5 dB at the operating frequency of the radar system .

The MPA output power as a function of the input power of the doubler-MPA module (see Section 4.4.2.1) is presented in Fig. B.6. In this case, the maximum input power at test point F is obtained with a doubler input power of 2 dBm.

Figure B.4: RPG low-noise amplifier: S-parameters measurement.

Figure B.5: Farran MPA amplifier: S-parameters measurement.

Figure B.6: Farran MPA amplifier: Output power measurement as a function of the doubler input power.

B.5 x6 frequency multiplier from Radiometer Physics GmbH

The x6 frequency multiplier has been characterized as a function of its input power. Fig. B.7 shows the output power curves versus input power levels from 6 to 12 dBm. As shown, the optimal input power is in the range from 9 to 12 dBm, according to the output power at the test point B (see Fig. 4.3).

Figure B.7: RPG x6 frequency multiplier: Output power measurement as a function of its input power.

The input return loss results are presented in Fig. B.8, which are better than 15 dB at the operating frequency of the radar system.

Figure B.8: RPG x6 frequency multiplier: S_{11} -parameter measurement.

B.6 HPA from Farran Technology, Ltd.

The MPA S-parameter results are shown in Fig. B.9. The MPA presents a gain of 23 dB, an input return loss of 17 dB and a low output return loss of 6 dB at the operating frequency of the radar system, according to the manufacturer datasheet.

Figure B.9: Farran HPA amplifier: S-parameters measurement.

The relationship between the output power and the current consumption as a function of the HPA input power is presented in Fig. B.10. Furthermore, the PAE results are also shown.

Figure B.10: Farran HPA amplifier: Output power, current consumption and PAE measurement results as a function of the HPA input power.

Appendix C

Active filters design

C.1 Introduction

In this Appendix, the design, fabrication and testing of two active bandpass filter are presented for the two radar systems of Parts I and II of this dissertation.

As indicated in Section 3, an active bandpass filter is included before the digitization of the IF signal in order to select a suitable range of distances of interest. The reason for the active bandpass filter is twofold. On the one hand, it reduces the low-frequency noise due to $1/f$ noise and the leakage noise due to the coupling between the transmitter and the receiver. On the other hand, it limits the contribution of thermal noise at higher frequencies.

The filter technology is selected as a function of the beat frequency of interest and the fractional bandwidth of the required filter. Fig. C.1 shows some filter technologies as a function of the fractional bandwidth and the center frequency [Mur11].

As regards the standoff imaging radar at 300 GHz, considering a maximum standoff distance of 10 m, and a chirp period of 150 μ s, the maximum beat frequency is 12 MHz using equation (2.1). The minimum beat frequency is 1 MHz, in order to reduce the $1/f$ noise and the leakage noise. Therefore, considering a bandwidth of 11 MHz centered at 6.5 MHz, the fractional bandwidth is 169%. On the other hand, regarding the cloud profiling radar at 94 GHz, the maximum range of interest is 20 km. This means that the maximum beat frequency is 2.4 MHz taking into account the operational parameters of Table 4.3. The minimum beat frequency is 40 kHz, which corresponds to a distance of 100 m. Therefore, considering a bandwidth of 2.4 MHz centered at 1.2 MHz, the fractional bandwidth is 200%.

Taking into account the high fractional bandwidth, an LC filter combined with RF low-noise amplifiers was selected for this application.

Figure C.1: Frequency range for some filter technologies as a function of the fractional bandwidth and center frequency. Graph extracted from [Mur11].

C.2 Filter I: Active filter for the imaging radar at 300 GHz

C.2.1 Simulation results

Circuitual simulation of this filter and its optimization has been carried out with the commercial software ADS from Agilent Technologies. The LC bandpass filter was designed with the Filter Design Guide tool from ADS. Subsequently, the LC bandpass was inserted between two RF low-noise amplifiers from Sirenza (SGA-3486Z model). Two amplifiers are required to achieve a gain higher than 40 dB in order to observe the receiver noise floor above the digitizer noise floor. The SGA-3486Z amplifier presents a gain between 19-23 dB and a noise figure of 3.2 dB. Another important requirement for the filter design is that the bandpass ripple should be minimized (up to 1 dB) in order to avoid linear distortion.

Fig. C.2 shows the schematic of the active bandpass filter based on two low-noise RF amplifiers and a seven-order LC bandpass filter, which is enough for filtering out thermal noise at higher frequencies.

Considering the non-ideal behavior of inductors due to the fact that real wires have a small resistance, which is frequency dependent, the S-parameters provided by the inductor manufacturer (Murata Manufacturing Co., Ltd.) have been considered in the simulation. The selected inductors present a extremely high quality factor and a low resistance. The schematic is shown in Fig. C.3.

The simulation results of the S-parameters of the designed bandpass filter are shown in Fig. C.4, taking into account the ideal (dashed line) and real (solid line) behavior of Murata inductors. The measured bandwidth at -1 dB is 0.8-14.6 MHz. The maximum gain is 46.8 dB considering ideal inductors, and 1.5 dB of insertion loss are obtained by considering the real behavior of inductors.

C.2.2 Characterization results

The designed filter has been fabricated using commercial surface-mount technology (SMT) components, and then assembled on custom-made technology circuits. This circuit has been designed by using a commercial CuClad substrate ($\epsilon_r=2.17$, $h=0.254$ mm). A photograph of the active filter is shown in Fig. C.7. The S-parameter measurement results of the active bandpass filter are shown in Fig. C.5. The maximum gain is 44.4 dB and the measured bandwidth at -1 dB is 1.11-13.5 MHz, which is in good agreement with the simulation results. The additional loss (~ 0.6 dB) can be related to the effect of manufacturing process tolerances of the amplifier, as specified by the manufacturer in the datasheet (± 2 dB of gain variation across the typical value).

Another parameter of interest is the power compression of the filter, in order to estimate the input 1 dB compression point at the filter input for preventing ADC saturation. To this end, the 1 dB compression point of the active filter is measured. The results are shown in Fig. C.6. The input 1 dB compression point of the active filter is -34 dBm, which is enough for preventing ADC saturation (14 dBm).

Figure C.2: Schematic of the active bandpass filter of the imaging radar at 300 GHz (filter I).

Figure C.3: Schematic of the active bandpass filter of the imaging radar at 300 GHz (filter I) taking into account the S-parameters of the inductors used.

Figure C.4: S-parameter simulation results of filter I taking into account the ideal (dashed line) and real (solid line) behavior of Murata inductors.

Figure C.5: S-parameter measurement results of filter I.

Figure C.6: Power compression measurement results of filter I.

Figure C.7: Photograph of the active bandpass filter of the imaging radar at 300 GHz (filter I).

C.3 Filter II: Active filter for the cloud profiling radar at 94 GHz

C.3.1 Simulation results

The same simulation procedure is carried out for the active filter of the cloud profiling radar at 94 GHz. Taking into account its bandwidth requirements, a fourth-order bandpass filter is enough for this system, since the required bandwidth is lower than the required bandwidth of filter I. In this case, only one RF low-noise amplifier is needed, since 20 dB is enough to observe the receiver noise floor above the digitizer noise floor. Fig. C.8 shows the schematic of the proposed circuit.

As indicated before, the non-ideal behavior of inductors can be considered in the simulation by means of their S-parameters. The schematic of this simulation is shown in Fig. C.10. As occurs before, 1.5 dB of insertion losses are obtained by considering the real behavior of inductors, as shown in Fig. C.10.

C.3.2 Characterization results

This active filter is fabricated in the same way as the previous active filter. A photograph of the active filter is shown in Fig. C.13. The S-parameter measurement results are shown in Fig. C.11. The maximum gain is 20.5 dB and the measured bandwidth at -1 dB is 0.005-4.3 MHz, which is in good agreement with the simulation results. As indicated before, the additional loss (~ 1.3 dB) can be related to the effect of manufacturing process tolerances of the amplifier, as specified by the manufacturer in the datasheet (± 2 dB of gain variation across the typical value).

The 1 dB compression point measurement of the active filter is shown in Fig. C.12. The input 1 dB compression point of the active filter is -14 dBm.

A comparison between simulation and measurement results is shown in Table C.1.

Table C.1: Comparison of simulation and measurement results of the active filters developed for the imaging radar at 300 GHz (I) and the cloud profiling radar at 94 GHz (II).

Filter	BW ($ S_{21} _{max}-1$ dB) (MHz)		BW ($ S_{11} < -10$ dB) (MHz)		Max. gain (dB)		P _{1dB_{in}} (dBm)
	Simul.	Meas.	Simul.	Meas.	Simul.	Meas.	Meas.
I	1.17-13.74	1.11-13.51	0.44-15.17	0.7-14.51	44.9	44.3	-34
II	0.007-4.48	0.005-4.3	0.008-4.95	0.005-4.7	21.8	20.5	-14

Figure C.8: Schematic of the active bandpass filter of the cloud profiling radar at 94 GHz (filter II).

Figure C.9: Schematic of the active bandpass filter of the cloud profiling radar at 94 GHz (filter II) taking into account the S-parameters of the used inductors.

Figure C.10: S-parameter simulation results of filter II taking into account the ideal (dashed line) and real (solid line) behavior of Murata inductors.

Figure C.11: S-parameter measurement results of filter II.

Figure C.12: Power compression measurement results of filter II.

Figure C.13: Photograph of the active bandpass filter of the cloud profiling radar at 94 GHz (filter II).

Appendix D

Receiver noise power level calculations of the cloud profiling radar at 94 GHz

The receiver noise power calculations of the developed cloud profiling radar are shown below. Fig. D.1 shows the block diagram of the receiver chain at 94 GHz. The parameters required for the calculation of the noise figure and temperature of the system at 94 GHz are included in Table D.1.

Figure D.1: Block diagram of the 100-GHz receiver.

Firstly, the noise power at the input of the active filter, $P_{n_{out}}$, is computed in order to define the active filter gain, taking into account the ADC noise power (-102 dBm, see Fig. 4.14). By using the Friis formula, the expression of the noise factor at the receiver input is as follows:

$$F = F_{isol} + \frac{F_{LNA} - 1}{G_{isol}} + \frac{F_{mix} - 1}{G_{isol}G_{LNA}} \quad (D.1)$$

The noise figure, the equivalent noise temperature and the receiver's gain are:

$$F(dB) = 4.63 \text{ dB} \quad (D.2)$$

Table D.1: Receiver parameters at 94 GHz.

Parameter	Value
Isolator insertion losses ($L_{isol}(\text{dB})$)	0.8 dB
LNA gain ($G_{LNA}(\text{dB})$)	25 dB
LNA noise figure ($F_{LNA}(\text{dB})$)	3.8 dB
Double balanced mixer DSB noise figure ($F_{mix}(\text{dB})$)	7.4 dB
Antenna noise temperature (T_a)	60 K*
Standard noise temperature (T_0)	290 K
Boltzmann constant (k)	$1.38 \cdot 10^{-23}$ J/K

$$T_e = T_a + T_0(F - 1) = 612.1 \text{ K} \quad (\text{D.3})$$

$$G_{Rx} = \frac{G_{LNA}}{L_{isol}L_{mix}} \Rightarrow G_{Rx}(\text{dB}) = 16.8 \text{ dB} \quad (\text{D.4})$$

The noise power level has to be integrated over the FFT bandwidth $B_{FFT}=8.1$ kHz, which corresponds to the equivalent noise bandwidth for Hamming windowing of the FFT bandwidth for rectangular windows ($B_{FFT}=1.36 \cdot B_{FFT-R}=1.36 \text{PRF}=1.36 \cdot 6 \text{kHz}=8.1$ kHz) [Har78]. From these data, the receiver noise power (P_{nout}) at the receiver output is:

$$P_{nout} = kT_e B_{FFT} G_{Rx} \Rightarrow P_{nout}(\text{dBm}) = -114.8 \text{ dBm} \quad (\text{D.5})$$

Therefore, the noise power at the active filter input is ~ 13 dB lower than the ADC noise power. If a power margin of 7-8 dB is defined to clearly exceed the ADC noise power, the active filter gain should be 20-21 dB.

Lastly, the noise power at the active filter output ($P_{n_{AFout}}$) is calculated taken into account its influence on the overall noise figure. The active filter gain G_{AF} is 20.5 dB and the noise figure F_{AF} is 3.2 dB, as indicated in Appendix C. Therefore, the noise figure and the equivalent noise temperature referred at the active filter output are:

$$F = F_{isol} + \frac{F_{LNA} - 1}{G_{isol}} + \frac{F_{mix} - 1}{G_{isol}G_{LNA}} + \frac{F_{AF} - 1}{G_{isol}G_{LNA}\frac{1}{L_{mix}}} = 4.67 \text{ dB} \quad (\text{D.6})$$

$$T_e = T_a + T_0(F - 1) = 619.9 \text{ K} \quad (\text{D.7})$$

Before calculating the noise power at the active filter output, the overall gain is:

$$G_{Rx-AF} = \frac{G_{LNA}G_{AF}}{L_{isol}L_{mix}} \Rightarrow G_{Rx}(\text{dB}) = 37.3 \text{ dB} \quad (\text{D.8})$$

Therefore, the noise power at the active filter output is:

$$P_{nout_{FA}} = kT_e B_{FFT} G_{Rx-FA} \Rightarrow P_{nout}(\text{dBm}) = -95.4 \text{ dBm} \quad (\text{D.9})$$

The Friis formula does not take into account the mismatch between components. This affects the noise power estimation, in particular, the mismatch between the LNA and the balanced mixer. To this end, a noise budget analysis in ADS of the receiver noise power, as shown in Fig. D.2, is carried out in order to take into account this mismatch between components by using the S-parameter measurements of the receiver components. The results of Fig. D.3 are 1.4 dB worse (higher noise) than the results obtained by means of Friis formula. As shown in Fig. 4.14, this result agrees with the measured noise power at the receiver output (0.2 dB of error) due to the fact that it provides more precise results than those obtained by the Friis formula.

Figure D.2: Schematic of the receiver chain for ADS budget analysis of noise power.

Figure D.3: Results obtained by an ADS noise budget analysis of the receiver noise power levels at the output of each component.

Appendix E

Automatic biasing sequencer design

E.1 Introduction

In this Appendix, the design, fabrication and testing of an automatic biasing sequencer is presented. The geometry of this device has been adapted to the geometric constraints of the developed doubler-MPA module (see Section 4.4.2.1), since the sequencer is located just above the doubler module, as shown in Fig. E.1.

Figure E.1: Photographs of the automatic biasing sequencer.

Some MMICs require the user to bias the device manually in order to prevent damage

to the device [Hit14]. Specific circuits have been developed in order to bias MMICs automatically, such as the HMC920LP5E active bias controller from Hittite. This circuit also provide a stable bias current over temperature and process variation and aging.

The developed doubler-MPA module presented in Section 4.4.2.1 is based on two Hittite MMICs: HMC598 doubler and HMC499 MPA. The HMC598 doubler requires three power supplies, one positive power supply (5 V) for the drain and two negative supplies (-1.25 V and -0.8 V) for the gate, that are adjusted between -2 V to 0 V to achieve $I_d=140$ mA typical. The HMC499 medium-power amplifier requires one positive power supply (5 V) for the drain and one negative supply (-1.4 V) for the gate, which is adjusted between -2 V to 0 V to achieve $I_d=200$ mA typical. To obtain the typical drain current consumption of both MMICs, the gate power supplies of the doubler are -1.25 V and -0.85 V, and the gate power supply of the amplifier is -1.4 V. The power-on sequence consists on the activation of the three gate power supplies and afterwards the two drain power supplies, starting with supplies corresponding to the MPA, since it is the last component of the developed module.

To this end, an adjustable dual voltage regulator LT3029 from Linear Technologies has been chosen to provide the drain power supply to the MMICs. In order to obtain the negative gate power supply, a buffer operational amplifier (LT6200 from Linear Technologies) combined with a potentiometer to adjust to the optimal biasing point between -2 V to 0 V have been used.

E.2 Drain power supply

The LT3029 is a dual low noise, low dropout micropower linear regulator. The device is capable of supplying 500 mA of output current per channel with a dropout voltage of 300 mV. A key feature of the LT3029 regulator is its low output noise of $20 \mu V_{RMS}$ between 10 Hz to 100 kHz bandwidth. This regulator is stable with output tantalum capacitors as low as $3.3 \mu F$. Internal protection circuitries include reverse battery protection, current limit with foldback and thermal shutdown. The adjustable version of the LT3029 has an output voltage range of 1.8 to 20 V. The output voltage is set by the ratio of two external resistors (R_1 and R_2 of the first channel, and R_3 and R_4 of the second channel), as shown in Fig. E.2. The device servos the output to maintain the ADJ pin voltage at 1.22 V referenced to ground. The output voltage can be calculated using the Eq. (E.1). The value of the top resistance should not be greater than 250 k Ω to minimize errors in the output voltage caused by the ADJ pin bias current.

$$V_{out} = 1.22V \left(1 + \frac{R_2}{R_1} \right) + I_{adj} \cdot R_2 \quad (E.1)$$

$$5V = 1.22V \left(1 + \frac{R_2}{R_1} \right) + 30nA \cdot R_2 \Rightarrow Si R_2 = 62K\Omega \Rightarrow R_1 = 20K\Omega \quad (E.2)$$

Therefore, it is necessary to use a 10 K Ω potentiometer (R_{12} and R_{122} in schematic of Fig. E.2) in series with 15 K Ω to achieve an output voltage variable between 4 V and 5.5 V.

A bypass capacitor (C_2 and C_4 in Fig. E.2) of 10 nF from OUT pin to the BYP pin has been added to lower output voltage noise. A good quality low leakage capacitor is used. A $3.3 \mu F$ output capacitor (C_3 and C_5 in Fig. E.2) will prevent oscillations on the output. The ESR of the output capacitor affects stability, most notably with small capacitors.

The SHDN pin is used to put the LT3029 regulators into a low power shutdown state. The output will be off when the SHDN pin is pulled low. The SHDN pin can be driven either by 5V logic or open-collector logic with a pull-up resistor.

E.3 Gate power supply

The gate power supply is obtained with the LT6200 operational amplifier from Linear Technologies, configured as a buffer amplifier, and a potentiometer to adjust the negative supply from 0 V to -2 V. The LT6200 device is a low-noise ($0.95 \text{ nV}/\sqrt{\text{Hz}}$ at 100 kHz) and rail-to-rail operational. The gain bandwidth product is 165 MHz. It also allows power and thermal shutdown through the SHDN pin. As shown in Fig. E.2, a 1 K Ω potentiometer (R5 with R7, R8 with R17 and R18 with R21 for each gate output voltage) is used as a power divider to adjust the output voltage between 0 V and -2 V. At the output, a power divider with fixed-value resistors (R12 with R6, R16 with R14 and R20 with R19 for each gate output voltage) is used to limit the minimum output voltage up to -2 V, in order to not damage the device, since this value is the minimum value indicated by Hittite.

E.4 Power-on sequencer

Power supply must follow a sequence to not damage the MMICs [Hit14]. This sequence must guarantee the correct order in power supply activation. First, the gate power supply signal must be activated and afterwards drain power supply. To this end, the designed circuit incorporates two power MOSFETs and an RC circuit that act as an interrupt in series with the LT3029 voltage regulator output (see Fig. E.3). This circuit is in ON state when and RC circuit reaches its final charge. While the RC circuit is charging the power MOSFETs are off, as shown in Fig. E.4. This delay can guarantee power on sequence in all cases. In this case, the potentiometer of the RC circuit is configured to generate a delay of 1 s and 4 s for the drain power supply of the HMC499 MPA and HMC598 doubler, respectively.

Power MOSFETs are used in active bias controller instead of Bipolar Junction Transistors (BJTs) [Tor04] due to their high-input impedance and to the fact that they are a majority carrier device, so they not do not suffer from minority carrier storage time effects, thermal runaway or second breakdown [Loc98]. The IRF7316 power MOSFET from International Rectifier is used in this circuit. It is a dual P-Channel MOSFET, based on Hexagonal Field Effect Transistor (HEXFETTM) technology and the gate threshold voltage is -1 V.

Figure E.3: Schematic of a voltage sequencer using a RC-circuit and two power MOSFETs.

Figs. E.5 and E.6 show the simulation results of the output voltages and the output noise voltage of the drain and gate power supplies, respectively. Both simulations have been

Figure E.4: Output voltage at the RC-circuit and gate outputs of the automatic biasing sequencer.

carried out with LTspice IV from Linear Technologies which provides a variety of custom design simulation tools and device models to quickly and easily evaluate circuits. Regarding the noise output voltage of the gate power supply, the obtained simulations results are better than other alternatives, such as adjustable negative regulators (for instance, LT1964 from Linear Technologies), whose output noise voltage is two orders of magnitude higher than the output noise voltage of the selected scheme.

Figure E.5: Output voltages of the automatic biasing sequencer.

Figure E.6: Noise voltage at the drain and gate outputs of the automatic biasing sequencer.

E.5 Measurements

The output power measurements with the WaveSurfer10 oscilloscope from Teledyne LeCroy, Inc. are shown in Fig. E.7. As indicated before, a delay of 1 s and 4 s for the drain power supply of the HMC499 MPA and HMC598 doubler, respectively, is configured through the potentiometers. Furthermore, the drain and gate power supplies of both MMICs are within the expected values.

Additionally, the noise voltage at the gate and drain outputs are measured with the frequency-domain noise-measurement option of the oscilloscope. The results are $5.95 \text{ nV}/\sqrt{\text{Hz}}$ and $1.46 \text{ nV}/\sqrt{\text{Hz}}$, respectively, which are in good agreement with simulation results.

Figure E.7: Output voltage measurements at the drain outputs of the automatic biasing sequencer.

Appendix F

Weathering protection for a Cassegrain antenna

F.1 Introduction

The weathering protection of the Cassegrain antenna is mandatory for the continuous outdoor measurement campaign that will be carried out in collaboration with AEMET. To this end, some alternatives have been explored in order to protect the Cassegrain antenna. Although there exist commercial radome alternatives, in this Appendix a cost-effective solution has been designed and then characterized in the anechoic chamber of the *Laboratorio de Ensayos y Homologación de Antenas* (LEHA)-UPM.

Instead of classical closing of the Cassegrain antenna with a drum that includes a removable window whose surface is the radome material, we have considered only closing the antenna feeder. Additionally, a rain drainage system is implemented by using four 15-mm diameter pipes that are fixed to the feeder by a tail-piece threaded adapter, as shown in Fig. F.1.

The radome material is Oratex 600 from *Better Aircraft Fabric, Inc.* They offer this material for aircraft covering systems, but it is also adequate for use as a low-loss antenna radome at W band. Some antenna manufacturers use this material at this band due to its power losses lower than 0.5 dB. They recommend that it should be fitted at an oblique angle, so that any reflection is diverted off axis so as not to degrade the feeder return losses. The Oratex specifications are listed in Table F.1.

In this Appendix the two developed radomes will be shown: Flat and inclined, as shown in Figs. F.2-F.3. In both cases, two sealing rings are added to the top and bottom of the radome structure in order to prevent drilling fluid ingress and to maintain proper pressure sealing. Furthermore, silica gel dryer is enclosed in the free space between the antenna feeder and the radome in order to minimize condensation growth. On the other hand, the inclination angle of the inclined radome is 20° to facilitate water runoff.

The compact-range anechoic chamber measurements of the feeder and the Cassegrain antenna considering both developed radomes are shown below.

Figure F.1: Details of the drainage systems and fixing screws for the radome on the feeder structure.

Table F.1: Oratex 600 specifications.

Parameter	Value
Material	Modified polyester
Material thickness	120 μm
Weight	92 - 122 g/m^2
Tensile strength lengthwise	750 - 1050 $\text{N}/50 \text{ mm}$
Tear propagation load lengthwise	18 - 24 N
Burst strength	500 - 700 kPa
Shrinkage lengthwise	(170 $^{\circ}\text{C}$) 8 - 12 %
Melting point	250 $^{\circ}\text{C}$
Air permeability	Airtight

Figure F.2: Photograph of the flat radome.

Figure F.3: Photograph of the inclined radome.

F.2 Feeder measurements

The antenna feeder was measured in the anechoic chamber with both radomes, as shown in Fig. F.4, in order to obtain the co-polar and cross-polar radiation pattern and two main parameters: beamwidth and cross-polar discrimination (XPD).

Figure F.4: Photograph of the antenna feeder under test in the anechoic chamber.

The measured co-polar and cross-polar radiation pattern of the feeder without radome, with flat radome and with inclined radome are shown in Figs. F.5-F.7, respectively. The HPBW is 12.3° and the XPD is ~ 25 dB in the three cases.

Fig. F.8 shows the radome power losses. The flat radome has 0.6 dB of power losses, and the inclined radome has practically no losses. This result would suggest that the power losses are due to mismatch of the feeder with the flat radome. To this end, the return losses in the three cases are measured, as shown in Fig. F.9. The return losses with the flat radome are slightly degraded in comparison with the inclined radome or without radome, which are practically identical. Therefore, the power losses depend directly on the incidence angle of the radome material.

Figure F.5: Measured co-polar and cross-polar radiation pattern of the feeder without radome.

Figure F.6: Measured co-polar and cross-polar radiation pattern of the feeder with flat radome.

Figure F.7: Measured co-polar and cross-polar radiation pattern of the feeder with inclined radome.

Figure F.8: Power losses of the developed radomes measured in the co-polar radiation pattern of the feeder.

Figure F.9: S_{11} results comparison of the Cassegrain antenna feeder with the developed radomes.

F.3 Cassegrain antenna measurements

The Cassegrain antenna was measured in the compact-range anechoic chamber with and without both radomes, in the same way as the antenna feeder, as shown in Fig. F.10 in order to obtain the co-polar and cross-polar radiation patterns and two main parameters: beamwidth and XPD.

Figure F.10: Photograph of the Cassegrain antenna under test in the anechoic chamber.

The measured co-polar and cross-polar radiation pattern of the feeder without radome, with flat radome and with inclined radome are shown in Figs. F.11-F.13, respectively. The HPBW is 0.18° at $\phi = 0^\circ$ and 0.3° at $\phi = 90^\circ$ in the three cases. The XPD is ~ 20 dB in the three cases also.

Fig. F.14 shows the Cassegrain power losses with the developed radomes. The flat radome has 0.4 dB of power losses, and the inclined radome presents only 0.2 dB. The Cassegrain antenna gain is 51.5 dB, which is much lower than the value of 59 dB indicated by the manufacturer (see Appendix A). The reasoning behind this result difference is currently being discussed with the manufacturer, since the test report of the purchased antenna indicates a gain of 59 dB at 94 GHz.

Additionally, the return losses in the three cases are measured, as shown in Fig. F.15. The return losses with the flat radome are slightly degraded in comparison with the inclined radome or without radome, which are practically identical. In all cases, the return losses are lower than 20 dB.

Figure F.11: Measured co-polar and cross-polar radiation pattern of the Cassegrain antenna without radome.

Figure F.12: Measured co-polar and cross-polar radiation pattern of the Cassegrain antenna with flat radome.

Figure F.13: Measured co-polar and cross-polar radiation pattern of the Cassegrain antenna with inclined radome.

Figure F.14: Power losses of the developed radomes measured in the co-polar radiation pattern of the Cassegrain antenna.

Figure F.15: S_{11} results comparison of the Cassegrain antenna with the developed radomes.

Appendix G

Photographs of the sky condition at the beginning of the measurements with the cloud profiling radar at 94 GHz

During the occasional radar measurements, a series of photographs at the measurement start were made from the roof of A building of the ETSIT. This complements the experimental analysis and is useful to observe in combination with the results of Section 4.8.

G.0.1 January 29th, 2016

Figure G.1: Sky condition on January 29th, 2016 at 10:21 AM.

G.0.2 February 5th, 2016

Figure G.2: Sky condition on February 5th, 2016 at 15:19 PM.

G.0.3 February 18th, 2016

Figure G.3: Sky condition on February 18th, 2016 at 10:39 AM.

G.0.4 May 12th, 2016

Figure G.4: Sky condition on May 12th, 2016 at 12:34 PM.

Bibliography

- [Abb11] M. Abbasi, S. Gunnarsson, N. Wadefalk, R. Kozhuharov, J. Svedin, S. Cherednichenko, I. Angelov, I. Kallfass, P. Leuther, and H. Zirath. “Single-Chip 220-GHz Active Heterodyne Receiver and Transmitter MMICs With On-Chip Integrated Antenna”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 59, No. 2, pp. 466–478, Feb 2011.
- [Abr11] J. Abril, E. Nova, A. Broquetas, A. Aguasca, J. Romeu, and L. Jofre. “Micrometric deformation imaging at W-Band”. In *IEEE MTT-S International Microwave Workshop Series on Millimeter Wave Integration Technologies (IMWS)*, pp. 65–68, Sept 2011.
- [Adh08] P. Adhikari. “Understanding Millimeter Wave Wireless Communication”. *Millimeter Wave Wireless Communication - Loea Corporation*, 2008.
- [Ahm13] S. Ahmed. “Personnel screening with advanced multistatic imaging technology”. In *Proc. SPIE 8715, Passive and Active Millimeter-Wave Imaging XVI, 87150B*, pp. 87150B–87150B–6, 2013.
- [Ahm15] S. S. Ahmed. “Advanced fully-electronic personnel security screening technology”. In *9th European Conference on Antennas and Propagation (EuCAP)*, pp. 1–4, April 2015.
- [Ahr09] C. D. Ahrens. *Meteorology Today: An Introduction to Weather, Climate, and the Environment*. Brooks/Cole, Cengage Learning, 2009.
- [Alm07] P. Almorox-Gonzalez, J. T. Gonzalez-Partida, M. Burgos-Garcia, C. de la Morena-Alvarez-Palencia, L. Arche-Andradas, and B. P. Dorta-Naranjo. “Portable High Resolution LFM-CW Radar Sensor in Millimeter-Wave Band”. In *International Conference on Sensor Technologies and Applications (SensorComm)*, pp. 5–9, Oct 2007.
- [Alm09] P. Almorox-Gonzalez, J. T. Gonzalez-Partida, M. Burgos-Garcia, B. P. Dorta-Naranjo, and J. Gismero. “Millimeter-Wave Sensor With FMICW Capabilities for Medium-Range High-Resolution Radars”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 57, No. 6, pp. 1479–1486, June 2009.
- [Ana99] Analog Devices. *A Technical Tutorial on Digital Signal Synthesis*. 1999.
- [App07] R. Appleby and H. Wallace. “Standoff Detection of Weapons and Contraband in the 100 GHz to 1 THz Region”. *IEEE Transactions on Antennas and Propagation*, Vol. 55, No. 11, pp. 2944–2956, 2007.
- [Atl54] D. Atlas. “The estimation of cloud parameters by radar”. *Journal of Meteorology*, Vol. 11, No. 4, pp. 309–317, 1954.

- [Ayd00] K. Aydin and S. Daisley. “Rainfall rate relationships with propagation parameters (attenuation and phase) at centimeter and millimeter wavelengths”. In *IEEE International Geoscience and Remote Sensing Symposium (IGARSS)*, pp. 177–179, 2000.
- [Ayd04] K. Aydin and J. Singh. “Cloud Ice Crystal Classification Using a 95-GHz Polarimetric Radar”. *Journal of Atmospheric and Oceanic Technology*, Vol. 21, No. 11, pp. 1679–1688, 2004.
- [Bai93] A. Bains. “An Overview of Millimeter Wave Communication for Military Applications”. In *Defence Science Journal*, pp. 27–36, 1993.
- [Bak11] S. Bakhtiari, S. Liao, T. Elmer, N. Gopalsami, and A. Raptis. “A real-time heart rate analysis for a remote millimeter wave I-Q sensor”. *IEEE Transactions on Biomedical Engineering*, Vol. 58, No. 6, pp. 1839–1845, June 2011.
- [Bal05] C. A. Balanis. *Antenna Theory: Analysis and Design*. Wiley-Interscience, 2005.
- [Bal06] L. Baldini and E. Gorgucci. “Identification of the Melting Layer through Dual-Polarization Radar Measurements at Vertical Incidence”. *Journal of Atmospheric and Oceanic Technology*, Vol. 23, No. 6, pp. 829–839, 2006.
- [Bal16] E. Baltas, D. Panagos, and M. Mimikou. “Statistical Analysis of the Raindrop Size Distribution Using Disdrometer Data”. *MDPI Hydrology*, Vol. 3, No. 1, p. 9, 2016.
- [Ban06] D. Banerjee. *PLL Performance, Simulation and Design*. Dog Ear Publishing, 2006.
- [Bar13] D. Barton. *Radar Equations for Modern Radar*. Artech House, 2013.
- [Bat03] R. Battaglia, M. Ferri, V. Dainelli, F. Sarullo, and M. Demeo. “Warden: W-band advanced radar for debris early notification form ISS”. In *IEEE Aerospace Conference*, pp. 1–81, March 2003.
- [Bat14] A. Battaglia, C. D. Westbrook, S. Kneifel, P. Kollias, N. Humpage, U. Löhnert, J. Tyynelä, and G. W. Petty. “G-band atmospheric radars: new frontiers in cloud physics”. *Atmospheric Measurement Techniques*, Vol. 7, No. 6, pp. 1527–1546, 2014.
- [Bat73] L. J. Battan. *Radar Characteristics of the Atmosphere*. University of Chicago Press, 1973.
- [Bel09] M. A. Belkin, Q. J. Wang, C. Pflugl, A. Belyanin, S. P. Khanna, A. G. Davies, E. H. Linfield, and F. Capasso. “High-Temperature Operation of Terahertz Quantum Cascade Laser Sources”. *IEEE Journal of Selected Topics in Quantum Electronics*, Vol. 15, No. 3, pp. 952–967, May 2009.
- [Ben09] A. Bennett, J. Nash, C. Gaffard, B. Moyna, M. Oldfield, P. Huggard, and C. L. Wrench. “Observations from the UK Met Office 94 GHz FMCW cloud radar”. In *8th International Symposium on Tropospheric Profiling (ISTP)*, 2009.
- [Ber14] D. Berry, H. Deng, R. Dobbs, P. Horoyski, M. Hyttinen, A. Kingsmill, R. MacHattie, A. Roitman, E. Sokol, and B. Steer. “Practical Aspects of EIK Technology”. *IEEE Transactions on Electron Devices*, Vol. 61, No. 6, pp. 1830–1835, June 2014.
- [Bil11] S. Bildik, S. Dieter, C. Fritzsche, M. Frei, C. Fischer, W. Menzel, and R. Jakoby. “Reconfigurable liquid crystal reflectarray with extended tunable phase range”. In *8th European Radar Conference*, pp. 404–407, Oct 2011.

- [Bla14] B. Blázquez, K. Cooper, and N. Llombart. “Time-Delay Multiplexing With Linear Arrays of THz Radar Transceivers”. *IEEE Transactions on Terahertz Science and Technology*, Vol. 4, pp. 232–239, 2014.
- [Ble17] D. Bleh, M. Rösch, M. Kuri, A. Dyck, A. Tessmann, A. Leuther, S. Wagner, B. Weismann-Thaden, H. P. Stulz, M. Zink, M. Rieße, R. Sommer, J. Wilcke, M. Schlechtweg, B. Yang, and O. Ambacher. “W-Band Time-Domain Multiplexing FMCW MIMO Radar for Far-Field 3-D Imaging”. *IEEE Transactions on Microwave Theory and Techniques*, pp. 1–11, 2017.
- [Bon15] S. Bony, B. Stevens, D. M. W. Frierson, C. Jakob, M. Kageyama, R. Pincus, T. G. Shepherd, S. C. Sherwood, A. P. Siebesma, A. H. Sobel, M. Watanabe, and M. J. T. Webb. “Clouds, circulation and climate sensitivity”. *Nature Geoscience*, Vol. 8, pp. 261–268, 2015.
- [Bor79] P. Borjesson and C. E. Sundberg. “Simple Approximations of the Error Function $Q(x)$ for Communications Applications”. *IEEE Transactions on Communications*, Vol. 27, No. 3, pp. 639–643, Mar 1979.
- [Bre11] P. Brennan, Y. Huang, M. Ash, and K. Chetty. “Determination of Sweep Linearity Requirements in FMCW Radar Systems Based on Simple Voltage-Controlled Oscillator Sources”. *IEEE Transactions on Aerospace and Electronic Systems*, Vol. 47, pp. 1594–1604, 2011.
- [Bri01] V. Bringi and V. Chandrasekar. *Polarimetric Doppler Weather Radar: Principles and Applications*. Cambridge University Press, 2001.
- [Bro05] G. M. Brooker. “Understanding Millimetre Wave FMCW Radars”. In *1st International Conference on Sensing Technology*, 2005.
- [Bro11] A. Brown, K. Brown, J. Chen, K. Hwang, N. Koliass, and R. Scott. “W-band GaN power amplifier MMICs”. In *IEEE MTT-S International Microwave Symposium Digest (MTT)*, pp. 1–4, June 2011.
- [Bru02] H. de Bruin. “Introduction: Renaissance of Scintillometry”. *Boundary-Layer Meteorology*, Vol. 105, No. 1, pp. 1–4, 2002.
- [Bru10] R. D. Brunner and A. H. Lynch. *Adaptive Governance and Climate Change*. American Meteorological Society, 2010.
- [Bru95] G. Brussaard, P. Watson, and P. A. Watson. *Atmospheric Modelling and Millimetre Wave Propagation*. Springer Science & Business Media, 1995.
- [Bry13] T. Bryllert, V. Drakinskiy, K. Cooper, and J. Stake. “Integrated 200-240-GHz FMCW Radar Transceiver Module”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 61, No. 10, pp. 3808–3815, 2013.
- [Bud93] J. Budge, M.C. and M. Burt. “Range Correlation Effects in Radars”. In *IEEE National Radar Conference*, pp. 212–216, 1993.
- [Bue07] S. A. Buehler, C. Jiménez, K. F. Evans, P. Eriksson, B. Rydberg, A. J. Heymsfield, C. J. Stubenrauch, U. Lohmann, C. Emde, V. O. John, T. R. Sreerekha, and C. P. Davis. “A concept for a satellite mission to measure cloud ice water path, ice particle size, and cloud altitude”. *Quarterly Journal of the Royal Meteorological Society*, Vol. 133, No. S2, pp. 109–128, 2007.
- [Bur93] M. Burt and J. Budge, M.C. “Range Correlation Effects on Phase Noise Spectra”. In *25th Southeastern Symposium on System Theory (SSST)*, pp. 492–496, 1993.

- [Cai15] J. Cai, X. Wu, and J. Feng. "Traveling-Wave Tube Harmonic Amplifier in Terahertz and Experimental Demonstration". *IEEE Transactions on Electron Devices*, Vol. 62, No. 2, pp. 648–651, Feb 2015.
- [Cao12] Q. Cao, M. Yeary, and G. Zhang. "Efficient Ways to Learn Weather Radar Polarimetry". *IEEE Transactions on Education*, Vol. 55, No. 1, pp. 58–68, Feb 2012.
- [Cha11] G. Chattopadhyay. "Technology, Capabilities, and Performance of Low Power Terahertz Sources". *IEEE Transactions on Terahertz Science and Technology*, Vol. 1, No. 1, pp. 33–53, Sept 2011.
- [Cha12] G. Chattopadhyay, N. Llombart, C. Lee, C. Jung, R. Lin, K. Cooper, T. Reck, J. Siles, E. Schlecht, A. Peralta, B. Thomas, and I. Mehdi. "Terahertz Array Receivers with Integrated Antennas". In *IEEE International Workshop on Antenna Technology (iWAT)*, pp. 319–322, 2012.
- [Cha13] G. Chattopadhyay, T. Reck, C. Jung-Kubiak, J. Siles, C. Lee, R. Lin, and I. Mehdi. "Silicon micromachining for terahertz component development". In *IEEE MTT-S International Microwave and RF Conference*, pp. 1–4, Dec 2013.
- [Cha14] G. Chattopadhyay, T. Reck, C. Jung-Kubiak, C. Lee, J. Siles, N. Chahat, K. Cooper, E. Schlecht, M. Alonso del Pino, and I. Mehdi. "Terahertz antennas with silicon micromachined front-end". In *8th European Conference on Antennas and Propagation (EuCAP)*, pp. 2647–2649, April 2014.
- [Cha15] C. W. I. Chan. *Towards Room-Temperature Terahertz Quantum Cascade Lasers: Directions and Design*. PhD Thesis, Massachusetts Institute of Technology, 2015.
- [Cha99] G. Chattopadhyay. *Dual Polarized and Balanced Receivers at Millimeter and Submillimeter Wavelengths*. PhD Thesis, Department of Electrical Engineering, California Institute of Technology, 1999.
- [Che07] A. Chenakin. "Frequency Synthesis: Current Solutions and New Trends". Technical report, Microwave Journal, May 2007. Vol. 50, No. 5, pp. 256-266.
- [Che09] A. Chenakin, S. Ojha, and I. Sihra. "An Innovative Approach in the Design of Fast-switching Microwave Synthesizers". In *IEEE AUTOTESTCON*, pp. 60–63, 2009.
- [Che11] V. Chen. *The Micro-Doppler Effect in Radar*. Artech House, 2011.
- [Che13] Z. Chen, S. Rea, H. Wang, B. Alderman, B. Zhang, and Y. Fan. "A power combined amplifier module in W-band". In *Millimeter Waves and THz Technology Workshop (UCMMT)*, Sept 2013.
- [Che98] Y. Chen, D. Ingram, R. Lai, M. Barsky, R. Grunbacher, T. Block, H. Yen, and D. Streit. "A 95-GHz InP HEMT MMIC amplifier with 427-mW power output". *IEEE Microwave and Guided Wave Letters*, Vol. 8, No. 11, pp. 399–401, Nov 1998.
- [Clo95] E. E. Clothiaux, M. A. Miller, B. A. Albrecht, T. P. Ackerman, J. Verlinde, D. M. Babb, R. M. Peters, and W. J. Syrett. "An Evaluation of a 94-GHz Radar for Remote Sensing of Cloud Properties". *Journal of Atmospheric and Oceanic Technology*, Vol. 12, pp. 201–229, 1995.

- [Clo99] E. Clothiaux, K. Moran, B. Martner, T. Ackerman, G. Mace, T. Uttal, J. Mather, K. Widener, M. Miller, and D. Rodriguez. "The Atmospheric Radiation Measurement Program Cloud Radars: Operational Modes". *Bulletin American Meteorological Society*, Vol. 16, pp. 819–827, 1999.
- [Coh95] L. Cohen. *Time-frequency Analysis: Theory and Applications*. Prentice-Hall, 1995.
- [Com96] Committee on Commercial Aviation Security; Panel on Passenger Screening; Commission on Engineering and Technical Systems; National Materials Advisory Board; Division on Engineering and Physical Sciences; National Research Council, Ed. *Airline Passenger Security Screening: New Technologies and Implementation Issues*. The National Academies Press, 1996.
- [Coo08a] K. Cooper, R. Dengler, G. Chattopadhyay, E. Schlecht, J. Gill, A. Skalare, I. Mehdi, and P. Siegel. "A High-Resolution Imaging Radar at 580 GHz". *IEEE Microwave and Wireless Components Letters*, Vol. 18, No. 1, pp. 64–66, 2008.
- [Coo08b] K. Cooper, R. Dengler, N. Llombart, T. Bryllert, G. Chattopadhyay, E. Schlecht, J. Gill, C. Lee, A. Skalare, I. Mehdi, and P. Siegel. "Penetrating 3-D Imaging at 4- and 25-m Range Using a Submillimeter-Wave Radar". *IEEE Transactions on Microwave Theory and Techniques*, Vol. 56, No. 12, pp. 2771–2778, 2008.
- [Coo09] K. Cooper, R. Dengler, N. Llombart, T. Bryllert, G. Chattopadhyay, I. Mehdi, and P. Siegel. "An Approach for Sub-Second Imaging of Concealed Objects Using Terahertz (THz) Radar". *Journal of Infrared, Millimeter, and Terahertz Waves*, Vol. 30, pp. 1297–1307, 2009.
- [Coo11] K. Cooper, R. Dengler, N. Llombart, B. Thomas, G. Chattopadhyay, and P. Siegel. "THz Imaging Radar for Standoff Personnel Screening". *IEEE Transactions on Terahertz Science and Technology*, Vol. 1, No. 1, pp. 169–182, 2011.
- [Coo12] K. Cooper, J. Trabert, and R. Dengler. "Carrier Noise-limited Penetration in THz Radar Imaging". In *IEEE MTT-S International Microwave Symposium Digest (MTT)*, pp. 1–3, 2012.
- [Coo13] K. B. Cooper, T. A. Reck, C. Jung-Kubiak, C. Lee, J. V. Siles, R. H. Lin, A. Peralta, E. Decrossas, E. T. Schlecht, G. Chattopadhyay, and I. Mehdi. "Transceiver Array Development for Submillimeter-wave Imaging Radars". In *Proc. SPIE 8715, Passive and Active Millimeter-Wave Imaging XVI, 87150A*, 2013.
- [Coo14] K. Cooper and G. Chattopadhyay. "Submillimeter-Wave Radar: Solid-State System Design and Applications". *IEEE Microwave Magazine*, Vol. 15, No. 7, pp. 51–67, Nov 2014.
- [Coo15] K. B. Cooper, T. Reck, R. J. Dengler, and N. Llombart. "Focal Plane and Synthetic Aperture Array Alternatives in a 340 GHz Imaging Radar". In *9th European Conference on Antennas and Propagation (EuCAP)*, pp. 1–4, April 2015.
- [CPI16] CPI Communications & Medical Products Division. "Extended Interaction Klystrons (EIKs)". <http://www.cpii.com/product.cfm/7/40>, 2016.
- [Cro05] T. W. Crowe, W. L. Bishop, D. W. Porterfield, J. L. Hesler, and R. M. Weikle. "Opening the terahertz window with integrated diode circuits". *IEEE Journal of Solid-State Circuits*, Vol. 40, No. 10, pp. 2104–2110, Oct 2005.
- [Cze14] M. G. Czerwinski and J. M. Usoff. "Development of the Haystack Ultrawideband Satellite Imaging Radar". *Lincoln Laboratory Journal*, Vol. 21, No. 1, pp. 28–44, 2014.

- [Dai09] H. Y. Dai, Y. Z. Li, Y. Liu, Y. I. Chang, X. S. Wang, and M. Q. Li. “Study on Polarimetric measurement and radar target identification in instantaneous polarization radar system”. In *IET International Radar Conference*, pp. 1–5, April 2009.
- [Dao13] M. T. Dao, D. H. Shin, Y. T. Im, and S. O. Park. “A Two Sweeping VCO Source for Heterodyne FMCW Radar”. *IEEE Transactions on Instrumentation and Measurement*, Vol. 62, No. 1, pp. 230–239, 2013.
- [Dav04] G. Davies and E. Linfield. “Bridging the terahertz gap”. *Physics World*, Vol. 17, No. 4, p. 37, 2004.
- [Del16] J. Delanoë, A. Protat, J.-P. Vinson, W. Brett, C. Caudoux, F. Bertrand, J. P. du Chatelet, R. Hallali, L. Barthes, M. Haeffelin, and J.-C. Dupont. “BASTA: A 95 GHz FMCW Doppler Radar for Cloud and Fog Studies”. *Journal of Atmospheric and Oceanic Technology*, Vol. 33, No. 5, pp. 1023–1038, 2016.
- [Den07] R. Dengler, K. Cooper, G. Chattopadhyay, I. Mehdi, E. Schlecht, A. Skalare, C. Chen, and P. Siegel. “600 GHz Imaging Radar with 2 cm Range Resolution”. In *IEEE MTT-S International Microwave Symposium*, pp. 1371–1374, 2007.
- [Den09] R. Dengler, K. Cooper, N. Llombart, G. Chattopadhyay, T. Bryllert, I. Mehdi, and P. Siegel. “Toward Real-time Penetrating Imaging Radar at 670 GHz”. In *IEEE MTT-S International Microwave Symposium Digest*, pp. 941–944, 2009.
- [Dhi17] S. S. Dhillon, M. S. Vitiello, E. H. Linfield, A. G. Davies, M. C. Hoffmann, J. Booske, C. Paoloni, M. Gensch, P. Weightman, G. P. Williams, E. Castro-Camus, D. R. S. Cumming, F. Simoens, I. Escorcia-Carranza, J. Grant, S. Lucyszyn, M. Kuwata-Gonokami, K. Konishi, M. Koch, C. A. Schmuttenmaer, T. L. Cocker, R. Huber, A. G. Markelz, Z. D. Taylor, V. P. Wallace, J. A. Zeitler, J. Sibik, T. M. Korter, B. Ellison, S. Rea, P. Goldsmith, K. B. Cooper, R. Appleby, D. Pardo, P. G. Huggard, V. Krozer, H. Shams, M. Fice, C. Renaud, A. Seeds, A. Stöhr, M. Naftaly, N. Ridler, R. Clarke, J. E. Cunningham, and M. B. Johnston. “The 2017 terahertz science and technology roadmap”. *Journal of Physics D: Applied Physics*, Vol. 50, No. 4, 2017.
- [Die12] S. Diebold, D. Goetzl, S. Ayhan, S. Scherr, P. Pahl, H. Massler, A. Tessmann, A. Leuther, O. Ambacher, T. Zwick, and I. Kallfass. “W-band MMIC radar modules for remote detection of vital signs”. In *7th European Microwave Integrated Circuits Conference (EuMIC)*, pp. 195–198, Oct 2012.
- [Dow90] D. R. Dowling and L. F. Radke. “A Summary of the Physical Properties of Cirrus Clouds”. *Journal of Applied Meteorology*, Vol. 29, No. 9, pp. 970–978, 1990.
- [Ede94] H. Edelsbrunner and E. P. Mücke. “Three-dimensional Alpha Shapes”. *ACM Transactions on Graphics*, Vol. 13, pp. 43–72, 1994.
- [Eme97] D. Emerson. “The work of Jagadis Chandra Bose: 100 years of millimeter-wave research”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 45, No. 12, pp. 2267–2273, Dec 1997.
- [Ess11a] H. Essen. “Monitoring food with millimeter waves”. <https://www.fraunhofer.de/en/press/research-news/2011/december/monitoring-food-with-millimeter-waves.html>, 2011.

- [Ess11b] H. Essen, F. Lorenz, S. Hantscher, P. Warok, R. Zimmermann, M. Schroder, W. Koch, M. Schikora, G. Luedtke, and K. Wild. “Millimeter-wave radar for runway debris detection”. In *Tyrrhenian International Workshop on Digital Communications - Enhanced Surveillance of Aircraft and Vehicles (TIWDC/ESAV)*, pp. 65–68, Sept 2011.
- [Eur12] European Commission. “Health effects of security scanners for passenger screening (based on X-ray technology)”. Technical report, DG Health & Consumers. Scientific Committee on Emerging and Newly Identified Health Risks (SCENIHR), 2012.
- [Eva05] K. F. Evans, J. R. Wang, P. E. Racette, G. Heymsfield, and L. Li. “Ice Cloud Retrievals and Analysis with the Compact Scanning Submillimeter Imaging Radiometer and the Cloud Radar System during CRYSTAL FACE”. *Journal of Applied Meteorology*, Vol. 44, No. 6, pp. 839–859, 2005.
- [Fab15] F. Fabry. *Radar Meteorology. Principles and Practice*. Cambridge University Press, 2015.
- [Fel11] D. Felguera-Martin, J. T. Gonzalez-Partida, P. Almorox-Gonzalez, M. Burgos-Garcia, and B. P. Dorta-Naranjo. “Interferometric inverse synthetic aperture radar experiment using an interferometric linear frequency modulated continuous wave millimetre-wave radar”. *IET Radar, Sonar Navigation*, Vol. 5, No. 1, pp. 39–47, January 2011.
- [Fox97] N. I. Fox and A. J. Illingworth. “The Retrieval of Stratocumulus Cloud Properties by Ground-Based Cloud Radar”. *Journal of Applied Meteorology*, Vol. 36, No. 5, pp. 485–492, 1997.
- [Fra16] Fraunhofer Institute for Applied Solid State Physics. “W-band radar modules for sensors, automation and security”. https://www.iaf.fraunhofer.de/content/dam/iaf/documents/Presse/infomaterial/gf1_folder_radarmodul_web.pdf, April 2016.
- [Fri11] F. Friederich, W. Von Spiegel, M. Bauer, F. Meng, M. Thomson, S. Boppel, A. Lisauskas, B. Hils, V. Krozer, A. Keil, T. Löffler, R. Henneberger, A. Huhn, G. Spickermann, P. Bolivar, and H. Roskos. “THz Active Imaging Systems With Real-Time Capabilities”. *IEEE Transactions on Terahertz Science and Technology*, pp. 183–200, 2011.
- [Fri95] A. S. Frisch, D. H. Lenschow, C. W. Fairall, W. H. Schubert, and J. S. Gibson. “Doppler Radar Measurements of Turbulence in Marine Stratiform Cloud during ASTEX”. *Journal of the Atmospheric Sciences*, Vol. 52, No. 16, pp. 2800–2808, 1995.
- [Fri97] A. S. Frisch, T. Uttal, C. W. Fairall, and J. B. Snider. “On the measurement of stratus cloud properties with a cloud radar and microwave radiometer”. In *IEEE International Geoscience and Remote Sensing (IGARSS)*, pp. 2090–2092, Aug 1997.
- [Fun11] A. Fung, J. Ward, G. Chattopadhyay, R. Lin, L. Samoska, P. Kangaslahti, I. Mehdi, B. Lambrigtsen, P. Goldsmith, M. Micovic, A. Kurdoghlian, K. Shinohara, I. Milosavljevic, and D. Chow. “Power combined Gallium nitride amplifier with 3 Watt output power at 87 GHz”. In *Proceedings of the ISSTT Symposium*, pp. 44–48, 2011.

- [Gal14] M. Galletti, D. Huang, and P. Kollias. “Zenith/Nadir Pointing mm-Wave Radars: Linear or Circular Polarization?”. *IEEE Transactions on Geoscience and Remote Sensing*, Vol. 52, No. 1, pp. 628–639, Jan 2014.
- [Gar13] A. Garcia-Pino, B. Gonzalez-Valdes, B. Mencia-Oliva, J. Grajal, O. Rubinos, and J. L. Besada. “Bifocal Reflector Antenna System for Radar Imaging at 300 GHz”. In *7th European Conference on Antennas and Propagation (EuCAP)*, pp. 3814–3818, 2013.
- [Gar14] A. Garcia-Pino, B. Gonzalez-Valdes, O. Rubinos, J. Grajal, A. Badolato, B. Mencia-Oliva, P. Garcia Soidan, and J. Besada-Sanmartin. “Bifocal Reflector Antenna for a Standoff Radar Imaging System With Enhanced Field of View”. *IEEE Transactions on Antennas and Propagation*, Vol. 62, No. 10, pp. 4997–5006, Oct 2014.
- [Gas13] A. J. Gasiewski, B. T. Sanders, and D. Gallaher. “CubeSat based sensors for global weather forecasting”. In *US National Committee of URSI National Radio Science Meeting (USNC-URSI NRSM)*, pp. 1–1, Jan 2013.
- [Gil11] A. S. Gilmour. *Klystrons, traveling wave tubes, magnetrons, crossed-field amplifiers, and gyrotrons*. Artech House, 2011.
- [Gon07] J. T. Gonzalez-Partida, P. Almorox-Gonzalez, M. Burgos-Garcia, and B. P. Dorta-Naranjo. “Ground SAR system with tunable distance limits and low sampling rate”. In *IET International Conference on Radar Systems*, pp. 1–5, Oct 2007.
- [Gon09] J. T. Gonzalez-Partida, P. Almorox-Gonzalez, M. Burgos-Garcia, B. P. Dorta-Naranjo, and J. I. Alonso. “Through-the-Wall Surveillance With Millimeter-Wave LFM CW Radars”. *IEEE Transactions on Geoscience and Remote Sensing*, Vol. 47, No. 6, pp. 1796–1805, June 2009.
- [Gon16a] B. Gonzalez-Valdes, Y. Alvarez, S. Mantzavinos, C. M. Rappaport, F. Las-Heras, and J. A. Martinez-Lorenzo. “Improving Security Screening: A Comparison of Multistatic Radar Configurations for Human Body Imaging”. *IEEE Antennas and Propagation Magazine*, Vol. 58, No. 4, pp. 35–47, Aug 2016.
- [Gon16b] B. Gonzalez-Valdes, Y. Álvarez, Y. Rodriguez-Vaqueiro, A. Arboleya-Arboleya, A. García-Pino, C. M. Rappaport, F. Las-Heras, and J. A. Martinez-Lorenzo. “Millimeter Wave Imaging Architecture for On-The-Move Whole Body Imaging”. *IEEE Transactions on Antennas and Propagation*, Vol. 64, No. 6, pp. 2328–2338, June 2016.
- [Gop01] N. Gopalsami and A. Raptis. “Millimeter-wave radar sensing of airborne chemicals”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 49, No. 4, pp. 646–653, Apr 2001.
- [Gop08] N. Gopalsami, S. Bakhtiari, T. Elmer, and A. Raptis. “Application of Millimeter-Wave Radiometry for Remote Chemical Detection”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 56, No. 3, pp. 700–709, March 2008.
- [Gop09] N. Gopalsami, H. T. Chien, A. Heifetz, E. R. Koehl, and A. C. Raptis. “Millimeter wave detection of nuclear radiation: An alternative detection mechanism”. *Review of Scientific Instruments*, Vol. 80, No. 8, 2009.
- [Gos97] E. E. Gossard, J. Snider, E. Clothiaux, B. Martner, J. S. Gibson, R. Kropfli, and A. Frisch. “The potential of 8-mm radars for remotely sensing cloud drop size distributions”. *Journal of Atmospheric and Oceanic Technology*, Vol. 14, No. 1, pp. 76–87, 1997.

- [Gra02] N. A. Grant. “Next generation mm-wave and submm-wave heterodyne hybrid/MIC receivers for passive radiometer meteorological satellite applications”. In *Proceedings of the Fourth IEEE International Caracas Conference on Devices, Circuits and Systems*, pp. I011–1–I011–5, 2002.
- [Gra15] J. Grajal, A. Badolato, G. Rubio-Cidre, L. Ubeda-Medina, B. Mencia-Oliva, A. Garcia-Pino, B. Gonzalez-Valdes, and O. Rubinos. “3-D High-Resolution Imaging Radar at 300 GHz With Enhanced FoV”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 63, No. 3, pp. 1097–1107, March 2015.
- [Gra16] M. Grabner, P. Pechac, and P. Valtr. “Attenuation due to hydrometeors in three millimeter wave bands”. In *10th European Conference on Antennas and Propagation (EuCAP)*, pp. 1–4, April 2016.
- [Gra17] J. Grajal, G. Rubio-Cidre, A. Badolato, L. Úbeda Medina, F. García-Rial, A. Garcia-Pino, and O. Rubiños. “Compact Radar Front-End for an Imaging Radar at 300 GHz”. *IEEE Transactions on Terahertz Science and Technology*, Vol. 7, No. 3, pp. 268–273, May 2017.
- [Gri91] H. Griffiths. “The Effect of Phase and Amplitude Errors in FM Radar”. In *IEEE Colloquium on High Time-Bandwidth Product Waveforms in Radar and Sonar*, pp. 9/1–9/5, 1991.
- [Gro10] E. Grossman, C. Dietlein, J. Ala-Laurinaho, M. Leivo, L. Gronberg, M. Gronholm, P. Lappalainen, A. Rautiainen, A. Tamminen, and A. Luukanen. “Passive terahertz camera for standoff security screening”. *Applied Optics*, Vol. 49, No. 19, pp. E106–E120, Jul 2010.
- [Guid08] “Guide to Meteorological Instruments and Methods of Observation”. Technical report 7, World Meteorological Organization (WMO), 2008.
- [Hae16] M. Haeffelin, S. Crewell, A. J. Illingworth, G. Pappalardo, H. Russchenberg, M. Chiriaco, K. Ebell, R. J. Hogan, and F. Madonna. “Parallel Developments and Formal Collaboration between European Atmospheric Profiling Observatories and the U.S. ARM Research Program”. *Meteorological Monographs*, Vol. 57, pp. 29.1–29.34, 2016.
- [Hag03] M. Hagen and S. E. Yuter. “Relations between radar reflectivity, liquid-water content, and rainfall rate during the MAP SOP”. *Quarterly Journal of the Royal Meteorological Society*, Vol. 129, pp. 477–493, 2003.
- [Hag08] M. Hagelen, G. Briese, H. Essen, T. Bertuch, P. Knott, and A. Tessmann. “A millimetre wave landing aid approach for helicopters under brown-out conditions”. In *IEEE Radar Conference*, pp. 1–4, May 2008.
- [Ham96] A. Hammoudeh, R. Stephens, and M. Al-Nuaimi. “Characterisation and modelling of scatter, attenuation and depolarisation of millimetre waves due to foliage”. In *26th European Microwave Conference*, pp. 916–919, Sept 1996.
- [Har78] F. Harris. “On the Use of Windows for Harmonic Analysis with the Discrete Fourier Transform”. *IEEE Proceedings*, Vol. 66, 1978.
- [Has12] J. Hasch, E. Topak, R. Schnabel, T. Zwick, R. Weigel, and C. Waldschmidt. “Millimeter-Wave Technology for Automotive Radar Sensors in the 77 GHz Frequency Band”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 60, No. 3, pp. 845–860, March 2012.

- [Hei08] P. L. Heinselman, D. L. Priegnitz, K. L. Manross, T. M. Smith, and R. W. Adams. “Rapid Sampling of Severe Storms by the National Weather Radar Testbed Phased Array Radar”. *Weather and Forecasting*, Vol. 23, No. 5, pp. 808–824, 2008.
- [Her16] R. Herschel, S. A. Lang, and N. Pohl. “MIMO imaging for next generation passenger security systems”. In *11th European Conference on Synthetic Aperture Radar (EUSAR)*, pp. 1–4, Jun 2016.
- [Hid06] A. Hideji, Y. Shinichi, K. Hiroshi, K. Youhei, T. Toshiaki, K. Hiroshi, T. Tamio, and N. Teruyuki. “Development of the monostatic FMCW Radar at 95 GHz”. *International Symposium on Antennas and Propagation (ISAP)*, 2006.
- [Hit14] Hittite Microwave Corporation. “MMIC Amplifier Biasing Procedure”. Technical report, Analog Devices, 2014. v02.0608.
- [Hit16] M. Hitzler, S. Saulig, L. Boehm, W. Mayer, W. Winkler, and C. Waldschmidt. “Compact bistatic 160 GHz transceiver MMIC with phase noise optimized synthesizer for FMCW radar”. In *IEEE MTT-S International Microwave Symposium (IMS)*, pp. 1–4, May 2016.
- [Hoa06] E. Hoare. “Automotive Millimetre-Wave Radar, Current Applications and Future Developments”. In *The Institution of Engineering and Technology Seminar on The Future of Civil Radar*, pp. 29–40, June 2006.
- [Hoc00] W. K. Hocking. “The Instruments of Meteorology”. <http://www.physics.uwo.ca/~whocking/p103/instrum.html>, 2000.
- [Hog03] R. J. Hogan, D. Bouniol, D. N. Ladd, E. J. O’Connor, and A. J. Illingworth. “Absolute Calibration of 94/95-GHz Radars Using Rain”. *Journal of Atmospheric and Oceanic Technology*, Vol. 20, No. 4, pp. 572–580, 2003.
- [Hor07] P. Horoyski, D. Berry, and B. Steer. “Compact, High Power Klystrons At 95 GHz”. In *IEEE International Vacuum Electronics Conference*, pp. 1–2, May 2007.
- [Hori14] “Horizon 2020 - Secure societies - Protecting freedom and security of Europe and its citizens”. Technical report, European Commission Decision C 1720, March 2014.
- [Hos15] M. Hossain, N. Weimann, M. Lisker, C. Meliani, B. Tillack, V. Krozer, and W. Heinrich. “A 330 GHz hetero-integrated source in InP-on-BiCMOS technology”. In *IEEE MTT-S International Microwave Symposium*, pp. 1–4, May 2015.
- [HRL13] HRL Laboratories, LLC. “Commercial GaN MMIC products”. <http://mmics.hrl.com/GaNPA/>, 2013.
- [Hro13] M. Hrobak, M. Sterns, M. Schramm, W. Stein, and L. P. Schmidt. “Design and Fabrication of Broadband Hybrid GaAs Schottky Diode Frequency Multipliers”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 61, No. 12, pp. 4442–4460, Dec 2013.
- [Hug08] P. G. Huggard, M. L. Oldfield, B. P. Moyna, B. N. Ellison, D. N. Matheson, A. J. Bennett, C. Gaffard, T. Oakley, and J. Nash. “94 GHz FMCW cloud radar”. In *Proc. SPIE 7117, Millimetre Wave and Terahertz Sensors and Technology, 711704*, 2008.

- [IAT15] “International Air Transport Association (IATA) Annual Review”. <http://www.iata.org/publications/Pages/index.aspx>, 2015.
- [Igl14] R. Iglesias, X. Fabregas, A. Aguasca, J. J. Mallorqui, C. Lopez-Martinez, J. A. Gili, and J. Corominas. “Atmospheric Phase Screen Compensation in Ground-Based SAR With a Multiple-Regression Model Over Mountainous Regions”. *IEEE Transactions on Geoscience and Remote Sensing*, Vol. 52, No. 5, pp. 2436–2449, May 2014.
- [Igl15] V. Iglesias, J. Grajal, M. A. Sánchez, and M. López-Vallejo. “Implementation of a Real-Time Spectrum Analyzer on FPGA Platforms”. *IEEE Transactions on Instrumentation and Measurement*, Vol. 64, No. 2, pp. 338–355, Feb 2015.
- [Ill07] A. J. Illingworth, R. J. Hogan, E. J. O’Connor, D. Bouniol, J. Delanoe, J. Pelon, A. Protat, M. E. Brooks, N. Gaussiat, D. R. Wilson, D. P. Donovan, H. K. Baltink, G.-J. van Zadelhoff, J. D. Eastment, J. W. F. Goddard, C. L. Wrench, M. Haeffelin, O. A. Krasnov, H. W. J. Russchenbergand, J.-M. Piriou, F. Vinit, A. Seifert, A. M. Tompkins, and U. Willen. “Cloudnet: Continuous Evaluation of Cloud Profiles in Seven Operational Models Using Ground-Based Observations”. *Bulletin of the American Meteorological Society*, Vol. 88, No. 6, pp. 883–898, 2007.
- [Im90] E. Im and K. Lellogg. “Spaceborne Radar For Rain And Cloud Measurements - A Conceptual Design”. *10th Annual International Geoscience and Remote Sensing Symposium*,, pp. 425–428, 1990.
- [Int14] “Airports Council International (ACI) Annual World Airport Traffic Report”. <http://www.aci.aero/Data-Centre/Annual-Traffic-Data>, 2014.
- [Inte11] “International Air Transport Association (IATA) - Checkpoint of the future - Executive summary”. <https://www.iata.org/whatwedo/security/Documents/cof-executive-summary.pdf>, 2011.
- [ITU 01] “ITU-R Recommendation P.372-7. Radio noise”. Technical report, 2001.
- [ITU 03] “ITU-R Recommendation P.838-2. Specific attenuation model for rain for use in prediction methods”. Technical report, 2003.
- [ITU 09] “ITU-R Recommendation P.840-4. Attenuation due to clouds and fog”. Technical report, 2009.
- [ITU 13] “ITU-R Recommendation P.676-10. Attenuation by atmospheric gases”. Technical report, 2013.
- [ITU 15] “ITU-R Recommendation P.453-11. The radio refractive index: its formula and refractivity data”. Technical report, 2015.
- [Jan07] M. Jankiraman. *Design of Multi-Frequency CW Radars*. SciTech, 2007.
- [Joa08] Y. Joannon, V. Beroulle, C. Robach, S. Tedjini, and J.-L. Carbonero. “Decreasing Test Qualification Time in AMS and RF Systems”. *IEEE Design Test of Computers*, Vol. 25, pp. 29–37, 2008.
- [Joy14] C. D. Joye, A. M. Cook, J. P. Calame, D. K. Abe, A. N. Vlasov, I. A. Chernyavskiy, K. T. Nguyen, E. L. Wright, D. E. Pershing, T. Kimura, M. Hyttinen, and B. Levush. “Demonstration of a High Power, Wideband 220-GHz Traveling Wave Amplifier Fabricated by UV-LIGA”. *IEEE Transactions on Electron Devices*, Vol. 61, No. 6, pp. 1672–1678, June 2014.

- [Kao13] T.-Y. J. Kao and J. Lin. “Vital sign detection using 60-GHz Doppler radar system”. In *IEEE International Wireless Symposium (IWS)*, pp. 1–4, April 2013.
- [Kap14] B. Kapilevich, S. W. Harmer, and N. J. Bowring. *Non-Imaging Microwave and Millimetre-Wave Sensors for Concealed Object Detection*. CRC Press, 2014.
- [Kar05] N. Karpowicz, H. Zhong, J. Xu, K.-I. Lin, J.-S. Hwang, and X.-C. Zhang. “Comparison between Pulsed Terahertz Time-domain Imaging and Continuous Wave Terahertz Imaging”. *Semiconductor Science and Technology*, Vol. 20, No. 7, pp. 293–299, 2005.
- [Kay66] A. F. Kay. “Millimeter wave antennas”. *Proceedings of the IEEE*, Vol. 54, No. 4, pp. 641–647, April 1966.
- [Kha09] S. Kharkovsky, M. T. Ghasr, and R. Zoughi. “Near-Field Millimeter-Wave Imaging of Exposed and Covered Fatigue Cracks”. *IEEE Transactions on Instrumentation and Measurement*, Vol. 58, No. 7, pp. 2367–2370, July 2009.
- [Kno07] J. Knorr. “Weather radar equation correction for frequency agile and phased array radars”. *IEEE Transactions on Aerospace and Electronic Systems*, Vol. 43, No. 3, pp. 1220–1227, July 2007.
- [Kno93] E. F. Knott. *Radar Cross Section*. Artech House, 1993.
- [Kok09] A. Kokhanovsky, C. Naud, and A. Devasthale. “Intercomparison of Ground-Based Radar and Satellite Cloud-Top Height Retrievals for Overcast Single-Layered Cloud Fields”. *IEEE Transactions on Geoscience and Remote Sensing*, Vol. 47, No. 7, pp. 1901–1908, July 2009.
- [Kol07] P. Kollias, E. E. Clothiaux, M. A. Miller, B. A. Albrecht, G. L. Stephens, and T. P. Ackerman. “Millimeter-Wavelength Radars: New Frontier in Atmospheric Cloud and Precipitation Research”. *Bulletin of the American Meteorological Society*, Vol. 88, pp. 1608–1624, 2007.
- [Kol11] P. Kollias, J. Rémillard, E. Luke, and W. Szyrmer. “Cloud radar Doppler spectra in drizzling stratiform clouds: 1. Forward modeling and remote sensing applications”. *Journal of Geophysical Research: Atmospheres*, Vol. 116, No. D13, 2011.
- [Kol14] P. Kollias, S. Tanelli, A. Battaglia, and A. Tatarevic. “Evaluation of EarthCARE Cloud Profiling Radar Doppler Velocity Measurements in Particle Sedimentation Regimes”. *Journal of Atmospheric and Oceanic Technology*, Vol. 31, No. 2, pp. 366–386, 2014.
- [Kra05] O. A. Krasnov and H. W. J. Russchenberg. “A synergetic radar-lidar technique for the LWC retrieval in water clouds: Description and application to the Cloudnet data”. *32nd Conference on Radar Meteorology*, 2005.
- [Kro03] V. Kroupa. *Phase Lock Loops and Frequency Synthesis*. Wiley, 2003.
- [Kro96] R. Krofli and R. Kelly. “Meteorological Research Applications of MM-Wave Radar”. In *Meteorology and Atmospheric Physics*, pp. 105–121, 1996.
- [L3 15] L3 Communications Security And Detection Systems. “Provision ATD”. <http://www.sds.1-3com.com/advancedimaging/provision-at.htm>, 2015.
- [Lea13] C. A. Leal-Sevillano, K. B. Cooper, J. A. Ruiz-Cruz, J. R. Montejo-Garai, and J. M. Rebollar. “A 225 GHz Circular Polarization Waveguide Duplexer Based on a Septum Orthomode Transducer Polarizer”. *IEEE Transactions on Terahertz Science and Technology*, Vol. 3, No. 5, pp. 574–583, Sept 2013.

- [Lea14a] C. A. Leal-Sevillano, K. B. Cooper, E. Decrossas, R. J. Dengler, J. A. Ruiz-Cruz, J. R. Montejo-Garai, G. Chattopadhyay, and J. M. Rebollar. "Compact Duplexing for a 680-GHz Radar Using a Waveguide Orthomode Transducer". *IEEE Transactions on Microwave Theory and Techniques*, Vol. 62, No. 11, pp. 2833–2842, Nov 2014.
- [Lea14b] C. Leal-Sevillano. *Contribution to the Analysis of Waveguides and Design of Passive Components at Millimeter and Submillimeter-Waves*. PhD Thesis, Technical University of Madrid, 2014.
- [Lee09] J. S. Lee and E. Pottier. *Polarimetric radar imaging: from basics to applications. Optical science and engineering*, CRC Press, 2009.
- [Lhe66] R. M. Lhermitte. "Probing Air Motion by Doppler Analysis of Radar Clear Air Returns". *Journal of the Atmospheric Sciences*, Vol. 23, No. 5, pp. 575–591, 1966.
- [Lhe87] R. Lhermitte. "A 94-GHz Doppler Radar for Cloud Observations". *Journal of Atmospheric and Oceanic Technology, American Meteorological Society*, Vol. 4, 1987.
- [Lhn01] U. Löhnert, S. Crewell, C. Simmer, and A. Macke. "Profiling Cloud Liquid Water by Combining Active and Passive Microwave Measurements with Cloud Model Statistics". *Journal of Atmospheric and Oceanic Technology*, Vol. 18, No. 8, pp. 1354–1366, 2001.
- [Li14a] L. Li, L. Chen, J. Zhu, J. Freeman, P. Dean, A. Valavanis, A. G. Davies, and E. H. Linfield. "Terahertz quantum cascade lasers with > 1 W output powers". *Electronics Letters*, Vol. 50, No. 4, pp. 309–311, Feb 2014.
- [Li14b] Y. Li, P. Hoogeboom, and H. Russchenberg. "Observations and modeling of fog by cloud radar and optical sensors". In *11th European Radar Conference (EuRAD)*, pp. 521–524, Oct 2014.
- [Li15] Y. Li, P. Hoogeboom, H. W. J. Russchenberg, and H. K. Baltink. "Retrieval of fog microphysical properties from cloud radar and optical sensors". In *IEEE International Geoscience and Remote Sensing Symposium (IGARSS)*, pp. 4860–4863, July 2015.
- [Li94] F. Li, E. Im, S. Durden, and W. Wilson. "Airborne and spaceborne cloud radar designs". In *International Geoscience and Remote Sensing Symposium*, pp. 672–674, Aug 1994.
- [Lie89] H. J. Liebe, T. Manabe, and G. A. Hufford. "Millimeter-wave attenuation and delay rates due to fog/cloud conditions". *IEEE Transactions on Antennas and Propagation*, Vol. 37, No. 12, pp. 1617–1612, Dec 1989.
- [Lin14] J. Lin and W. Wu. "Vital sign radars: Past, present, and future". In *15th IEEE Annual Wireless and Microwave Technology Conference (WAMICON)*, pp. 1–4, June 2014.
- [Liu11] L. Liu, J. L. Hesler, R. M. Weikle, T. Wang, P. Fay, and H. Xing. "A 570-630 GHz Frequency Domain Terahertz Spectroscopy System based on a Broadband Quasi-optical Zero Bias Schottky Diode Detector". *International Journal of High Speed Electronics and Systems*, Vol. 20, No. 03, pp. 629–638, 2011.
- [Liu13] L. Liu, R. Pathak, L.-J. Cheng, and T. Wang. "Real-Time Frequency-Domain Terahertz Sensing and Imaging of Isopropyl Alcohol-Water Mixtures on a Microfluidic Chip". *Sensors and Actuators B: Chemical*, Vol. 184, pp. 228–234, 2013.

- [Loc98] R. Locher. “Introduction to Power MOSFETs and their Applications”. Technical report, Fairchild Semiconductor, October 1998. Application Note 558, Rev. B1.
- [Loe09] T. Loeffler. “3D-Terahertz-Tomography for Material Inspection and Security”. In *SENSOR+TEST Conferences*, pp. 179 – 184, SynView GmbH, Glashütten, Germany, 2009.
- [Loh97] S. Lohmeier, S. Sekelsky, J. Firda, G. Sadowy, and R. McIntosh. “Classification of particles in stratiform clouds using the 33 and 95 GHz polarimetric cloud profiling radar system (CPRS)”. *IEEE Transactions on Geoscience and Remote Sensing*, Vol. 35, No. 2, pp. 256–270, Mar 1997.
- [Luk10] E. P. Luke, P. Kollias, and M. D. Shupe. “Detection of supercooled liquid in mixed-phase clouds using radar Doppler spectra”. *Journal of Geophysical Research: Atmospheres*, Vol. 115, No. D19, 2010.
- [Luk13] E. P. Luke and P. Kollias. “Separating Cloud and Drizzle Radar Moments during Precipitation Onset Using Doppler Spectra”. *Journal of Atmospheric and Oceanic Technology*, Vol. 30, No. 8, pp. 1656–1671, 2013.
- [Luu12] A. Luukanen, J. Ala-Laurinaho, J. Häkli, D. Gomes-Martins, T. Kiuru, P. Koivisto, M. Leivo, A. Rautiainen, J. Säily, A. Tamminen, H. Toivanen, R. Tuovinen, and A. Räisänen. “Towards video rate imaging at submillimetre-waves - Finnish developments of passive multi-band imaging and holographic submm-wave beam steering at VTT”. In *Asia Pacific Microwave Conference Proceedings*, pp. 782–784, Dec 2012.
- [Mae15] T. Maesaka, K. Iwanami, S. I. Suzuki, Y. Shusse, and N. Sakurai. “Cloud Radar Network in Tokyo Metropolitan Area for Early Detection of Cumulonimbus Generation”. *37th Conference on Radar Meteorology*, 2015.
- [Mai12] F. Maier, J. Bendix, and B. Thies. “Simulating Z–LWC Relations in Natural Fogs with Radiative Transfer Calculations for Future Application to a Cloud Radar Profiler”. *Pure and Applied Geophysics*, Vol. 169, No. 5, pp. 793–807, 2012.
- [Mar07] F. S. Marzano. “Predicting Antenna Noise Temperature Due to Rain Clouds at Microwave and Millimeter-Wave Frequencies”. *IEEE Transactions on Antennas and Propagation*, Vol. 55, No. 7, pp. 2022–2031, July 2007.
- [Mar12] J. Martinez-Lorenzo, Y. Rodriguez-Vaqueiro, C. Rappaport, O. Lopez, and A. Pino. “A compressed sensing approach for detection of explosive threats at standoff distances using a Passive Array of Scatters”. In *IEEE Conference on Technologies for Homeland Security (HST)*, pp. 134–139, Nov 2012.
- [Mar15a] MarketsandMarkets. “Millimeter Wave Technology Market by Product (Telecommunication Equipment, Imaging & Scanning Systems), Frequency Band (V-Band & E-Band), Component (Antennas & Transceivers & Amplifiers), Application, and Geography - Global Forecast to 2022”. http://www.marketsandmarkets.com/Market-Reports/millimeter-wave-technology-market-981.html?utm_source=PRnewswire&utm_medium=Referral&utm_campaign=PaidPR, 2015.
- [Mar15b] A. Martinez, M. Lort, A. Aguasca, and A. Broquetas. “Submillimetric motion detection with a 94 GHz ground based synthetic aperture radar”. In *IET International Radar Conference*, pp. 1–5, Oct 2015.
- [Mar16] A. Martinez, M. Lort, A. Aguasca, and A. Broquetas. “Micrometric deformation imaging at W-Band with GBSAR”. In *European Journal of Remote Sensing*, 2016.

- [Mas09] S. Masuda, T. Ohki, K. Makiyama, M. Kanamura, N. Okamoto, H. Shigematsu, K. Imanishi, T. Kikkawa, K. Joshin, and N. Hara. “GaN MMIC amplifiers for W-band transceivers”. In *European Microwave Integrated Circuits Conference (EuMIC)*, pp. 443–446, Sept 2009.
- [Meg15] S. Meguellati. “Precision optical device of freeform defects inspection”. *Proceedings of SPIE, Optical Systems Design: Optical Fabrication, Testing, and Metrology V*, Vol. 9628, 2015.
- [Meh12] G. Mehdi and J. Miao. “Millimeter wave FMCW radar for Foreign object debris (FOD) detection at airport runways”. In *9th International Bhurban Conference on Applied Sciences and Technology (IBCAST)*, pp. 407–412, Jan 2012.
- [Mei02] W. M. Meijninger, W. Kohsiek, and H. A. de Bruin. “The applicability of the scintillation method over heterogeneous areas”. In *Symposium on Boundary Layers and Turbulence*, p. 8, American Meteorological Society, 2002.
- [Men13a] B. Mencia-Oliva, J. Grajal, O. Yeste-Ojeda, G. Rubio-Cidre, and A. Badolato. “Low-Cost CW-LFM Radar Sensor at 100 GHz”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 61, No. 2, pp. 986–998, 2013.
- [Men13b] B. Mencia-Oliva, J. Grajal, A. Badolato, A. Garcia-Pino, J. L. Besada Sanmartin, and O. A. Yeste-Ojeda. “Experimental Radar Imager with Sub-cm Range Resolution at 300 GHz”. In *IEEE Radar Conference (RADAR)*, pp. 1–6, 2013.
- [Men14] B. Mencia-Oliva. *Desarrollo y aplicaciones de radares de alta resolución en bandas milimétricas*. PhD Thesis, Technical University of Madrid (Spain), 2014.
- [Mer15] D. Merk, H. Deneke, B. Pospichal, and P. Seifert. “Investigation of the adiabatic assumption for estimating cloud micro- and macrophysical properties from satellite and ground”. *Atmospheric Chemistry and Physics Discussions*, Vol. 15, No. 4, pp. 5129–5173, 2015.
- [MeT17] MeTeK. “Doppler Cloud Radar MIRA-35”. <http://metek.de/product-group/doppler-cloud-radar/>, 2017.
- [Mic05] M. Micovic, A. Kurdoghlian, H. Moyer, P. Hashimoto, A. Schmitz, I. Milosavljevic, P. Willadsen, W.-S. Wong, J. Duvall, M. Hu, M. Wetzel, and D. Chow. “GaN MMIC technology for microwave and millimeter-wave applications”. In *IEEE Compound Semiconductor Integrated Circuit Symposium*, Oct 2005.
- [Mic06] M. Micovic, A. Kurdoghlian, P. Hashimoto, M. Hu, M. Antcliffe, P. Willadsen, W.-S. Wong, R. Bowen, I. Milosavljevic, A. Schmitz, M. Wetzel, and D. Chow. “GaN HFET for W-band Power Applications”. In *International Electron Devices Meeting (IEDM)*, pp. 1–3, Dec 2006.
- [Mic12] M. Micovic, A. Kurdoghlian, A. Margomenos, D. Brown, K. Shinohara, S. Burnham, I. Milosavljevic, R. Bowen, A. Williams, P. Hashimoto, R. Grabar, C. Butler, A. Schmitz, P. Willadsen, and D. Chow. “92–96 GHz GaN power amplifiers”. In *IEEE MTT-S International Microwave Symposium Digest (MTT)*, pp. 1–3, June 2012.
- [Mil00] N. L. Miles, J. Verlinde, and E. E. Clothiaux. “Cloud Droplet Size Distributions in Low-Level Stratiform Clouds”. *Journal of the Atmospheric Sciences*, Vol. 57, No. 2, pp. 295–311, 2000.
- [Mil16] Millitech, Inc. “Antenna and Quasioptical Products”. <http://www.millitech.com/MMW-AntennaQuasioptical-Millitech.htm>, 2016.

- [Mill97] “Millimeter Wave Propagation: Spectrum Management Implications”. Technical report 70, Office of Engineering and Technology, New Technology Development Division. Federal Communications Commission, 1997.
- [Mor98] K. Moran, B. E. Martner, M. J. Post, R. Kropfli, D. Welsh, and K. . B. Widener. “An Unattended Cloud-Profiling Radar for Use in Climate Research”. *Bulletin of the American Meteorological Society*, Vol. 79, pp. 443–455, 1998.
- [Mot07] H. Mott. *Remote sensing with polarimetric radar*. John Wiley & Sons, Inc., 2007.
- [Mou12] J. E. Moulder. “Risks of Exposure to Ionizing and Millimeter-Wave Radiation from Airport Whole-Body Scanners”. *Radiation Research*, Vol. 177, No. 6, pp. 723–726, 2012.
- [Muo12] A. Muñoz-Acevedo, L. Rolo, M. Sierra-Castañer, and M. Paquay. “Influence of atmospheric conditions on mmWave CATR facilities: Formulation and compensation scheme”. In *6th European Conference on Antennas and Propagation (EUCAP)*, pp. 3681–3685, March 2012.
- [Mur11] Murata Manufacturing Co. *Application Manual P11E-1: Ceramic Filters*. 2011.
- [Mus97] T. Musch and B. Schiek. “A High Precision Analog Frequency-Ramp Generator using a Phase-Locked-Loop Structure”. *27th European Microwave Conference*, Vol. 1, pp. 62–68, 1997.
- [Nas04] A. Nashashibi, K. Sarabandi, S. Oveisgharan, M. Dobson, W. Walker, and E. Burke. “Millimeter-wave measurements of foliage attenuation and ground reflectivity of tree stands at nadir incidence”. *IEEE Transactions on Antennas and Propagation*, Vol. 52, No. 5, pp. 1211–1222, May 2004.
- [Nas16] A. Y. Nashashibi, A. A. Ibrahim, S. Cook, and K. Sarabandi. “Experimental Characterization of Polarimetric Radar Backscatter Response of Distributed Targets at High Millimeter-Wave Frequencies”. *IEEE Transactions on Geoscience and Remote Sensing*, Vol. 54, No. 2, pp. 1013–1024, Feb 2016.
- [Nav16] Navtech Radar. “AdvanceGuard”. <http://www.navtechradar.com/security-surveillance/>, 2016.
- [Nor08] D. Norenberg, S. Crewell, U. Lohnert, T. Rose, and A. Martellucci. “A Novel Ground-Based Microwave Radiometer for High Precision Atmospheric Observations between 10 and 90 GHz”. In *IEEE International Geoscience and Remote Sensing Symposium (IGARSS)*, pp. 919–922, July 2008.
- [Now16] S. Nowok, R. Herschel, R. Zimmermann, A. Shoykhetbrod, S. A. Lang, and N. Pohl. “3D imaging system based on a MIMO approach at 360GHz for security screening”. In *Progress in Electromagnetic Research Symposium (PIERS)*, pp. 671–675, Aug 2016.
- [Oka08] S. Oka, H. Togo, N. Kukutsu, and T. Nagatsuma. “Latest trends in millimeter-wave imaging technology”. *Progress In Electromagnetics Research Letters*, Vol. 1, pp. 197–204, 2008.
- [Oreg17] “Oregon Scientific”. <http://weather.oregonscientific.com/>, 2017.
- [Ovt99] M. Ovtchinnikov and Y. L. Kogan. “Evaluation of Cloud Water Retrieval Using Radar Measurements in Stratocumulus Clouds”. *Ninth ARM Science Team Meeting Proceedings*, March 1999.

- [Oze02] S. Ozev, A. Orailoglu, and H. Haggag. “Automated Test Development and Test Time Reduction for RF Subsystems”. In *IEEE International Symposium on Circuits and Systems (ISCAS)*, pp. 581–584, 2002.
- [Pac09] P. Pace. *Detecting and Classifying Low Probability of Intercept Radar*. Artech House, 2009.
- [Pag09] A. Pagels, M. Hagelen, G. Briese, and A. Tessmann. “Helicopter Assisted Landing System - Millimeter-Wave against Brown-Out”. In *German Microwave Conference*, pp. 1–3, March 2009.
- [Pak98] A. Pakhomov, Y. Akyel, O. Pakhomova, B. Stuck, and M. Murphy. “Current State and Implications of Research on Biological Effects of Millimeter Waves: A Review of the Literature”. *Bioelectromagnetics*, Vol. 19, pp. 393–413, 1998.
- [Pal14] S. Paloscia, E. Santi, and P. Pampaloni. “Radiometric features of land surfaces at 37GHz”. In *XXXIth URSI General Assembly and Scientific Symposium (URSI GASS)*, pp. 1–2, Aug 2014.
- [Per02] F. Perez, A. Asensio, J. Gismero, J. I. Alonso, J. M. Monje, F. Casanova, R. Cortijo, and J. F. Perez-Ojeda. “ARIES: a high-resolution shipboard radar”. In *Proceedings of the IEEE Radar Conference*, pp. 148–153, 2002.
- [Per13] G. Perez-Palomino, P. Baine, R. Dickie, M. Bain, J. A. Encinar, R. Cahill, M. Barba, and G. Toso. “Design and Experimental Validation of Liquid Crystal-Based Reconfigurable Reflectarray Elements With Improved Bandwidth in F-Band”. *IEEE Transactions on Antennas and Propagation*, Vol. 61, No. 4, pp. 1704–1713, April 2013.
- [Per15] G. Perez-Palomino, M. Barba, J. A. Encinar, R. Cahill, R. Dickie, P. Baine, and M. Bain. “Design and Demonstration of an Electronically Scanned Reflectarray Antenna at 100 GHz Using Multiresonant Cells Based on Liquid Crystals”. *IEEE Transactions on Antennas and Propagation*, Vol. 63, No. 8, pp. 3722–3727, Aug 2015.
- [Pet08] D. T. Petkie, C. Casto, F. C. D. Lucia, S. R. Murrill, B. Redman, R. L. Espinola, C. C. Franck, E. L. Jacobs, S. T. Griffin, C. E. Halford, J. Reynolds, S. O’Brien, and D. Tofsted. “Active and passive imaging in the THz spectral region: phenomenology, dynamic range, modes, and illumination”. *Journal of the Optical Society of America B*, Vol. 25, No. 9, pp. 1523–1531, Sep 2008.
- [Pop11] J. Popp, V. Tuchin, A. Chiou, and S. Heinemann. *Handbook of Biophotonics: Photonics for Health Care*. Wiley, 2011.
- [Pre16] J. du Preez and S. Sinha. *Millimeter-wave antennas configurations and applications*. Springer, 2016.
- [Pro62] J. R. Probert-Jones. “The radar equation in meteorology”. *Quarterly Journal of the Royal Meteorological Society*, Vol. 88, pp. 485–495, 1962.
- [Pru10] H. Pruppacher and J. Klett. *Microphysics of Clouds and Precipitation. Atmospheric and Oceanographic Sciences Library 18*, Springer Netherlands, 2 Ed., 2010.
- [Qua09] H. Quast and T. Löffler. “Towards Real-time Active THz Range Imaging for Security Applications”. In *International Conference on Electromagnetics in Advanced Applications (ICEAA)*, pp. 501–504, 2009.

- [Rad14a] Radiometer Physics. “FMCW Cloud Doppler Radar”. <http://www.radiometer-physics.de/products/microwave-remote-sensing-instruments/94-ghz-fmcw-doppler-cloud-radar/>, 2014.
- [Rad14b] Radiometer Physics. “Microwave Scintillometer”. http://www.radiometer-physics.de/download/PDF/Scintillometer/RPG-MWSC-160_Instrument_Manual.pdf, 2014.
- [Ran07] M. Rangwala, J. Lee, and K. Sarabandi. “Design of FMCW millimeter-wave radar for helicopter assisted landing”. In *Geoscience and Remote Sensing Symposium (IGARSS)*, pp. 4183–4186, July 2007.
- [Rap13] T. S. Rappaport, S. Sun, R. Mayzus, H. Zhao, Y. Azar, K. Wang, G. N. Wong, J. K. Schulz, M. Samimi, and F. Gutierrez. “Millimeter Wave Mobile Communications for 5G Cellular: It Will Work!”. *IEEE Access*, Vol. 1, pp. 335–349, 2013.
- [Rap15] T. S. Rappaport and S. Deng. “73 GHz wideband millimeter-wave foliage and ground reflection measurements and models”. In *IEEE International Conference on Communication Workshop (ICCW)*, pp. 1238–1243, June 2015.
- [Rec12] T. Reck, C. Jung-Kubiak, C. Lee, G. Chattopadhyay, I. Mehdi, and N. Llombart-Juan. “Terahertz Antenna for Arrays of Hundreds of Pixels”. In *IEEE Antennas and Propagation Society International Symposium (APSURSI)*, pp. 1–2, 2012.
- [Rec15] T. Reck, C. Jung-Kubiak, J. Siles, C. Lee, R. Lin, G. Chattopadhyay, I. Mehdi, and K. Cooper. “A Silicon Micromachined Eight-Pixel Transceiver Array for Submillimeter-Wave Radar”. *IEEE Transactions on Terahertz Science and Technology*, Vol. 5, No. 2, pp. 197–206, March 2015.
- [Rei14] J. Reichardt. “Cloud and Aerosol Spectroscopy with Raman Lidar”. *Journal of Atmospheric and Oceanic Technology*, Vol. 31, No. 9, pp. 1946–1963, 2014.
- [Ric10] M. A. Richards, J. A. Scheer, and W. A. Holm. *Principles of Modern Radar: Basic Principles*. SciTech Publishing, 2010.
- [Rob12] D. A. Robertson, P. N. Marsh, D. R. Bolton, R. J. C. Middleton, R. I. Hunter, P. J. Speirs, D. G. Macfarlane, S. L. Cassidy, and G. M. Smith. “340-GHz 3D Radar Imaging Test Bed with 10-Hz Frame Rate”. In *Proceedings of SPIE, Passive and Active Millimeter-Wave Imaging XV*, 2012.
- [Rob13] D. A. Robertson, S. L. Cassidy, and D. R. Bolton. “Nonlinearity and Phase Noise Effects in 340 GHz 3D Imaging Radar”. In *Proceedings of SPIE, Passive and Active Millimeter-Wave Imaging XVI*, 2013.
- [Rob16a] D. A. Robertson, D. G. Macfarlane, and T. Bryllert. “220 GHz wideband 3D imaging radar for concealed object detection technology development and phenomenology studies”. In *Proceedings of SPIE, Passive and Active Millimeter-Wave Imaging XIX*, 2016.
- [Rob16b] D. A. Robertson, D. G. Macfarlane, and T. Bryllert. “A 220 GHz 3D imaging radar with sub-cm³ voxel resolution for security applications”. In *European Radar Conference (EuRAD)*, pp. 398–401, Oct 2016.
- [Roe15] K. Roe and C. Gregory. “Wave-based sensing and imaging for security applications”. In *9th European Conference on Antennas and Propagation (EuCAP)*, pp. 1–5, April 2015.

- [Rsc16] M. Rösch, A. Tessmann, A. Leuther, R. Weber, G. Moschetti, B. Aja, M. Kotiranta, H. Massler, V. Kangas, M. G. Perichaud, M. Schlechtweg, and O. Ambacher. “Low noise amplifiers for MetOp-SG”. In *Global Symposium on Millimeter Waves (GSMM) ESA Workshop on Millimetre-Wave Technology and Applications*, pp. 1–4, June 2016.
- [Rub14] G. Rubio-Cidre, A. Badolato, L. Ubeda-Medina, B. Mencia-Oliva, J. Grajal, A. Garcia-Pino, B. Gonzalez-Valdes, and O. Rubinos Lopez. “Characterization of a 300 GHz imaging radar for standoff detection”. In *International Radar Conference*, pp. 1–6, Oct 2014.
- [Rub15] G. Rubio-Cidre, A. Badolato, L. Ubeda-Medina, J. Grajal, B. Mencia-Oliva, and B.-P. Dorta-Naranjo. “DDS-Based Signal-Generation Architecture Comparison for an Imaging Radar at 300 GHz”. *IEEE Transactions on Instrumentation and Measurement*, Vol. 64, No. 11, pp. 3085–3098, 2015.
- [Ruf06] D. Ruffieux, J. Nash, P. Jeannet, and J. Agnew. “The COST 720 temperature, humidity, and cloud profiling campaign: TUC”. *Meteorologische Zeitschrift*, Vol. 15, No. 1, pp. 5–10, Feb 2006.
- [Sam06] L. Samoska. “Towards Terahertz MMIC Amplifiers: Present Status and Trends”. In *IEEE MTT-S International Microwave Symposium Digest*, pp. 333–336, June 2006.
- [Sam08] L. Samoska, W. R. Deal, G. Chattopadhyay, D. Pukala, A. Fung, T. Gaier, M. Soria, V. Radisic, X. Mei, and R. Lai. “A Submillimeter-Wave HEMT Amplifier Module With Integrated Waveguide Transitions Operating Above 300 GHz”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 56, No. 6, pp. 1380–1388, June 2008.
- [Sam11] L. A. Samoska. “An Overview of Solid-State Integrated Circuit Amplifiers in the Submillimeter-Wave and THz Regime”. *IEEE Transactions on Terahertz Science and Technology*, Vol. 1, No. 1, pp. 9–24, Sept 2011.
- [Sas14] K. Sasaki, T. Sakai, T. Nagaoka, K. Wake, S. Watanabe, M. Kojima, N. Hasanova, H. Sasaki, K. Sasaki, Y. Suzuki, M. Taki, Y. Kamimura, A. Hirata, and H. Shirai. “Dosimetry Using a Localized Exposure System in the Millimeter-Wave Band for in vivo Studies on Ocular Effects”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 62, No. 7, pp. 1554–1564, July 2014.
- [Sau87] H. Sauvageot and J. Omar. “Radar Reflectivity of Cumulus Clouds”. *Journal of Atmospheric and Oceanic Technology*, Vol. 4, No. 2, pp. 264–272, 1987.
- [Sau92] H. Sauvageot. *Radar Meteorology*. Artech House, 1992.
- [Sch10] J. Schellenberg, E. Watkins, M. Micovic, B. Kim, and K. Han. “W-band, 5W solid-state power amplifier/combiner”. In *IEEE MTT-S International Microwave Symposium Digest (MTT)*, pp. 1–1, May 2010.
- [Sch13] J. Schellenberg, B. Kim, and T. Phan. “W-Band, Broadband 2 W GaN MMIC”. In *IEEE MTT-S International Microwave Symposium Digest (IMS)*, pp. 1–4, June 2013.
- [Sch15] S. Scherr, B. Göttel, S. Ayhan, A. Bhutani, M. Pauli, W. Winkler, J. C. Scheytt, and T. Zwick. “Miniaturized 122 GHz ISM band FMCW radar with micrometer accuracy”. In *European Radar Conference (EuRAD)*, pp. 277–280, Sept 2015.

- [Sch16] J. Schellenberg, A. Tran, L. Bui, A. Cuevas, and E. Watkins. “37 W, 75-100 GHz GaN power amplifier”. In *IEEE MTT-S International Microwave Symposium (IMS)*, pp. 1–4, May 2016.
- [Sch93] J. Scheer and J. L. Kurtz. *Coherent Radar System Performance Estimation*. Artech House, 1993.
- [Sek02] S. M. Sekelsky and E. E. Clothiaux. “Parallax Errors and Corrections for Dual-Antenna Millimeter-Wave Cloud Radars”. *Journal of Atmospheric and Oceanic Technology*, Vol. 19, No. 4, pp. 478–485, 2002.
- [Sek98] S. M. Sekelsky, L. Li, J. Calloway, R. E. McIntosh, M. A. Miller, E. E. Clothiaux, S. Haimov, G. G. Mace, and K. Sassen. “Comparison of Millimeter-Wave Cloud Radar Measurements for the Fall 1997 Cloud IOP”. *Proceedings of Annual ARM Science Team Meeting*, pp. 671–675, 1998.
- [She01] D. Sheen, D. McMakin, and T. Hall. “Three-dimensional millimeter-wave imaging for concealed weapon detection”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 49, No. 9, pp. 1581–1592, Sep 2001.
- [She05] D. M. Sheen, D. L. McMakin, W. M. Lechelt, and J. W. Griffin. “Circularly polarized millimeter-wave imaging for personnel screening”. pp. 117–126, 2005.
- [She09] D. Sheen, D. McMakin, T. Hall, and R. Severtsen. “Active Millimeter-wave Stand-off and Portal Imaging Techniques for Personnel Screening”. In *IEEE Conference on Technologies for Homeland Security (HST)*, pp. 440–447, 2009.
- [Shu04] M. D. Shupe, P. Kollias, S. Y. Matrosov, and T. L. Schneider. “Deriving Mixed-Phase Cloud Properties from Doppler Radar Spectra”. *Journal of Atmospheric and Oceanic Technology*, Vol. 21, No. 4, pp. 660–670, 2004.
- [Shu08] M. D. Shupe, P. Kollias, M. Poellot, and E. Eloranta. “On Deriving Vertical Air Motions from Cloud Radar Doppler Spectra”. *Journal of Atmospheric and Oceanic Technology*, Vol. 25, No. 4, pp. 547–557, 2008.
- [Shu16] M. D. Shupe, J. M. Comstock, D. D. Turner, and G. G. Mace. “Cloud Property Retrievals in the ARM Program”. *Meteorological Monographs*, Vol. 57, pp. 19.1–19.20, 2016.
- [Sie02] P. H. Siegel. “Terahertz technology”. *IEEE Transactions on Microwave Theory and Techniques*, Vol. 50, No. 3, pp. 910–928, Mar 2002.
- [Sie03] P. H. Siegel. “THz Technology: An Overview”. *International Journal of High Speed Electronics and Systems*, Vol. 13, No. 02, pp. 351–394, 2003.
- [Sie07] P. H. Siegel. “THz Instruments for Space”. *IEEE Transactions on Antennas and Propagation*, Vol. 55, No. 11, pp. 2957–2965, Nov 2007.
- [Sie10] P. Siegel, T. Löffler, D. Mittleman, K. Mizuno, and X. Zhang. “Guest Editorial - Terahertz Technology: Bridging the Microwave-to-Photonics Gap”. In *IEEE Transactions on Microwave Theory and Techniques*, 2010.
- [Sil13] G. A. Siles, J. M. Riera, P. G. del Pino, B. Mencia-Oliva, and J. Grajal. “Estimation of atmospheric attenuation at 99 GHz using a total power radiometer”. In *7th European Conference on Antennas and Propagation (EuCAP)*, pp. 2518–2522, April 2013.

- [Sil15] J. V. Siles, C. Lee, R. Lin, G. Chattopadhyay, T. Reck, C. Jung-Kubiak, I. Mehdi, and K. B. Cooper. “A High-Power 105-120 GHz Broadband On-Chip Power-Combined Frequency Tripler”. *IEEE Microwave and Wireless Components Letters*, Vol. 25, No. 3, pp. 157–159, March 2015.
- [Sin03] S. Singh and M. Singh. “Explosives detection systems (EDS) for aviation security”. pp. 31 – 55, 2003.
- [Sko08] M. Skolnik. *Radar Handbook, Third Edition*. McGraw-Hill, 2008.
- [Sko89] N. Skou. *Microwave radiometer systems: Design and analysis*. Artech House, 1989.
- [Spi01] S. Spitz, A. Prata, J. Harrell, R. Perez, and W. Veruttipong. “A 94 GHz spaceborne cloud profiling radar antenna system”. In *Aerospace Conference*, pp. 2/685–2/694, 2001.
- [Ste03] G. Stephens and D. Vane. “The CloudSat Mission”. In *IEEE International Geoscience and Remote Sensing Symposium (IGARSS)*, pp. 1326–1328, July 2003.
- [Ste09] B. Steer, A. Roitman, P. Horoyksi, M. Hyttinen, R. Dobbs, and D. Berry. “High power millimeter-wave Extended Interaction Klystrons for ground, airborne and space radars”. In *IEEE Radar Conference*, pp. 1–3, May 2009.
- [Ste13] B. Stevens and S. Bony. “What Are Climate Models Missing?”. *Science*, Vol. 340, No. 6136, pp. 1053–1054, 2013.
- [Tan13] A. Tang, K. B. Cooper, R. J. Dengler, N. Llombart, and P. Siegel. “Compact 680 GHz Sub-Millimeter Wave Imaging Radar Systems”. In *International Microwave Symposium (IMS), Seattle, Washington, June 2-7, Workshop WSN-6*, 2013.
- [Tan14] S. Y. Tan. *Meteorological Satellite Systems. Springer Briefs in Space Development*, Springer-Verlag New York, 2014.
- [Tay15] Z. Taylor, J. Garritano, S. Sung, N. Bajwa, D. Bennett, B. Nowroozi, P. Tewari, J. Sayre, J.-P. Hubschman, S. Deng, E. Brown, and W. Grundfest. “THz and mm-Wave Sensing of Corneal Tissue Water Content: In Vivo Sensing and Imaging Results”. *IEEE Transactions on Terahertz Science and Technology*, Vol. 5, No. 2, pp. 184–196, March 2015.
- [Tes01] T. Teshirogi and T. Yoneyama. *Modern Millimeter-wave Technologies. Wave Summit Series*, IOS Press, 2001.
- [Tho14] B. Thomas, M. Brandt, A. Walber, H. Gibson, M. Philipp, G. Sonnabend, M. Benzazaa, R. Gonzalez, M. Bergada, J. Martinez, M. G. Perichaud, T. Narhi, U. Klein, S. d’Addio, and V. Kangas. “Millimeter and sub-millimeter wave radiometer instruments for the next generation of polar orbiting meteorological satellites - MetOp-SG”. In *39th International Conference on Infrared, Millimeter, and Terahertz waves (IRMMW-THz)*, pp. 1–3, Sept 2014.
- [Tho16] Y. de Thonel d’Orgeix, R. Farré, S. Haddad, D. Sanson, D. Mouneyrac, J. M. Goutoule, T. Decoopman, A. Lemasson, J. Tailhades, R. Hammett, M. Rösch, A. Leuther, B. Aja, B. Alderman, J. Powell, V. Kangas, S. D’Addio, M. G. Périchaud, and P. Piironen. “Ultra low noise V-band down-converters for MetOp-SG”. In *Global Symposium on Millimeter Waves (GSMW) ESA Workshop on Millimetre-Wave Technology and Applications*, pp. 1–4, June 2016.
- [Ton07] M. Tonouchi. “Cutting-edge terahertz technology”. In *Nature Photonics*, pp. 97–105, 2007.

- [Tor04] G. Torregrosa-Peñalva. *Caracterización eléctrica y térmica de amplificadores MMIC de potencia*. PhD Thesis, Technical University of Madrid, 2004.
- [Tri15] F. Tridon and A. Battaglia. “Dual-frequency radar Doppler spectral retrieval of rain drop size distributions and entangled dynamics variables”. *Journal of Geophysical Research: Atmospheres*, Vol. 120, No. 11, pp. 5585–5601, 2015.
- [Tru12] R. True. “The evolution of microwave and millimeter wave tubes”. In *IEEE MTT-S International Microwave Symposium Digest (MTT)*, pp. 1–3, June 2012.
- [Tsa11] G. Tsaknakis, A. Papayannis, P. Kokkalis, V. Amiridis, H. D. Kambezidis, R. E. Mamouri, G. Georgoussis, and G. Avdikos. “Inter-comparison of lidar and ceilometer retrievals for aerosol and Planetary Boundary Layer profiling over Athens, Greece”. *Atmospheric Measurement Techniques*, Vol. 4, No. 6, pp. 1261–1273, 2011.
- [Tsu04] J. Tsui. *Digital Techniques for Wideband Receivers, 2nd Edition*. Artech House, 2004.
- [Ula90] F. Ulaby and C. Elachi. *Radar Polarimetry for Geoscience Applications*. Artech House, 1990.
- [Univ16] “University of Wyoming. Department of Atmospheric Science.”. <http://weather.uwyo.edu/upperair/sounding.html>, 2016.
- [Vai17] Vaisala. “Radiosonde RS92”. <http://www.vaisala.com/en/meteorology/products/soundingsystemsandradiosondes/radiosondes/Pages/RS92.aspx>, 2017.
- [Van10] J. Vankka and K. Halonen. *Direct Digital Synthesizers: Theory, Design and Applications*. Springer, 2010.
- [Wal03] M. Walden and A. Garrod. “A European low cost MMIC based millimetre-wave radar module for automotive applications”. In *33rd European Microwave Conference*, pp. 1195–1198 Vol.3, Oct 2003.
- [Wal99] W. Wall, A. Carramiñana, L. Carrasco, and P. Goldsmith. *Millimeter-Wave Astronomy: Molecular Chemistry and Physics in Space*. Springer Netherlands, 1999.
- [Wav06] WaveBand Corporation. “Development of a 94 GHz radar system for dedicated bird detection at airports and airfields”. Technical report AFRL-SN-RS-TR-2006-21, 2006.
- [Weh95] D. Wehner. *High-Resolution Radar*. Artech House, 1995.
- [Wei53] H. K. Weickmann and H. J. aufm Kampe. “Physical Properties of Cumulus Clouds”. *Journal of Meteorology*, Vol. 10, No. 3, pp. 204–211, 1953.
- [Wid12] K. Widener, N. Bharadwaj, and K. Johnson. *Scanning ARM Cloud Radar (X/Ka/W-SACR) Handbook*. US Department of Energy, Office of Science, 2012.
- [Wil07] B. S. Williams. “Terahertz quantum-cascade lasers”. *Nature Photonics*, Vol. 1, 2007.
- [Wil91] M. Wilson and T. Tozer. “Spurious Reduction Techniques For Direct Digital Synthesis”. In *IEEE Colloquium on Direct Digital Frequency Synthesis*, pp. 4/1–4/5, 1991.

- [Wol91] D. Wolaver. *Phase-Locked Loop Circuit Desing*. Prentice Hall, 1991.
- [Woo07] D. Woolard, J. Jensen, R. Hwu, and M. Shur. *Terahertz Science and Technology for Military and Security Applications*. World Scientific, 2007.
- [Wu05] L. Wu, S. S. Peng, X. H. Ying, and Z. C. Xu. “Remote sensing of atmospheric water vapor content and liquid water content with dual-channel millimeter wave radiometer”. In *Asia-Pacific Microwave Conference Proceedings*, Dec 2005.
- [Wu09] W. Wu, J. Long, and R. Staszewski. “A Digital Ultra-fast Acquisition Linear Frequency Modulated PLL for MM-wave FMCW Radars”. In *IEEE International Symposium on Radio-Frequency Integration Technology (RFIT)*, pp. 32–35, 2009.
- [Xia16] W. Xia and L. Huang. “Target vibration estimation in SAR based on phase-analysis method”. *EURASIP Journal on Advances in Signal Processing*, p. 94, 2016.
- [Yam06] J. Yamaguchi, T. Takano, T. Nakajima, T. Takamura, H. Ohno, Y. Nakanishi, K. Akita, Y. Kawamura, H. Futaba, and S. Yokote. “Sensitivity of FMCW 95 GHz cloud radar for high clouds”. In *Asia-Pacific Microwave Conference*, pp. 1841–1846, Dec 2006.
- [Yin11] J. Yinon. *Counterterrorist Detection Techniques of Explosives*. Elsevier, 2011.
- [Yuj03] L. Yujiri, M. Shoucri, and P. Moffa. “Passive millimeter wave imaging”. *Microwave Magazine, IEEE*, Vol. 4, No. 3, pp. 39 – 50, 2003.
- [Zhu09] B. Zhu, Y. Chen, K. Deng, W. Hu, and Z. S. Yao. “Terahertz Science and Technology and Applications”. In *Progress In Electromagnetics Research Symposium*, 2009.
- [Zie13] G. Zieger, D. Born, S. Anders, E. Heinz, K. Peiselt, A. Brömel, V. Zakosarenko, T. May, and H. G. Meyer. “A passive submillimeter video camera for security applications”. In *38th International Conference on Infrared, Millimeter, and Terahertz Waves (IRMMW-THz)*, pp. 1–2, Sept 2013.
- [Zou08] R. Zoughi and Kharkovsky. “Microwave and millimetre wave sensors for crack detection”. *Fatigue and Fracture of Engineering Materials and Structures*, Vol. 31, No. 8, pp. 695–713, 2008.
- [Zrn99] D. S. Zrnica and A. V. Ryzikov. “Polarimetry for weather surveillance radars”. *Bulletin of the American Meteorological Society*, Vol. 80, p. 389, 1999.