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UWB Millimeter-Wave FMCW Radar using Hilbert Transform Methods

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Abstract— **We present design and realization of a broadband FMCW Radar working in the Millimeter-Wave (MMW) region. The usable frequency range lies between 91 GHz and 102 GHz. The quadrature receiver signal is synthesized from a single homodyne receiver channel using Hilbert transform methods. This ultrawide bandwidth of 11 GHz is necessary for the purpose of a high resolution imaging task. Due to the fact that MMWs propagate easily through common clothing it is feasible to image objects like concealed weapons worn beneath the cloth. Imaging of humans in the MMW region is one possibility to enhance the capabilities of nowadays security checkpoints, e. g. at airports.**

I. Introduction

High resolution imaging heavily depends on broadband imaging sensors no matter whether one applies passive or active systems, direct imaging or synthetic aperture focusing methods. The resolution along at least one image axis, in most cases the range or propagation delay axis, is directly proportional to bandwidth and does not depend on the actual frequency domain. The selection of the frequency domain can be based on considerations with respect to the available technology and can be further chosen according to the desired propagation characteristics of the electromagnetic waves.

The MMW region (30 GHz . . . 300 GHz) and the THz region (300 GHz . . . 10 THz) provide fairly well conditions for short range, high resolution and ultrawideband imaging applications. Above ca. 300 GHz the use of spectroscopic information is possible.

For security applications dealing with the imaging of concealed objects, which are metallic materials, ceramic materials or explosives , the spectroscopic properties of the THz region could be a major advantage. The technology of THz sensors yet is not suitable for environments outside the laboratory and also is still very expensive [1], which is not the case for the MMW region. This fact makes the MMW region a good candidate. But it should be kept in mind that it cannot provide the spectroscopic information which could be used to identify certain materials unambiguously. Today one can also find fully developed devices and systems up to 200 GHz including all components needed for a broadband radar, e. g. sources, mixers, LNAs, power amplifiers and antennas.

The imaging of concealed objects, which in our case are mainly dielectric objects, requires the sensor to have high sensitivity and dynamic range even though a short range application with ranges below approx. 3 m is intended.

The system is supposed to operate in an indoor environment which requires a source in order to illuminate the person under surveillance no matter whether an active or passive radar (radiometer) is applied.

In order to implement a measurement system for the MMW range we developed and realized an ultrawideband FMCW radar which provides the bandwidth and dynamic range needed for high resolution images.

II. System Concept

A. Homodyne Radar Setup

Figure 1 illustrates the schematic of the MMW FMCW Radar. Basically a homodyne radar setup was chosen. The

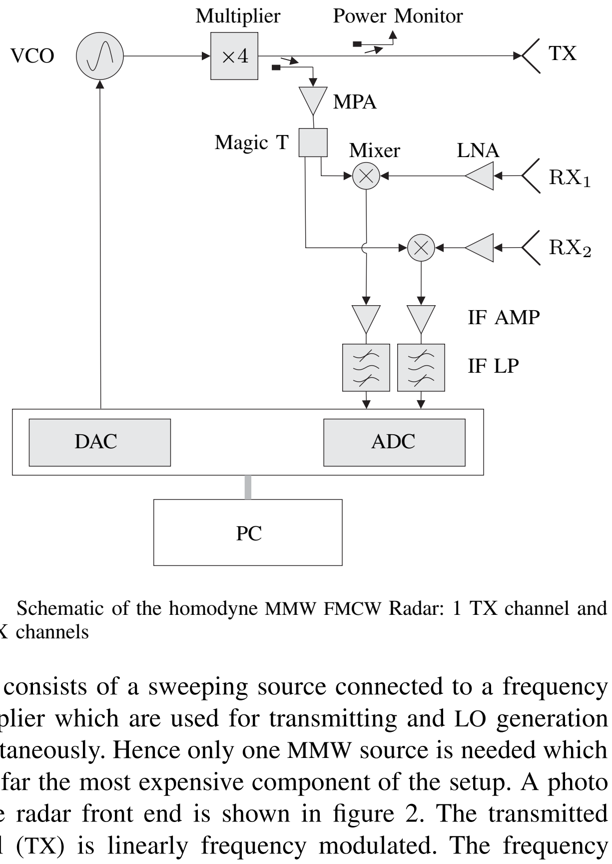

Fig. 1. Schematic of the homodyne MMW FMCW Radar: 1 TX channel and two RX channels

radar consists of a sweeping source connected to a frequency multiplier which are used for transmitting and LO generation simultaneously. Hence only one MMW source is needed which is by far the most expensive component of the setup. A photo of the radar front end is shown in figure 2. The transmitted signal (TX) is linearly frequency modulated. The frequency swept signals returned from the object are delayed copies of the transmitted signal. The delay is given by the round trip propagation time to the object and back.

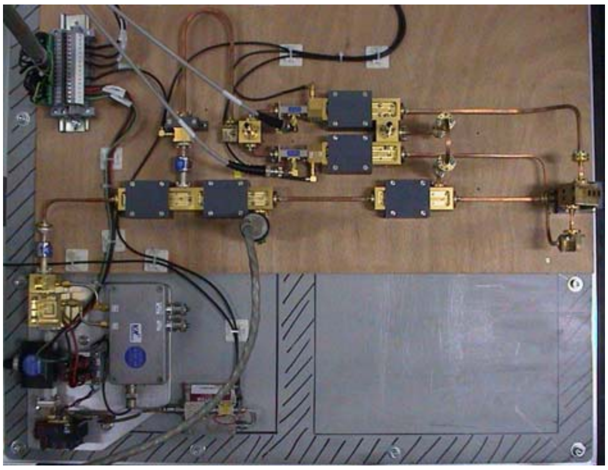

Fig. 2. Photo of the radar front end

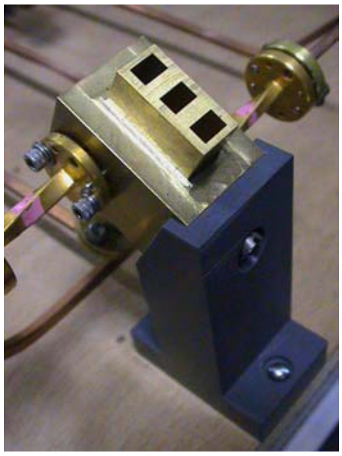

Fig. 3. Milled horn antenna block consisting of three antennas: 1 TX, 2 RX

The received signal (RX) is down-converted to IF. This results in an instantaneous difference in frequency between the transmitted and received signal. The IF signal is the so called beat frequency f_{b} which is linearly proportional to the range r to the object. The range resolution $\Delta r = {c_0}/{2B}$ is only depending on the usable system bandwidth B.

1) TX Signal: The TX and LO signal is generated by a voltage controlled oscillator (VCO) operating between 22.5 GHz and 25.5 GHz. It drives a frequency multiplier which has a multiplication factor of four. The usable output frequency range of the multiplier stage lies within 90.5 GHz and 102 GHz providing approx. 20 dBm output power. Most of this high amount of power is needed for the LOs. The pumping power at the mixers LO port should be within 10 dBm to 13 dBm. In order to reduce the power ripple of the multiplier to the LO input port of the mixer a broadband medium power amplifier (MPA) operating in saturation at approximately 17 dBm output power is used.

2) TX/RX Antenna: The TX signal is transmitted by a linear polarized horn antenna. For TX and RX separate antennas are used in order to avoid the effects of the return loss. Another reason for separation is that we consider to have two receiver channels. These two channels either can be used for interferometric imaging approaches or for measurements with two orthogonal polarizations. Figure 3 shows a photo of the antenna which is milled out of a single brass block.

3) RX Signal: The RX signal is amplified by a low noise amplifier (4.5 dB noise figure, 20 dB gain) before downconverting to IF. The IF signal is a low frequency low pass signal (0 Hz to 300 kHz). The upper frequency limit is only depending on the maximum expected range. For distances up to three meters we use 16 bit AD conversion equipment with a maximum sampling rate of about 1.5 MS/sec.

4) Data Acquisition: Because the FMCW radar transmits and receives simultaneously, signal generation and data acquisition have to be synchronized. In order to drive the VCO we use a 16 bit DA converter which is synchronized by a common clock with the AD conversion of the received signal and the further data processing.

B. Hilbert Transform Receiver

The realization of broadband quadrature mixers in W-Band is nontrivial because a broadband 90° phase shifter with sufficient accuracy cannot be realized easily. Also the calibration process and the removal of DC offsets on the I and Q channels require intense computational efforts.

The homodyne measurement system acquires a real valued beat frequency signal which is band limited and supposed to have a causal impulse response. It is considered to represent the real part $u_{\mathrm{s,Re}}(t)$ of the complex analytical signal, $u_{\mathrm{s}}(t) = u_{\mathrm{s,Re}}(t) + ju_{\mathrm{s,Im}}(t)$ (t equals the time axis). It is desirable to compute the analytical signal in order to obtain phase information which is necessary for coherent imaging and radar purposes.

The relationship between the real and imaginary components of $u_{\mathrm{s}}(t)$ is defined by the causality principle and can be established by using the Hilbert transform. The Hilbert transform can either be performed in the time domain by using a correlation filter or in the frequency domain by means of a multiplication with the spectral response of the Hilbert operator which is $\mathcal{H}(f) = -j\mathrm{sgn}(f)$ [2]. The analytical signal is obtained by

$$u_{\mathrm{s},\mathcal{H}}(t) = \mathcal{F}_f{}^{-1}\{\mathcal{F}_t\{u_{\mathrm{s,Re}}(t)\} \cdot \mathcal{H}(f)\}. \tag{1}$$

$\mathcal{F}$ stands for the Fourier Transform.

The impulse response of a Hilbert transformer has an infinite extend with respect to the time. A physical measurement system can only measure a time limited part of the signal. That means we measure the time windowed (rectangular window) real part of the analytical signal and thus the Hilbert transform only can provide an approximate solution for the imaginary part. The errors introduced by using the Hilbert transform can be reduced by using an appropriate window function before applying the Fourier transform [3] in equation 1.

III. Calibration

Similar to a network analyzer calibration the amplitude and phase response of the radar system have to be calibrated in order to obtain a flat amplitude response and to establish a fixed phase center. Due to the radar functionality of the homodyne measurement system the calibration has to be done in free space. Hence one has to use a reference calibration object like a corner reflector with a known and precise free space reflection coefficient S_{11}, e. g. a trihedral. By measuring the response of the reference object $u_{\mathcal{H},\mathrm{s},\mathrm{ref}}(t)$ and the response of the empty room $u_{\mathcal{H},\mathrm{s},\mathrm{emp}}(t)$ a calibration procedure for the measured data $u_{\mathcal{H},\mathrm{s},\mathrm{m}}(t)$ can be implemented by [4]

$$u_{\mathrm{s,cal,m}}(t) = \frac{u_{\mathcal{H},\mathrm{s},\mathrm{m}}(t) - u_{\mathcal{H},\mathrm{s},\mathrm{emp}}(t)}{u_{\mathcal{H},\mathrm{s},\mathrm{ref}}(t) - u_{\mathcal{H},\mathrm{s},\mathrm{emp}}(t)} \cdot S_{11,\mathrm{ref}} \quad . \tag{2}$$

IV. Linearization

The FMCW radar performance heavily depends on the linearity of the transmitted linear frequency modulated signal. Especially the range resolution is affected, which gets worse with increasing range. Frequency sweep non-linearities are often the limiting factor in FMCW radar range resolution [5].

As seen in figure 1 the chirp TX signal is generated by the VCO. The frequency output can be controlled by the tuning voltage which is supplied by the DA conversion equipment. The frequency vs. tuning voltage characteristics can be measured by a static setup, e. g. by means of a spectrum analyzer or frequency counter. This data can be used to supply a predistorted tuning voltage ramp to the VCO. One has to note that this method cannot correct for dynamic non-linearities.

V. Results

All following results have been obtained with the parameter settings documented in table I:

frequency range	$f_{\min} \dots f_{\max}$	91 GHz. . . 102 GHz
bandwidth	B	11 GHz
sweep time	T	2 ms
RX AD sampling rate	$f_{\mathrm{RX,s}}$	1 MHz
RX samples/sweep	$N_{\mathrm{RX,s}}$	1000
VCO DA sampling rate	$f_{\mathrm{VCO,s}}$	1 MHz
VCO linearization method	predistortion	
VCO samples/sweep	$N_{\mathrm{VCO,s}}$	1000
time between each sweep		5 ms
range resolution	Δr	13.6 mm
unambiguous range	r_{amb}	13.65 m
maximum beat frequency (allowed by low pass filter)	$f_{\mathrm{b,max}}$	300 kHz
maximum range	$r_{\max}$	8.18 m

TABLE I

Parameter settings used for operation

A. Linearization Using Predistortion

By using the above explained predistortion method based on a static lookup-table a linearity of about 0.1 % could have been obtained.

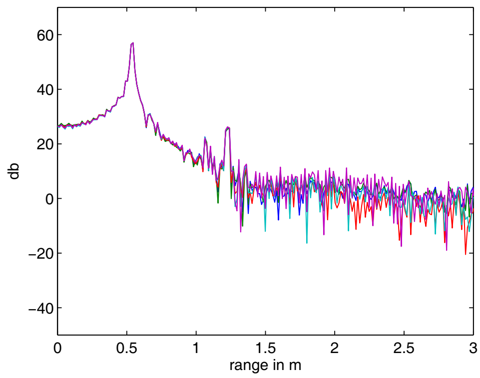

Fig. 4. Several calibrated range profiles of a trihedral at about 0.5 m distance to the antenna at range zero

B. Performance of the Calibrated Data

Fig. 4 shows a superposition of several calibrated range profiles of a trihedral which was also used as the reference calibration object but it was positioned with a small spacial offset. The peak's 3dB resolution width for the trihedral's position at about 0.5 m is very close to the theoretical limit of approx. 14 mm. The dynamic range is dependent on the sidelobe levels of the peak. In the case of a rectangular window it would be about 60 dB measured from the peak down to the lowest sidelobe level at a distance of ca. 3 m. It can be further increased by using some windowing function (e. g. a Kaiser–Bessel window), i. e. the dynamic range yet is not limited by noise within the range of reasonable window function. The non-symmetric behavior of the sidelobe spectrum comes from the errors of the Hilbert transform when applied on non-causal signals. These errors may be further reduced by using a window function before performing the Hilbert transform [3].

The calibrated amplitude response of the calibration reference has an ripple of ± 1 dB.

VI. Conclusions

We have developed an ultrawideband, homodyne MMW FMCW Radar with more than 10 GHz bandwidth and approx. 20 dBm TX power. It is designed for short range imaging applications with ranges up to 3 m. It is showing a dynamic range better than 60 dB which is expected to be sufficient for this type of application.

References

[1] J. F. Federici, B. Schulkin, F. Huang, D. Gary, R. Barat, F. Oliveira, and D. Zimdars, "Thz imaging and sensing for security applications – explosives, weapons and drugs," *Semiconductor Science and Technology*, vol. 20, no. 7, p. S266, 2005.

[2] J. G. Proakis and D. G. Manolakis, *Digital Signal Processing*, 3rd ed., ser. Principles, Algorithms, and Applications. New Jersey: Prentice Hall, 1996.

[3] D. Lipka, "A modified hilbert transform for homodyne system analysis," *AEÜ*, vol. 42, no. 3, pp. 190–192, 1988.

[4] F. C. Smith, B. Chambers, and J. C. Bennett, "Calibration techniques for free space reflection coefficient measurements," *Science, Measurement and Technology, IEE Proceedings-*, vol. 139, no. 5, pp. 247–253, 1992, 1350-2344.

[5] S. O. Piper, "Homodyne fmcw radar range resolution effects with sinusoidal nonlinearities in the frequency sweep," 1995, pp. 563–567.