Broadband Millimeter-wave FMCW Radar for Imaging of Humans

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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. We use a homodyne radar setup. Thus only one MMW source is necessary which is used for TX and LO generation simultaneously. The complex RX radar signal is calculated by a Hilbert transform in order to avoid a broadband MMW IQ mixer.**

A free space calibration procedure is used to obtain a flat amplitude response and a fixed phase center. Static non-linearities of the transmitted chirp signal are compensated by predistortion of the VCO's tuning voltage characteristic. A microwave coaxial delay line combined with a time domain resampling method corrects dynamic non-linearities.

The ultra wide 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 sensor (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 has been chosen. The

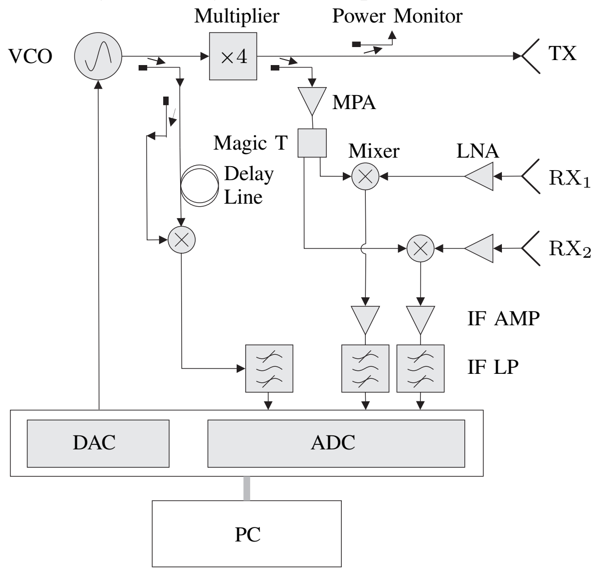


Fig. 1: Schematic of the homodyne MMW FMCW Radar: 1 TX channel and two RX channels

radar consists of a sweeping source connected to a frequency

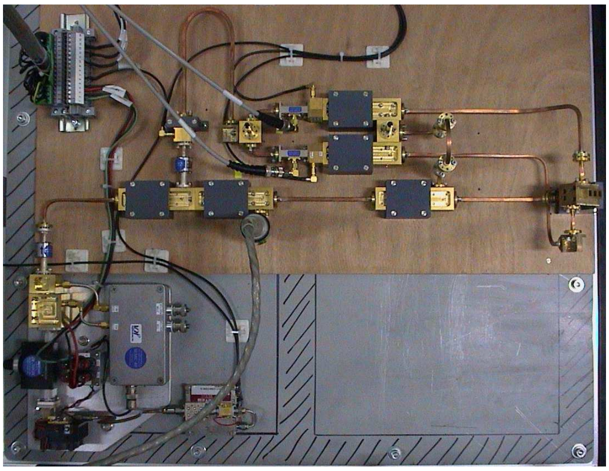


Fig. 2: Photo of the radar front end

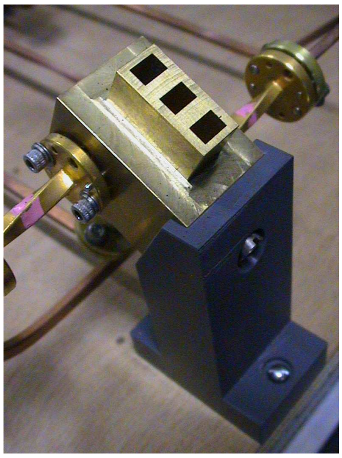


Fig. 3: Milled horn antenna block consisting of three antennas: 1 TX, 2 RX

multiplier which is 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.

The received signal (RX) is down-converted to baseband. This results in an instantaneous difference in frequency between the transmitted and received signal. The baseband 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 power is needed for the LOs. The pumping power at the mixers LO port is required to be within 10 dBm to 13 dBm. In order to reduce the power ripple of the multiplier at 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 linearly 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 similar built 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 baseband. The baseband 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 to 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 considerable computational efforts.

The homodyne measurement system acquires a real valued beat frequency signal which is band limited and assumed to have a causal impulse response. It is considered to represent the real part $u_{\mathrm{s,Re}}(t)$ of the complex analytical signal, $u_\mathrm{s}(t) = u_{\mathrm{s,Re}}(t) + j u_{\mathrm{s,Im}}(t)$. 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 $u_\mathrm{s}(t)$ can be derived by applying the causality principle and can be calculated by the Hilbert transform. The Hilbert transform can either be realized in time domain by using a correlation filter or in 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 extent with respect to time. A physical measurement system can only measure a time limited portion 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 $u_{\mathcal{H},\mathrm{s,ref}}(t)$ and the response of the empty room $u_{\mathcal{H},\mathrm{s,emp}}(t)$ a calibration procedure for the measured data $u_{\mathcal{H},\mathrm{s,m}}(t)$ can be implemented by [4]

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

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].

A. Predistortion

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

B. Resampling Method

Dynamic non-linearities are efficiently compensated for by a software resampling method [6]. If the transmitter generates an ideal linear chirp a static target causes a linear phase behavior of the beat frequency signal. The phase is proportional to the targets distance R with respect to the sensor, that is $\varphi_\mathrm{b} = 2\pi\frac{B}{T}\tau$, $(\tau = 2R/c_0)$. Any non-linearities in the chirp will cause non-linearities in the beat frequency phase. These phase errors can be equalized by resampling the measured signal in the way, that sampling is not performed at fixed time intervals Δt, but at fixed beat frequency signal phase increments $\Delta\varphi_\mathrm{b}$. A certain upsampling of the data is necessary to enable a convenient computation of a new time/sampling axis.

This method may be realized by using a delay line with a known length l_d and propagation velocity. The delay line produces an isolated target signature at distance $l_d/2$. Assuming that the frequency sweep is linear during a small time span Δt_i the instantaneous frequency of the beat signal equals

$$f_{\mathrm{b},i} = \frac{\Delta f_i}{\Delta t_i}\tau = \frac{1}{2\pi}\frac{\Delta\varphi_{\mathrm{b},i}}{\Delta t_i}, \tag{3}$$

where Δf_i is the frequency increment. Hence the actual transmitted bandwidth B can be calculated by

$$B = \sum_{i=1}^{N_\mathrm{s}-1} \frac{\Delta\varphi_{\mathrm{b},i}}{2\pi l_d/c_0}, \tag{4}$$

which is important because the range axis is related to the bandwidth.

In order to avoid attenuation and dispersion effects of rectangular waveguides the delay line is realized by a coaxial cable for the microwave signal, i. e. before multiplying the VCOs output signal.

Figure 4 shows the result obtained by this method. We use a delay line of length $l_d = 4.6\,\mathrm{m}$. Its propagation velocity is 77% of the free space velocity. After downconversion the beat frequency signal is Hilbert transformed in order to obtain the phase information. The non-linear phase behavior of the beat frequency is shown in 4a and the unfocused range profile in 4b. The range profile of the resampled beat frequency signal can be seen in figure 4c. The resampling indices are stored in a file and are used for linearization of the radars beat frequency signals.

V. Results

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

frequency range	$f_\mathrm{min} \cdots f_\mathrm{max}$	91 GHz…102 GHz
bandwidth	B	11 GHz
sweep time	T	1.25 ms
RX AD sampling rate	$f_\mathrm{RX,s}$	1 MHz
RX samples/sweep	$N_\mathrm{RX,s}$	1250
VCO DA sampling rate	$f_\mathrm{VCO,s}$	1 MHz
VCO linearization method	predist. & delay line	
VCO samples/sweep	$N_\mathrm{VCO,s}$	1250
range resolution	Δr	13.64 mm
unambiguous range	r_amb	17.05 m
maximum beat frequency (limited by low pass filter)	$f_\mathrm{b,max}$	300 kHz
maximum range	$r_\mathrm{max} = \frac{r_\mathrm{amb}}{2}$	8.53 m

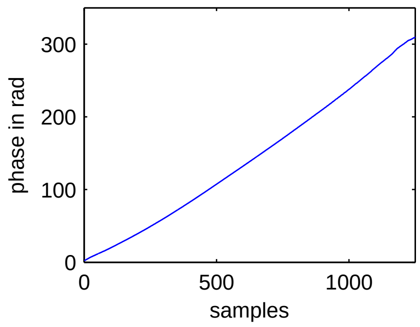
TABLE I: Parameter settings used for operation

A. Linearization Using Predistortion and Resampling

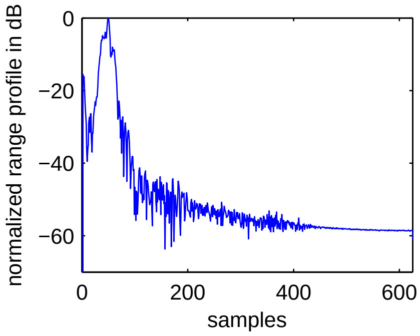
By using the predistortion and resampling method as explained above an effective linearity of about 0.1 % was achieved.

B. Performance of the Calibrated Data

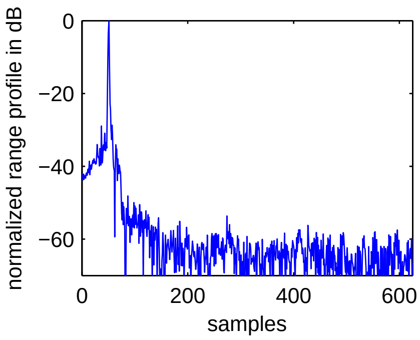
Fig. 5 shows the calibrated data of a trihedral which was also used as the reference calibration object when positioned with a small spacial offset. The peak's 3dB resolution width for the trihedral's position at about 0.75 m is very close to



(a) unwrapped non-linear beat frequency phase

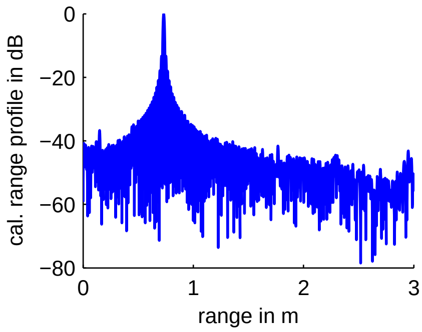


(b) range profile before resampling

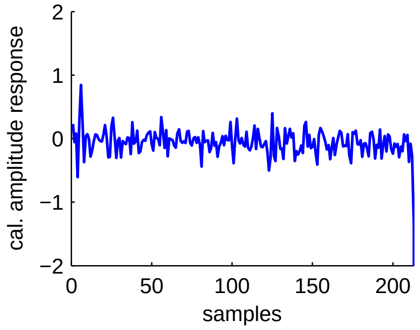


(c) linearized range profile

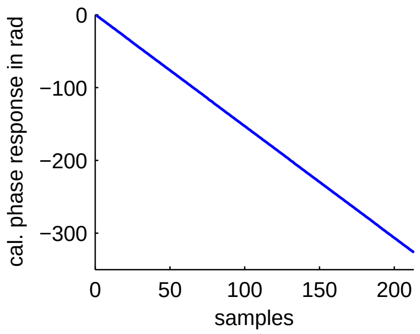
Fig. 4: Linearization by resampling method by means of a coaxial delay line with length $l_d = 4.6\,\mathrm{m}$ and propagation velocity $c_d = 0.77c_0$



(a) range profile



(b) amplitude response



(c) phase response

Fig. 5: Linearized and calibrated data of a trihedral at about $0.75\,\mathrm{m}$ distance to the antenna at range zero, range gated at ca. $3\,\mathrm{m}$

the theoretical expectation of approx. 14 mm. The dynamic range is dependent on the sidelobe levels of the peak. In the case of a rectangular window it was 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 a suitable 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 functions. The non-symmetric behavior of the sidelobe spectrum results from the errors produced by the Hilbert transform when applied on non-causal signals. These errors may be further reduced by using a window function before calculating the Hilbert transform [3].

The calibrated amplitude response of the calibration reference has a ripple of $\pm 1\,\mathrm{dB}$ as shown in figure 5b.

VI. Conclusions

We have developed an ultrawideband, homodyne MMW FMCW Radar with more than 10 GHz bandwidth and approx. 16 dBm TX power. It is designed for short range imaging applications with ranges up to 3 m. The complex radar signal required for calibration is obtained by applying a Hilbert transform. The radar is showing a dynamic range better than 60 dB which is expected to be sufficient for this type of application. Predistortion and a resampling method are implemented in order to correct for static and dynamic non-linearities, respectively.

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