NOVEL 24 GHZ FMCW FRONT-END WITH 2.45 GHZ SAW REFERENCE PATH FOR HIGH-PRECISION DISTANCE MEASUREMENTS

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ABSTRACT

A new 24 GHz non-linear FMCW radar is reported, which features a high-precision 2.45 GHz SAW (Surface Acoustic Wave) reference and adaptively compensates for phase errors by software. Kernel functions are implemented at a 2.45 GHz IF level reducing the expense of critical RF components in order to realize a fully planar front-end.

INTRODUCTION

Wireless sensing is essential for many future tasks in traffic control, industrial automation and identification [1]. Meanwhile, cost-effective K-band devices open up the commercial application of high resolution radar sensors. Especially for measuring speed or distance, microwave sensors can support or replace existing sensor techniques like ultrasonics or laser. Presently used FMCW radar sensors are based on waveguide modules with a Gunn VCO (Voltage Controlled Oszillator), which has limitations in terms of temperature stability, sweep bandwidth and FM linearity. The accuracy of a FMCW radar is severely restricted by the quality of the transmitted signal. Sophisticated system concepts are required, to achieve precise distance measurements. The new patented FMCW radar concept presented here avoids these constraints with a reasonable hardware expense.

CONSIDERATIONS ON FMCW RADAR

In FMCW radar a VCO frequency is swept linearly over the bandwidth Δf during the measuring period T. This signal returns with a time delay of τ_m. Mixing the delayed echo with the momentary transmit signal supplies the target signal

$$s_m(t) = \cos[2\pi f(t) \cdot t - 2\pi f(t-\tau_m) \cdot (t-\tau_m)]. \quad (1)$$

With the transmit signal

$$f(t) = f_0 + \frac{\Delta f}{2T} t \quad (2)$$

the beat frequency of the target signal of a single reflection is

$$f_m = \frac{\Delta f}{T} \tau_m. \quad (3)$$

The frequency f_m is determined by performing the FFT of $s_m(t)$. A non-linear sweep causes f_m to become a function of time $f_m = f_m(t)$, which spreads the spectrum of the target signal, equivalent to a decrease in accuracy and dynamic range of the measurement. Conventional linearization principles are either based on dynamic feedback loops [2] or on predistorted voltage ramps considering the voltage frequency characteristic stored in the memory of a digital controller [3, 4].

The novel patented FMCW radar front-end reported here operates with a non-linear sweep and linearizes the target signal supplementary by software. The compensation of systematic as well as statistic phase errors (i.e. phase noise) is based on measuring the target signal of an exactly known "distance standard" [5]. This standard is represented by a SAW delay line with $\tau_r = a \cdot \tau_m$, where a is an arbitrary factor. Assuming that the frequency sweep Δf_i is linear during a small time span Δt_i, the instantaneous frequencies of the reference signal and the target signal equal

$$f_{ri} = \frac{\Delta f_i}{\Delta t_i} \tau_r = \frac{\Delta \varphi_r}{2\pi \Delta t_i}, \quad (4)$$

$$f_{mi} = \frac{\Delta f_i}{\Delta t_i} \tau_m = \frac{\Delta \varphi_m}{2\pi \Delta t_i}. \quad (5)$$

The relation

$$\frac{f_{ri}}{f_{mi}} = \frac{\tau_r}{\tau_m} = \frac{\Delta \varphi_r}{\Delta \varphi_m} = a, \quad (6)$$

where $\Delta\varphi_r$ and $\Delta\varphi_m$ are the phase increments within an arbitrary time span, is independent of the actual sweep rate $\frac{\Delta f_i}{\Delta t_i}$ as long as neither the reference nor the target path is dispersive. Each interval between two zero crossings of the reference

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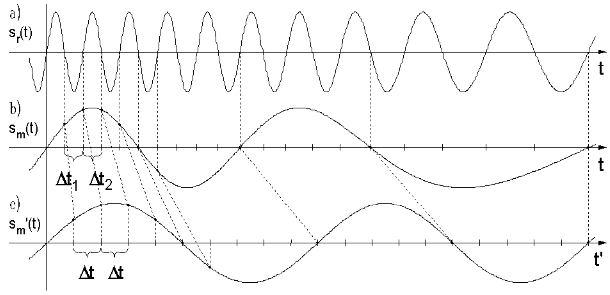

Figure 1: Principle of the linearization procedure

signal $s_r(t)$ corresponds to a reference phase increment of $\Delta\varphi_r = \pi$ and hence to a target phase increment of $\Delta\varphi_m = \frac{\pi}{a}$. If $a > 1$, the linearization can be performed by sampling the target signal $s_m(t)$ (Fig. 1b) at each zero crossing of $s_r(t)$ (Fig. 1a). Rearranging the data samples in a new array (Fig. 1c) returns a modified data set $s'_m(t)$. This rearrangement is equivalent to a full compensation of FM linearity errors. In case of using the reference and target data sets of the same sweep, the phase noise is compensated as well. This sampling procedure can be done either by software or by hardware.

SENSOR CONCEPT

By moving critical system components, i.e. the VCO and the reference delay line to a 2.45 GHz IF, a complete planar design of the 24 GHz frontend is realized, based on the heterodyne concept depicted in Fig. 2. A control voltage $m(t)$ sweeps the frequency of the 2.45 GHz VCO monotonely over the bandwidth Δf. One part of the VCO signal feeds the reference block comprising the 2.45 GHz SAW delay line and a mixer yielding the reference signal $s_r(t)$. The other part is up-converted with a 21.7 GHz LO signal, bandpass filtered, amplified and fed through a directional coupler to the antenna. The transmit/receive hybrid diplexes the delayed echo signal from the antenna to the down-converter which is pumped by the LO. The resulting delayed 2.45 GHz IF signal is subsequently mixed with the IF transmit signal providing the sensor signal $s_m(t)$ for further digital signal processing.

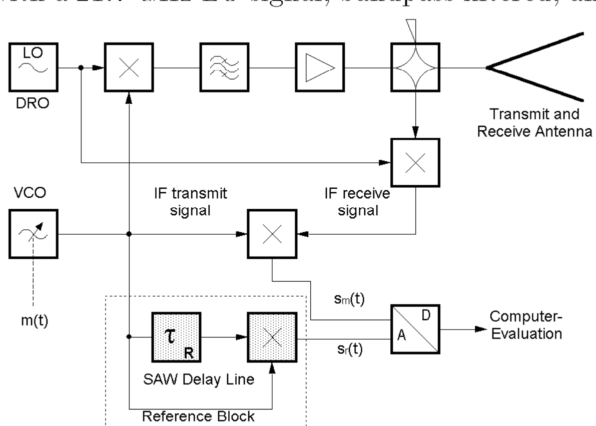

Figure 2: 24 GHz FMCW sensor with 2.45 GHz SAW reference

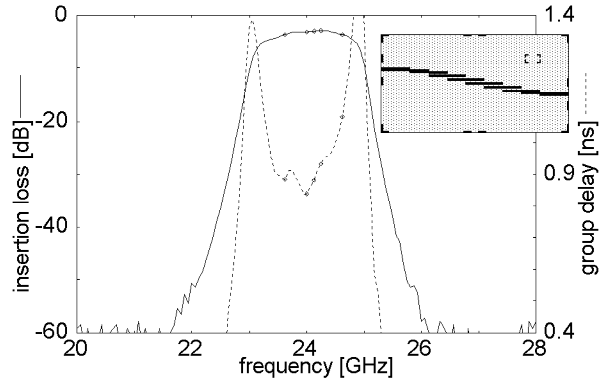

Figure 3: Filter layout and measured filter response

HARDWARE REALIZATION

The complete RF-circuitry, except for the 24 GHz amplifier, is designed in microstrip technique. The bandpass filter, the DRO and the amplifier will be discussed in the following.

The bandpass filter, stringent in a heterodyne concept in order to reject spurious intermodulation products like the LO and the lower sideband, turns out to be the component restricting the choice of the substrate. Experimental investigations reveal, that the substrate thickness mainly affects the stop band rejection, while the dielectric constant mainly influences the insertion loss and the reproducibility of the filter. Thus, the best performance is achieved with a thin substrate, keeping the wave better within the substrate, which is equivalent to a better stopband rejection, and with a low dielectric constant, which is equivalent to low losses and high reproducibility due to the larger strip line structure. The filter, depicted in Fig. 3, yields an excellent bandpass characteristic with an insertion loss of less than 5 dB and a stopband rejection of about 55 dB.

A 21.7 GHz DRO, oscillating at a higher order resonator mode [6, 7], operates as LO and achieves an output power of +12 dBm with a phase noise of -100dBc/Hz @ 100kHz (Fig. 4). The active device is a PHEMT with a matched gain of about 10 dB at 21.7 GHz. The SMD transistor performance

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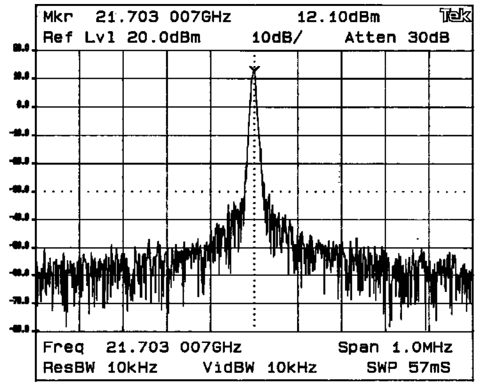

Figure 4: Measured 21.7 GHz DRO spectrum

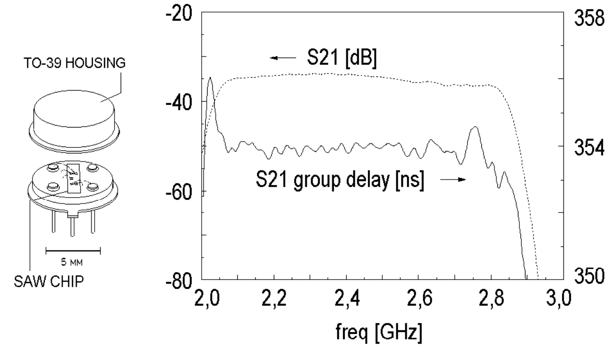

Figure 6: Package outline of the SAW delay line (left) and measured S21 and group delay (right)

exhibits reasonable scattering parameter tolerances sufficient for a minimum-tuning DRO design but still critical for the required 24 GHz amplifier. Thus, an external flip-chip MMIC amplifier with about 20 dB gain is employed [8].

The amplifier, flip-chip mounted on an Al_2O_3 substrate, is assembled in a separate housing using standard SMA connectors. The substrate interconnections are coplanar waveguides. As depicted in Fig. 5, the measurements of the amplifier S-parameters on-chip, flip-chip bonded and fully assembled show very good agreement.

The complete 2.45 GHz IF circuitry is designed with components taken from the high-volume communication market, except for the 2.45 GHz SAW delay line, where a sophisticated design was made for [9]. In order to achieve a high frequency, a dispersive split finger IDT design, operating at the 3rd harmonic, was made. The fabrication of the SAW devices requires a submicron lift-off patterning process. The delay line has a time delay of 354 ns, offers a very low time delay ripple of less than 500 ps within a broad bandwidth and fits into a small TO-39 package (Fig. 6).

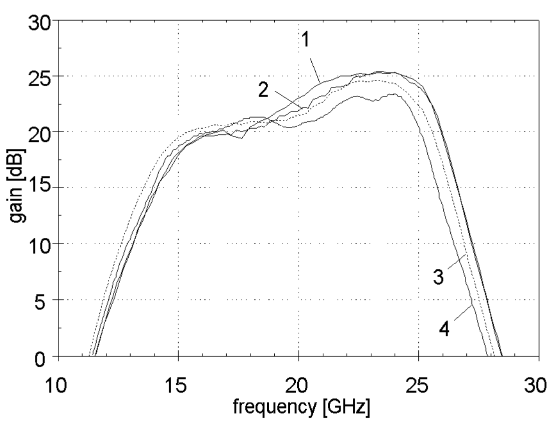

Figure 5: Measured gain (S21) of the amplifier:1: on-chip; 2+3: flip-chip; 4: complete amplifier module

MEASUREMENTS

Fig. 7 shows the 24 GHz FMCW spectrum with 250 MHz bandwidth. Distance measurements were made in a multitarget scenario with three main reflectors. Fig. 8 depicts the FFT of the target signal before (left) and after (right) the linearization using the same data set. In normal operation, the VCO is swept using a linear voltage ramp. However, the SAW reference method linearizes correctly even if arbitrary monotone waveforms are used. Special attention is paid to the group delay ripple of the reference delay and of the 24 GHz bandpass filter. Any group delay ripple of the 2.45 GHz SAW delay line will lead to erroneous sampling, while the dispersion of the 24 GHz bandpass filter is equivalent to a frequency dependent target delay. Both effects basically will spread the target spectrum, but the measurements reveal, that none of the group delay ripples is affecting the measurement accuracy significantly.

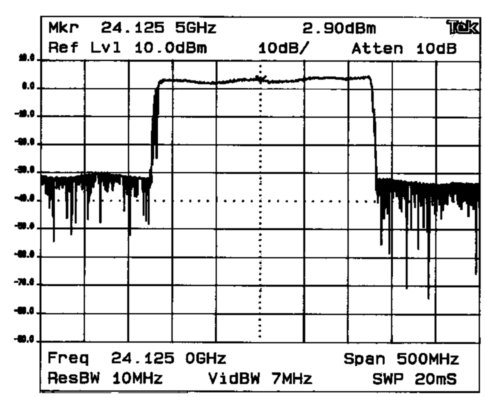

Figure 7: Measured 24 GHz FMCW spectrum

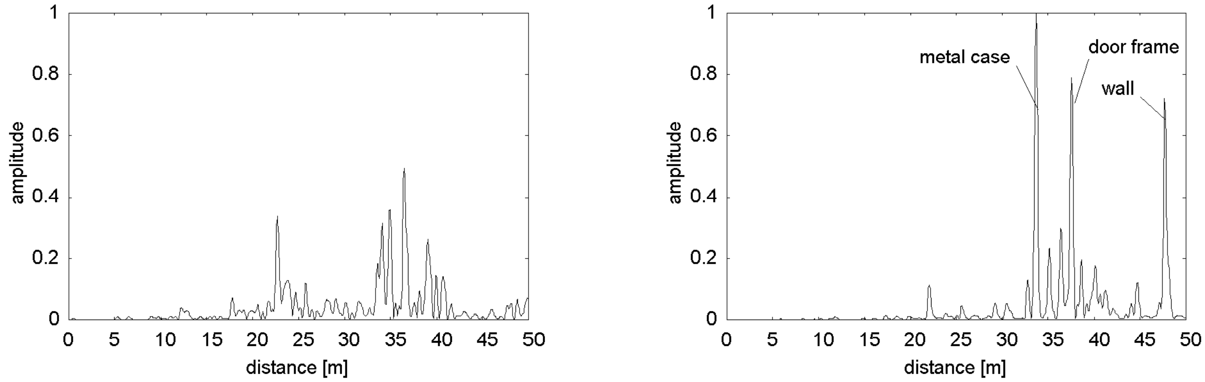

Figure 8: Range profile of a scenario with three reflectors before (left) and after (right) sweep linearization

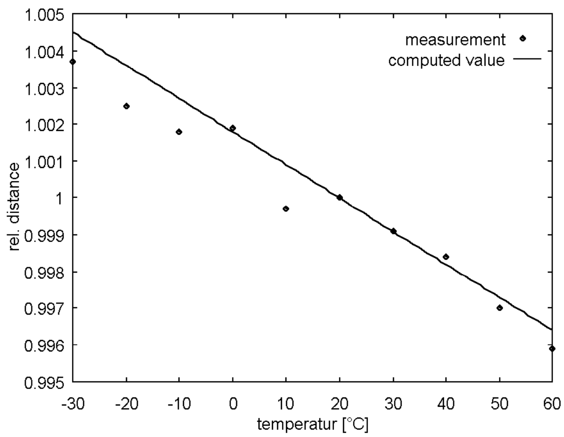

Figure 9: Measured and theoretically estimated temperature drift of the distance

Due to the temperature behaviour of the SAW delay line (+90ppm/°C), a software based temperature compensation must be performed. The comparison of the theoretical deviation of the relative distance, caused by the temperature characteristic of the SAW delay line and the relative distance measured (Fig. 9), allows a very precise temperature compensation.

CONCLUSION

A novel non-linear heterodyne 24 GHz FMCW radar sensor featuring a highly efficient compensation of phase errors has been built. The supplementary software linearization, based on the measurement of a high-precision 2.45 GHz SAW reference path, proved to give excellent accuracy at lower cost than presently used hardware methods. The modularity of the heterodyne concept provides a significant simplification for the design and leads to a complete 24 GHz planar sensor.

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