Design of a 24 GHz FMCW Radar System based on Sub-Harmonic Generation

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Abstract: This paper presents a novel frequency modulated continuous wave (FMCW) radar system based on sub-harmonic generation where a 24 GHz frequency divider-by-10 is used as an active reflector tag. A practical prototype is designed and fabricated on a GF45nm-SOI technology for the 24 GHz building blocks, while a GF0.18 μ m 7WL BiCMOS technology was used for the 2.4 GHz receiver (RX) and baseband. System measurement results show that as opposed to conventional primary radars, the proposed system is immune to strong multi-path interferences resulting from direct reflections of the interrogating signal. When measured in lab environment, the system achieves a ranging precision of 3.7 mm. Moreover, when measured in an indoor environment, ranging results show a ranging precision and accuracy of 5.8 and 22.3 cm, respectively, which outperforms other FMCW radars in the literature.

1. Introduction

Accurate indoor localization systems are highly needed by many applications such as automated vehicles, automotive industries, safe evacuation systems and warehouse management with demanding specifications on accuracy and precision. FMCW radar systems have been proven to be a good candidate for indoor localization systems with a precision of few tens of centimetres [1] and ranging accuracy of few centimetres [2]. However the accuracy of FMCW radar systems is degraded in indoor environments due to multi-path interferences caused by static or moving scatterers.

In order to enhance the robustness of FMCW radar systems in indoor environments against multi-paths, many techniques were adopted. The use of directive antennas was proposed in [1]. Another technique showed that the use of multi-bands for ranging improves the robustness of FMCW radar systems against interferers [3]-[4], however; the added redundancy increases the cost and the processing time. In [5], FMCW radar with spatial diversity was proposed to reduce multi-path interferences resulting from ground reflections in automotive radars. Other approaches to mitigate multi-path interferences are the multiple-input multiple-output (MIMO) radar with digital beam forming (DBF) proposed in [6] and the switched-beam antenna array implemented in [7]. However, the proposed solutions in [5]-[7] come on the expense of more complexity since the base-station (BS) requires either multiple numbers of RXs, mixers or antennas. The use of an artificial on-chip target (OCT) in order to mitigate short range leakage caused from direct reflections of very close objects in front of the radar antennas was investigated in [8]-[10]. This technique is suitable for automotive applications where the car bumpers are usually very close to the radar antennas causing severe interferences. However, in other indoor environments, this technique will not be able to mitigate reflections of moving objects as it depends on designing the delay time of the OCT to be the same as the round-trip delay time of the direct reflections.

The use of modulated backscatter transponders with FMCW radars was also proposed in order to reduce multi-path

Fig. 1. Proposed FMCW radar system

effects [11]. However, the proposed active tag has a power consumption of 75 mW, which is relatively high for battery operated devices. The use of nonlinear transmission lines (NLTL) as passive harmonic tags was proposed in [12]-[13]. In this system, the transmitter (TX) and the RX of the radar front-end are at two different frequencies. The interrogating signal is transmitted at frequency f_0 and then reflected at the second ($2f_0$) or third ($3f_0$) harmonic by the harmonic tag which eliminates multi-path interferences from direct reflections of the interrogating signal. However NLTLs suffer from conversion loss during harmonic generation. Additionally, the low output power of these tags limits the localization range and a high transmitting power of 4 W is required from the radar BS [13].

In [14], the use of frequency dividers as sub-harmonic generators in FMCW radar systems is proposed in order to eliminate multi-path interferences of the direct reflections of the interrogating signal and to provide a conversion gain of 48 dB which helps to increase the range of the radar system.

In this work, a first prototype of the FMCW radar system based on sub-harmonic generation is implemented, where the two ISM bands at 24 and 2.4 GHz are utilized for the transmitted and received signals respectively. In this system, the 24 GHz frequency divider-by-10 developed in [14] is used as a sub-harmonic active reflector tag. The rest of this paper is organized as follows; in section 2, an overview of the proposed sub-harmonic FMCW radar system is presented. In section 3, the design and measurement of both the active reflector tag and the FMCW radar frontend are presented.

Then in section 4, measurements of the proposed sub-harmonic FMCW radar system in lab and indoor environments are conducted and compared to state-of-the-art FMCW radars. Finally, conclusions and future work are presented in section 5.

2. System Overview

Fig. 1 shows a top level overview of the proposed sub-harmonic FMCW radar system. At the TX side, a chirp is generated at a starting frequency f_0 which is swept linearly over a bandwidth B_{fm} in a time T_{fm} , as expressed by:

$$s_{TX}(t) = A_{TX} \cos(2\pi f_0 + \frac{B_{fm}}{2T_{fm}} t^2 + \varphi_{TX}) \quad (1)$$

where A_{TX} is the signal amplitude at the TX output and φ_{TX} is an arbitrary starting phase.

The transmitted chirp $s_{TX}(t)$ is then propagated at the speed of light c , till it reaches the sub-harmonic active reflector tag after a time-of-flight (ToF) τ . There, the 10th sub-harmonic is generated by the frequency divider, sent back and received by the RX after the same ToF τ as expressed by:

$$s_{RX}(t) = A_{RX} \cos(2\pi \frac{f_0}{10} + \frac{B_{fm}/10}{2T_{fm}} (t - 2\tau)^2 + \varphi_{RX}) \quad (2)$$

where A_{RX} and φ_{RX} are the signal amplitude and phase at the RX input respectively. The received signal is then mixed with a locally generated 10th sub-harmonic of the transmitted chirp generating a baseband signal whose frequency f_D is equal to the frequency difference between the locally generated and received chirps and related to the round-trip time-of-flight (RToF) by:

$$f_D = \frac{0.1 B_{fm} 2\tau}{T_{fm}} \quad (3)$$

and then the range d_0 can be evaluated by:

$$d_0 = \frac{f_D c T_{fm}}{0.2 B_{fm}} \quad (4)$$

Since the transmitted and received chirps occupy two different frequency bands, any direct reflections by passive scatterers are not demodulated to a baseband frequency thus leading to better robustness towards multi-path propagation as compared to conventional primary FMCW radars.

3. System Design

3.1. Sub-Harmonic Active Reflector Tags

A 24 GHz frequency divider-by-10 was designed and fabricated in GF45nm-SOI technology to be used as a sub-harmonic generator for the active reflector tag [14]. The chip is bonded onto a 4-layer Rogers-4350B printed circuit board (PCB) as shown in Fig. 2. A passive band-pass filter (BPF) and a rat-race balanced-to-unbalanced (balun) converter were designed on the PCB at the 24 GHz input. Voltage regulators and current sources are used to provide the required bias point to the frequency divider. The chip consumes a total current of 7.4 mA from a 1 V supply.

The phase noise (PN) of the 2.4 GHz output of the frequency divider is measured at -128 dBc/Hz at 1 MHz offset frequency (Fig. 3a). The divider locking range was measured again on PCB after the BPF was used and compared to the previous on-wafer measurements (Fig. 3b).

Fig. 2. Active reflector tag
(a) PCB (4.8 cm x 4.5 cm), (b) Bonded die photo of the frequency divider

Fig. 3. Frequency divider PCB measurements
(a) Phase noise of the frequency divider-by-10 measured on PCB, (b) Divider locking range on PCB with band-pass filter

Fig. 4. Sub-harmonic FMCW radar front-end diagram

The PCB measurement still shows a good input sensitivity of -50 dBm at 24 GHz. It is worth noting that since the bond wire inductances were taken into account when designing the on-chip radio frequency (RF) input matching network, the ripples in the on-wafer-probed sensitivity curve of the chip are now smoothed in the PCB measured version.

3.2. FMCW Radar Front-end

Fig. 4 shows a simplified block diagram of the proposed sub-harmonic FMCW radar frontend where it is here partitioned into three chips:

Chip 1 is a 24 GHz voltage-controlled oscillator (VCO) that was fabricated also on GF45nm-SOI technology (Fig. 5a). Measured on PCB, the VCO chip has a PN of -119 dBc/Hz at 10 MHz offset frequency, an output power of -10.7 dBm and a 24-26.25 GHz tuning range.

Chip 2 is the same frequency divider chip used in the active reflector tag. This chip provides the frequency divided-by-10 local oscillator signal for mixing with the 2.4 GHz received signal.

Finally, chip 3 is a 2.4 GHz FMCW radar RX and fractional-N phase locked loop (Frac-N PLL) fabricated on a GF0.18 μ m 7WL BiCMOS technology (Fig. 5b). This chip is a modified version of a previous design reported in [15] where on the RX side, it includes the same integrated low-noise amplifier (LNA), mixer, baseband variable gain amplifier (VGA) and automatic gain control (AGC) loop. While on the TX side, the integrated VCO in [15] is replaced by the external 24 GHz VCO (Chip 1) followed by the 24 GHz frequency divider-by-10 (Chip 2) such that together with the integrated multi-modulus divider (MMD), sigma-delta modulator (SDM), phase frequency detector (PFD) and charge pump (CP), a 24 GHz Frac-N PLL is created. For chirp generation, the fractional (and integer) parts of the PLL frequency control word are incremented gradually using a chirp counter at a rate that is faster than the PLL bandwidth, such that a smooth frequency ramp is generated at the PLL output. In order to mitigate a frequency glitch issue when the output frequency approaches integer multiples of the 20 MHz reference frequency during chirp generation, integer pipelining stages are used as explained in [16].

A summary of the chips specifications is shown in tables 1, 2 and 3. The three chips are bonded on a 4-layer Rogers PCB with dimensions 10 cm x 7 cm as shown in Fig. 6a. A rat-race balun similar to that used in the active reflector tag is used to convert the differential output of the VCO (Chip 1) to a single-ended output to feed the TX antenna. Additionally, an RF directional coupler is designed on PCB in order to couple the VCO output to the input of the frequency divider-by-10 (Chip 2). The PCB also includes the PLL loop filter, the baseband low-pass filter (LPF) and the analogue-to-digital converter (ADC) for digitizing the down-converted baseband signal.

Another 8-layer FR-4 PCB (Fig. 6b) with a field programmable gate array (FPGA) and a micro-controller is stacked on top of the radar frontend PCB in order to perform digital signal processing and to provide digital controls.

Fig. 5. Bonded chip photos
(a) 24 GHz VCO chip, (b) 2.4 GHz FMCW RX and PLL chip

Table 1 Summary of Chip 1 specifications

Parameter	Value	Unit
Process	45 SOI	nm
P _{dc}	12.3	mW
Supply	1	V
Area	0.64	mm ²
Tuning Range	24-26.5	GHz
Efficiency	2.15	GHz/mW
P _{out}	-10.7	dBm
PN @ 10 MHz	-119	dBc/Hz

Table 2 Summary of Chip 2 specifications

Parameter	Value	Unit
Process	45 SOI	nm
P _{dc}	7.4	mW
Supply	1	V
Area	0.64	mm ²
Centre Frequency	24	GHz
Input Sensitivity Level	-50	dBm
Relative Bandwidth ^a	33	%
Efficiency ^b	3.58	GHz/mW
P _{out}	-2	dBm
PN @ 1 MHz	-128	dBc/Hz

$$^a \text{Relative BW} = \frac{f_{\max} - f_{\min}}{(f_{\max} + f_{\min})/2} \times 100 \%, \text{ at } P_{\text{in}} = -20 \text{ dBm}$$

$$^b \text{Efficiency} = f_{\max}/P_{\text{dc}}$$

Table 3 Summary of Chip 3 specifications

Parameter	Value	Unit
Process	180 BiCMOS	nm
P _{dc}	150	mW
Area	2.4	mm ²
Frequency Band	2.4	GHz
RX Bandwidth	100	MHz
RX Conversion Gain	82.0	dB
RX Noise Figure	7.3	dB
RX IP1dB	-24	dBm
PLL Freq. Resolution ^a	0.6	Hz
MMD Divide Ratio (N)	120-125	
CP Current	800	μ A
SDM Order	3 rd	
SDM Word Width	25	bit

^a Relative to MMD input

In locked state, the PN of the generated 24 GHz VCO signal and the divided-by-10 locally generated signal at 2.4 GHz are measured as shown in Fig. 7, having around 20 dB difference as expected due to the division factor of 10. Fig. 8 shows an example of the VCO control voltage (V_{ctrl}) taking the expected triangular shape during the 24 GHz chirp generation.

Fig. 6. FMCW radar base-station
(a) Radar front-end PCB, (b) Digital PCB with FPGA and micro-controller

Fig. 7. Measured phase noise of both the fundamental and the local sub-harmonic signal

Fig. 8. VCO control voltage during chirp generation

4. System Measurements

4.1. Lab Measurements

Ranging experiments using the proposed sub-harmonic FMCW radar prototype are performed in laboratory conditions using RF cables and power dividers/combiners in order to easily mimic a direct reflection of the interrogating signal. At the TX side of the FMCW radar frontend, a power divider is used in order to split the power equally into two

paths, one that is transmitted through a long RF cable to the active tag, and the other is transmitted through a short cable back to the RX side of the FMCW radar front end in order to mimic a strong direct reflection from a passive scatterer as shown in Fig 9a. The directly reflected signal is added through a power combiner to the received subharmonic signal generated from the active tag. For the sake of comparison, a similar setup is used for a conventional primary FMCW radar using the same hardware from [15] at 2.4 GHz as shown in Fig. 9b. Fig. 10 shows the reconstruction of the digitized baseband signal from the sub-harmonic FMCW radar system indicating that chirp generation, transmission, reception and demodulation are successful resulting in a baseband signal with a signal-to-noise ratio of 32 dB at the RX output. Fig. 11 shows the final ranging result using both systems. The proposed sub-harmonic FMCW radar system can successfully detect the range to the active tag even with the presence of a strong direct reflection of the interrogating signal (represented by the short RF cable in Fig. 9) while the conventional FMCW radar failed to do so.

Fig. 9. Multi-path measurement setup
(a) Sub-harmonic radar ranging, (b) Conventional primary radar ranging

Fig. 10. Reconstruction of the digitized baseband signal

Fig. 11. Multi-path ranging measurement
(a) Sub-harmonic radar ranging, (b) Primary radar ranging

Fig. 12. CDF of the sub-harmonic radar ranging error

A thousand consecutive measurements were performed in order to evaluate the system ranging precision. The corresponding cumulative distribution function (CDF) of the ranging error is plotted in Fig. 12 where the ranging precision is measured at 3.7 mm.

4.2. Indoor Measurements

Ranging measurements were conducted indoors as shown in Fig. 13a, where omnidirectional and horn antennas were used for the 2.4 and 24 GHz channels, respectively. The distance between the active tag and the BS was measured at several locations. At each location, one hundred ranging measurements were performed, showing an overall ranging precision of 5.8 cm. Fig. 13b shows the measured range calculated by the proposed FMCW radar vs. the actual range acquired from a laser meter. The root-mean-square (RMS) ranging error is measured at 22.3 cm, which shows that the system accuracy is suitable for indoor ranging applications.

In order to further illustrate the system immunity to direct reflections vs. conventional primary radars, another ranging measurement was conducted with a corner reflector as shown in Fig. 14a and b. The proposed system could successfully detect the same desired range as opposed to the primary radar, thus showing its suitability for challenging indoor FMCW radar applications (Fig. 14c and d).

Fig. 13. FMCW radar indoor measurement results
(a) Radar ranging measurement setup (b) Measured range vs. actual range

Fig. 14. FMCW radar measurement results
(a) Subharmonic radar measurement setup with corner reflector (b) Primary radar measurement setup with corner reflector (c) Sub-harmonic radar ranging with and without corner reflector (d) Primary radar ranging with and without corner reflector

Table 4 Comparison with state-of-the-art FMCW radars

Parameter	[4]	[6]	[17]	[17]	[18]	This	Unit
Frequency	2.4	5.8	5.8	5.8	5.8	24/2.4	GHz
Precision ^a	56	N/A	5.96	26	11.4	5.8	cm
Accuracy ^b	40	37.0	26.6	74.2	55	22.3	cm
Measurement Type	Positioning	Positioning	Ranging	Ranging	Ranging	Ranging	
Architecture	Secondary	MIMO	Active Tag	Active Tag	Secondary	Subharmonic Tag	
Antenna Type	Omni-directional	Four-element Array	Directional	Omni-directional	Directional	Directional/Omni-directional	

^a Defined as the standard deviation in a set of ranging measurements taken at the same range

^b Defined as the standard deviation of all ranging measurement errors at all ranges

Compared to the state-of-the-art, Table 4 shows that the proposed system outperforms other reported systems in the literature in terms of ranging precision and accuracy. The performance of the proposed system may be further improved by combining the VCO and the frequency divider on one chip leading to a more compact PLL and therefore less susceptibility to noise and interference on the V_{ctrl} and supply voltages [19]–[20].

5. Conclusion and Future Work

A novel FMCW radar system that is based on sub-harmonic generation was investigated. Integrated chips required for both RF front-end and the sub-harmonic low-power active reflector tags were fabricated and the application PCBs were manufactured. System measurements were carried out showing that the proposed system has sub-centimetre precision in lab conditions and is immune to interferences caused by direct reflections of the interrogating signal as opposed to conventional primary FMCW radar systems. Future work includes the integration of the VCO and the frequency divider on a single chip, as well as development of more compact 24 GHz patch antennas for the active reflector tags.

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7. References

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