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A HETERODYNE SIX-PORT FMCW RADAR SENSOR ARCHITECTURE BASED ON BEAT SIGNAL PHASE SLOPE TECHNIQUES

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Abstract—A Heterodyne six-port FMCW collision avoidance radar sensor configuration based on beat signal phase slope techniques is presented in this paper. Digital IF circuitry has been used in order to avoid problems related to DC offset and amplitude and phase imbalance. Simulations show that the velocity and range to the target is obtained simultaneously, with very good accuracy. Results are compared to other techniques and system architectures.

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1. INTRODUCTION

The paper presents new six-port architecture using frequency modulated continuous wave (FMCW) technique, to obtain range and speed information simultaneously. This information is embedded in the spectrum of the beat signal, which is the mixing product of received signal with transmitted one. Beat frequency measurement is performed conventionally, by means of frequency counting or Fast Fourier transform (FFT) techniques. For accurate measurement, these conventional techniques require a long data acquisition time which must cover several cycles of the beat signal. For data acquisition time covering less than one cycle of the beat signal for beat frequency measuring, the phase slope method can be useful. This method was used for range and speed measurement in FMCW radar in [1, 2] and for Doppler frequency measurement in [3–6]. The implementation of the phase slope method requires an interferometer which can be seen in millimeter-wave domain as an I/Q-mixer. For cost efficiency and compactness six-port interferometers or six-port I/Q-mixer have been used [1–6] in the homodyne architecture in radar sensor receivers. Although homodyne FM-CW radar is known for its simplicity, it incorporates several problems: DC offset, amplitude and phase imbalance, FM-AM conversion noise. DC offset and amplitude and phase imbalance are caused by the six-port imbalance and the non-linear analog devices (diodes, transistors, etc) used for the I/Q-mixer. The FM-AM conversion noise is produced by unwanted envelope components resulting from transmission line effects, mixing, and frequency response limitations of the FM modulator [7]. In fact DC offset, amplitude and phase imbalance and FM-AM conversion noise cause unacceptable measuring errors. To overcome the problem related to the DC offset, amplitude and phase imbalance, six-port calibration is performed conventionally, in the frequency domain with network analyzer and important data processing means are necessary in final determination of speed and range [4] and [5]. In order to avoid this complicated calibration with network analyzer, baseband data processing techniques, known as baseband analytical calibration were proposed in [1] and [2]. Both methods do not reduce the FM-AM conversion noise.

In present paper, a heterodyne architecture using a digital IF I/Q mixer is proposed for the phase slope method implementation. The heterodyne architecture reduces the FM-AM conversion noise and low frequency noise as demonstrated in [7] and [8], improving thereby the receiver sensitivity. The digital IF I/Q mixer avoid DC offset and amplitude and phase imbalance.

2. THE HETERODYNE RADAR SENSOR CONFIGURATION

FM-AM conversion noise is the largest noise component in a homodyne radar receiver [7]. The FM-AM conversion noise is produced by unwanted envelope components resulting from transmission line effects, mixing, and frequency response limitations of the FM modulator. Given that the FM-AM conversion noise and its harmonics lie in the same frequency band as the detected beat signal, the sensitivity of the radar receiver is degraded. In order to improve the sensitivity of the receiver, the frequency range of the beat signal must be separated from that of FM-AM conversion. This is achieved by a heterodyne down-conversion. In a heterodyne configuration the beat signal is in the IF band and the FM-AM conversion noise, remaining in baseband, can be filtered by the IF band-pass filter. The heterodyne configuration alone cannot solve the remaining problems, namely DC offset, amplitude and phase imbalance. They are produced on one hand by the imbalance structure of the analog passive multi-ports (such as baluns, couplers and six-ports), and, on the other hand, by the fact that the analog non-linear components used for mixers (diodes, transistors), which generate the DC component by rectifying the millimeter-wave signals, are not ideally identical. In order to overcome the problems related to DC offset and amplitude and phase imbalance, the heterodyne down-conversion must be performed in conjunction with digital IF circuitry. The DC offset problem is solved in such a way that, in place of analog mixers, digital multipliers are used. They generate only the lower and the upper sidebands, DC and other mixing components are not generated. The amplitude imbalance results from unequal power splitting in analog configurations. In digital domain, in place of power splitting, data are duplicated. The same data go to both I and Q channels. In this manner the problem of amplitude imbalance is solved.

Figure 1 shows the simplified operating principle of the heterodyne FMCW radar with a vector receiver using a six-port double-balanced mixer [9] and a single sideband mixer in the receiver. The millimeter-wave voltage controlled oscillator (VCO) is modulated by a linearly swept frequency signal which gives a linear frequency modulated signal at its output. A portion of the generated FMCW signal is coupled to the six-port mixer at port 5 and serves as reference signal. The main part of generated FMCW signal is amplified and transmitted towards the target. The signal reflected from target is received by radar antenna and amplified using a low noise amplifier (LNA). The amplified receiving signal is first downconverted by the first mixer. Because of the fact that the used six-ports are not very wide band in frequency,

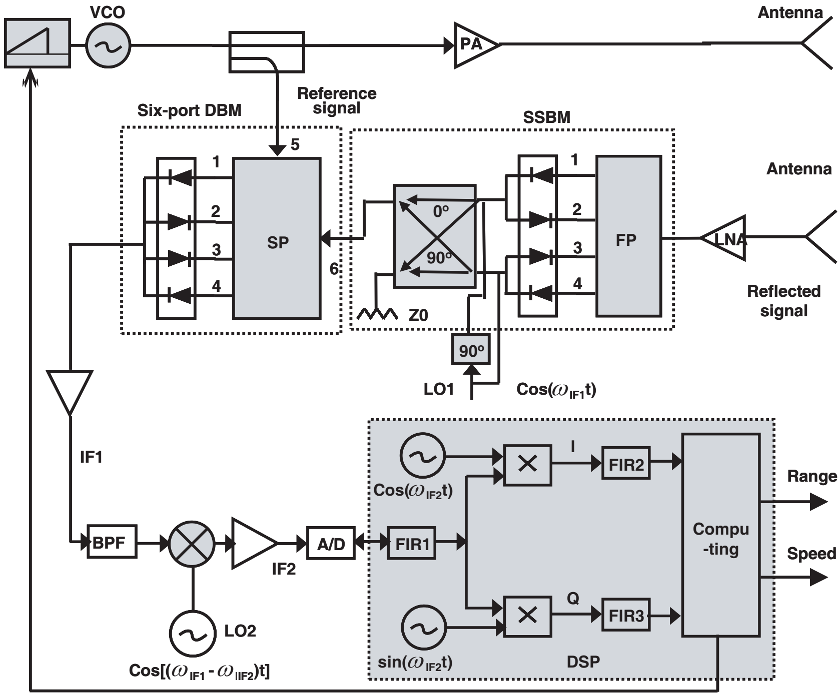

Figure 1. Block diagram of the heterodyne FMCW radar sensor with a six-port double-balanced mixer a five-port single sideband mixer.

the resulting IF frequency cannot be high. Therefore, both sideband signals move closer together in frequency, forcing the selectivity of the filter to be high enough, in order to separate the two sideband signals. Due to the fact that it is impossible to design a bandpass filter with very sharp cut-offs at millimeter-wave frequencies, a single sideband mixer has been used. Unlike conventional mixers, single sideband mixers achieve sideband rejection through phase cancellation, not filtering, so the frequency spacing between, the desired and undesired sidebands can be small. This means that down-conversion can be accomplished without filtering, and in fewer stages, saving the cost of extra mixers, amplifiers, local oscillators, and filters. The output signal of the single sideband mixer is mixed by the six-port double-balanced mixer with the reference signal to give the IF signal. The IF signal is amplified, A/D converted, and then downconverted to baseband with a digital I/Q mixer whose sine and cosine function are numerically generated and have their frequencies equal to the IF frequency. The outputs of the I/Q mixer are processed to obtain relative speed and range of target.

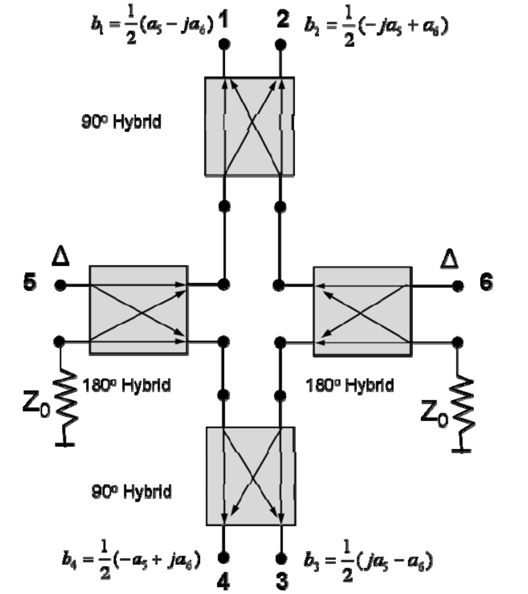

Figure 2. Block diagram of the six-port junction.

Figure 2 shows the block-diagram of the six-port used for the double-balanced mixer, a passive circuit composed of two 90° and two 180° hybrid couplers. The relationship between output and input normalized waves is highlighted. It is to be noted that various millimeter-wave six-port front-end architectures, fabrication technologies, and modulation schemes were proposed in recent years [12, 13].

3. RANGE AND SPEED MEASUREMENT

In order to implement the phase slope method for beat frequency measuring in FMCW radar, DC offset, amplitude and phase imbalance and FM-AM noise must be negligible. Two solutions are offered to us for solving the problem. The first one is to use simple homodyne architecture (hardware) and spend much time in signal processing (software). The second one is to use a complex heterodyne or double heterodyne architecture and spend little time in signal processing. In [1] and [2] we proposed the first kind of solution, namely a simple homodyne architecture with a complex signal processing to overcome the problems related to DC offset and amplitude imbalance. In the present paper, we propose the second kind of solution, namely, heterodyne architecture combined with digital IF circuitry. The heterodyne architecture reduces FM-AM noise and the digital IF circuitry keeps DC offset and amplitude and phase imbalance at minimum. In fact, after analog to digital conversion, the same data go to both I and Q channels, so that the problem of amplitude imbalance is avoided. The DC offset is not generated due to the fact that, in place of analog mixers, digital multipliers are used. They produce only, lower and upper sidebands. The lower sideband is, in case, of direct conversion a baseband signal without DC offset. The upper sideband can be filtered easily by the Finite Impulse Response Filters (FIR). The local oscillator for digital I/Q mixer is a numerically controlled oscillator whose sine and cosine functions are generated accurately, i.e., same amplitude and 90° phase shift between them, so that the phase imbalance is avoided.

The complex beat signal at the digital I/Q mixer outputs (Figures 1) is:

$$\bar{S} = S_I + jS_Q. \tag{1}$$

Due to the fact that DC offset and amplitude and phase imbalance are negligible, the I/Q signals can be written as follows:

$$S_I = A\cos(2\pi f_b t + \varphi_0) = A\cos\varphi, \tag{2}$$

$$S_Q = A\sin(2\pi f_b t + \varphi_0) = A\sin\varphi. \tag{3}$$

From (2) and (3) the beat signal instantaneous phase is found to be:

$$\varphi = 2\pi f_b t + \varphi_0 = Arc\tan\left(\frac{S_Q}{S_I}\right). \tag{4}$$

From (4) the beat frequency can be determined as follows:

$$f_b = \frac{1}{2\pi}\frac{d\varphi}{dt}. \tag{5}$$

Figure 3 shows the waveform of the FMCW signal. It is the same as that used in [2]. The top part of this figure shows plots of frequency versus time for signal transmitted by radar (solid line), and return signal from target received by radar (dotted line). The bottom part of figure shows frequency of output signal generated by radar mixer with transmitted and received input signal frequencies.

The frequency shift due to the range of target moving away from the radar is [8]:

$$f_R = f_b(up) - f_d. \tag{6}$$

$f_b(up)$ and f_d are, respectively, the beat frequency at rising chirp and the Doppler frequency. It is to be noted that range can be calculated

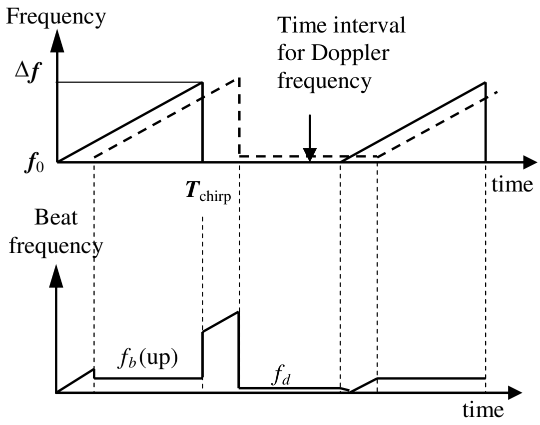

Figure 3. FMCW waveforms.

using f_R as follows [8]:

$$R = \frac{cf_R}{2f_{chirp}\Delta f}. \tag{7}$$

From f_d one can find the relative speed [8]:

$$v_r = \frac{cf_d}{2f_0} \tag{8}$$

where:
c is the speed of light in vacuum.
f_0, the center frequency,
Δf, the bandwidth of the FMCW signal and
$F_{chirp} = 1/T_{chirp}$ is the chirp cycle.

4. SIMULATION RESULTS

Design and simulation consist of three sections as follows: (i) design and electromagnetic simulation of multi-ports: six-port junction, five-port junction and a Wilkinson power divider, (ii) large-signal harmonic balance simulation of the six-port double-balanced mixer, and the single sideband mixer, and (iii) envelope simulation of the heterodyne radar system according to Figure 1.

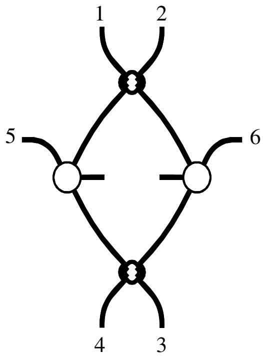

Figure 4. Layout of the six-port junction.

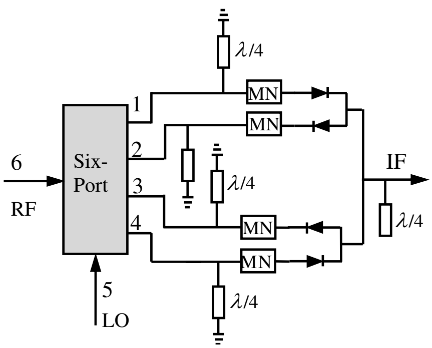

Figure 5. Block schematic of the six-port double-balanced mixer.

4.1. Six-port Double-balanced Mixer

The six-port junction was simulated with electromagnetic simulation tool software of Agilent Technologies: Momentum of Advanced Design System (ADS). The circuit was designed in MHMIC technology on a 127 µm alumina substrate with a relative permittivity of 9.9. Figure 4 shows the layout of the six-port junction. The outer dimensions are 6.5 mm by 6.5 mm.

In order to perform advanced system simulations, S parameters of six-port model have been saved in a data file. This data file is then used as data input to large-signal harmonic balance simulation of mixer along with Spice model of Microwave Device Technology (MDT) flip-chip Schottky diodes.

Figure 5 shows the block schematic of the six-port double-balanced mixer. In order to provide a ground return at the IF frequencies, four $\lambda/4$ short-circuit stubs were used. As known, these stubs are open circuits for RF and LO signal. One $\lambda/4$ open-circuit stub is connected to the mixer output preventing LO and RF leakage in IF path. In addition, matching networks (MN) are inserted at the inputs of all Schottky diodes.

Figure 6 shows the spectrum at the IF output port. As seen, all the harmonics are efficiently suppressed. Their levels are at least 130 dB under the IF signal level. This fact is to be explained that, beside suppression of harmonics due to the balanced structure of the mixer, the six-port itself acts as a filter. Figure 5(b) represents the zooming of the Figure 5(a) in the neighborhood of the IF frequency. The levels of all spurious products of the form mf_{RF}–mf_{LO} are at least 70 dB under the IF signal level and their filtering with an extra filter is no more necessary.

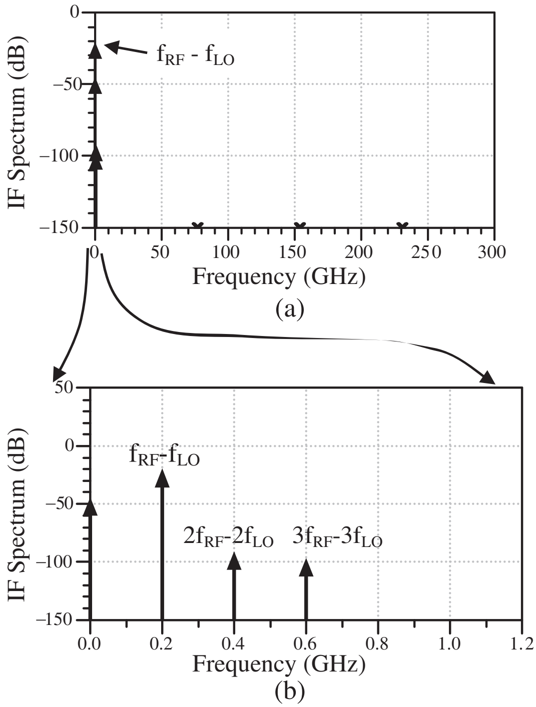

Figure 6. IF output spectrum of the six-port double-balanced mixer.

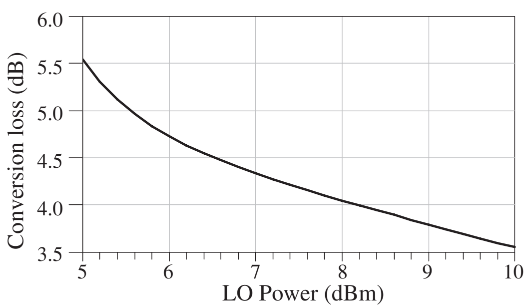

Figure 7. Conversion loss versus LO power of the six-port double-balanced mixer.

Figure 7 shows the conversion loss value versus the LO power. In order to improve the conversion loss, the LO power must be increased. A conversion loss of 3.6 dB is obtained with 10 dBm LO power.

4.2. The Single Sideband Mixer

A diode single sideband and image-reject mixer has been analyzed in [10]. This configuration is only suitable for low microwave frequency applications because at millimeter-wave frequencies the diodes must be matched and the ring configuration of the diodes does not allow matching the diodes easily. Furthermore air-bridges which cannot be designed with good performances at millimeter-wave frequencies are needed in place of path crossings.

In order to overcome theses problems, we propose to use a multiport in the designing of a single sideband mixer. For this aim, the configuration of the diode single sideband mixer analyzed in [10] has been modified as follows: (i) the monopolar excitation of LO signal has been replaced by a differential excitation, then, (ii) LO and RF ports have been interchanged. Figure 8 shows the modified diode ring single sideband mixer configuration. The analysis of LO and RF phase relationships (at the diodes) yields to a multi-port configuration.

Figure 9 shows the block-diagram of the multi-port single sideband and image-reject mixer. In order to provide a ground return, matching networks (MN) with short-circuit stubs are inserted at the inputs of all Schottky diodes.

Figure 10 shows the block-diagram of resulting five-port. Commercial full-wave software (High Frequency Structure Simulator (HFSS) version 10) of Ansoft Corporation was used for the design and simulation of the five-port. The circuit was designed for MHMIC technology on a 127 μm substrate with a relative permittivity of 9.9.

Figure 11 shows the lay-out of the five-port. The outer dimensions are, approximately 4.5 mm by 4.5 mm. In order to perform advanced system simulations, S parameters of five-port model were saved in a data file. This data file was then used as data input to large-signal harmonic balance simulation of mixer along with Spice model of Microwave Device Technology (MDT) flip-chip Schottky diodes.

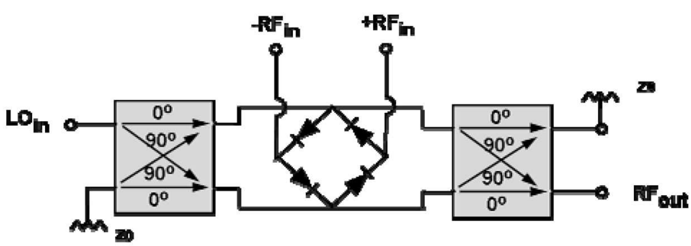

Figure 8. Modified diode ring single sideband mixer configuration.

Figure 12 shows the conversion gain value versus the LO voltage. The sideband suppression is maximal at 1.3 Volts of the LO voltage.

Figure 13 shows the spectrum at the output port. The level of the upper sideband is 40 dB under the level of the lower sideband signal. Furthermore, the spurious products of the form $f_{RF} \pm m f_{LO}$ are at least 40 dB under the low sideband signal level and their filtering with an extra filter is no more necessary.

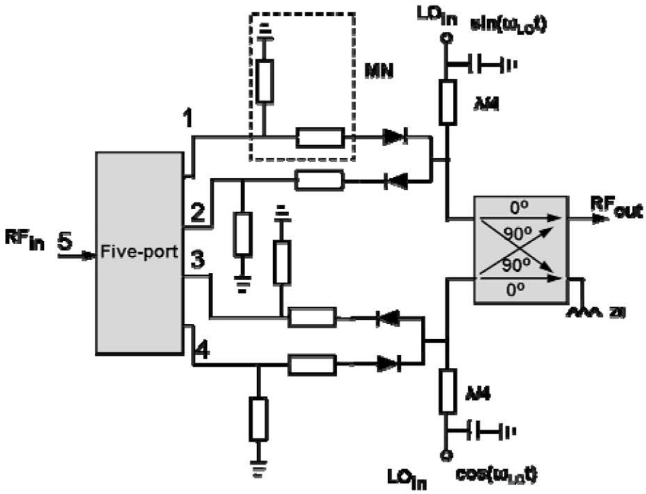

Figure 9. Block-diagram of the multi-port single sideband and image-reject mixer.

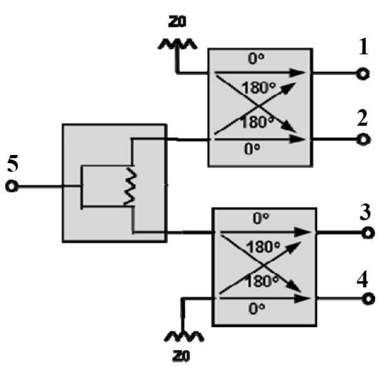

Figure 10. Block-diagram of resulting five-port.

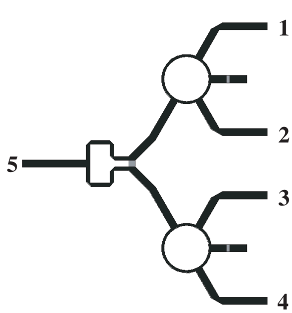

Figure 11. Lay-out of the five-port.

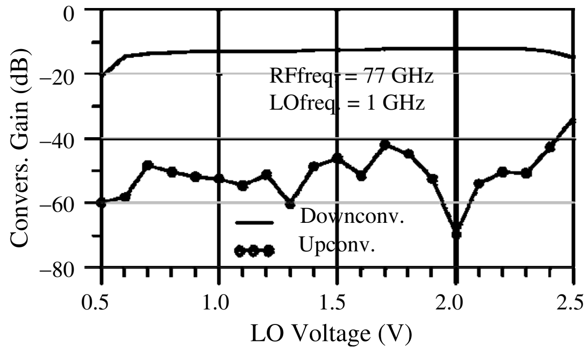

Figure 12. Conversion gain value versus the LO voltage.

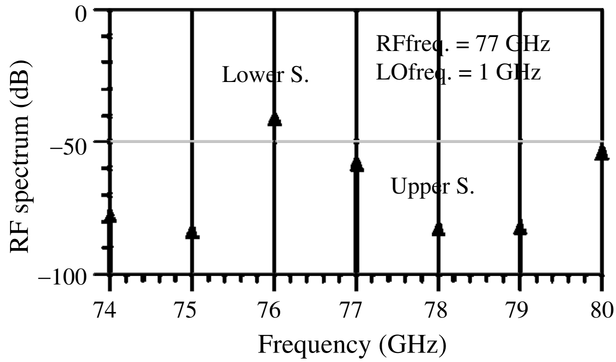

Figure 13. Spectrum at the output port of the single sideband mixer.

4.3. The Heterodyne Radar System

The heterodyne radar system according to Figure 1 has been simulated on envelope simulator of ADS. ADS system models of VCO, amplifiers, mixers, multipliers and filters have been used. In addition, models have been used for various path lengths taking into account: signal attenuation (radar equation [8]), propagation delay (between radar and target), and Doppler effect. The radar cross-section of target has been set to $1\,\mathrm{m}^2$. The simulated radar parameters have been for the transmitters as follows: $\Delta f = 200.0\,\mathrm{MHz}$, $f_0 = 77.0\,\mathrm{GHz}$, $f_{chirp} = 1/T_{chirp} = 83.333\,\mathrm{kHz}$, VCO power has been 10 dBm, PA gain has been set at 10 dB, the coupling factor of the directional couplers has been -10 dB. $f_{LO1} = f_{IF1} = 1\,\mathrm{GHz}$, $f_{LO2} = 950\,\mathrm{MHz}$ and $f_{IF2} = f_{LO1} - f_{LO2} = 40\,\mathrm{MHz}$. In addition, transmitter and receiver antenna gains are 20 dBi each.

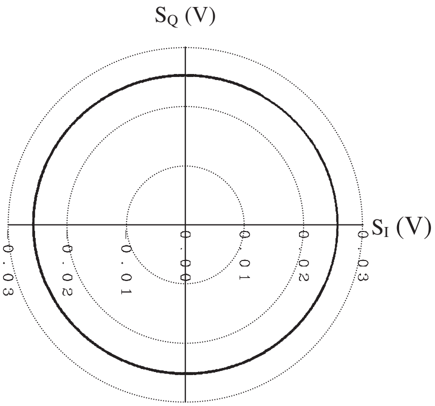

Figure 14. Complex beat signal for the speed of 20 m/s.

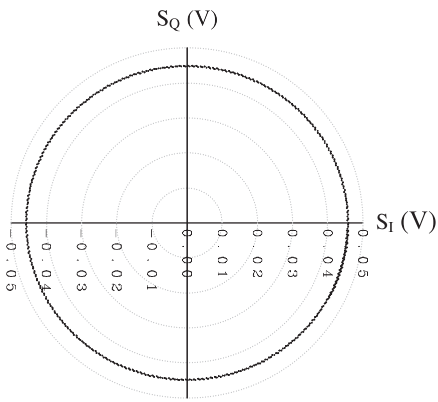

Figure 15. Complex beat signal for the range of 25 m.

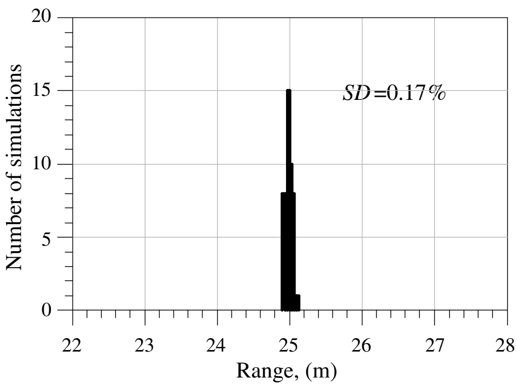

Figure 16. Histograms of the range for a target situated at 25 m from the radar.

Figures 14 and 15 show the complex beat signals in polar form for a relative target speed of 20 m/s and target range of 20 m, respectively. All these curves describe circles centered at origin of polar coordinates systems, showing that errors related to FM/AM noise, DC offset, amplitude and phase imbalance are negligible.

In order to do a statistical evaluation of range and speed obtained by simulations, a range resolution equal to about half of the 77 GHz wavelength and a speed resolution equal to half of the 77 GHz wavelength per second were chosen. Up to 50 simulations for each range and speed value were performed.

Range measurement histograms are also obtained, as show in Figures 16 and 17 (for distances of 25 m and 45 m, respectively). The standard deviations are 0.044 m (0.17%) and 0.097 m (0.21%) for the 25 m and 45 m, respectively.

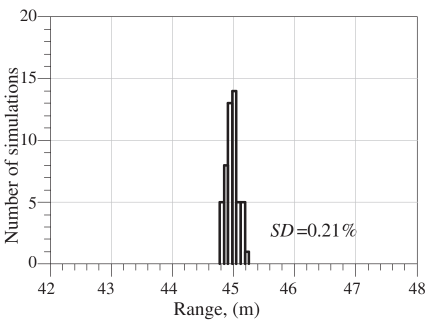

Figure 17. Histograms of the range for a target situated at 45 m from the radar.

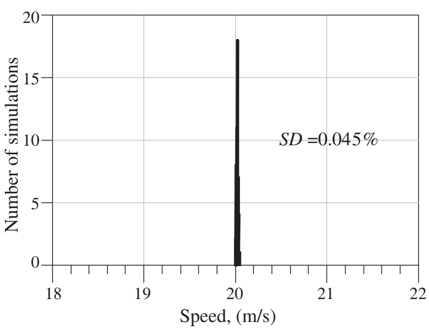

Figure 18. Histograms of the speed for a target speed of 20 m/s.

Table 1. Statistical simulation results comparison.

	SD for the speed of 20 m/s	SD for the range of 25 m	SD for the range of 45 m
Heterodyne FMCW radar (phase slope based range and speed measurement)	0.009 m/s (0.045%)	0.044 m (0.17%)	0.097% (0.21%)
Homodyne FMCW rader in [2] (phase slope based range and speed measurement)	0.003 m/s (0.015%)	0.07 m (0.28%)	0.283 m (0.6%)
Rader sensor in [5] (phase defference between two properly spaced frequencies based range measurement)	–	2.8%	1.8%

Figure 18 shows histograms for speed of 20 m/s. These histograms show negligible dispersions of range and speed values. In addition, Table 1 shows standard deviations (SD) in comparison with those obtained in [2] and [6].

Present results show that method of determining range by evaluating the slope of the instantaneous phase of beat signal is more accurate than past methods of determining range from phase difference between two CW signals properly spaced in frequency. Moreover, these results show that the heterodyne radar system with a single sideband mixer is more accurate is range measuring than the homodyne configuration from [2], due to the fact that, in heterodyne configurations FM-AM conversion noise is completely suppressed.

5. CONCLUSION

A Heterodyne FMCW collision-avoidance radar sensor, based on phase slope techniques for range and speed measuring, has been presented. Range and speed are determined by evaluating the slope of instantaneous phase of beat signal. Simulation results show that the heterodyne radar sensor with single sideband mixer offer good performances. Range measurements derived from slope of instantaneous phase of baseband beat signal is found to be more accurate than method of determining range from phase difference between two CW signals properly spaced in frequency. Moreover, the results show that the heterodyne radar system with a single sideband mixer is more accurate is range measuring than the homodyne configurations using the same phase slope method.

ACKNOWLEDGMENT

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