SIX-PORT FMCW COLLISION AVOIDANCE RADAR SENSOR CONFIGURATIONS

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ABSTRACT

This paper presents two six-port 77GHz FMCW collision avoidance radar sensor configurations. The first one uses a six-port double balanced mixer for direct frequency conversion and conventional frequency counting to determine the range and velocity of the target. The second one uses a six-port phase/frequency discriminator. Here velocity and range are determined in the same manner by evaluating the beat frequencies which are the slope of the instantaneous phase of the beat signals. Unlike other six-port collision avoidance sensors, measurement data processing techniques are developed to overcome the problems related to the DC current offset and amplitude imbalance, avoiding the need of six-port calibration. Computer simulation shows very good results for both configurations.

***Index Terms*—** Collision avoidance radar, six-port junction, FMCW.

1. INTRODUCTION

Commercial high volume application of microwave and millimeter-wave technology in the automobile area requires cost efficient and high reliable sensors. Under this aspect six-port homodyne receivers have been introduced in the design of radar sensors to replace the expensive heterodyne receivers and phase detectors [1] - [3]. In these sensors the six-port junction is used as a phase/frequency discriminator. The Doppler frequency is read by measuring the rate of phase variation of the beat signal and the range by measuring the phase difference between two properly spaced frequencies. These techniques of range measuring by measuring the phase difference between discrete frequencies limit the maximum measurable range, the so called maximum unambiguous range. For accurate measurement, the six-port must be calibrated and this is the major disadvantage. In [1] - [3], these radar sensors are considered as alternatives to FMCW sensors. In this paper we will not consider the techniques of phase/frequency discrimination as an alternative to the conventional FMCW techniques, but like frequency counting and FFT, as a method of measuring beat frequencies. Two configurations of six-port radar sensors will be presented. The first one uses a six-port double balanced mixer presented in [4]. The six-port does not suffer from amplitude and phase imbalance, and hence does not need any calibration, since there is only one output signal. The major advantage of this configuration is that it delivers a clean signal, without harmonics and some unwanted intermodulation products, whose frequency can be measured by a frequency counter. The second configuration utilizes a six-port phase/frequency discriminator to measure beat frequencies. Like Doppler frequency, they are read by measuring the rate of phase variation of the beat signals. The problem of ambiguous range measuring does not occur because we are dealing with the slope of the phase variation. Some techniques are developed to overcome the problems related to the DC current offset and amplitude imbalance, avoiding the need of six-port calibration.

2. RADAR SENSOR CONFIGURATIONS

Fig. 1a shows the block diagram of the radar with the six-port double-balanced mixer and Fig. 1b the block diagram of the radar sensor with the six-port phase/frequency discriminator. The millimeter-wave voltage controlled oscillator (VCO) is modulated by a triangle signal which gives a linear frequency modulated signal at its output. A portion of the emitted signal is coupled to the six-port mixers and serves as reference. The received signal is injected to the port 6 of the circuits after low noise amplification. The output of the mixer, in fig. 1a, is connected to a frequency counter which sends the counted frequencies to a DSP for the calculation of the relative speed and the range of the target. The outputs of the mixer, in Fig. 1b are A/D converted and a DSP is used to obtain the relative speed and the range of the target.

Fig. 2 shows the block diagram of the six-port junction for the double-balanced mixer [4]. It is composed of two 180^o and two 90^o hybrid couplers. The block diagram of the six-port junction for the I-Q mixer or phase/frequency discriminator is shown in Fig.3:

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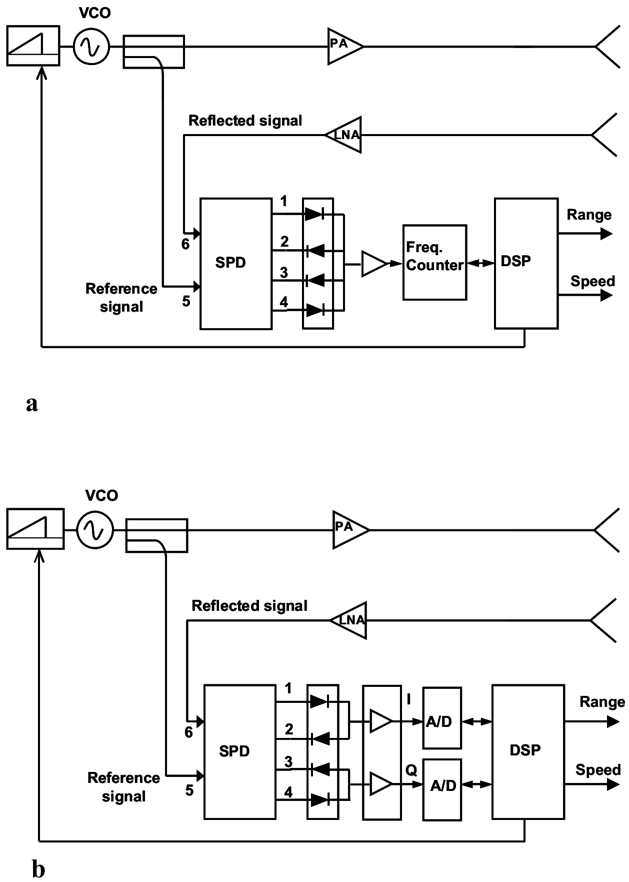

Figure 1. Radar configurations

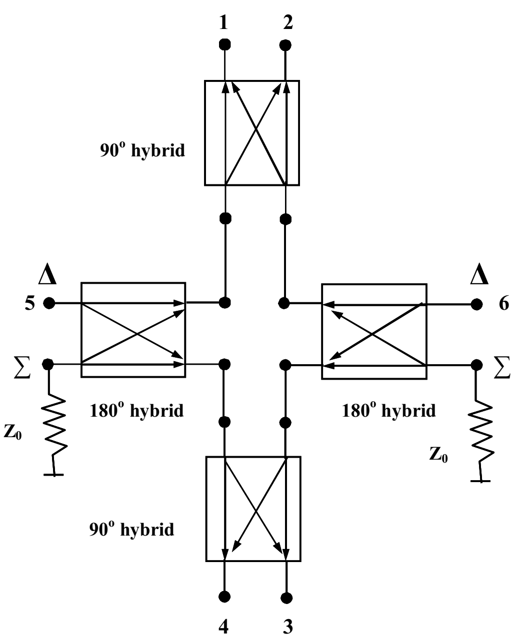

Figure 2. Six-port for the double-balanced mixer in Fig. 1a

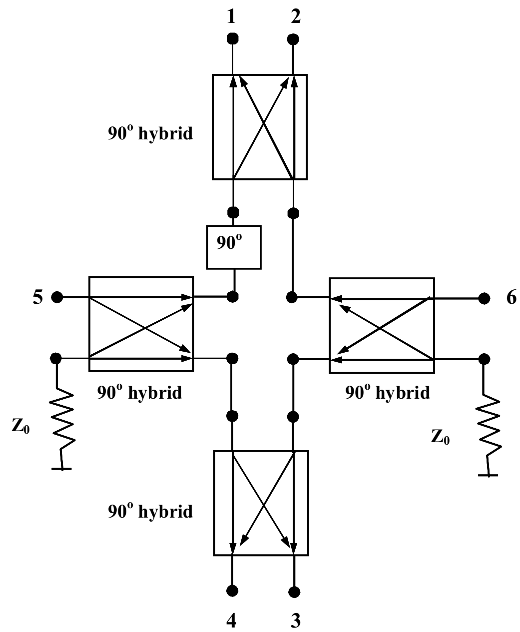

Figure 3. Six-port for the I-Q mixer in Fig. 1b

3. METHOD FOR DERTERMINING THE BEAT FREQUENCIES, RANGE AND SPEED

The radars transmit linear modulated continuous wave signals, which are positive modulated and negative modulated alternatively. Fig. 4a shows the linear FM of sending signals and receiving signals reflected by a moving target. In this figure the solid line represents the sending signals and the doted line the receiving signals. The frequency difference of sending and receiving signals is shown in Fig. 4b.

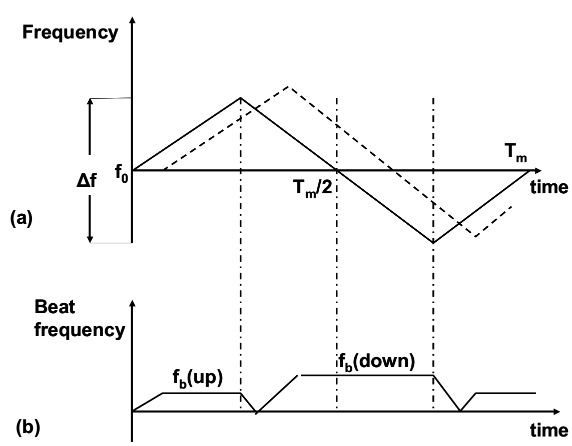

Figure 4. Frequency Modulated Continuous Wave

In [5] it is shown that the frequency shifts due to the range and the relative speed of the moving target are respectively:

$$f_R = \frac{1}{2}\left[f_b(up) + f_b(down)\right] \quad (1)$$

$$f_d = \frac{1}{2}\left[f_b(up) - f_b(down)\right] \quad (2)$$

Where $f_b(up)$ and $f_b(down)$ are respectively beat frequencies at the raising and falling chirp. A negative f_d means that the target is moving towards the radar and a positive f_d means that the target is moving away from the target. It is to be noted that the range can be calculated using f_R as follows [5]:

$$R = \frac{cf_R}{4 f_m \Delta f} \quad (3)$$

From f_d one can find the relative speed [5]:

$$v_r = \frac{cf_d}{2 f_0} \quad (4)$$

Where, c is the speed of light in vacuum. For the definition of f_0, Δf, and f_m , see Fig. 2a.
For the radar in Fig. 1a, $f_b(up)$ and $f_b(down)$ are read from the frequency counter and the range and the relative speed of the target are determined from (3) and (4).
For the radar in Fig.1b, the procedure for determining the range and the relative speed of the target is in the following:
The complex beat signal is:

$$\overline{S} = S_I + jS_Q \quad (5)$$

In the ideal case that power is equally splitted between the I- and Q- branches, that the phase difference between them is exactly 90^o, that there is no DC offset, and that good filtering is reached, S_I and S_Q are sinusoidal with the same amplitude:

$$S_I = A\cos(2\pi f_b t + \varphi_0) = A\cos\varphi \quad (6a)$$

$$S_Q = A\sin(2\pi f_b t + \varphi_0) = A\sin\varphi \quad (6b)$$

The instantaneous phase is then found to be:

$$\varphi = 2\pi f_b t + \varphi_0 = Arc\tan\frac{S_Q}{S_I} \quad (7)$$

And the beat frequency can be calculated from:

$$f_b = \frac{1}{2\pi}\frac{d\varphi}{dt} \quad (8)$$

In the reality, there is a DC offset and amplitude and phase imbalance. The phase imbalance can be kept at minimum by good designing, since at millimeter-wave frequencies phase shifters are realized from transmission line sections, their lengths have to be adjusted correctly. In this case S_I and S_Q become:

$$S_I = A_{0I} + A_I\cos(2\pi f_b t + \varphi_0) = A_{0I} + A_I\cos\varphi \quad (9)$$

$$S_Q = A_{0Q} + A_Q\cos(2\pi f_b t + \varphi_0) = A_{0Q} + A_Q\sin\varphi \quad (10)$$

At the first look, we can say that, to eliminate the DC offset, we have to differentiate the S_I and S_Q signals. But these signals are not very clean, they are noisy and there is an amount of intermodulation products in them, so that differentiation will yield to bad results. Since S_I and S_Q signals are known, we can calculate the DC offsets A_{0I} and A_{0Q} by evaluating their first moments:

$$A_{0I} = \frac{1}{T}\int^{+T} S_I dt \quad (11)$$

$$A_{0Q} = \frac{1}{T}\int^{+T} S_Q dt \quad (12)$$

The DC offsets must be removed from the S_I and S_Q signals. We obtain:

$$S_I' = S_I - A_{0I} = A_I\cos\varphi \quad (13)$$

$$S_Q' = S_Q - A_{0Q} = A_Q\sin\varphi \quad (14)$$

Before we can use equation (7) we must have both in phase and quadrature components of the same amplitude. We have to determine A_I and A_Q and normalize each component. The amplitudes are calculated from the second moments as follows:

$$A_I = \sqrt{2}\sqrt{\frac{1}{T}\int^{+T}(S_I - A_{I0})^2 dt} \quad (15)$$

$$A_Q = \sqrt{2}\sqrt{\frac{1}{T}\int^{+T}(S_Q - A_{Q0})^2 dt} \quad (16)$$

The period T must be a multiple of the beat signal periods. Since the periods are not known, T is fixed so that it covers several cycles to minimize the errors. Now that all the coefficients in (9) and (10) are calculated the instantaneous phase is found to be:

$$\varphi = 2\pi f_b t + \varphi_0 = Arc\tan\left(\frac{\dfrac{S_Q - A_{0Q}}{A_Q}}{\dfrac{S_I - A_{0I}}{A_I}}\right) \quad (17)$$

This procedure is summarized as follows:

1. Compute A_{0I}, A_{0Q}, A_I and A_Q from (11) to (16)

2. Compute $\varphi(up)$ and $\varphi(down)$ from (17) and then

3. $f_b(up)$ and $f_b(down)$ from (8)

4. Compute f_R and f_d from (1) and (2) and

5. The range and the relative speed from (3) and (4)

4. SIMULATION RESULTS

Simulations are performed using Advanced Design System software of Agilent Technologies (ADS). The electromagnetic simulator Momentum of ADS is used for the six-port circuits. The six-port circuits are designed in MHMIC technology on a 125 µm substrate having a relative permittivity of 9.9. The parameters according to Fig. 4 are as follows: $\Delta f = 200.0$ *MHz*, $f_0 = 77.0$ *GHz*, $f_m = 41.666$ *kHz.*
Fig. 5a shows the beat signal corresponding to the rising chirp and Fig.5b the beat signal corresponding to falling chirp for the radar sensor with the six-port double-balanced mixer (Fig. 1a) for the following parameters: *Relative speed=50 m/s, range =20 m.*

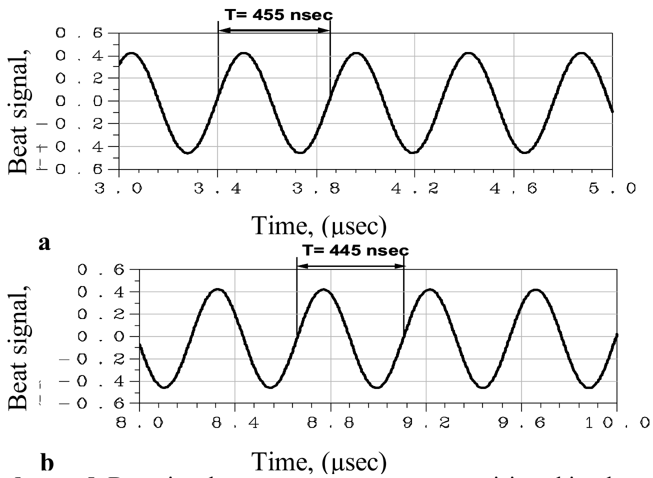

Figure 5. Beat signal wave forms a. signal at the rising chirp, b. Signal at falling chirp

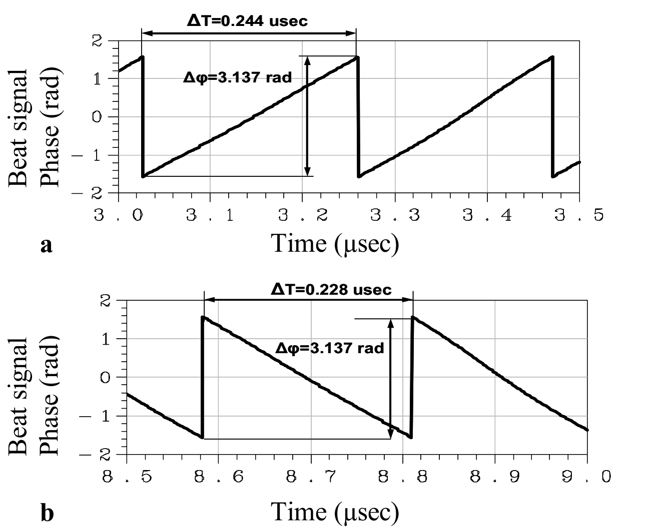

Figure 6. Beat signal phase a. phase at the rising chirp, b. phase at falling chirp

Fig. 6a shows the instantaneous phase of the beat signal corresponding to the rising chirp obtained from (17), and Fig. 6b the instantaneous phase of the beat signal corresponding to falling chirp obtained from (17) for the radar sensor with the six-port phase/frequency (fig. 1b) for the following parameters: *Relative speed=50 m/s, range =20 m.*

Table 1 shows relative speeds and ranges obtained from the simulation. According to this table, the maximum deviation is, for de relative speed 6.7%, and for the range 2.7%. The results are therefore acceptable. Noticeably more accurate results, for the relative speed, could be obtained by allocating a time interval at which no modulation is performed. The beat frequency at this time interval would be composed only of Doppler frequency which can be measured with accuracy.

Table 1. Results

Parameters: Speed [m/s] / range [m]	Radar fig. 1a Speed [m/s]/ range [m] simulated	Radar fig.1b Speed [m/s] / range [m] simulated
50/10	48.017 / 9.97	51.8 / 10.54
50/15	51.33 / 14.99	52.5 / 14.98
50/20	48.10/ 20.01	53.37/ 19.46

6. CONCLUSIONS

Two six-port FMCW radar configurations are presented. Both offer acceptable results. The proposed method of correcting of DC offset and amplitude imbalance for the six-port FMCW with phase/ frequency discriminator avoids the use of the complicated calibration method, which is the major disadvantage of the six-port radar sensors.

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