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A METHOD OF MULTI-CHANNEL DATA RECEPTION AND PROCESSING NOISE FOR SHORT DISTANCE DETECTION DOPPLER RADAR

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Abstract

In short distance detection Doppler radar signal processing, the communication and processing problems of various measurement and control signals such as mass data which are high speed A/D converted intermediate frequency signals of transmission and reception terminals and azimuth angle, altitude are suggested. In this paper we suggest one method for time sharing data communication and noise processing by determining threshold by using the transmission line of limited bits in connection of analog circuits consisting of electromagnetic transmission and reception antenna and various control, measuring circuits and FPGA+DSP type digital signal processing circuits and showed experiment results.

Keywords:
Doppler, DSP+FPGA, Digital Signal Processing System

1. INTRODUCTION

In digital area increasing operating ability, memory capacity of FPGA and multi-DSP, digital signal processing system of radar were developed rapidly [1] [15]. Those are composed of parallel processing, high processing density, efficient modularized design and so on [2] [3]. Parallel processing of mass data in parallel processing is main problem, high speed data communication between modules are discussed mainly. In particular, the instantaneous frequency accurate estimation is mainly discussed. In general, the performance of parallel processing system is determined by processing speed of every processing unit and communication speed of them. The data communication of multi-DSP parallel system is supported by peripheral interface of DSP. To improve processing speed multi-DSP such as TMS320C6678 are used and main part of processing such as FFT, IFFT, Multiplexer are implemented by FPGA [17] [18]. Those are typically HPI, EMIF, McBSP in TMS serial of TI corporation, Serial RapidIO of C6455 serial, LinkPort in TigerSHARC serial of ADI corporation. [1], [4] - [6], [9] - [11]. Also we can compose multi-DSP parallel processing system by special communication IC without using of peripheral interface [6]. These are divided into direct link mode, bus direct link mode and indirect link mode according to link mode of DSP.

In radar signal processing the main problem is also to improve the accuracy of tracking target in noise environment such as reflective noise by various object and threshold processing are discussed as well as signal processing by probability characteristics [7], [12] - [14], [17] - [20].

Here many problems based on wavelet are introduced. Instantaneous frequency estimation of various signal whose phase is different, adaptive linear prediction, instantaneous frequency estimation with filtering are introduced and its precision and robustness are discussed. In previous works referred determining threshold problem depending on SNR and these formula doesn't work well in some cases such as strong reflect noise [8] [9] [14]. For this problem signal processing methods in MIMO system are suggested.

In this paper we suggest one method for time sharing data transmitting and noise processing by determining threshold in connection of analog circuit composed with several transmission and reception antenna and control, measuring circuits and FPGA+DSP type system in short distance Doppler radar system.

In this paper we discuss a short distance Doppler radar system, which detect the position and move velocity of short distance object. It emits and receives electromagnetic wave in several transmitting and reception antenna and process intermediate frequency signal, azimuth angle and altitude angle and transmits FPGA+DSP high digital signal processing system.

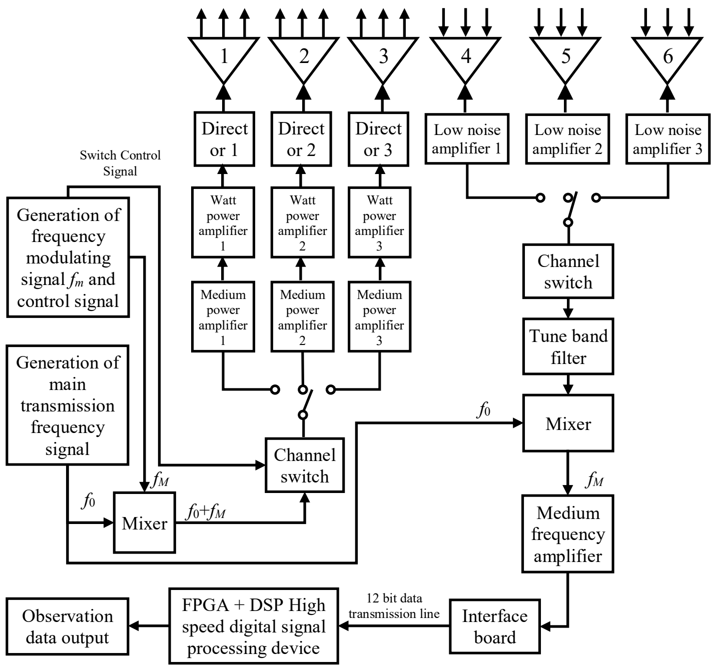

Fig.1. Short Distance Detection Doppler Radar System

As shown in Fig.1 ultra-high frequency signal in GHz emitted from transmit antenna has $(f_0+f_m)_i$. Here f_0 is constant, main frequency and f_m is sawtooth type linear modularization frequency. Those are reflected from object and received to reception antenna with frequency of $(f_0+f_{md}+f_d)_i$ (i=1,2,..,3), where f_m means time delay of f_m corresponding to distance from here to object, f_d means Doppler frequency, which shows movement of object. This signal is converted to intermediate frequency $(f_{md}+f_d)_i$ in mixer.

The time delay character of f_{md} in contradistinction to f_m, Doppler frequency f_d, azimuth angle, altitude angle is converted to digital signal in interface circuit board and transmitted to FPGA+DSP type high digital signal processing device. We can get the information of distance and movement of objects.

Also we proposed one method for reasonable time sharing transmitting in mass data high speed communication system by 12bit data transmit line and processing noise by threshold determination.

In section 2, we proposed one method for time sharing data communication. In section 3 we discussed one method for noise processing method by determining threshold.

2. TIME SHARING TRANSMITTING OF DETECTION SIGNALS.

The reception signal $r(t)$ of $(f_{md}+f_d)_i$ in MHz and signal information such as azimuth angle and altitude angle are converted to 12bit digital signal in interface circuit and transmitted to FPGA+DSP device.

Then reception signal $r(t)$ is converted to 60MSPS 12bit digital signal by AD9237BCP-65, other signals are converted to 12bit digital signal below 125KSPS and transmitted by using 8-channel ADC AD7938-6 and logic level conversion circuit. These are shown in Fig.2.

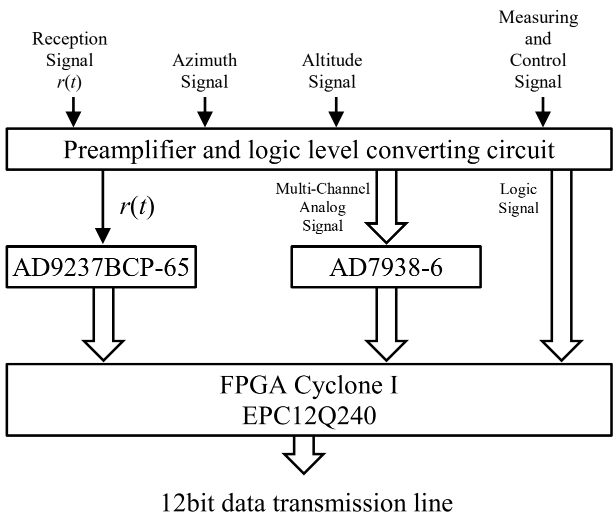

Fig.2. Signal Reception Converting Unit (interface board)

AD9237BCP-65 and AD7938-6 are controlled by Cyclone I EPC12Q240. Before converting to digital signals analog signals are inputted to these ADC devices through preamplifier and logic control signals are directly inputted to FPGA through level conversion circuit. We can write FPGA Verilog code to setup the reasonable data communication of high speed 12bit_60MSPS and low speed 125KSPS in FPGA.

In Fig.3 AD9237BCP-65 and AD7938-5 control conversion synchronization control signal and control word register control. We assign 12×512 FIFO1 and AD7938-6 logic signal data reception for eight channel 12bit digital data emitted from AD7938-6 and logic control signal.

We adjust writing control signal in fifo1 to the sum of frequency of AD7938-6 transmission synchronization and logic control synchronization and restrict 12-bit output signal number of AD7938-6 and logic control signal number to constant N_1, N_2 ($N_1 + N_2 = N$, $N < 512$) in AD7938-6 logic signal data reception. By the control of time sharing controller the FIFO1 outputs data of N number according to AD9237BCP-65 conversion synchronization signal.

In time sharing controller the write conversion control of AD9237BCP-65 controller and FIFO1 output signal are performed in a reasonable period about FIFO2 according to N settled as to receive the efficient information of the output signals of AD7938-6 and logic signals without losses, meanwhile according to control of short distance searching controller consuming most of the time to treat $r(t)$.

In result the 12bit FIFO2 output digital signals were transmitted to FPGA+DSP device.

Let the data reception time in multi-channel AD7938-6 to T_d, conversion rate to F_s (below 125KSPS), conversion rate of AD9237BCP-65 to F_f. Then data reception time is:

$$T_f = T_d \cdot F_s / F_f$$

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For example, let F_s = 125KSPS, F_f = 60MSPS, N = 500, we get T_d = 2ms. From this method it was reduced to 0.0083ms so that we can perform multi-channel data reception time sharing without information loss of detection object.

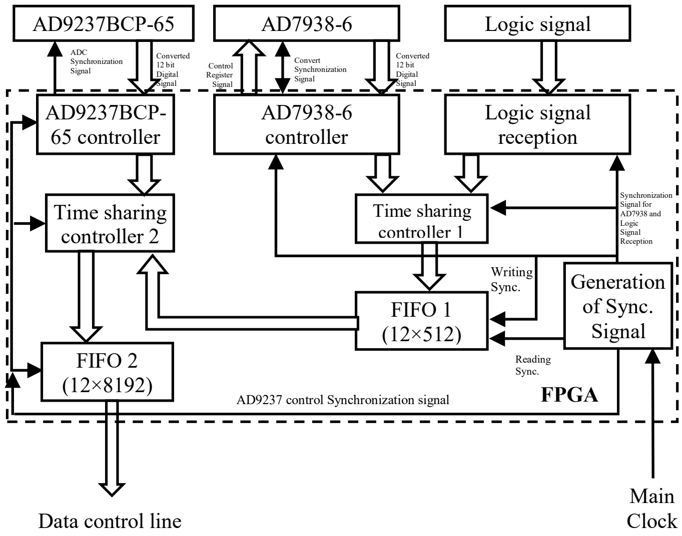

Fig.3. FPGA Verilog Code Configuration

3. NOISE PROCESSING BY DETERMINING THRESHOLD

3.1 DETERMINING THRESHOLD

It is important that we must estimate the exact frequency in short distance Doppler radar because signal $r(t)$ has many noise effect.

We can use Hilbert transformation to estimate the frequency of digital signal series frame $S_r = (s_{r1}, s_{r2}, ..., s_{rN})$ to which reception signal $r(t)$ are high speed A/D converted.

$$H = Hilbert(S) \tag{1}$$

In $H = (h_1, h_2, ..., h_N)$, for each i

$$h_i = H_i e^{j\phi_i} \tag{2}$$

Let the time interval of inputted discreet signal be Δt, then instantaneous angular velocity at i can be shown as below,

$$\omega_i = \frac{\phi_i - \phi_{i-1}}{\Delta t} \tag{3}$$

Now noised angular velocity can be get as below,

$$\omega_{ri} = \omega_i + n_i \tag{4}$$

where, noise n_i is very irregular so that we cannot expect exact possibility distribution.

We suggest one method, which set threshold of noise out range of variation range and estimate frequency value by square least method, where ratio of threshold value is already set. Considering change of noise is large, frame is divided to several segment. In each interval we decide threshold and process noise.

Let reception signal be $S_r = (s_{r1}, s_{r2}, ..., s_{rN})$, let the length of noise interval be dl, now we decide the signal series of partial interval as below,

$$s_{rk+i}, i = 1,...,dl,\ k = 1,...,K,\ K = \frac{N}{dl}$$

The instantaneous angular velocity corresponding upper signal series is below,

$$\omega_{rk+i},\ \ i = 1,...,dl,\ \ k = 1,...,K,\ \ K = \frac{N}{dl} \tag{5}$$

The threshold corresponding upper angular is decided as follows, where the ratio of noise out of variation range is already set so that the number of ω_{ri} $2t$ in dl interval is already set. We can estimate signal distribution $N_\omega(j)$, $j = 1,\ldots,M$ in phase change range $[-2\pi, 2\pi]$ as below. We divide $[-2\pi, 2\pi]$ into $M = \left[\frac{4\pi}{\Delta\omega}\right]$ intervals by $\Delta\omega$ and judge where ω_{ri} in Eq.(4) is put. If put in $[(j-1)\Delta\omega, j\Delta\omega)$, find $N_\omega(j)$, $j = 1,\ldots,M$ in interval dl by adding one for $N_\omega(j)$. Then we get angular velocity distribution series $N_\omega(j)$, $j = 1,\ldots,M$. The value out of change range is put $j = 0,1,\ldots, j = M, M-1,\ldots$ i.e. $[0,\Delta\omega)$, $[\Delta\omega, 2\Delta\omega),\ldots[(M-1)\Delta\omega, M\Delta\omega)$, $[(M-2)\Delta\omega,(M-1)\Delta\omega),\ldots$ so both sides of $N_\omega(j)$. From this we can set j_1, j_2 as $\sum_{j=1}^{j1} N_\omega(j) = t$, $\sum_{j=j2}^{M} N_\omega(j) = t$, lower and upper threshold of ω_{ri} is set as below,

$$th_1 = j_1 \times \Delta\omega\ ,$$

$$th_2 = j_2 \times \Delta\omega$$

For these series ω_{ri}, we get value in $[th_1, th_2]$ and remove ω_{ri} out of this range. We calculate average of value in $[th_1, th_2]$ and replace as removed ω_{ri}. We can repeat these steps in every intervals in Eq.(5).

3.2 COMPARISON EVALUATIONS BY LEAST SQUARE METHOD

The least square assumption for the estimating of following $\Omega_r = (\omega_{r1}, \omega_{r2}, ..., \omega_{rN})$ noise treated after setting the threshold is carried by splitting as the following part interval series.

$$\omega_{ri}, i = 1,...,bl, p = 1,...,P, P = \frac{N}{bl} \tag{6}$$

We use the least square method based on dealing with frequency assumption problem in the linear frequency modulated signal and bl, the length of partial interval in Eq.(6) is settled by considering the entire frame length and linear frequency modulating character.

Evaluating reference equation for the least square assumption is as follows:

$$J = \sum_{i=1}^{bl} [\omega_{ri} - (c_1 i + c_0)]^2 : \min$$

In other words, the assumption to partial series of Eq.(6) noise treated by setting the threshold is carried out by the method obtaining the coefficient c_1, c_0, where upper square evaluating reference equation is minimized.

It is carried out as the method repeats this processing till the end, in which method we obtain upper coefficient c_1, c_0 in the first partial interval of the length bl and basing on the obtained c_0 get c_1 of next interval in next interval where the length from middle position of this interval is also bl.

For the least square assumption the interpolation coefficient matrix is calculate as follows:

$$sq_2 = \sum_{i=1}^{bl} i^2, sq_1 = \sum_{i=1}^{bl} i$$

$$SC = \begin{bmatrix} sq_2 & sq_1 \\ sq_1 & bl \end{bmatrix}^{-1}$$

First of all, the least square assumption in initial partial interval $s_1, s_2, ..., s_{bl}$ is considered as follows:

$$\begin{pmatrix} c_1 \\ c_0 \end{pmatrix} = SC \cdot \begin{pmatrix} \sum_{i=1}^{bl} i \cdot \omega_{ri} \\ \sum_{i=1}^{bl} \omega_{ri} \end{pmatrix} \tag{7}$$

About the upper coefficient c_1, c_0 the signal series in the first partial interval $s_1, s_2, ..., s_{bl}$ can be expressed by following linear approximated equation.

$$\hat{\omega}_i = c_1 \cdot i + c_0 \quad i = 1,...\left[\frac{bl}{2}\right] \tag{8}$$

Next partial interval is:

$$\omega_{ri}, i = \left[\frac{bl}{2}\right] + 1, ..., \left[\frac{bl}{2}\right] + bl \tag{9}$$

The least square assumption in next partial interval the constant c_1 is $c_0^1 = c_1 \cdot \left[\frac{bl}{2}\right] + c_0$ (here c_1, c_0 is the values of Eq.(8)) and first order coefficient c_1^1 is calculated as follows:

$$c_1^1 = \frac{\left(\sum_{i=1}^{bl} i\omega_{r\left[\frac{bl}{2}\right]+i} - c_0^1 sq_1\right)}{sq_2} \tag{10}$$

Based on upper coefficients the least square assumption in signal series partial interval $\omega_{ri}, i = \left[\frac{bl}{2}\right] + 1, \ldots, bl$ is obtained as follows:

$$\hat{\omega}_{i+\left[\frac{bl}{2}\right]} = c_1^1 \cdot i + c_0^1 \; i = 1, \ldots \left[\frac{bl}{2}\right] \quad (11)$$

Repeating upper process we can obtain the least square assumption Eq. i in signal series partial interval

$$\omega_{ri}, i = m \cdot \left[\frac{bl}{2}\right] + 1, \ldots, (m+1)\left[\frac{bl}{2}\right] \quad i = 1, \ldots \left[\frac{bl}{2}\right] \quad (12)$$

Thus, the instant frequency assumption series about frame signal series $\Omega_r = (\omega_{r1}, \omega_{r2}, \ldots, \omega_{rN})$ in entire interval can be obtained.

$$\hat{\Omega} = (\hat{\omega}_1, \hat{\omega}_2, \ldots, \hat{\omega}_N) \quad (13)$$

We compare the noise treating result under the condition where the length of signal frame series is $N = 2\times10^6$ and is in following strong noise environment with the result applying the least square method mentioned in preceding research.

The waveform of reception signal $S_r = (s_{r1}, s_{r2}, \ldots, s_{rN})$ is equal as Fig.4.

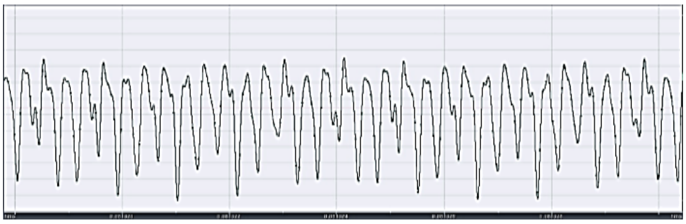

Fig.4. Received signal waveform

The result of Eq.(3) using the upper signal and Hilbert conversion, $\Omega_r = (\omega_{r1}, \omega_{r2}, \ldots, \omega_{rN})$ is equal as Fig.5.

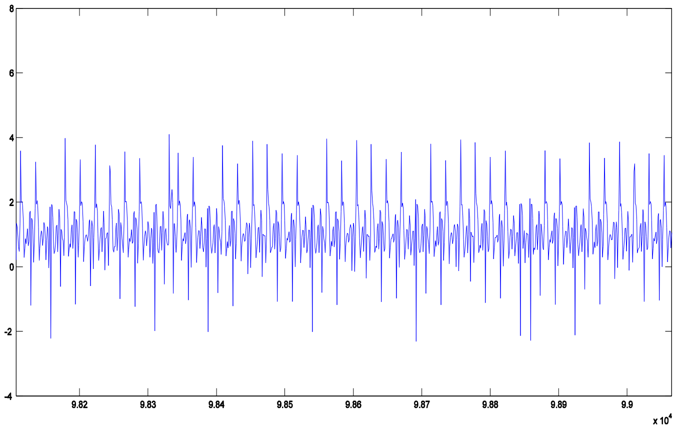

Fig.5. Angular velocity curve of noise-mixed signal

Applying the method mentioned in preceding research and 2 and 3 in this, the following result curving line is obtained. Here t = 10, dl = 100, bl = 2000.

As shown in Fig.6 we can reduce the noise assumption error to 9% by the least square method using threshold value determine method proposed in this paper.

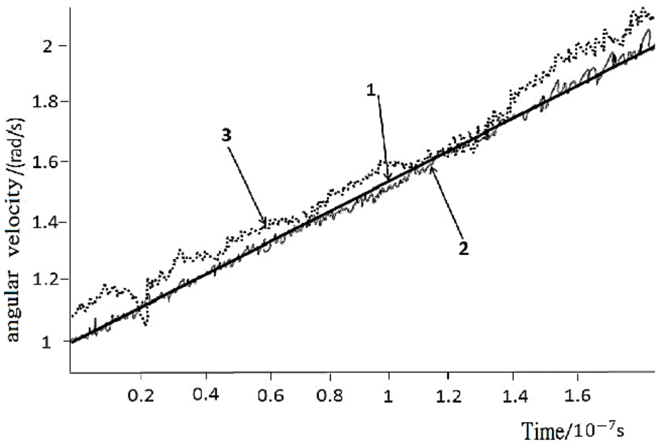

Fig.6. Result analysis graph (1: exact model, 2: suggested method, 3: square least method)

4. CONCLUSION

In Doppler searching system the integrated process of high frequency signals in medium frequency, various low frequency signals and logic control signals is carried out at the same time. Considering such condition and using the time sharing method proposed in this paper in terms of transmission and processing we can achieve the signal processing performance without the losses of the efficient information under the condition of the limited bit resolution. Also we can decrease the device manufacturing cost and improve the convenience in maintaining control. And we can improve the searching performance considerably by treating the noises is not able to predict the exact probability distribution under the various noise conditions with the proposed threshold process method. It is considered that such method can be introduced several parts such as digital radar signal processing system, sonar signal processing and real time image signal processing.

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