Radar Pulse Compression Techniques

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Abstract: - The aim of this paper is to provide an introduction to principle behind the pulse compression radar. Pulse compression is an important signal processing technique used in radar system to reduce the peak power of a radar pulse by increasing the length of pulse, without sacrificing the range resolution associated with a shorter pulse. In this paper the pulse compression technique will be described and compared.

Key-Words: - Pulse compression, radar, polyphase code, Doppler shift, sidelobe, autocorrelation function

1 Introduction

For radar the peak to mean ratio of the transmitted power is comparatively low as is common for phased array radar using solid-state transmitting modules. Therefore, in order to use the available mean radar power more efficiently long pulses should be transmitted in practically all modes of operation and for most acquisition ranges. To achieve required range resolution the transmitted pulses must be modulated with the time compression on receive.

The modulated transmitted pulses should satisfy the following requirements:

- Specified time resolution of the compressed pulse.
- Small signal to noise ratio losses for the main lobe of the compressed pulse over the range of likely target Doppler shifts.
- Minimum level of the range side lobes of the compressed signal over the range of possible Doppler shifts.
- The employed pulse modulation should allow for a flexible modification of the transmitted pulse length while preserving the required time resolution in order to adapt the radar power resources to target detection and tracking requirements.
- As far as possible, the adopted modulation technique should be easily realizable in hardware and computationally efficient pulse compression algorithms should be employed.

For multifunction radar, a comprehensive set of different pulse modulated signals should be developed at research and development stage which are best suited to target detection and acquisition for diverse tasks and environments.

For long range target detection and in the absence of a significant clutter the radiated pulses should be comparatively long to more efficiently use the available radar mean power. High range resolution is not needed in this mode. In the filter synthesis discussed below the time resolution equal to 150 m was used for this mode. The maximum level of range side lobes for the long range detection mode can be –35 dB or less.

The detection at long range in intensive extended clutter requires the radiation of coherent pulse Doppler sequences. The initial target detection can be accomplished by unmodulated pulse Doppler sequences with a high pulse repetition frequency but for the following target acquisition, resolving of range and Doppler ambiguities, and estimation of target coordinates modulated pulses of a small length must be used.

The tracking at small and medium ranges at a large elevation or in the absence of an extended clutter requires the pulse signals ensuring a range resolution about 30 m. Similar signals will be required for an MTI canceller operation when a ground or precipitation clutter is present.

2 Pulse modulation techniques

Since the amplitude of the radiated signal should be constant to achieve high efficiency of power amplifiers of the transmission modules and the signal processing is digital, discrete phase modulation methods must be applied. The phase coded radiated waveform consists of a number of subpulses the subpulse length determining the time resolution, and the code length defines the total length of the transmitted pulse and affects the range side lobe level.

Many phase code techniques can be found in the literature though there is obviously no universal solution satisfying the multiplicity of requirements. A formal method allowing design a phase coded waveform meeting a specified set of requirements is as yet not devised. Therefore, the codes appropriate for the task are searched for using modeling and specifically designed software [3].

A most readily implemented phase modulation technique is a binary phase coding (0^0, 180^0). The well-known classes of binary coding are a Barker code and pseudorandom sequences of maximum length. Unfortunately, the behavior of these codes for Doppler-shifted signals is unsatisfactory and the number of available codes is limited wherefore they are not well suited for generating the waveforms with flexibly adaptable lengths.

The polyphase codes are commonly based on previously developed analogue modulation techniques of linear and nonlinear frequency modulation.

For instance, a well-known Frank code [1] is derived as the samples at Nyquist rate from a stepped linear frequency modulated waveform. A Frank code compression filter is computationally efficient but its peak to maximum range time-sidelobe ratio deteriorates if the signal is band-limited in the preceding receiver stages. Moreover, the Frank waveform is not tolerant to Doppler shifts which cause considerable distortions of the compressed main lobe and a significant loss of signal energy.

As an attempt to alleviate the drawbacks of Frank coding a class of polyphase codes was proposed [2] known as P1-P4 codes. Codes P1 and P2 of this class are tolerant to receiver bandwidth limitations. However, they are based on a stepped approximation to a linear chirp just as the Frank code and share with it the same degree of main lobe distortions and an unacceptable loss of signal energy if a significant Doppler shift is present in the signal.

Codes P3 and P4 on the other hand were derived as the samples from a linear frequency modulated (LFM) signal and inherited its tolerance to Doppler shifts. The maximum of the compressed signal is shifted along the time axis with the Doppler shift as is the case with an LFM signal (which is due to the same causes) but the compressed waveform is distorted only slightly and there is no significant loss of signal power. In addition, a P4 code is tolerant to receiver bandwidth limitations. Therefore, P4 codes were selected for the researched radar along with some codes based on nonlinear frequency modulation.

A polyphase code P4 of length $N = T / \tau$, where T is the length of the radiated pulse, is generated by sampling the LFM phase at time intervals equal to subpulse length τ:

$$\varphi_n = \frac{\pi}{N} n^2 - \pi n \bmod 2\pi \text{ for } 0 \le n < N . \quad (1)$$

The code length N and therefore the length of the radiated pulse can be chosen arbitrarily while preserving a specified time resolution. In this way an important feature can be realized, that is a flexible adaptation of the radar power resource.

The range side lobe level for a polyphase coded waveform depends on the code length N and diminishes as the code length is increased. The analysis has shown that for the code lengths required in the researched radar the resulting side lobe levels usually are too high if a matched compression filter is employed.

The compression filtering with a weighting window will be used to reduce the range side lobes. The weighted filtering reduces the amount of high-frequency components in the signal spectrum and thereby decreases the side lobes in time domain. At the same time the filter becomes mismatched and some loss of the signal to noise ratio is inevitable.

In some cases, e.g. for detection of a target at small and medium ranges, the resulting loss causes no significant deterioration in radar performance and is acceptable. At long ranges, however, the additional signal to noise ratio losses may lead to an increased dwell time and unfavorably influence the radar energy balance.

Employment of the polyphase codes (based on nonlinear frequency modulation (NLFM)) in radar systems was increased recently. The NLFM methods were proposed early in radar history but their use was limited due to problems with practical implementation of these techniques with analogue technologies.The situation has obviously changed with the advent of digital processing.

A complex phase modulated signal can be represented as:

$$u(t) = a(t)\exp[j\theta(t)] \quad (2)$$

where $\theta(t)$ is a law of the phase dependence on time that can be practically arbitrary if a digital synthesis method is used.

The problem of the NLFM filter design is to find a law $\theta(t)$ that meets the filter performance requirements including an acceptable level of range side lobes. The general approach to its solution is to generate a signal waveform which has a spectrum where the high-frequency components are reduced as compared with an LFM signal which has an approximately rectangular spectrum. By analogy with a linear frequency modulated filter using a weighting window such a modification of the signal spectrum should result in reduced range side lobes but since the filter is now matched no signal to noise ratio loss will occur.

The NLFM filters for the researched radar were designed by employing a technique based on an approach discussed in [4]. The starting point is the equation relating the representations of the signal in the frequency and time domains. This equation can be derived from (3.2) using Fourier transform and stationary phase integration:

$$\int_{-\infty}^{t} a^2(\tau)d\tau = \frac{1}{2\pi}\int_{-\infty}^{\omega(t)} |U(\Omega)|^2 d\Omega . \qquad (3)$$

To design a signal a desired signal spectrum $|U(\Omega)|^2$ should be specified and by solving the equation (1.3) the frequency as a function of time $\omega(t)$ corresponding to this spectrum should be derived. By integrating $\omega(t)$ phase as a function of time $\theta(t)$ can be obtained. The samples from $\theta(t)$ give a discrete phase code with the desired signal spectrum.

Evidently, if a reduced range side lobe level is desired the signal spectrum $|U(\Omega)|^2$ should be some bell-shaped function. There exists no rigorous method that could be used to derive the signal spectrum $|U(\Omega)|^2$ which would result in a desired compressed waveform. The selection of a suitable spectrum $|U(\Omega)|^2$, however, can be facilitated by taking into account the results obtained with weighting windows in the LFM filters.

It can be expected that an NLFM spectrum $|U(\Omega)|^2$ of a shape similar to the frequency response of a weighting window filter having the desired performance would give rise to an NLFM phase code with a similar performance. It turns out that it is often the case.

Of course, an NLFM filter obtained in this way will be matched to the signal and therefore will show no signal to noise ratio loss as opposed to its weighting window prototype.

One sample per a subpulse is enough for a normal operation of a polyphase code filter, however, in some cases it proves to be beneficent to increase the number of samples available per a subpulse since it may improve the filter main lobe response.

An example of the signal spectrum $|U(\Omega)|^2$ selected for the design which in this case is a Cosine-on-pedestal function with the parameters Cosine power = 3.000, Pedestal = 0.040000 and the frequency response of the synthesized filter is given in Fig.1.

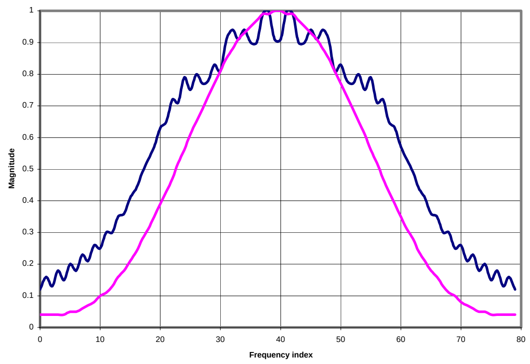

Fig.1: The initial spectrum and the frequency response of the synthesized filter

The frequency dependence on time $\omega(t)$ for this code obtained by solution of equation (1.3) and representing the resulting law of nonlinear frequency modulation is shown in Fig. 2.

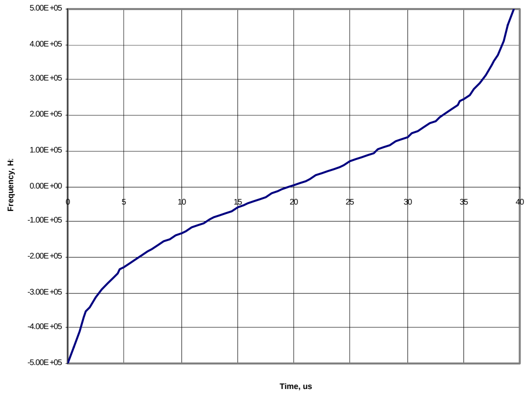

Fig. 2: Modulation frequency as a function of time

The steeper frequency parts of the function $\omega(t)$ at the beginning and at the end exist only for a short time and therefore introduce less power within the whole spectrum. This leads to the desired bell shape of the spectrum $|U(\Omega)|^2$.

3 The pulse signals used by the radar

A general inventory of the signals which will be required in the radar can be set up using the radar operational requirements and taking into account the signal compression techniques discussed above.Table 1 lists the main categories of signals intended for use in radar.

Table 1

Signal purpose	Range	Range resolution	Filter parameters	Main requirements
Search and tracking in absence of clutter and for use with MTI	Short and medium range	0.2 µs	N =100, T=20 µs N =250, T=50 µs	Low side lobes, small SNR losses
Search and tracking in absence of clutter	Long range	1 µs	N =40, T=40 µs N =100, T=100 us	Small SNR losses, low side lobes
Tracking in intensive clutter, Doppler filtering	Long range	1 µs	N =16, T=15 µs	Small SNR losses, low side lobes
Search in intensive clutter	Long range	3 – 5 µs	No compression filter	

The nonlinear frequency modulated signals were designed using the suggestions on the selection of the initial signal spectrum proposed in [5].

The column "Filter parameters" in the Table lists the code lengths N and the lengths of the transmitted signals T. In practice the signals of differing length can be used by radar to achieve an optimal allocation of power resources depending on an environment scenario.

The problem of the signal losses in a compression filter requires some explanation. The signal losses arise when a filter is mismatched with the received signal, e.g. when a weighting window is applied. Often it is the only source of signal losses which is considered.

However, for air surveillance radar the losses associated with the filter mismatch caused by a target Doppler shift can be no less significant. It is interesting to note that the signal losses at a near-zero Doppler frequency are often of no consequence since only clutter enters the radar receiver at those frequencies.

These considerations were accounted for in the filter synthesis. The signal to noise ratio for all possible Doppler shifts must be monitored and the signal to noise ratio loss averaged over the Doppler frequency range was calculated. These data were then used as an optimization parameter when assessing the resulting performance. As an example, the plots of the signal main lobe loss (the lower curve) and signal to noise ratio loss as a function of the Doppler shift are shown in Fig.3 for a P4 LFM polyphase code with the parameters N =100, T=20 µs and a Hamming weighting window.

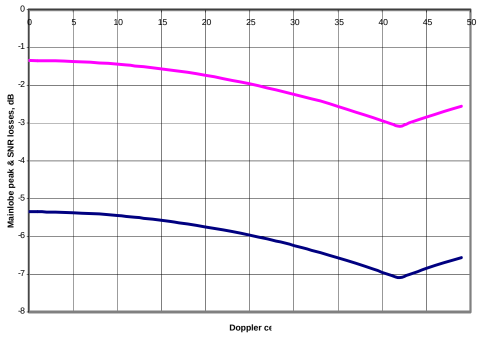

Fig. 3: The signal mainlobe loss and signal to noise ratio loss vs Doppler shift

The problem of required side lobe level should be considered in some more detail. For the Doppler shifts near zero it is desirable to reduce the mean side lobe level and so to diminish the clutter power received by the side lobes. On the other hand, for a Doppler shift near the signal frequency of expected targets the reduction of the maximal side lobes could be of more interest so as to diminish the probability of a false detection due to the presence of a high-RCS airborne target. In both cases the side lobe level of -30 … -40 dB will be as a rule acceptable.

If one sample per a subpulse is used in a polyphase code signal a range cell adjacent to the mainlobe maximum should be formally considered as belonging to a side lobe [3]. This approach is feasible if only near-zero Doppler shifts are considered. As was discussed above the main lobe of a polyphase code signal is shifted along the time axis when the Doppler frequency varies. Therefore, there will be a number of Doppler frequencies for which the signal amplitudes in two adjacent range cells will have near or equal amplitudes. It seems that such conflicts should be resolved by a dedicated signal processing algorithm along with other possible situations when a target signal is split between two range cells and it has no direct bearing on the problem of range side lobes.

4 Performance of compression filters

The performance of the synthesized filters listed in Table 1 is discussed below. For every signal from the Table the data are given for a filter pair corresponding to higher and lower bounding values of code lengths N and the lengths of the transmitted pulse T listed in Table for that type.

A polyphase code P4 signal and a filter with a weighting window are suitable for search and tracking at small (up to 35 km) and medium (up to 75 km) ranges when no significant clutter is present. At these ranges the available signal to noise ratio is typically large enough and the SNR losses inherent to a filter of this type usually do not deteriorate the radar power balance. At the same time, filters of this type can ensure a small level of side lobes.

In selection of the type and parameters of a weighting window a balance should be found between the achieved level of side lobes and the width of the main lobe and SNR losses. As the weighting coefficients are increased to reduce the side lobe level the effective filter bandwidth is reduced resulting in a broadening of the main lobe.

There exist many weighting window types with differing properties. At the present development phase we employed a Hamming weighting which is often used in the compression filter synthesis. Several different weighting schemes may be implemented in the radar as well so that the one best fitted to a particular task can be selected at run-time.

The autocorrelation function of the filter with the parameters N =100, T=20 μs corresponding to the lower bound for this category of filters is shown in Fig.4. The filter has the following parameters:

Weighting Window	Hamming
Code length N	100
Uncompressed pulse length T	20.0 us
Subpulse length	0.200 us
Max Doppler shift	29.25 kHz
Average Mainlobe Losses	-6 dB
Average SNR Losses	-2 dB.

Here and later in the text the average main lobe losses and SNR losses are obtained by averaging over the range of possible Doppler shifts.

This filter can be used to search for and track the targets at large elevations and also at the near-horizon angles when there is no intensive clutter or together with an MTI canceller when a significant ground clutter is present.

The autocorrelation function of the filter with the parameters N =250, T=50 μs for the upper bound (corresponds to 75 km) for this category of filters is shown in Fig.5. The filter has the following parameters:

Weighting Window	Hamming
Code length N	250
Uncompressed pulse length T	50.0 us
Subpulse length	0.200 us
Max Doppler shift	29.25 kHz
Average Mainlobe Losses	-5.9 dB
Average SNR Losses	-1.9 dB.

In some scenarios the SNR losses inherent to a weighting window filter (-1.9 dB for the discussed filter) may be unacceptable for target detection at the ranges near the upper bound of the medium-range distances. A filter based on NLFM techniques can be profitably employed in such situations. The autocorrelation function of such a filter for the same parameters N =250, T=50 μs is shown in Fig.6. The filter has the following parameters:

Signal Spectrum	Cos on pedestal
Code length N	250
Uncompressed pulse length T	50.0 us
Subpulse length	0.200 us
Max Doppler shift	29.25 kHz
Cosine power	3.0
Pedestal	0.0183
Average Mainlobe Losses	-0.7 dB.

This filter has average SNR losses which are lower by 1.2 dB but an average level of side lobes higher by 2.3 dB than its weighting window LFM counterpart.

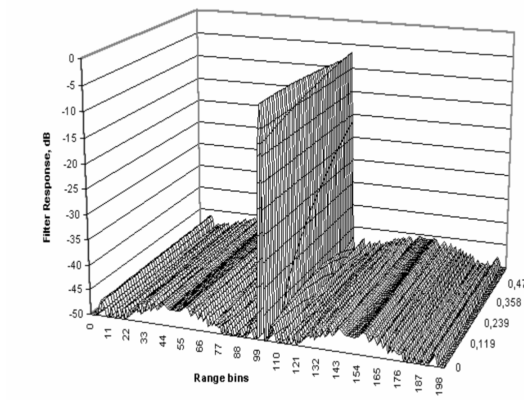

Fig.4: The autocorrelation function of the filter N =100, T=20 us, P4 code, Hamming window

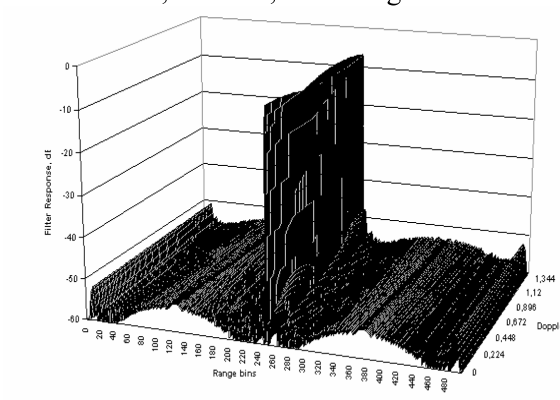

Fig.5: The autocorrelation function of the filter N =250, T=50 us, P4 code, Hamming window

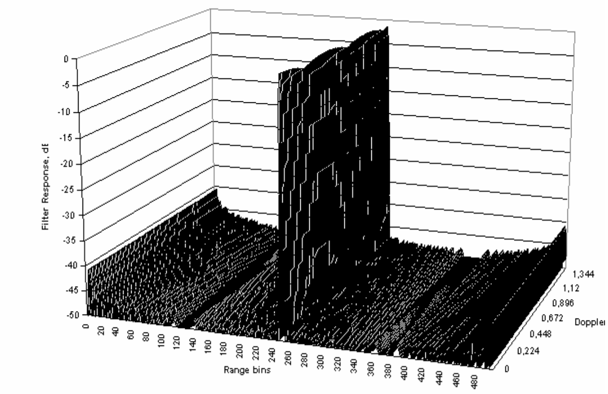

Fig.6: The autocorrelation function of the filter N =250, T=50 us, nonlinear frequency modulation

To avoid undesirable SNR losses in search and tracking tasks at long ranges 60 – 150 km when no intensive external clutter is present a matched NLFM filter is preferable.

The autocorrelation function of such a filter for the parameters N =40, T=40 μs is shown in Fig.7. The filter has the following parameters:

Signal Spectrum	Cos on pedestal
Code length N	40
Uncompressed pulse length T	40.0 us
Subpulse length	1.0 us
Max Doppler shift	29.25 kHz
Cosine power n	3.0
Pedestal k	0.04
Average Mainlobe Losses	-0.5 dB.

The autocorrelation function of the corresponding filter N =100, T=100 μs designed by the same method for the upper bound of the long ranges 150 km is shown in Fig.8. The filter has the following parameters:

Signal Spectrum	Cos on pedestal
Code length N	100
Uncompressed pulse length T	100.0 us
Subpulse length	1.0 us
Max Doppler shift	29.25 kHz
Cosine power n	3.0
Pedestal k	0.0234
Average Mainlobe Losses	-0.9 dB.

If an intensive precipitation clutter is present at long ranges the near-horizon target detection and acquisition will be performed by a bank of Doppler filters. The coherent sequences of unmodulated pulses at a comparatively high pulse repetition frequency (several tens of kHz) will be used for search task.

The ambiguity resolution and target acquisition at these ranges require range resolutions of the order of 1 us, which means that the pulse Doppler sequences using modulated pulses and a subsequent pulse compression should be employed. For obvious reasons, it is desirable to achieve both a small SNR loss and a low level of range side lobes in the compressed waveform.

The pulse length in a pulse Doppler sequence can not be large, for a coherent sequence at 6.67 kHz pulse repetition frequency its value is T=15 us. Consequently the code length is small, N =16 for the discussed example.

As was mentioned above, the range side lobe level for a polyphase coded waveform depends on the code length and diminishes as the code length is increased and obtaining a satisfactory side lobe level in the discussed case may present a problem. We designed two filters for the purpose. One of them is a Hamming weighted LFM filter based on the P4 polyphase code. Its autocorrelation function is shown in Fig.9. The filter has the following parameters:

Weighting Window	Hamming
Code length N	16
Uncompressed pulse length T	15.0 us
Subpulse length	0.94 us
Max Doppler shift	29.25 kHz
Sample rate	1
Windowing Enabled	
Matched/mismatched noise ratio	4.01 dB
Average ML	-5.8 dB
ML at Doppler=0	-5.35 dB
Average ML SNR loss	-1.8 dB.

The second filter is designed as an NLFM matched filter. Four samples per a subpulse were used in the design. Its autocorrelation function is shown in Fig.10. The filter has the following parameters:

Spectrum	Cos on pedestal
Code length N	16
Uncompressed pulse length T	15.0 us
Subpulse length	0.937 us
Max Doppler shift	29.25 kHz
nfm slope	0.25
Cosine power n	3.0
Pedestal k	0.1321
Average ML Loss	-0.22 dB.

As is seen from the presented data the second filter has the SNR losses lower by 0.7 dB but a somewhat larger average side lobe level.

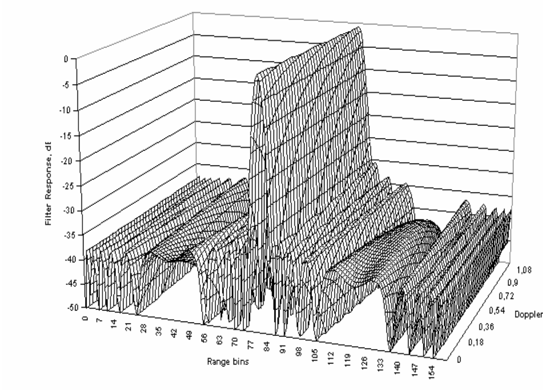

Fig.7: The autocorrelation function of the filter N =40, T=40 us, nonlinear frequency modulation

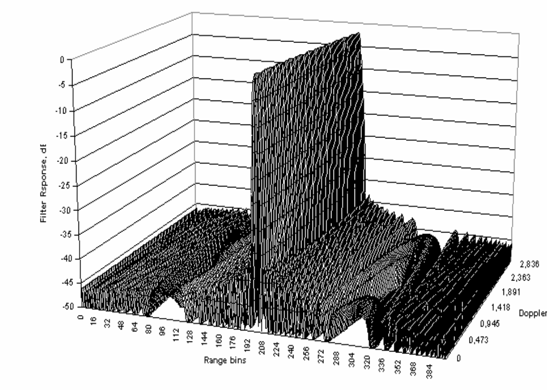

Fig.8: The autocorrelation function of the filter N =100, T=100 us, nonlinear frequency modulation

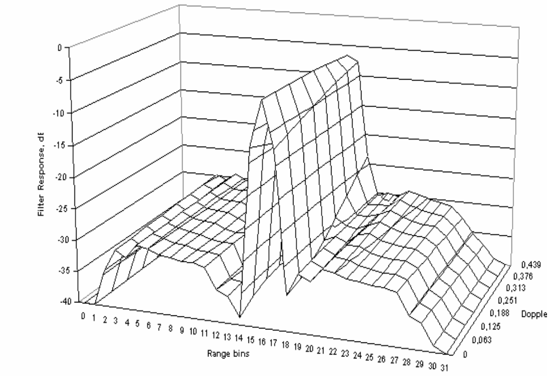

Fig.9: The autocorrelation function of the filter N =16, T=15 us, P4 code, Hamming window

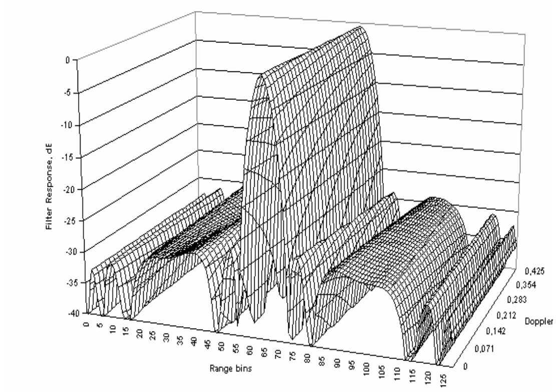

Fig.10: The autocorrelation function of the filter N =16, T=15 us, nonlinear frequency modulation

4 conclusions

In this paper has been provided an introduction to principle behind the pulse compression radar. Then the pulse compression techniques has been described and compared. The problem of the signal losses in a compression filter has been analyzed and explained. We showed that the signal losses arise when a filter is mismatched with the received signal, e.g. when a weighting window is applied, and showed that for air surveillance radar the losses associated with the filter mismatch caused by a target Doppler shift can be no less significant. These considerations have been accounted for in the filter synthesis. The signal to noise ratio for all possible Doppler shifts monitored and the signal to noise ratio loss averaged over the Doppler frequency range calculated. After simulation we find that use of polyphase code in small and medium range and use NLFM and weighted LFM for long range.

References:

[1] Frank,R.,"*polyphase codes with good nonperiodic correlation properties*", June 1963,IEEE/IT,9,43-45

[2] Kretschmer, F.F., Lewis, B.L, "*Anew class of polyphase pulse compression codes and techniques*", IEEE Tran. Aerosp. Electron. Sys., May 1981,17,pp.364-372

[3] Nunn, C.J., Welch, L.R., "*Multi-parameter local optimization for the design of superior matched filter polyphase pulse compression codes*", IEEE International radar conference,2000

[4] W.D. Wirth, *"Radar techniques using array antennas"*, the institution of electrical engineers, london,2001

[5] Kretschmer, F.F., Welch, L.R., "*Sidelobe reduction techniques for polyphase pulse compression codes*",IEEE International radar conference,2000