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# PULSE COMPRESSION IN SEARCH RADAR

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## Abstract

Pulse compression is a widely used method to maintain the range resolution of radar while increasing the average power per pulse that can be placed on a target. Using all these methods a wideband modulation is added to the transmitted pulse and a filter matched to this modulation is used in the receiver. There are several kinds of these additional modulations, and each has different properties, therefore simulation of these methods is essential during design of modulation. In this article an overview of pulse compression techniques is given, the benefits and disadvantages of these methods are discussed. Biphase coded systems are detailed, because they are easy to implement with digital hardware. At the end, our simulation hardware and software tools are described, giving some illustrations from the results.

*Keywords:* radar, pulse compression, sidelobe reduction.

## 1. Introduction to Pulse Compression

Pulse compression has been extensively used since the 1950s to overcome the practical problem of extending the operating radar range while maintaining the required range accuracy and resolution. When a radar transmits a pulse to a target, the pulse energy falls proportional to the square of distance. After the reflection from target the reflected pulse energy falls also proportional to the square of distance, which results in an aggregate signal attenuation proportional to the 4th power of target distance. This can be written as:

$$P_r = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 R^4}, \tag{1}$$

where $P_r$ is the received power, $P_t$ is the transmitted power, $G$ is the radar antenna gain, $\lambda$ is the wavelength, $\sigma$ is the target radar cross-section (RCS), $R$ is the target distance. In the case where only additive noise is present, the maximal $S/N$ ratio at the receiver can be achieved with a matched filter (2).

$$H_{\mathrm{opt}}(\omega) = k\frac{F^*(\omega)}{S(\omega)}e^{-j\omega t_0}, \tag{2}$$

where $k$ is a constant, $S(\omega)$ is the power spectrum of the noise and interference, and $t_0$ is the delay of the filter. If the additive noise is white noise, then $h(t)$ impulse response of the filter equals $f^*(-t-t_0)$, where $f(t)$ is the transmitted waveform. Consequently, the ideal matched filter is a correlator, so the filter output will be the autocorrelation function of the used signal. The achievable maximal $S/N$ ratio at the moment of decision is proportional to the used signal energy (3)

$$\left(\frac{S}{N}\right)_{\max} = \frac{E}{N_0}, \tag{3}$$

where $N_0 = kT_{\mathrm{eff}}$.

When only one pulse is used for a decision, the $S/N$ ratio can be written as:

$$\left(\frac{S}{N}\right)_{\max} = \frac{P_t G^2 \lambda^2 \sigma \tau}{(4\pi)^3 R^4 N_0}, \tag{4}$$

where $\tau$ is the pulse duration time. Our objective is to maximize the $S/N$, which can be made by increasing the values in the numerator, or by decreasing the items of the denominator. The latter is very difficult, only the noise temperature of the system can be reduced. In the numerator we could increase transmitter power, antenna gain or wavelength, but these changes are very expensive. Moreover, improving one parameter usually damages another one. Only increasing $\tau$ seems easy. It is not surprising that a longer pulse has higher energy, if the transmitter power is fixed. The only problem is that a longer pulse results in lower range resolution, because the long reflected pulses overlap each other if targets are close to each other. The idea of pulse compression is that adding a suitable broadband modulation to a long pulse results in a much shorter pulse at the output of the matched filter (*Fig. 1*).

This means that we can increase the average transmitted power while maintaining the range resolution. The price of it is the higher system complexity and Doppler sensitivity. Of course, there are limits of increasing the compression ratio as we will show in the next paragraph.

## 2. Properties of Pulse Compression Systems

It is essential to introduce some further parameters of pulse compression systems. They make it easier to discuss this topic. The main properties are:

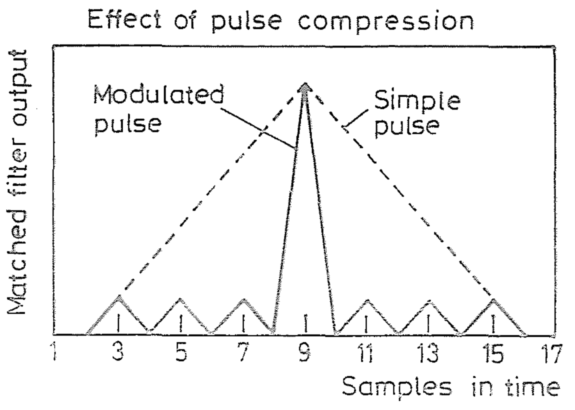


*Fig. 1.* Matched filter outputs of simple and modulated (biphase coded Barker 7 signal) pulses

- Dynamic range; It describes the ratio of the largest and smallest detectable reflected signal amplitude.
- Compression ratio ($CR$); It can be defined as the ratio between the pulse widths before and after the matched filter. It also equals *Signal bandwidth * Transmitted pulse duration time.*
- Detection range ($R$); It is the maximal distance of a detectable target.
- Maximal radial speed of a detectable target (Doppler sensitivity) ($v_{\text{rad, max}}$); Faster targets cannot be detected, because their reflected signal frequency is out of the receiver's bandwidth. Radial speed is the radial component of target speed (if the target moves merely to the radar's direction, then its $v_{\text{rad}}$ equals its speed).
- Range resolution ($\delta_R$); It is the smallest distance of two identical targets when they just can be distinguished from each other.
- System complexity, cost.

For example, an early warning airborne radar has to detect airplanes and missiles up to $n^*100$ km distances and up to 1000 m/s speeds. Of course, it can be realised only by a complex and sophisticated system.

There are further important parameters, which are essential when designing a system, they are: carrier frequency ($f_c$), pulse duration time ($\tau$), signal bandwidth ($B$), modulation type, transmitted power ($P_t$), pulse repetition time ($T_{pr}$). There are the following relations between system parameters:

$$\tau_{\max} = \frac{N_d c}{4 f_c v_{\text{rad, max}}}, \tag{5}$$

where $c$ is the speed of light, $N_d$ is the number of Doppler channels in the receiver. Doppler channels are used to increase the speed range of detectable targets, each channel makes a correlation by a frequency shifted version of the matched filter, so the whole speed range is covered by a number ($N_d$) of compression filters. Doppler effect causes mismatch between the compression filter and the received signal, so the compressed pulse amplitude falls and also the sidelobes of it grow. *Eq.* (5) tells us that too long pulses cannot be used, because this degradation is proportional to $\tau f_{\mathrm{Doppl}}$.

$$\delta_R = \tau v_{\mathrm{rad,\,max}}. \tag{6}$$

It means that the range resolution should not be smaller than the maximal target displacement during the pulse duration time, since it would also destroy the pulse compression effect.

$$B = \frac{c}{2\delta_R}. \tag{7}$$

That is to say that larger bandwidth results in better range resolution. This is not surprising, because larger bandwidth means higher compression ratio, so a thinner compressed pulse appears at the filter output.

$$CR = \tau B. \tag{8}$$

It is the inherent property of pulse compression, as we have written above.

$$T_{pr} \geq \frac{2R}{c}. \tag{9}$$

The background of it is that the reflected pulse has to arrive even from the most far detectable target before the next pulse is transmitted. Thus, $T_{pr}$ cannot be too small. In (9) $R$ means the maximal unambiguous range, while in (10) $R$ is limited by the quality of detection.

$$P_r = \frac{\mathrm{const}}{R^4}, \tag{10}$$

where 'const' depends on the antenna gain, pulse duration, pulse repetition time, losses, desired $S/N$ ratio, antenna turn round time, etc. A detailed form of it can be seen at (1). This means that higher detection range requires higher transmitting power. For example, if the carrier frequency, the maximal detectable target speed and detection range are given, then (5)..(10) determine the other parameters.

## 3. Pulse Compression as ECCM

ECCM (electronic counter-countermeasure) capability is required in several systems. This property is essential in military applications. There are several types of jamming a radar, the simplest way of it is continuously transmitting a narrow band signal at the system's carrier frequency. This interference can be eliminated by adaptively fitting a minimum of the radar's antenna pattern to the jammer's direction. More effective ECM (electronic countermeasure) is when imitative targets are generated onto the radar screen, because these imitative targets are very difficult to distinguish from the real ones. The method is the following: Synchronize onto the pulse repetition frequency, and antenna turn round cycle of radar, calculate the reflected signal delays of imitative targets, and transmit these pulses to the radar as if they were its own reflected pulses.

There are several protection techniques that make jamming difficult. Continuous jammers can be eliminated easily by adaptively fitting a zero value of our antenna pattern at the jammer's direction. Varying the pulse repetition time is good against imitative target jammers. This method makes it hard to synchronize onto the radar pulses. It is also good for varying the blind frequencies of an MTI receiver, because the blind frequencies are proportional to $1/T_{pr}$, so the hidden targets can be detected by the next pulse. Unfortunately, there is a limit of this process because $T_{pr}$ cannot be adjusted during the *coherent pulse interval (CPI)* if we use coherent signal processing on the received pulses.

Using pulse compression also improves ECCM skills of radar system. The substance of it is that these systems use filters in the receiver matched to the transmitted pulse's modulation. So if the jammer uses pulses that are not exactly matched to our receiver, then only a part of the jammer pulse energy will get through the filter. Consequently, the system has jammer rejection property. This rejection is proportional to the mismatch between the receiver's filter and the jammer's signal. So it is worth using a signal that is difficult to imitate. Actually, a better protection can be achieved, if our transmitted waveform is changed pulse by pulse. The biphase coded methods (*Figs. 2, 3*) have large flexibility for realization. In this case the transmitted binary code is varied, and the matched filter follows these changes by updating the contents of the reference code register in its binary correlator circuit. Although in the case of pulse-integration in the receiver the carrier frequency and $T_{\mathrm{pr}}$ cannot be changed, there is an opportunity to change the transmitted code if the coherent processing of pulses is realized after the matched filter. Summing up, using pulse compression improves ECCM skills of the radar system, especially, in the case of code modulated systems.

## 4. Pulse Compression Techniques

There are several methods for expanding the transmitted signal bandwidth. Generally, the most difficult thing is the implementation of the compression filter. One family of pulse compression waveforms is where the carrier frequency is varied during the pulse length. They are the *Linear FM* and

*Non-Linear FM* signals. Another group is the *Phase-Coded* techniques. It means that the transmitted pulse is divided into subpulses, and a digital code is added to it by phase shift keying modulation.

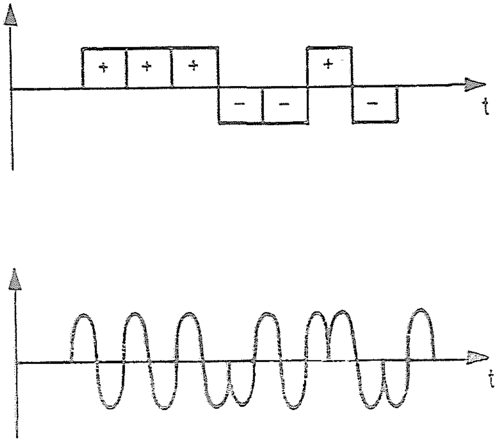


*Fig. 2.* Biphase coded signal (Barker 7 code)

The simplest of them is *Biphase code modulation* (*Fig. 2*), where a binary code is added to the transmitted pulse using BPSK modulation. The subpulses have a 180 ° or 0 ° phase shift in the carrier according to the code items. This method is simple and flexible, which makes it widely used. The received signal is processed by a character filter, which is matched to one element of the code. Then the signal is digitized and a binary correlator is used as the second stage of the matched filter (*Fig. 3*).

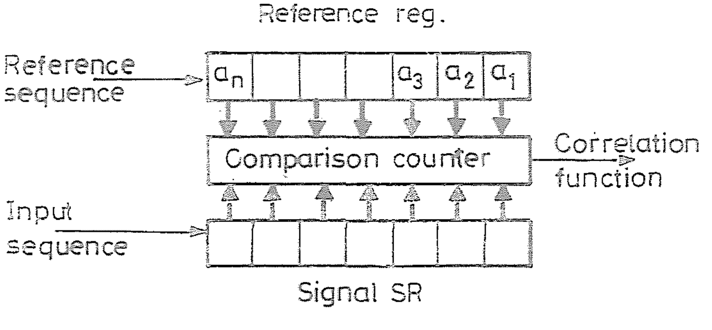


*Fig. 3.* Compression filter structure for biphase coded waveforms

The reference code is loaded into the reference register, the digitized received signal is continuously loaded into the input shift register. The comparison counter counts the bit similarities between the two registers. When only noise is present, about the half of the bits will agree. When a reflected pulse is received, there will be one peak at the correlator output when all the bits agree. This structure is suboptimal, but the required very high sampling rates can be achieved only this way.

There are polyphase modulations, where more than two phase states are allowed. Some types of them are worked out well, e.g. P4, Welti, codes. An interesting quadriphase code modulation was developed by J. W. Taylor and H. J. Blinchikoff. They derive their codes from binary codes with a *Binary To Quadratic Transformation*, and use *Minimum Shift Keying* modulation. The result is high spectral efficiency, and lower spectral sidelobes [1]. Polyphase modulations require more complex signal processing in the receiver than biphase codes.

## 5. Finding the Proper Codes for Pulse Compression

When the desired pulse duration time (detection range) and signal bandwidth (range resolution) are given, the exact waveform has to be designed. In this paragraph the design of binary coded waveforms is discussed. While receiving the reflected signal the autocorrelation function of the used code can be observed at the correlator output. Unfortunately, it contains sidelobes near to the correlation peak (*Fig. 4*).

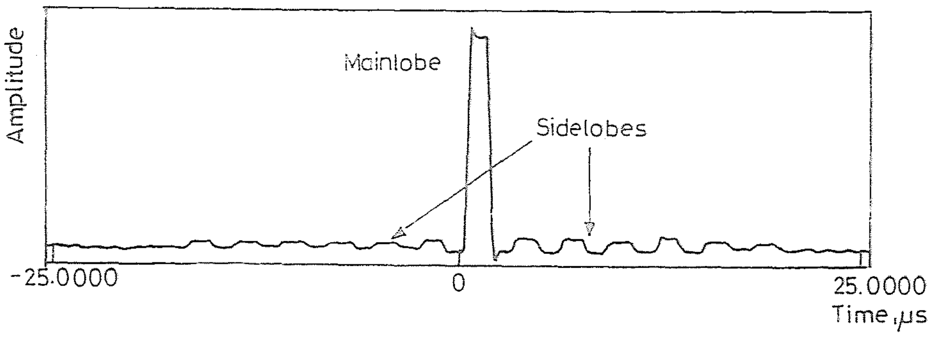


*Fig. 4*. Typical output of binary correlator as a function of time (Barker 13 code, hardware simulation, one sample per code bit)

A good code has sidelobes as low as possible in its autocorrelation function. There are two major parameters that describe a code's properties. They are:

- Relative *Peak Sidelobe Level (PSL)* = highest sidelobe amplitude / main lobe amplitude

| | | | | | |
|---|---|---|---|---|---|
| $N = 5$ Barker code (frame) | $+$ | $+$ | $+$ | $-$ | $+$ |
| $N = 4$ Barker code | $++-+$ | | | | |
| $5 \times 4$ Barker code | $++-+$ | $++-+$ | $++-+$ | $--+-$ | $++-+$ |

*Fig. 5.* Combined Barker codes

- Relative *Integrated Sidelobe Level (ISL)* = total energy in the sidelobes / energy in the mainlobe

They are similar, but not the same. We have to decide in advance which is more important for us, then optimize our system for that property. In spread spectrum telecommunications there are very similar requirements for the used codes, especially, in the DS CDMA systems. The main difference is that in radar systems a single pulse is transmitted, so its aperiodical autocorrelation function must be used, while in telecommunication the code is transmitted continuously, so its periodical auto- and crosscorrelation properties play a more important role. The theory of codes with good periodical autocorrelation properties is worked out well [2]. There are Maximal Linear, Gold, JPL, Nonlinear, etc. codes.

Unfortunately, finding the codes with the best aperiodical autocorrelation properties (e.g. *Minimum Peak Sidelobe-MPS codes*) is more difficult. Presently the only way to find them is trying all the code variations for a given length, calculating their ISL or PSL and choosing the best ones from them [3], [4]. The number of code variations is proportional to $2^N$ where $N$ is the code length. The number of multiplications required to calculate the autocorrelation function is proportional to $N^2$. So the number of multiplications is proportional to $2^n N^2$. It means that there is no chance using this method for codes longer than 55..60 in the near future even with the best computers. For example, in 1975 it took 50 days to find the codes with best PSL for code length of 40 with a special, optimized hardware [5]. In 1995 optimum codes up to length 48 were reported on a radar conference [6]. Unfortunately, it seems that there does not exist a more effective algorithm for finding the codes with lowest sidelobes. Radar systems may require code lengths up to 400. Such long codes can be found in several ways. We can carry out an extensive code search with a computer, trying many random code variations for a given length and selecting the best ones. This method requires very long computational time (several weeks or months) if we want to find usable codes. Another way is cascading short optimum codes like *Combined Barker Codes* (*Fig. 5*). Combined short codes are also called *Kronecker Product Sequences.* Their sidelobe level equals the frame short codes' one, but the compression ratio is higher.

The third way is choosing codes from code groups, which have known good autocorrelation properties. They are also called pseudorandom codes. Although their periodical autocorrelation properties usually have exact the-

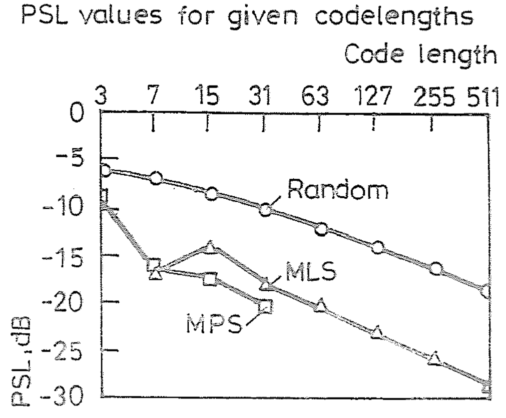


*Fig. 6.* PSL levels of different types of codes as a function of code length

ories, finding the ones with best aperiodical autocorrelation properties is also a search, but there are much less variations to try. These code groups are *Maximal Length Sequences, Gold Codes, Legendre Sequences, Extended Legendre Sequences, Skew Symmetric Sequences.* They are not optimal, but acceptable. Making calculations with *Maximal Length Sequences* we found that codes with good and uniform periodical autocorrelation properties had very different aperiodical autocorrelation properties according to the code sequences and the starting point of cutting out one period (startvector of MLS generator). Unfortunately, many times the codes with best PSL are not identical with ones with best ISL. It seems that it is very difficult to find a code set with large number of codes and uniform properties.

Another method for generating long codes is taking full random bit sequences with a random number generator. The original idea is that theoretically the autocorrelation function of a white noise process is a Dirac pulse. We used a software random generator and made calculations. The result was that by increasing the code length the relative sidelobe level decreased and also the deviation of maximal sidelobe amplitudes decreased. Comparing to the *Maximal Length Sequences* random codes have worse properties both in PSL and ISL, so they are not worth using for pulse compression. This can be seen in *Fig. 6.*

## 6. Sidelobe Reduction Methods

If lower sidelobes than those produced by the best known code for a given length are required, then sidelobe reduction filtering can be used. Two kinds of such filters are known. In the first case a mismatched filter is used for pulse compression, so the crosscorrelation function will appear at the output instead of the autocorrelation function. In the second case a separated sidelobe reduction filter is used after the matched filter. This structure is called nonmatched filter (*Fig. 7*).

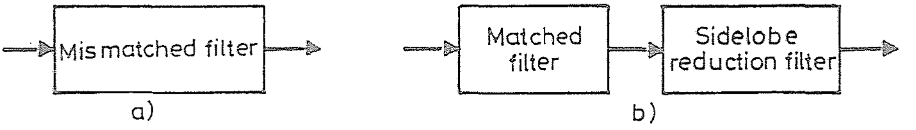


*Fig. 7.* Mismatched (a) and nonmatched (b) sidelobe reduction filtering

These filters can be optimized for PSL or ISL. ISL optimized filters can be derived from the desired discrete crosscorrelation functions by solving matrix-vector equations [7]. Designing PSL optimized filters is a little bit more difficult, in this case a linear programming algorithm gives the optimal filter coefficients. In this linear programming model the target function is the maximization of the mainlobe, while the contracts hold the sidelobe absolute values under a limit. In [8] it is shown that this LP model results in the best PSL value.

*Fig. 8* shows a result of this PSL optimized filtering. A. W. RIHACZEK and R. M. GOLDEN have developed a special filter structure for sidelobe reduction of Barker codes [9]. Its main advantage is that it requires only two or three multiplications, so a less complicated hardware can be used. Originally this filter was optimized for its power spectra density, but in [10] there is an algorithm for PSL-optimizing this structure. A longer filter gives better sidelobe reduction. An infinitely long filter would make all the sidelobes disappear. This is called an inverse filter, because its transfer function equals the reciproque of the Fourier transform of the used modulation signal. (Of course, this is true only within the signal bandwidth.) These methods cause higher noise level ratio, but if original sidelobes disturb us more than the increased noise, then this is no problem. The most important property of these methods is the mismatch loss, which is the loss in $S/N$ ratio, caused by mismatch. J. RUPRECHT and M. RUPF have investigated several binary codes comparing their PSL, ISL values and their mismatch losses in the case of inverse filtering (case of excellent sidelobe suppression) [11]. The minimal mismatch loss is produced by the Barker 13 code (0.21 dB), and they have published codes with mismatch losses less than 1 dB for almost every examined codelengths. Further disadvantages of sidelobe reduction are increased Doppler-sensitivity and larger hardware complexity. While in matched filters binary waveforms can be processed with simple binary correlators, sidelobe reduction filters require many multiplications with real number coefficients.

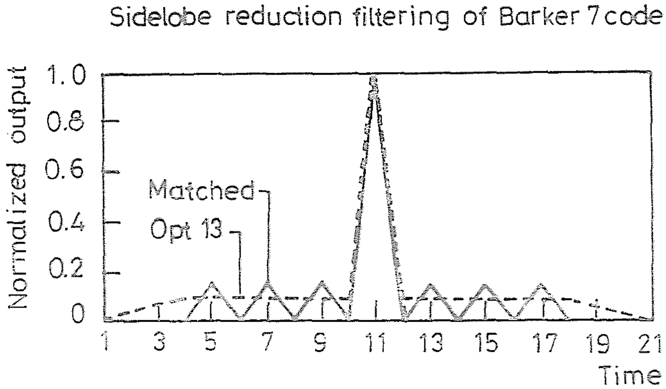


*Fig. 8.* Matched and PSL-optimized mismatched filter responses of Barker 7 code; mismatched filter length of 13

## 7. Doppler Response of Biphase Codes

In the presence of Doppler mismatch the main lobe level falls and the sidelobes rise. If we calculate the waveforms at the receiver output as a function of Doppler shift, then we get a 3-dimensional function. A normalized version of this function equals $|\chi(\tau, \nu)|$, where $\chi(\tau, \nu)$ is the so-called ambiguity function of the waveform (*Fig. 9*). Biphase modulations are very Doppler-sensitive, and this sensitivity is proportional to the code length. In addition, the sidelobe level of MPS codes rise to a higher level in the case of Doppler mismatch, while MLS codes do not show such behavior. This is caused by the property of the ambiguity function that the total volume under $|\chi(\tau, \nu)|^2$ equals unity, independent of the signal waveform [12]. As the volume under the mainlobe is small, the main part of this unit volume has to be under the sidelobes. So an average sidelobe level can be calculated from this volume if the time duration and bandwidth of the transmitted signal is known. This average sidelobe level approximately equals the sidelobe level of MLS codes, while MPS codes have less sidelobes which can be kept only in the narrow surroundings of zero Doppler shift. Sidelobe reduction techniques increase Doppler sensitivity of the system, because they can produce their low sidelobes only in the narrow surroundings of zero Doppler shift. *Fig. 10* shows this behavior.

## 8. Experimental System

At our department a biphase pulse compression system has been developed. *Fig. 11* shows the structure of this system. The desired code can be loaded

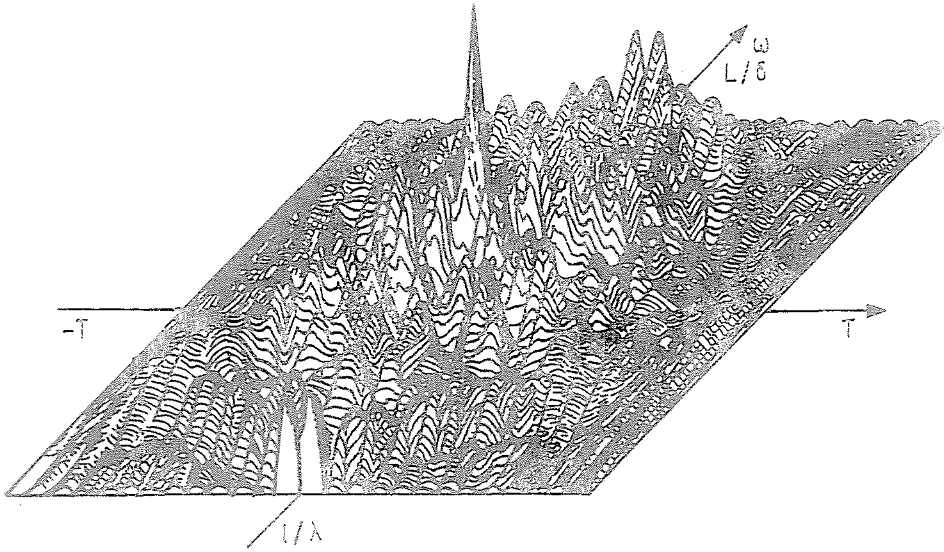


*Fig. 9.* $|\chi(\tau,\nu)|$ of Barker 13 code [12]

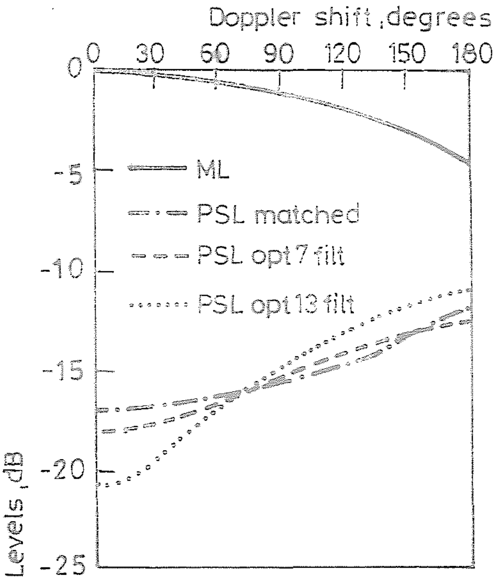


*Fig. 10.* PSL and Mainlobe Levels of Barker 7 code as a function of Doppler shift (in degrees during the pulse length) in the case of matched and PSL-optimized mismatched filtering; mismatched filter lengths of 7 and 13

from a PC to an external FIFO memory through a parallel port. This memory transmits its contents at every start impulse, which is generated by a frequency divider from the master oscillator's 30 MHz signal. The analogue interface converts digital data stream to an analogue baseband signal, and after that a mixer converts it to the IF band. In the receiver part a second mixer restores the baseband signal. This signal is fed to the compression filter, whose construction can be seen in *Fig. 12*.

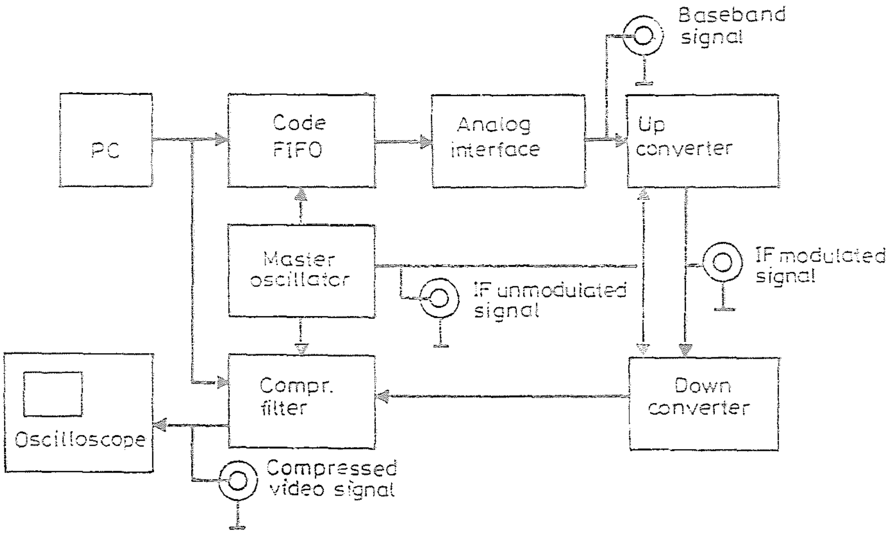


*Fig. 11.* Structure of biphase coded pulse compression system.

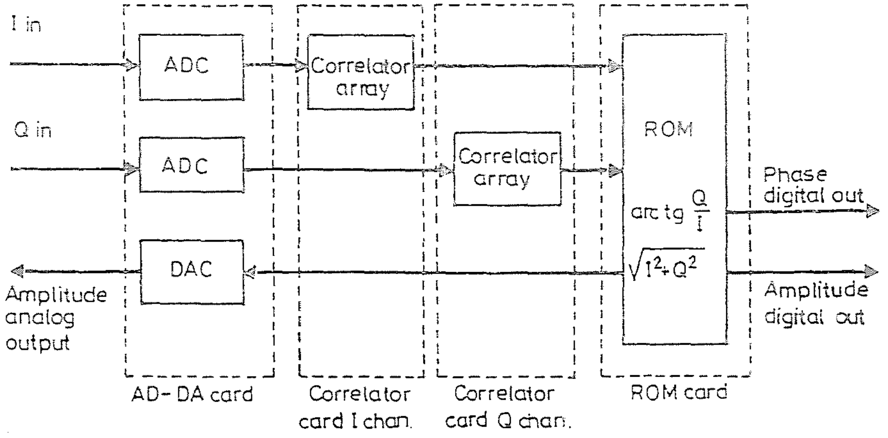


*Fig. 12.* Realization of compression filter

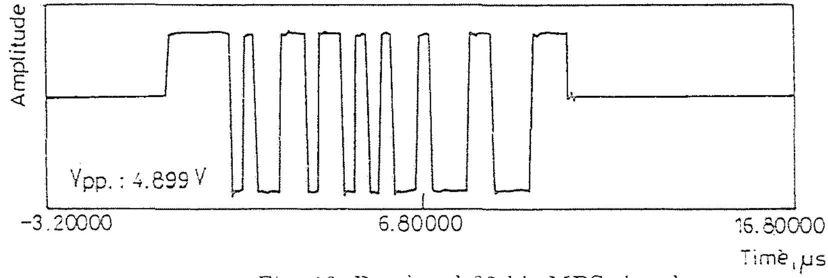


*Fig. 13.* Baseband 32 bit MPS signal

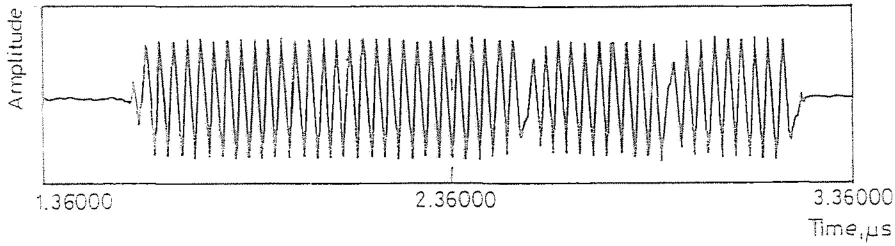


*Fig. 14.* Barker 5 modulated pulse at IF

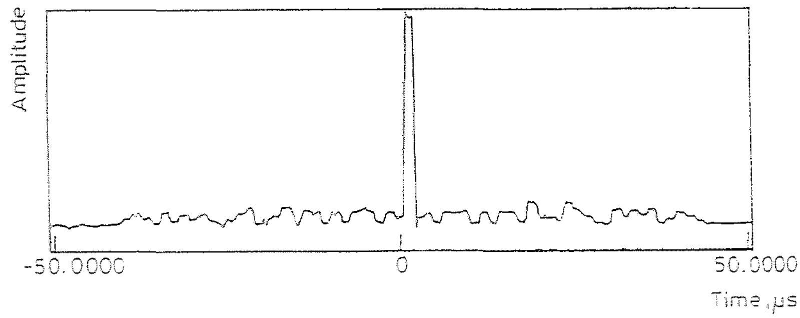


*Fig. 15.* Compressed 32 bit MPS pulse

The received signal is divided into $I$ and $Q$ channel. Each channel contains an analogue to digital converter and a correlator array, which makes the compression. The output signal is derived from the compressed waveforms of $I$ and $Q$ channel. This square root function is realized by a table stored in a ROM. This structure is fast and simple. Correlators are realized with special binary correlator ICs. These circuits contain eight simple binary correlators of length 32, which can be configured parallel or serially. This facility allows adjusting processed data width 1 up to 8 bits, depending on the desired correlator length. Configuration can be set by writing bytes into

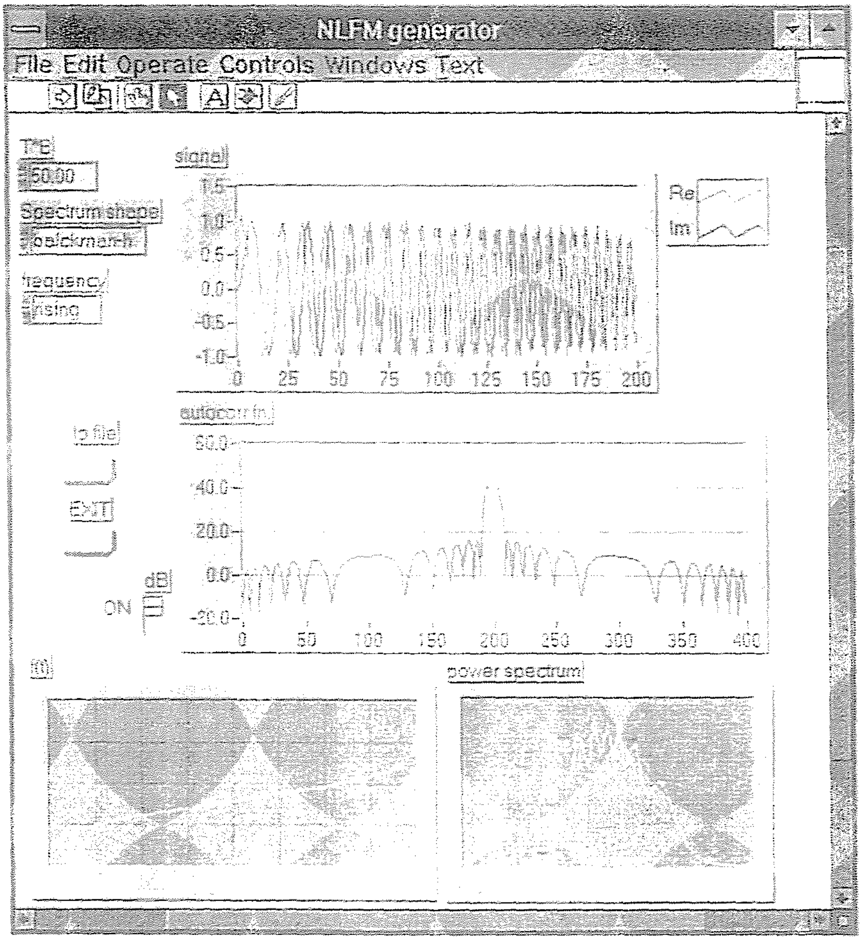


*Fig. 16.* Window of Nonlinear FM waveform generator module

configuration registers of the ICs. In our system all the configuration, the reference code and the mask registers are loaded from a PC through parallel ports. The internal configuration of correlators is optimized to the code-length by the loader program of the PC. The filter's clock signal is obtained from the master oscillator by a frequency divider. We intend to develop this system to make more complex measurements. The added facilities will be adjustable phase noise and Doppler shift at the second mixer's local oscillator and additive Gaussian white noise. It will allow making measurements investigating the effects of these degrading factors. In *Figs. 4* and *13–15* some typical measured waveforms of the system are shown.

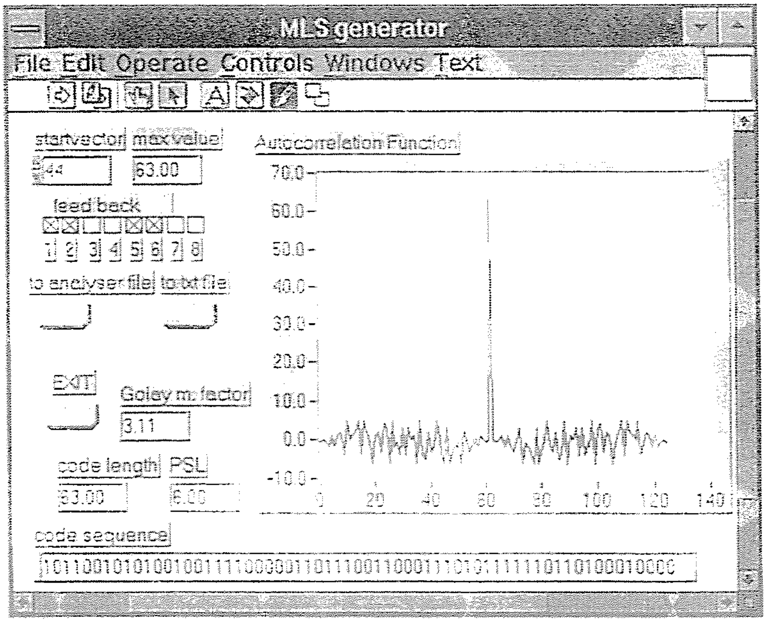


*Fig. 17.* Window of MLS code generator module

## 9. Software Simulation

The hardware described above works only with biphase codes, and this time there is no facility to simulate the effect of Doppler shift with it. Therefore, there was a need for a software tool that is able to simulate all pulse compression methods, and also makes adding degrading factors possible. It was developed in a graphical programming environment, which made the work effective, because the graphical user interface could be constructed easily. This environment also makes the software most clearly arranged. The software contains two main parts. The first was made for generating waveforms and save them into files. At the present Linear FM, windowed LFM, NonLinear FM (*Fig. 16*) and Maximal Length Sequences (*Fig. 17*) can be generated, moreover, any biphase code can be written in, binary signals can be transformed into quadriphase waveforms with BTQ transformation, and also two separate signals can be linked together. The user can adjust the parameters of the desired signal type (e.g. compression ratio, window type, MLS generator feedback structure, etc.), then the software generates the waveform, calculates and plots its autocorrelation function. If it is acceptable for the user, then it can be saved into a file.

The second part of the software package is the signal analyzer (*Fig. 18*), where the user can load a waveform from a file generated by the first program part, then the effects of noise and Doppler bshift can be investigated.

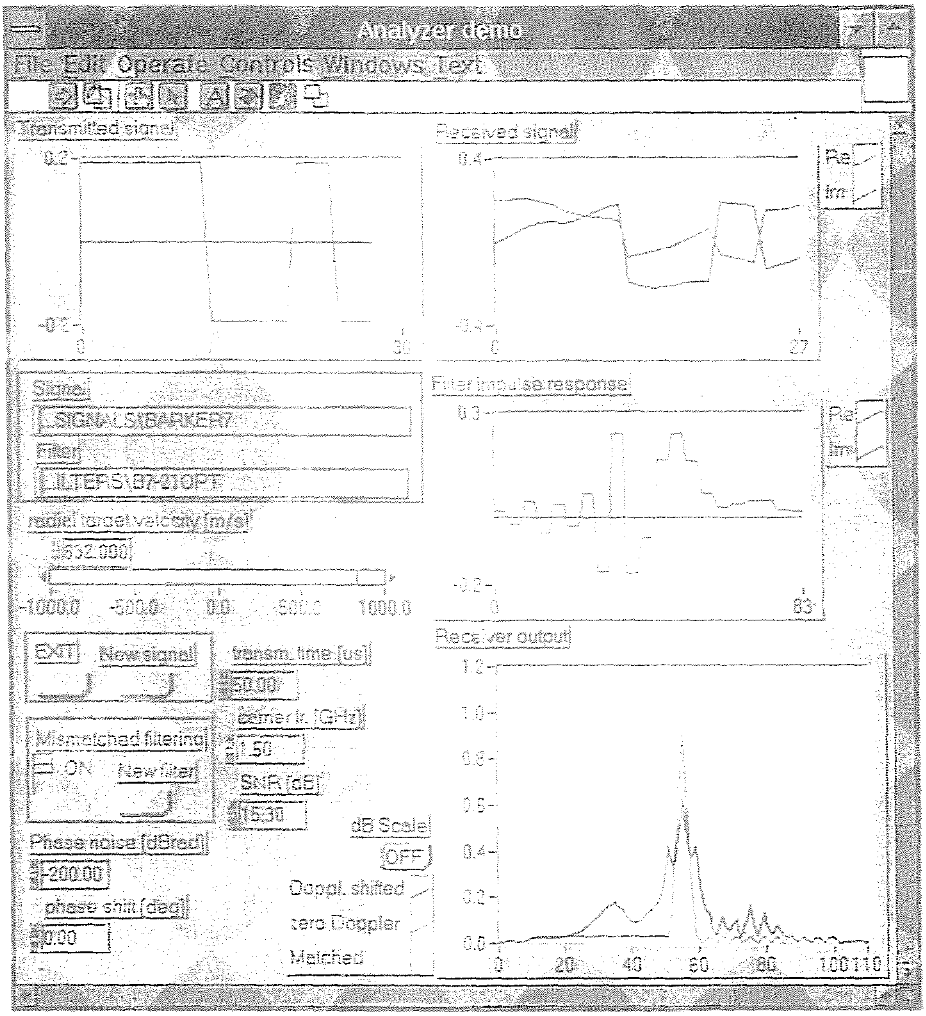


*Fig. 18.* Window of signal analyzer module simulating a 7 bit Barker waveform

The adjustable parameters are transmitter carrier frequency, pulse duration time, target radial velocity, $S/N$ ratio at receiver input and phase of incident waveform. The program plots the original waveform, the received waveform, the crosscorrelation function of them (compression filter output signal) and the autocorrelation function of the original waveform (ideal compressed signal). The last two ones make it possible to compare the degraded and the ideal compressed signals. This tool is suitable only for matched filtering yet, but there is an additional program which makes using mismatched filtering possible. It was developed to investigate the effect of Doppler shift on the Peak Sidelobe Level of biphase codes. This program calculates and plots PSL as a function of phase winding caused by Doppler shift during the pulse. The resolution and the maximal phase winding can be adjusted. It should be mentioned that PSL-optimized filter coefficients were derived by linear programming with EXCEL solver, because integrating this mismatched filter design method into our software would have been difficult. Some user's windows of these program items illustrate the operation of this software in *Figs. 16–18*.

## 10. Summary

The achieved hardware and software simulation tools were described. The first is suitable only for measuring biphase modulations. It will be upgraded to a more complex system, which will make the addition of Gaussian white noise, phase noise and Doppler shift possible. The simulation software already provides these facilities. It also allows simulation of any waveforms. Further possible upgrades are integrating into it, as different radar cross-section models, clutter models and propagation anomalies.

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