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AN ADAPTIVE FILTER DESIGN FOR SIDE LOBE REDUCTION IN PULSE COMPRESSION RADAR SYSTEMS

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ABSTRACT

Pulse compression techniques are widely used and active research topic in radar systems. There were few demerits like masking with the existing methods. In order to overcome these limitations , in this project a R-G filter is used to suppress the sidelobes of the radar coming out from the matched filter and BLMS algorithm is used to calculate the filter coefficients. A weighting function is utilized to shape the sidelobe energy in an iterative manner that will yield more sidelobe reduction. Comparison of the (R-G) filter using BLMS algorithm and the matched filter shows that at the expense of insignificant loss in signal-to-noise ratio (LSNR), adequate mainlobe-to-peak-sidelobe ratio (MSR) can be achieved.

KEYWORDS: LSNR, MSR, BLMS Algorithm, Radar Range Resolution

INTRODUCTION

High range-resolution is important for many radar applications. Pulse compression technique is employed to achieve high range resolution with enough low power [1]. Radar range resolution depends on the Bandwidth of the received signal. Bandwidth of a pulse is inversely proportional to the pulse duration. So, short pulses are better for range resolution. But large Bandwidth can increase system complexity and interference. A long pulse can have same Bandwidth of short pulse if it is modulated in frequency or phase. The modulated long pulse is compressed by the matched filter. To modulate the long pulse various types of encodings are used.

In Biphase coding, Changes in phase can be used to increase the signal bandwidth of a long pulse. A long pulse is divided into some sub pulses each of same width. An increase in Bandwidth is achieved by changing the phase of each sub pulse. The phase of the transmitted signal alternates between 0 and 180 degrees according to the binary sequences [2].

When a target echo signal passes a matched filter, the filter output consists of a spike-like main lobe and some noise-like side lobes. These side lobes can mask the main lobe of weak target echo signals. To prevent this problem, there is a need to design a binary code whose autocorrelation function mainlobe-to-peak-sidelobe ratio (MSR) is maximized for a given code length.

Furthermore, a side lobe reduction filter is used to achieve an adequate MSR. The filter coefficients vary from code to code, and change as a function of code and filter length. There are several techniques used to generate a mismatched filter to minimize side lobes. There are two methods which have been usually used for side lobe reduction. Designing a mismatched filter directly from codes. Employing an additional weighting network after the matched filter.

One possibility is conventionally called the (R-G) filters which were introduced by Rihaczek and Golden [3]. The (R-G) filters have advantages of simple filter structure and relatively suitable performance.

In this work an adaptive (R-G) filter is used and the Block LMS algorithm is used to calculate the filter unknown coefficients. The filter yields a satisfactory mean square sidelobe level (MSSL) and an adequate MSR.

DESIGN ISSUES

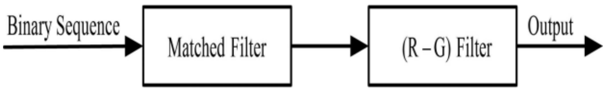

Figure 1: Filter Structure in General

The overall filter structure is shown in Figure 1. The (R-G) filter [4] in Figure 1 is employed to suppress the sidelobe level. Consider the input waveform at the receiver front end x (t):

$$x(t) = \sum_{t=0}^{N-1} x_i \, p_r (t - iT_c) \tag{1}$$

where $\{x_i\}_0^{N-1}$ is a binary sequence with the length of N, T_c denotes the interval of chip time, and $P_r(t)$ is a time-limited pulse waveform in [0, T_c]. Let r(t) denote the matched filter output. so, for $t = ßT_c + \tau$, $0 \le ß \le (2N-1)$, $0 \le \tau \le T_c$,

$$r(t) = \int_{-\infty}^{+\infty} x(\gamma) x(NT_c - t + \gamma) d\gamma$$

$$= O_ß u_p(0,\tau) + O_{ß-1} u_p(\tau, Tc) \tag{2}$$

that has both discrete and continuous convolution parts, with

$$o_\beta = \sum_{i=\max\{0,\beta-N+1\}}^{\min\{N-1,\beta\}} x_i x_{N-1-\beta+i} \tag{3}$$

$$u_p(a,b) = \int_a^b p_r(\gamma) p_r(T_c - a - b + \gamma) d\gamma \tag{4}$$

The continuous part i.e., u_p(a, b) essentially determines the output pulse shape, whereas the discrete part i.e., $o_ß$ reveals the relative output amplitude. Therefore, they can be evaluated separately [5] and the designing can be performed in discrete domain. The true output of matched filter can be also be written as

$$r(t) = \int x(u) x(u-t) du = \int x(u) x(-(t-u) du$$

$$= x(t) * x(-t) \tag{5}$$

where * represents time convolution. By taking the Fourier transform of (5), energy density spectrum is given by

$$R(f) = \left| X(f) \right|^2 = \left| p_r(f) \left(\sum_{i=0}^{N-1} x_i e^{-j2\pi f i T_c} \right) \right|^2$$

$$= \left| p_r(f) \right|^2 \left| \left(\sum_{i=0}^{N-1} x_i e^{-j2\pi f i T_c} \right) \right|^2$$

$$= \left| p_r(f) \right|^2 \left(\sum_{k=-(N-1)}^{(N-1)} c_k e^{j2\pi f k T_c} \right) \tag{6}$$

where $\{c_k\}$ coefficients are the autocorrelation function values, i.e., $c_k = o_{k+N-1}$(and hence $c_0 = N$).

$$r(t) = p_t(t) * \left((N\delta(t) + \sum_{\substack{k=-(N-1)\\k\neq 0}}^{(N-1)} c_k \delta(t - kT_c) \right) \tag{7}$$

where $p_t(t)$ is the autocorrelation function of the $p_r(t)$.

The second term in (7) is due to sidelobes, so this true correlation function (r (t)) may not have some required properties for a given application. Therefore one method is try to synthesize the objective correlation function d (t) by resorting to a filtering step H (f) after the matched filter step:

$$d(t) = p_t(t)$$

$$D(f) = |p_r(f)|^2 = R(f)H(f)$$

$$H(f) = \frac{|p_r(f)|^2}{R(f)}$$

$$= \frac{1}{c_0} \frac{1}{1 + \sum_{\substack{k=-(N-1)\\k\neq 0}}^{N-1} \frac{c_k}{c_0} e^{-j2\pi f k T_c}} \tag{8}$$

$$= \frac{1}{c_0} \frac{1}{1 + \frac{1}{c_0} G(f)}$$

Hence, the transfer function of sidelobe reduction filter is suggested as the following parameterized form:

$$H(f) = a_0 + a_1 G(f) + a_2 G(f)^2 + a_3 G(f)^3 ...$$

$$= a_0 + \sum_{i=1}^{\infty} a_i G(f)^i \tag{9}$$

In order to find the unknown coefficients, the first m terms of H (f) in (9) are transformed to impulse response form by tacking inverse Fourier transformation. Therefore, the impulse response of (R-G-m) filter is given by

$$h_m(t) = a_0\delta(t) + a_1 \sum_{\substack{k=-(N-1)\\k\neq 0}}^{(N-1)} c_k \delta(t - kT_c) + ...+$$
$$a_m \sum_{\substack{k=-(N-1)\\k\neq 0}}^{(N-1)} c_k \delta(t - kT_c) * ...* \sum_{\substack{k=-(N-1)\\k\neq 0}}^{(N-1)} c_k \delta(t - kT_c) \tag{10}$$

Finally, the output of the (R-G-m) filter in discrete domain can be written as [4]

$$s_m[n] = \sum_{k=-(N-1)}^{(N-1)} C_k h_m[n-k] = \sum_{k=-(N-1)}^{(N-1)} C_k a_i h_{m_i}[n-k]$$

$$|n| \leq (m+1)(N-1) \tag{11}$$

Equation (11) can be rewritten as

$$s_m[n] = a_0 s_{m_0}[n] + a_1 s_{m_1}[n] + \cdots + a_m s_{m_m}[n]$$

$$= \sum_{i=0}^{m} a_i\, s_{m_i}[n] \qquad (12)$$

So, from (11) and (12):

$$s_{m_i}[n] = \sum_{k=-(N-1)}^{(N-1)} C_k h_{m_i}[n-k],$$

$$2 \le i \le m,\ |n| \le i\ (N-1) \qquad (13)$$

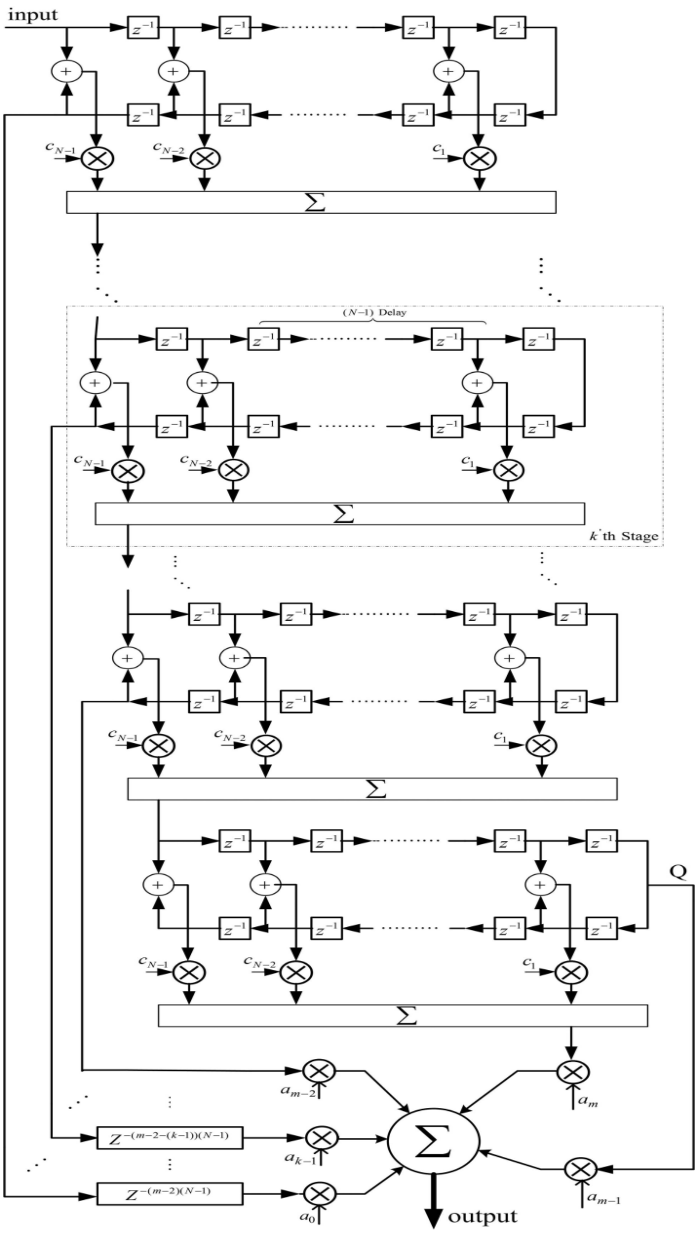

Figure 2: (R-G-m) Filter Block Diagram

BLOCK LMS ALGORITHM

Let the given input signal samples $\{u(1), U(2), u(3), ...\}$ and the desired signal samples $\{d(1), D(2), d(3), ...\}$ correlated with $\{u(1), U(2), u(3), ...\}$.

The algorithm [6] of Block LMS is as follows:

Step 1: Initialize the algorithm with an arbitrary parameter vector *w(0)*, for example *w(0) = 0.*

Step 2: Iterate for $k = 0,1,2,3,...k_{max}$(k is the block index) and initialize $\varphi = 0$.

Step 3: Iterate for $i = 0,1,2,3,\ldots(L-1)$. Generate a new data pair, $(u(kL+i), d(kL+i))$.

Filter Output: $y(kL+i) = w(k)^T \left(u(kL+i) = \sum_{j=D}^{M-1} wj(k)u(kL+i-j)\right)$

Output Error: $e(kL+i) = d(kL+i) - y(kL+i)$.

$\text{Accumulate} \leftarrow \varphi + \mu e(\text{kL} + \text{i})u(kL+i)$

Step 4: parameter adaptation: $w(K+1) = w(k) + \varphi$.

Step 5: Complexity of the algorithm: *2M+1* multiplication and 2M+M/L addition per iteration.

Block Lms algorithm by approximating the gradient by time average the criterion $J = Ee^2(n) = E\left(d(n) - w(n)^T u(n)\right)$ has the gradient with respect to the parameter vector *w (n),* $\nabla_{w(n)} J = -2Ee(n)u(n)$. The adaptation of parameter in the Block LMS algorithm is

$w(k+1) = w(k) + \mu \sum_{i=0}^{L-1} u(kL+1)ew(k) = (kL+i)$ and denoting $\mu_B = \mu L$ the adaptation can be rewritten

$$w(k+1) = w(k) + \mu_B \left[\frac{1}{L} \sum_{i=0}^{L-1} u(kL+1)ew(k)(kL+i)\right] = w(k) - \mu_B \frac{1}{2} \nabla_{w(k)} J$$

where, $\nabla_{w(k)} J = -\frac{1}{L} \sum_{i=0}^{L-1} u(kL+1)ew(k)(kL+i)$, which shows that expectation in the expression of the gradient is replaced by time average.

The problem is to find a_k, such that Error will be a least possible value. By taking the partial derivative of mean square error e (n) with respect to a_k and equating it to zero, the filter coefficients a_k will be obtained.

PERFORMANCE EVALUATION

To demonstrate the acceptability of the filter's performance, the (R-G) filter is compared with the matched filter for three cases. In the first case, performances of the suggested filter and the matched filter are compared for a point target. The second case has several targets distributed in range. In this case the accuracy of the matched filter is expected degrade significantly because of a masking problem. The Doppler's effect is evaluated in the third case. Furthermore, the filter is compared with some other filters in individual cases.

Point Target

The waveform used for this case is the length N = 40 minimum peak sidelobe (MPS) biphase coded signal which is 5A5C3BC444 in hex representation.

By considering a point target scenario, we compare the matched filter with the (R-G-2) filter using BLMS algorithm. Figure 3 shows the outputs of standard matched filter (SMF) and $(\text{R-G-2})_{\text{BLMS}}$ filter.

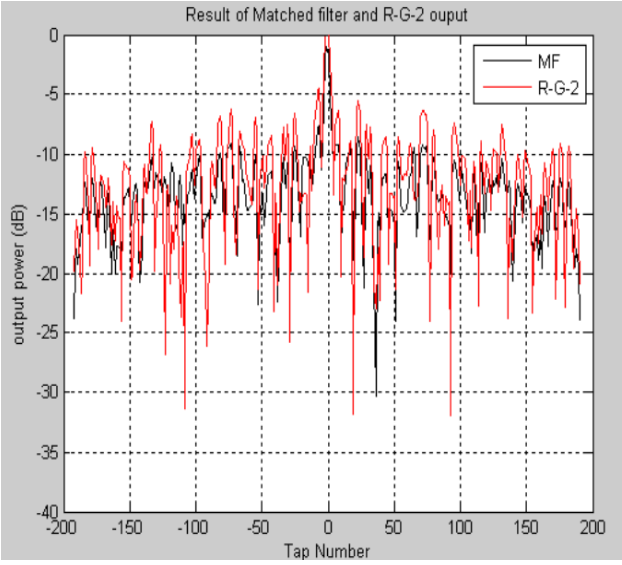

Figure 3: Output of SMF and (R-G-2)$_{BLMS}$ Filter for 40 Bit Binary Code

Doppler Effect

Figure 4. shows the MSR results in which Doppler frequency varies from 0 to 100 kHz . As we can see in Figure 4, the sensitivity of the (R-G) filter using Block LMS algorithm is raised by increasing the stages. However, the MSR results of the (R-G)$_{BLMS}$ filters are much better than the matched filter and (R-G-3)$_{WF}$ in the range 0 to 70 kHz.

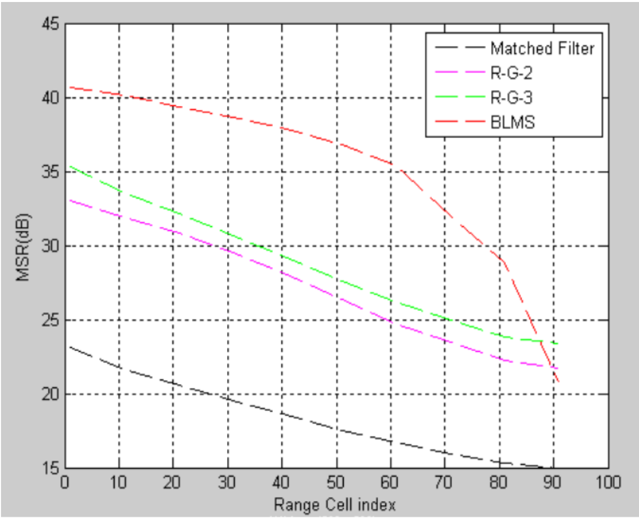

Figure 4: MSR Results of Matched Filter and One, Two (R-G)$_{WF}$ Filters and (R-G)$_{BLMS}$ Filter

Comparison with Other Filters

In Table 1, the performance (MSR) of the matched filter, (R-G-2) filter using BLMS algorithm and (R-G-2) Filter using wiener filtering, weighted wiener filtering, least squares and linear programming methods are compared.

Table 1: Results of SMF and (R-G) Filters in a Point Target Scenario for 40 Bit Binary Code

Filter	Filter Length	MSR(dB)
Matched Filter	40	22.50
(R-G-2)$_{BLMS}$	157	40.05
(R-G-2)$_{WF}$	157	33.25
(R-G-2)$_{WWF}$	157	34.28
(R-G-2)$_{LS}$	157	33.25
(R-G-2)$_{WLS}$	157	34.27
(R-G-2)$_{LP}$	157	34.38

CONCLUSIONS

A m-stage adaptive filter is designed for pulse compression radar systems. This filter yields a remarkable reduction in sidelobe levels. The number of stages is chosen based on the required value of MSR. The filter coefficients were obtained by using several algorithms. one of the best from the previous works is wiener filtering, which gives the MSR value of 33.25dB for the 2nd stage (R-G) filter. In this work, by using BLMS algorithm, more side lobe reduction can be achieved and MSR value is raised to 40.05dB.

The weighting function in an iterative algorithm may be modified in order to have more reduction in the sidelobe level. Such filters can be used in tracking applications or in target recognition systems.

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