## A NEW APPROACH TO THE PROBLEM OF DOPPLER RESISTANT
## BARKER SIDELOBES REDUCING NETWORKS

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## RESUME

Les signals codifiés par la phase sont utilisés souvent dans les modernes radar avec beaucoup de résolution; ils souffrent la présence des lobes secondaires en distance que l'on peut réduire trés difficilement, surtout parce qu'ils changent leurs positions et niveaux lorsque l'effet Doppler est présent.

Ceci est important surtout lorsque deux (ou plusieurs) cibles sont proches, parce que celle-là plus-grande peut cacher, par ses lobes secondaires, l'autre.

Ceci a lieu pour les radar de poursuite où il faut détecter une cible trés petite proche d'une bien plus grande.

Cet article décrit une solution au probléme précédent qui peut etre utiliseé dans les radar de poursuite.

La connaissance de la position de la cible poursuitée permet de mesurer la phase statique du signal interferent.

Par consequent il est possible de tourner le code comprimé: ceci signifie détecter la cible si le déplacement Doppler n'est pas présent; de toute façon l'influence de l'effet Doppler résulte grandement réduite et il est possible de réaliser un approprié circuit de réduction du niveau des lobes secondaires qui est résistant a l'effet Doppler et aussi capable ae faire dégrader tres peu la rapport pic/lobes.

## SUMMARY

Phase coded waveforms are often used in modern high resolution radars; they suffer from the presence of range sidelobes which can be reduced with difficulty, chiefly as they change their positions and amplitudes when Doppler effect in present.

This is especially important when two (or more) targets are closely spaced, as the stronger can mask the other with its sidelobes. This is the case of tracking radars where a very small target must be detected very close to a stronger one.

In this paper a solution to this problem is described which can be usefully adopted in tracking radars.

Knowledge of the range of the tracked echo allows us to measure the static phase of the interferred echo. In this way it is possible to rotate the compressed code; if no Doppler shift is present this means to detect the target; anyhow the influence of the doppler effect is greatly reduced and a proper Doppler resistant sidelobes reducing network can be designed able to change the peak to sidelobes ratio very slightly.

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## 1. INTRODUCTION

In modern high resolution radars phase coded waveforms are widely adopted to reach high compression ratios; their autocorrelation functions are characterized by the presence of high range sidelobes; echoes due to small targets can be masked by the sidelobes of stronger echoes.

This effect is enhanced by the targets Doppler shifts as the sidelobes are strongly affected by the frequency mismatch between the compression filter transfer function and the echoes spectra.

In many receivers the autocorrelation filters are followed by proper networks (sidelobe reducing networks- SRN) able to increase the peak to sidelobe ratios (PSR) at the expense of small mismatching losses.

These networks are usually designed in absence of Doppler shift trying to match the ripple structure of the autocorrelation spectra ( /1/ for Barker codes) or using linear programming with bounds due to the problem which must be solved. The peak to sidelobes ratios thus obtained, even greater than 40 dB, are strongly reduced by the presence of Doppler shifts. These problems have been deeply studied for Barker codes/2/ /3/, but can be used for any other phase coded waveform.
It has been suggested/4/to insert a side-lobes reducing network after the envelope detector as, in this case, the Doppler dependance of the sidelobes amplitudes is significatively smaller; even if this can be considered true for a single target, the envelope non linearity can cause a difficult detection of smaller targets masked by strong sidelobes thus nullifying the effect of the SRN.

In this paper it is shown that the knownledge of the phase due to the propagation delay allows to process only the in-phase component of the received signal thus reducing the influence of the Doppler shift, due to the target motion.

This can be easily obtained in tracking radars, as the positions of the targets are known and an estimation of their phases is possible.

Using linear programming a special SRN can be designed optimized not only at zero doppler shift, but also at the maximum assumed frequency.

Sensitivity losses are evaluated and the behaviour of the SRN is analyzed.

## 2. SIDELOBES REDUCING NETWORKS

Nearly all the approaches known in literature to the problem of reducing the sidelobes of a Barker autocorrelation, are based on an optimization of the weights of a FIR filter in absence of Doppler shift/1/ /2/ /3/ /4/ ; this is very limiting in those radars involved in the detection of small high speed targets in presence of stronger mobile targets such as tracking radars.

In fact the peak to sidelobes ratio worsens as the echo speed increases.

Ref. /4/suggests not to use a F.I.R. filter on the in-phase and quadrature channels of the receiver, but to insert if after the detector.

This approach is less Doppler sensitive as the reduction of the in-phase component is partially compensated by the growth of the quadrature component,

thus the changes of the sidelobes in the envelope of the compressed signal are smaller than those present in its cartesian components and the effectiveness of the SRN is enhanced.

It has been shown/5/that better SRN can be designed using a linear programming approach which is based on a wide bandwidth optimization.

Even if this approach leads to a less Doppler sensitive network the non linear behaviour of the detector limits its usefulness to cases of strong interfering echoes with a limited amplitude difference.

A a point of fact the output of the envelope detector is (neglecting noise)

$$e = \sqrt{l_i^2 + p^2 + 2l_i p \cos\theta} \qquad (1)$$

and assuming $\theta$ uniformly distributed it's obtained /6/

$$\bar{e} = \frac{2}{\pi}(l_i + p) E\left(\frac{2\sqrt{l_i p}}{l_i + p}\right) \qquad (2)$$

where $E(k)$ is the complete second kind elliptic function.

In eq.s (1) and (2) $l_i$ is the strongest interfering lobe and p is the peak of the smaller signal $\theta$ is the phase displacement between the signals.

Assuming $l_i$ much greater that p, and this is often the case of tracking radars where high dynamic ranges are required,

$$\bar{e} \simeq l_i + \frac{1}{4} \frac{p^2}{l_i} \simeq l_i \qquad (3)$$

So the average output of the detector is the interfering sidelobe.

Obviously this is no more true while increasing p, but as a limit eq. (3) shows that no sidelobe reducing network, even if Doppler resistant can allow the detection of the weaker signal as the stronger has captured it, this even if its signal to noise ratio is high.

Using a linear network a better behaviour can be obtained.

This can be accomplished properly rotating the compressed echo and processing only its real part.

In absence of Doppler shift the code subpulses are aligned, so the knowledge of the phase $\phi_o$ (fig. 1) allows to match the echo using only one component:

$$\sigma(t) = R_e\left[u(t)e^{-j\phi_o}\right] \qquad (4)$$

where $\sigma(t)$ is the real signal to be processed, $u(t)$ is the complex compressed received signal.

A Doppler shift causes the growth of a signal component in quadrature with the phase $\phi_o$ ; if only the in-phase component described by eq.(4) can be processed the Doppler sensitivity will be greatly reduced at least in the bandwidth of interest. (fig.2).

The normalized peak of the compressed code is in this case

$$V_p = \sum_{o}^{12} k \cos k \Delta\phi \qquad (5)$$

being $\Delta\phi$ the Doppler shift per subpulse. It is observed that in the bandwidths of interest the change of $V_p$ is small; if $\Delta\phi = 0.06$ rad/subpulse there is a peak loss of .8 dB. This fact suggests that the peak at the output of a F.I.R. sidelobes reducing network can be maximized at the highest Doppler frequency of

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interest, so remaining more or less constant.

The SRN is a transversal filter whose transfer function is:

$$H(z) = \sum_{o\,i}^{N_w} w_i z^{-i} \qquad (6)$$

The weights $w_i$ can be computed using a proper linear programming. The output of such a filter is

$$\chi(t) = \sum_{o\,j}^{2N} \sum_{o\,i}^{N_w} b_j w_i \Delta_\tau \left[ t-(i+j)\tau \right] \qquad (7)$$

being $b_j$ the amplitudes of the compressed Barker code (of lenght N) and $\Delta_\tau(t)$ a rectrangular pulse ($\tau$ sec long).

The statements of the programming problem are:

$$\begin{cases} \chi_{peak} = \chi(h_{peak}) = max \quad at \ \Delta\phi = \Delta\phi_{max} \\[2mm] |\chi(h\tau)| \leq 1 \quad for \ \Delta\phi = 0 \\[2mm] |\chi(h\tau)| \leq 1 \quad for \ \Delta\phi = \Delta\phi_{max} \end{cases} \qquad (8)$$

In system (8) h assumes all the values between 0 and $2N + N_w$ except $h_{peak}$ and $h_{peak} \pm 1$ and $h_{peak} \pm 2$. This is necessary to account for the peak widening.

In fig. 3 the peak to sidelobes ratios which can be reached changing the lenght of the F.I.R. sidelobes reducing network are shown.
In fig. 4 the peak losses, the peak to sidelobes ratio and the enhancement of the closest lobe are shown versus the phase Doppler shift $\Delta\phi$ ; the peak to sidelobes ratio is referred to those lobes for which the linear programming applies. The SRN filter under examination has $N_w = 41$ and $\Delta\phi_{max} = 0.06$ rad/subpulse.

When two signals interfere the estimation of their propagation phase ($\phi_{o1}$ and $\phi_{o2}$) allows to rotate the signals according to the approach just described; in such a way we will process two real signals

$$s_1(t) = Re \left[ \underline{u}(t) e^{-j\phi_{o1}} \right]$$

$$s_2(t) = 2 s_1(t) \cos\left( \phi_{o1} - \phi_{o2} \right) \qquad (9)$$

$\underline{u}(t)$ is the complex signal received; in this case the sum of the two echoes, is:

$$\underline{u}(t) = \underline{u}_1(t) + \underline{u}_2(t) \qquad (10)$$

and, as the phase difference $\phi_{o1} - \phi_{o2}$ is uniformly distributed, after a few passages it is obtained:

$$\overline{s_1(t)} = \sigma_1(t)$$

$$\overline{s_2(t)} = \sigma_2(t) \qquad (11)$$

so overcoming the problems pointed out by eq. 3 if $\sigma_1(t_i)$ represents the sidelobe of a strong echo and $\sigma_2(t_i)$ is the peak of the weaker echo.

The estimation procedure is rather complex in search radars; on the contrary it is very easy in tracking radars where the range of the tracked targets is known.

## 3. APPLICATION TO TRACKING RADARS

In tracking radars the sidelobes problem is very important during both the acquisition and the tracking phases.

In acquisition the radar must inizialise the AGC and the tracking loops, measuring the range of the target in a rough way.

This is done in a range window of proper lenght, whose position is known by other sensors (search radars, infrared cameras and so on).

The choice of the target which must be tracked is made knowing the distances of the possible different targets from the centre of the window (other choices are possible but they are always based on measures of distances); it is clear that there is a significant probability of detecting the sidelobes if their level is high, thus the need of a sidelobes reducing wetwork during this inizialization phase arises.

During tracking, even if the position of the tracked target is known, it is necessary to measure position and speed of other small targets closely spaced to the first one and this accounts for the need of a SRN in this phase.

In both cases the position of the target which must be detected is roughly known and a simple phase estimator can be designed.

This estimator is a phase detector able to measure the phase of the target in a prefixed range gate, which is the centre of the acquisition window in the former case and a gate properly displaced in the second case; the displacement is a few subpulses.

The phase estimator input is the sum of a code subpulse and the noise. In fig. 5 the proposed solution is sketched.

For the acquisition phase a real time phase detector can be designed which is able to measure the phase of the echo without knowing its position /7/ ; the SRN is different from that previously defined as a different Doppler behaviour must be compensated.

As the tracking computations are implemented by a microprocessor in modern tracking radars /7/ , there is no need to simplify the weights as pointed out in /3/ for real time search solutions; moreover the SRN could be different in the different phases of operations.
Due to the non linear behaviour of the phase detector the sensitivity losses have been evaluated by simulation.

First of all a threshold has been computed when only noise is present to have a fixed probability of false alarm; the value chosen is $10^{-2}$ as an integration follows able to reach the wanted $10^{-6}$.

Then a signal has been superimposed and the results obtained are shown in fig. 6 (full line) where the conventional detection curve is represented by a dashed line.

In the presented results we have assumed a known phase ($\phi_o = \emptyset$); further analysis is in progress to account for the phase estimation noise, which will degrade the performances.

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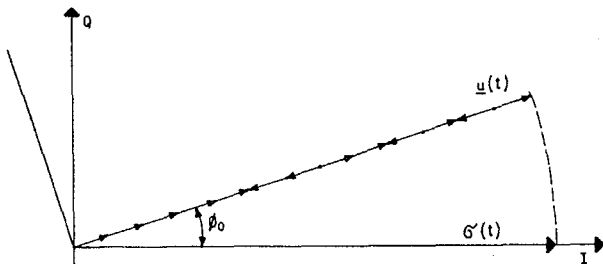
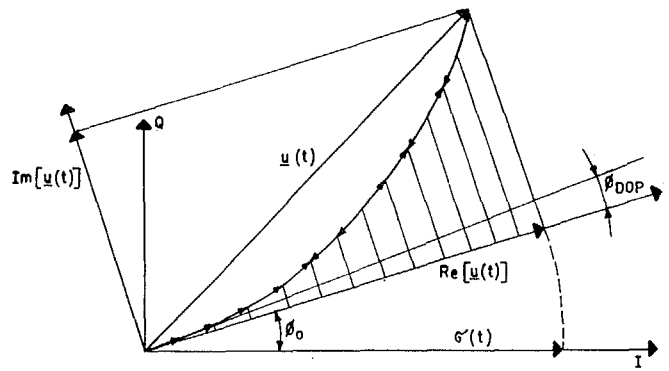Fig. 1 - 13 Barker code without Doppler shift.
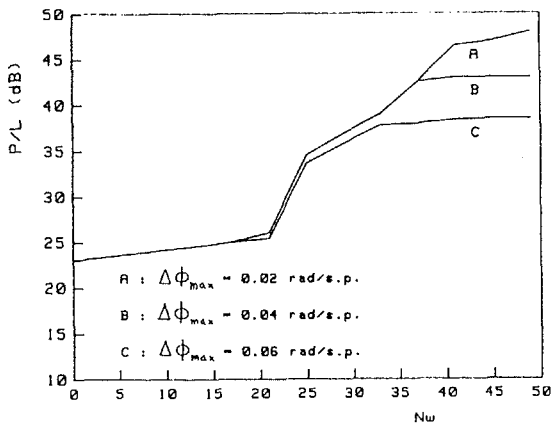


Fig. 2 - As Fig. 1 with Doppler phase shift.
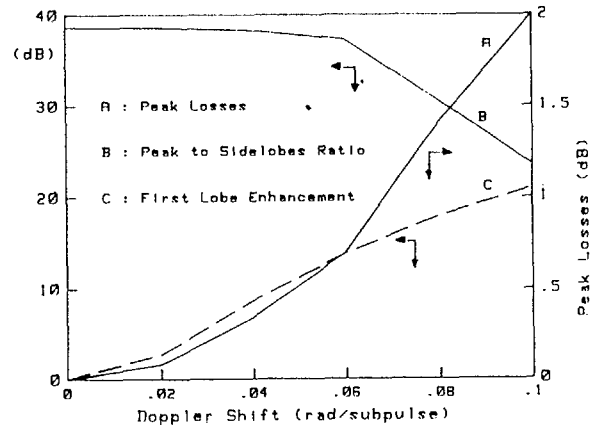


Fig. 3 - Performances of different SRN.



Fig. 4 - Performances of the SRN with $N_w$ = 41.

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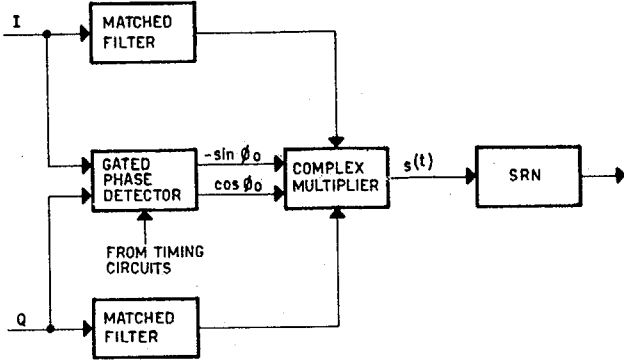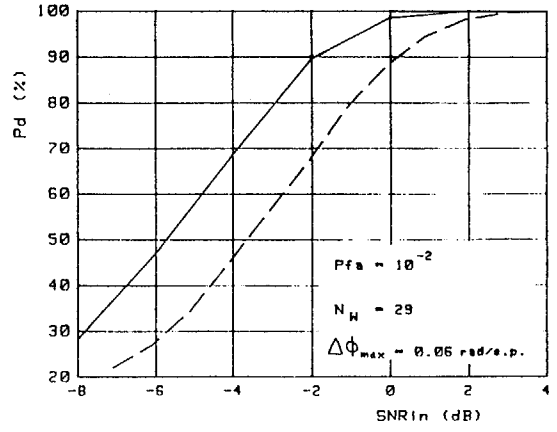


Fig. 5 - Block diagram for traking radars.



Fig. 6 - Detection Probability for single target.