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Un eliminateur adaptif des intérférences naturelles à boucle ouverte: l'ACEC 3

An open-loop adaptive MTI: the ACEC 3

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RESUME

On décrit la structure et les caractéristiques essentielles d'un eliminateur adaptif d'echos étendus, (ACEC) pour applications radars. L'ACEC est constitué par un filtre adaptif, capable de suivre les évolutions et de supprimer en temps réel la fréquence doppler des interférences naturelles (clutter).

Le présent article est divisé en quatre sections: dans la première section on rappelle les caractéristiques génerales des systèmes et des algorithmes adaptifs qui ont déjà été ou peuvent être utilisé dans le domaine des radars. On étudie ensuite le probleme de l'élimination adaptive des intérferences pour les radars, au égard tout spécialement aux contraintes operationnelles de l'équipement. Dans la troisième partie de l'article on présente le schéma à blocs de l'ACEC et on décrit son principe de fonctionnement. On analyse enfin,dans la derniere section, les prestations de l'ACEC pour différents environnement interferentiels (échos de terre, de mer ou de pluie), et pour différentes conditions d'exploitation (presence de bruit blanc, dechalage de la fréquence de répetition radar (PRF staggering)).
Le résultat le plus important est que l'ACEC peut supprimer le clutter mobile avec une adaptation automatique à sa vitesse moyenne sans appreciable degradation de la cancellation .

SUMMARY

The purpose of the present work is to describe the structure and performance of the ACEC (Adaptive Canceler for Extended Clutter). The ACEC is basically an Adaptive MTI filter, able to correct in realtime the mean Doppler frequency of the clutter echoes.

The paper is divided in four sections.
In the first one, the general features of adaptive systems in radar field and the main adaptation algorithms are described. The second section is dedicated to the problem of adaptive clutter cancellation in an MTI radar and its operational requirements. In the third section the block diagram of the ACEC system is introduced and its working principle is described. The last section describes ACEC's performance in various clutter environments (i. e. ground, sea, rain) and in various operational conditions (i. e. presence of white noise, PRF staggering).
The main result is that ACEC can cancel moving clutter with automatic adaptation to its average speed with a very small loss in Improvement Factor.

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0. INTRODUCTION

In radar systems the problem of elimination of unwanted echoes has great importance.
Such echoes are generally referred to as clutter (ground clutter, sea clutter, weather clutter) and chaff.
Anticlutter receivers are generally of the MTI (Moving Target Indicator) type, and utilize the presence of Doppler shift in signals originating from moving targets to filter clutter echoes. A conventional MTI canceller is basically a filter that accepts only signals having a Doppler frequency greater than a fixed value. The design of an optimum MTI filter, i. e. a filter that maximizes the signal-to-clutter ratio at the output, would require the knowledge of clutter characteristics. They are not known a priori, so conventional MTI are designed on the basis of average properties of clutter.
The development of more and more advanced techniques, specially in the field of digital signal processing, makes possible the application of adaptive algorithms to MTI field.
In such a way the main shortcomings of conventional MTI filters can be eliminated.
The algorithms and general features of adaptive systems in radar field, with reference to MTI systems and antenna systems, are described in chapter 1.
The special problem of adaptive clutter cancellation in a radar and the relative operational requirements are object of chap. 2. In chap. 3 the structure and working principle of the devised system, the ACEC, are described and chap. 4 deals with ACEC's performance in various clutter environments.

1. SOME ADAPTIVE SYSTEMS IN RADAR

Adaptive systems are of great interest in modern radars because of the need of coping with more and more difficult environments and, specially, achieving a very high cancellation of unwanted signals such as clutter and jammer.
As a matter of fact, adaptive systems are capable of matching themselves automatically to the variable characteristics of their environment in order to achieve some optimum operational criterion.
A distinction has to be made between open-loop adaptive systems and closed-loop adaptive systems.

In the former case, the adaptation, that is the modification of the system parameters, is obtained only through the system inputs and not through the output. The adaptation process consists of two phases: the evaluation of some unknown or variable parameters, relative to the environment, that is to the inputs, and then the matching of the system, based on the estimated quantities, to implement the selected optimum condition (for example, the maximization of the output SNR).
In closed-loop systems, the adaptation is obtained through both inputs and output of the system.
In these systems, the adaptation philosophy is different i. e. there is not an estimation phase of the unknown or variable, external parameters, but system parameters are changed according to a fixed algorithm so as to realize in an asymptotic way the optimum condition.
For these systems, the stability problem is also to be taken into account.

1. 1 Adaptive Antenna Systems

In radar and telecommunication field, the adaptive systems that have been more widely analyzed and implemented are adaptive antenna systems /1/ - /5/.
These systems are of great interest not only by their specific value, but also for the great resemblance to the adaptive MTI (moving target indication) systems.
The goal of adaptive array antennas is to detect the desired signal in the presence of interference coming from any unknown direction other than that of the desired signal.
In practice the maximum signal to noise ratio is wanted at the output.
An array antenna may be arranged as shown in Fig. 1: $n_1, n_2, \ldots, n_N$ are the complex signals (components in phase and quadrature), relative to interference and noise at the output of the receivers, connected to the various elements of the array antenna; $s_1, s_2, \ldots s_N$ are phasors determined by the direction of the desired signal:

$$S_k = \exp\left(j 2\pi k \frac{d \sin\theta}{\lambda}\right)$$

where: d= element spacing
θ= angle of desired signal direction from boresight
λ= wavelength

α defines the level and the time variation of the desired signal, $w_1, w_2, \ldots, w_N$ are complex weights.
It is shown /1/ that the weights realizing the maximization of the SNR at the output are given by:

$$\underline{W} = \mu \underline{M}^{-1} \underline{S}^* \qquad (1)$$

where μ = an arbitrary constant

$$\underline{W} = \begin{bmatrix} w_1 \\ w_2 \\ \vdots \\ w_N \end{bmatrix}, \quad \underline{S}^* = \begin{bmatrix} S_1^* \\ S_2^* \\ \vdots \\ S_N^* \end{bmatrix}$$

(* stands for conjugate)

$$\underline{M} \equiv [m_{ik}] \equiv [E(n_i^* n_k^T)] =$$
= interference covariance matrix

If in addition to the vector $\underline{S}^*$, also the interference covariance matrix $\underline{M}$ is known a priori, the optimum weights could be determined at the design stage and the system would be non-adaptive.
Conversely, the matrix $\underline{M}$ is usually unknown and may also be time dependent, therefore adaptivity is introduced.

1. 1. 1 Open-loop adaptive arrays

Fig. 2 shows the block diagram of an open-loop adaptive array.
The system inputs, which are the outputs of the receivers connected to the various antennas, are processed so as to obtain an estimation of the interference covariance matrix, on the assumption that both the power and the duration of desired signal are negligible compared with those of the interference signals.

The estimated matrix is then inverted and multiplied by the vector $\underline{S}^*$ so as to obtain an estimation of the optimum weights. The weights, thus obtained, are

used to multiply the signals coming from the various antennas.
Then, by adding the products, the output is obtained.
A considerable disadvantage of this configuration consists in the complexity of the processing and, therefore, of the required hardware. In fact, to estimate the covariance matrix in stationary conditions, N complex multiplications and as many average operations are required.
The inversion of the matrix $\underline{M}$ involves, then, for $N > 2$, many complex multiplications and divisions (for example, for N=3, if the $\underline{M}$ symmetrical properties are not utilized, for the inversion 21 complex multiplications and 9 divisions are necessary).
Finally, the multiplication of the matrix $\underline{M}^{-1}$ by the vector $\underline{S}^*$ involves N^2 multiplications and N(N-1) complex additions.

1. 1. 2 Closed-loop adaptive arrays

The block diagram of a closed-loop adaptive array, shown in Fig. 3, is less complex.
The algorithm used to process the weights to make them tend towards the optimum values is that of the gradient: it seeks for the minimum output mean power by increasing the weights at each step in the direction of the negative gradient of $E|Z|^2$ and proportionally to its amplitude.
Thus we assume:

$$\underline{W}(t+\Delta t) = \underline{W}(t) - K \nabla_{\underline{W}} E\left[|z(t)|^2\right]$$

and, by approximating the expected value $E\left[|z(t)|^2\right]$ to the measured value $|z(t)|^2$, we have

$$\underline{W}(t+\Delta t) = \underline{W}(t) - K \nabla_{\underline{W}} |z(t)|^2$$

$$\nabla_{\underline{W}} |z(t)|^2 = 2 z(t) \nabla_{\underline{W}} z^*(t) =$$

$$= 2 z(t) \nabla_{\underline{W}} (\underline{W}^{T*} \underline{V}^*) = 2 z(t) \underline{V}^*$$

$$\underline{W}(t+\Delta t) = \underline{W}(t) - 2K z(t) \underline{V}^*(t)$$

This, in the continuous case, becomes

(2) $$\frac{d\underline{W}}{dt} = -2K' z(t) \underline{V}^*(t)$$

This corresponds to the diagram of Fig. 3 wherein each weight, in addition to a constant contribution due to the desired signal, is composed of a term given by the integration of (2). It is also possible to demonstrate a posteriori that the diagram of Fig. 3 leads to the weight optimization /2/.
In the steady state we find:

$$\underline{W} = \left(\underline{M} + \frac{1}{G}\underline{I}\right)^{-1} \underline{S}^* \qquad (\underline{I} = \text{identity matrix})$$

that, for large G, approximates to (1).

1. 2 Adaptive MTI with known signal doppler frequency

The problem of the adaptive MTI is very similar to that of adaptive array antennas.
It is a question of filtering the desired signal out of the undesired signals, such as clutter, by taking advantage of their different doppler frequencies.
In fact, it is possible to compare the doppler phase shift between two succeeding clutter samples of an MTI system, defined by:

$$\varphi_D = 2\pi f_D T$$

Where:

T = radar repetition period
f_D = clutter doppler frequency

with the jammer direction, or more precisely the quantity $2\pi d \sin\theta/\lambda$ (phase shift due to the difference of the optical paths, relative to two adjacent antenna elements) in the antenna systems.
If the doppler frequency of the desired signal fs were known, it should be possible to realize an open-loop or a closed-loop adaptive MTI according to diagrams identical to those of Figs. 2 and 3.
Fig. 4 shows the diagram of a closed-loop adaptive MTI.
In this case, the vector of the reference signals $\underline{S}^*$ is defined by:

$$S_K = \exp(jK2\pi f_s T) = \exp(jK\varphi_s)$$

The optimum weights are still given by:

(1) $$\underline{W} = \mu \underline{M}^{-1} \underline{S}^*$$

However the doppler frequency of the desired signal is usually unknown and may vary in time. It is therefore necessary to resort to a different approach.
This is suggested by the consideration of the SCS systems.

1. 3 Adaptive array antennas and SCS (Sidelobe Cancellation System)

The SCS diagram is shown in Fig. 5.
It consists of a main directional antenna and of N-1 omnidirectional secondary antennas with gain equal to the sidelobe average level of the radiation pattern of the main antenna. The desired signal received by each secondary antenna is negligible compared with the desired signal received by the main antenna.
The purpose of the secondary antennas is to provide signals correlated with the interference signal, received by the side-lobes of the main antenna, thus permitting cancellation.
The overall radiation pattern has side-lobes with nulls in the direction of the interference signals.
We may consider the SCS as a particular case of a closed-loop adaptive array. In fact, we select as reference signal vector the following signal vector:

$$\underline{S} = \begin{bmatrix} 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix} \Big\} N-1$$

This corresponds to the assumption that the goal is the maximization of the ratio: power of the desired signal

in the main channel/total power of the interference at the output.
The optimum condition

$$\underline{M}\,\underline{W} = \mu\,\underline{S}^*$$

becomes:

$$(3)\qquad \begin{bmatrix} p_1 & \underline{R}^{T*} \\ \underline{R} & \underline{M}' \end{bmatrix} \begin{bmatrix} w_1 \\ \underline{W}' \end{bmatrix} = \begin{bmatrix} \mu \\ \underline{0} \end{bmatrix}$$

where:

$$\underline{R} = \begin{bmatrix} \overline{n_2^* n_1} \\ \overline{n_3^* n_1} \\ \vdots \\ \overline{n_N^* n_1} \end{bmatrix}, \quad \underline{M}' = \begin{bmatrix} \overline{n_2^* n_2} & \dots & \overline{n_2^* n_N} \\ \overline{n_3^* n_2} & \dots & \overline{n_3^* n_N} \\ \vdots & & \vdots \\ \overline{n_N^* n_2} & \dots & \overline{n_N^* n_N} \end{bmatrix}$$

$$p_1 = \overline{n_1^* n_1}, \quad \underline{W}' = \begin{bmatrix} w_2 \\ w_3 \\ \vdots \\ w_N \end{bmatrix}$$

Equation(3) may be written as two separate equations:

$$p_1 w_1 + \underline{R}^{T*}\underline{W}' = \mu$$
$$w_1 \underline{R} + \underline{M}'\underline{W}' = \underline{0}$$

Since μ is arbitrary and, when the optimum is achieved, necessarily $w_1 \neq 0$, we can assume $w_1=1$; for the remaining weights, we have:

$$(4)\qquad \underline{W}' = \underline{M}'^{-1}\underline{R}$$

The block diagram obtained from that of Fig. 3, by assuming $w_1=1$, $s_2=s_3=\dots, s_N=0$, is the SCS of Figure 5.
It should be noted that an SCS could also be realized in open loop by estimating $\underline{M}'$, $\underline{R}$ and by determining the weights according to equation (4).
Note that an SCS system continues to operate, even if at lower performances when the main directional antenna is replaced by a non directive antenna. Then the overall radiation pattern is composed of a number of lobes, separated by nulls in the directions of the interference signals.

1.4 Adaptive MTI with unknown signal doppler frequency

An adaptive MTI, with unknown signal doppler frequency, may be implemented in a similar manner to an SCS-type antenna system with a non-directive main antenna.
This is equivalent, as shown for antenna systems, to selecting as reference signal vector the following vector:

$$\underline{S} = \left.\begin{bmatrix} 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix}\right\} N-1$$

So doing we are not filtering the desired signal, in the presence of uncorrelated noise: in this case we would rather have a flat frequency response in the doppler range.
If $\underline{M}$ is a diagonal matrix, we have:

$$\underline{W} = \underline{M}^{-1}\underline{S}^* = K\,\underline{I}\,\underline{S}^* = \begin{bmatrix} K \\ 0 \\ \vdots \\ 0 \end{bmatrix}$$

and this involves just a flat frequency response of the MTI. As for antenna systems here also two arrangements may be considered, that of the closed-loop adaptive MTI (Fig. 6) and that of the open-loop adaptive MTI (Fig. 7). Note that the closed-loop adaptive MTI is much simpler, as hardware, than the open-loop one, but, in the digital realization, there are stability problems, which do not occur with open loop systems.

2. OPERATIONAL REQUIREMENTS OF AN ADAPTIVE MTI

The problem of clutter cancellation in radar requires the knowledge of clutter characteristics; they are not known a priori, so conventional MTI filters are designed on the basis of average properties of clutter.
The clutter properties of interest are:

a) The mean Doppler frequency, due to relative motion with respect to the radar. In practical situations, such as shipborne radars or moveable clutter (such as rain and chaff) this is the main cause of performance degradation in conventional MTI's that are designed assuming zero mean Doppler frequency.
Let us consider, as our example, an X-band tracking radar with low-elevation search capabilities. For a relative velocity of the order of 20 m s^{-1} (40 knots) the mean Doppler frequency of clutter is of the order of 1500 Hz, that generally is a large fraction (even more than 50%) of radar's pulse-repetition frequency (PRF). In similar conditions, a conventional MTI cannot cancel clutter without greatly reducing the detectability of targets.

b) The clutter power and the clutter spectral width- An MTI filter is required to have an Improvement Factor (defined as the ratio of clutter-to-signal ratio at the input and clutter to signal ratio at the output, with signal averaged over all possible Doppler frequencies) of the order of clutter to noise ratio. In conventional MTI's the width and shape of the passband are designed in such a way to obtain(possibly) a sufficient Improvement Factor for every operational condition of the system of interest.
On the contrary, an adaptive approach would allow to maintain constant, at the greatest possible extent, the output clutter to noise ratio, with a filter passband as large as possible to ensure a good visibility of targets.
In this paper only the problem of mean Doppler frequency correction will be examined.

The operational requirements that have been assumed are, therefore:

- to cancel a single clutter with an unknown Doppler shift; the clutter is supposed to be extended in Range. The Improvement Factor (I. F.) must be sufficient to keep the clutter residues lower than thermal noise level.
- to guarantee a good behaviour (i. e., no strong degradation of I. F.) at the trailing and leading edge of the extended clutter.

- to provide an extended clutter Doppler estimation insensitive to the Doppler of a point target also in the case of low clutter to signal ratio.

The closed loop approach should lead to the filter structure described in fig. 6, where the number of adaptive loops is equal to the number of MTI samples minus one (e. g. two loops for a double canceller). With such systems it is possible to cancel the clutter regardless of its Doppler frequency. However, some shortcomings resulted from analysis and simulations on a digital computer. These are:

- some stability problems, that require special circuitry to avoid instability when the clutter power variates
- a not good clutter cancellation at the leading edge of a bank of clutter; this is due to the high time constant (of the order of ten range-cells) of the adaptive loops
- a degradation of clutter Doppler estimation when a strong target signal is present.

For these reasons, open-loop systems have been investigated.

3. AN OPEN-LOOP ADAPTIVE MTI: THE ACEC

3.1 General Features

The ACEC (Adaptive Canceller for Extended Clutter) is an open-loop adaptive system with a single degree of freedom. This means that a single (scalar) parameter, namely the mean Doppler shift, is estimated and corrected.
The working principle of this adaptive system can be explained as follows.
Consider the case of an adaptive system which should cancel a single interference (jammer or clutter).
In this instance, the number of inputs may be assumed equal to two. Examine the implementation type open-loop SCS (one weight= 1 and the other one adaptive) that in this instance involves a not excessive hardware complexity.
The input(complex) signals of interest are v_1 and v_2; the "reduced" covariance matrix $\underline{M}'$ consists of a single element , as well as the vector of the cross-correlations $\underline{R}$.
With: (where the superalignment denotes expected value)

(5) $$M' = \overline{v_2^* v_2} = \overline{|v_2|^2}$$

(6) $$R = \overline{v_2^* v_1}$$

if the weight of the signal v_1 is assumed to be 1, the weight w_2 of the signal v_2 is, according to (4):

(7) $$w_2 = -\frac{\overline{v_2^* v_1}}{\overline{|v_2|}}$$

In the case of arrays, we have a system capable of cancelling a single interference and to receive signals from any direction, other than the interference one. The diagram of such a system is reported in Fig. 8. In the case of MTI, we have the single open-loop adaptive MTI whose diagram is reported in Fig. 9.
This MTI allows attenuation of a single clutter signal, with any average doppler frequency and detection of the target, provided its doppler frequency is not coincident with that of the clutter signal.
To simplify relation (7) and, consequently, to reduce the required hardware for w_2 determination, we introduce in (5), (6), (7) the quantities relevant to the clutter: ρ_1 correlation coefficient, σ^2 average power, and average doppler phase φ_D .
With v'_2 being the clutter return obtainable under the same conditions wherein v_2 is obtained, but with null average doppler, we have:

$$v_2 = v_2' \exp(j\varphi_D)$$

$$\overline{v_1^* v_2'} = \rho_1 \sigma^2$$

thus

$$\overline{v_1 v_2^*} = \rho_1 \sigma^2 \exp(-j\varphi_D)$$

$$\overline{v_2^* v_2} = \sigma^2$$

$$w_2 = \rho_1 \exp(-j\varphi_D)$$

In general ρ_1 is slightly less than 1; thus the expression may be approximated by replacing ρ_1 with 1.
In this instance, it is sufficient to evaluate the complex quantity $v_1 v_2^*$ and, through a coherent limitation, to extract the quantity $\exp(-j\varphi_D)$. It is to be noted that, for the single canceler, the weights 1 and $\exp(-j\varphi_D)$ are also the optimum ones, in the sense of the maximization of the Improvement Factor /6/.
When a single clutter is present, if an improvement higher than that obtainable through a single canceller is required, it is possible to apply to a cancellation system with several coincident zeroes, which is easily obtainable from that of Fig. 9 and is shown in Fig. 10.
This system consists of many single cancellers in cascade, each one with one weight equal to 1 and the other weight (adaptive) equal to the weight determined in the estimation device, only one of which is required and is the same as that in the diagram in Fig. 9.

3.2 Description of ACEC

In this section the block diagram of ACEC is described in a greater detail. ACEC has been patented (see reference /9/) and part of this description has been taken from ref. /9/.
As already said, in ACEC the adaptivity is restricted to the clutter mean Doppler shift $\varphi_D = 2\pi f_D T$ and a single adaptive complex weight $\exp(j\varphi_D)$ has to be estimated. This weight, going to the complex multipliers located after each delay line, provides the shift of the MTI frequency response and in particular of its stopband central frequency from zero to $f_D = \hat{\varphi}_D / 2\pi T$, the estimated clutter mean doppler frequency.
As, in general, the clutter Doppler frequency is time and space dependent, the weight has to be estimated in real time and must provide a good clutter attenuation also in the first and in the last range-cell of a clutter bank, as already stated.

Moreover the insensitivity of the clutter Doppler estimate to the target Doppler must be guaranteed for the target range-cell (to allow subclutter visibility) and, if possible, for the contiguous cells (to avoid false alarms).
A possible solution for a weight estimator which fulfils such requirements is shown in fig. 11.
The complex input video signals relative to the same range cells of two consecutive sweeps, v_1 and v_2, after complex conjugation of v_1, go to a complex multiplier which yields the complex signal $v_1^* v_2$. This signal enters a delay line, consisting of N taps each of length τ (range cell duration). So N+1 samples of $v_1^* v_2$, relative to as many contiguous range cells, become available. All such samples, except the central one, are summed to give an estimation of $E[v_1^* v_2]$. The sum signai is coherent-limited originating the adaptive weight. In formulas we have:

$$w = \exp(j\hat{\varphi}_0) = \exp\left(j\angle \hat{E}[v_1^* v_2]\right) = \exp\left(j\angle \sum_{N\,cells} v_1^* v_2\right)$$

With such a scheme of weight estimator, the block diagram of fig. 10 should be modified inserting in points a and b two delays $N\tau/2$ long. So doing when v_1 sample comes to the complex multiplier the corresponding sample(i. e. relative to the same range cell) of $v_1^* v_2$ does not affect the weight estimation. This fact guarantees the subclutter visibility of a point target. But when the target is not in the middle of the delay line it may affect the estimation of the clutter Doppler and give rise to a false alarm. However if N is large enough ($\geqslant 4$) and C/S ratio is not less than 0 dB the clutter Doppler estimation is still good also near the target and CFAR characteristic is not lost.
For what concerns the good clutter cancellation in the leading and in the trailing edge of a bank, i. e. in its first and last range cells, one can see from fig. 11 that since N/2 samples of clutter contribute in these situations to clutter Doppler estimation, if C/N ratio is not too low the canceller performances are still good.

4. PERFORMANCES OF ACEC

In this section the performances of ACEC systems, for single and double MTI's, are related. Of main interest are:
1) the characteristics of the Improvement Factor versus clutter to noise ratio CNR and clutter correlation coefficient ρ (or equivalently clutter spectral spread σ)
2) the I. F. degradation due to prf staggering
3) MTI insertion loss in clutter free environments.

4.1 ACEC, single and double canceller, Improvement Factor characteristics

By simulation on digital computer the mean clutter power out of an ACEC system has been evaluated and the Improvement Factor obtained for both single and double canceller.
For the single canceller case the Improvement Factor of an ACEC system with a weight estimator on 2 and 4 samples and the I. F. of an ideal adaptive single canceller are reported, in table I, for clutter correlation coefficient ranging from 0. 95 to 0. 9999, in the absence of white noise i. e. with CNR = ∞

ρ	I_2(dB)	I_4(dB)	I (dB) ideal adapt.
0. 95	11. 5	12. 6	13
0. 99	18. 6	19. 6	20
0. 995	21. 6	22. 4	23
0. 999	28. 6	29. 6	30
0. 9995	31. 6	32. 4	33
0. 9999	38. 6	39. 6	40

Table I - Improvement Factor of ACEC, single canceller, with weight estimation on 2 and 4 samples, in absence of white noise, compared with the I. F. of an ideal adaptive single canceller.

The Improvement degradation for the two samples estimation is, for any value of ρ, of the order of 1. 5 dB. With a four samples estimation the Improvement degradation is reduced to 0. 5 dB.
In addition a good cancellation in the leading edge of a clutter bank has been confirmed by the simulations.
Since the clutter doppler estimator performances and therefore the adaptive canceller Improvement Factor are influenced by the presence of white noise, the ACEC I. F. dependence on the power clutter-to-noise ratio CNR has been evaluated and reported in figures 12 and 13. Comparing these data with those of an ideal adaptive single canceller (see last column of table I), I. F. degradations of an ACEC system (single canceller) with weight estimation on two and four samples are obtained (tables II andIII). It is to be noted that for high ρ and low CNR the Improvement degradation is very high. However this doesn't matter since in spite of the low I. F. the clutter residues are at a lower level than noise.

ρ	ΔI(dB) CNR=10dB	ΔI(dB) CNR=20dB	ΔI(dB) CNR=30dB	ΔI(dB) CNR=40dB
0. 95	5. 1	2. 4	1. 9	1. 9
0. 99	8. 9	3. 2	1. 7	1. 5
0. 995	11. 4	4. 8	2. 3	1. 8
0. 999	16. 8	7. 1	2. 9	2. 2
0. 9995	22. 1	13. 2	5. 4	2. 3
0. 9999	28. 7	16. 8	7. 4	1. 8

Table II - I. F. degradation, due to adaptivity, Δ I, for ACEC single canceller with two samples estimation.

ρ	ΔI(dB) CNR=10dB	ΔI(dB) CNR=20dB	ΔI(dB) CNR=30dB	ΔI(dB) CNR=40dB
0. 95	2. 6	0. 8	0. 6	0. 6
0. 99	5. 4	1. 5	0. 6	0. 6
0. 995	4. 7	1. 2	0. 7	0. 7
0. 999	12. 4	4. 6	1. 6	1. 2
0. 9995	16. 1	7. 6	1. 9	0. 6
0. 9999	20. 9	10. 9	2. 4	0. 4

Table III - I. F. degradation, due to adaptivity, for ACEC single canceler with four samples estimation

For what concerns the number of samples for the weight estimation, simulations have shown that the I. F. increase that is obtained using 8 samples instead of 4 samples is not very much significant.
Consider now the double canceler case. The Improvement Factor has been evaluated by simulation for an ACEC system with a 4 samples weight estimator. For the clutter-to-noise ratio, CNR, the values of 20,30, 40 and 50 dB and for the correlation coefficient,

ρ, the values of 0.95, 0.99, 0.995 and 0.999 were assumed. The resulting I. F. appear in table IV together with the Improvement of an ideal adaptive double canceler.

ρ	I (dB) CNR=20 dB	I (dB) CNR=30 dB	I (dB) CNR=40 dB	I (dB) CNR=50 dB	I (dB) ideal adapt.
0.95	22.1	21.3	21.5	21.4	23.2
0.99	25.9	32.2	34.4	35.6	37.0
0.995	26.6	37.4	39.8	41.2	43.0
0.999	27.7	35.3	44.7	52.7	57.0

Table IV - I. F. of ACEC, double canceler, with four samples estimation, in presence of clutter and noise, compared with the I. F. of an ideal adaptive double canceller

The Improvement degradation, due to adaptivity, when CNR is not less than the corresponding Improvement of the ideal adaptive double canceler, does not exceed 3 dB.

4.2 I. F. degradation due to prf staggering

The question that arises in the consideration of adaptive staggered MTI filters is whether or not I. F. degradations due to prf staggering and to adaptivity simply add or interact in some way.
It is well known that prf staggering is equivalent to a non-uniform sampling and causes a broadening and a distortion of clutter spectrum. This yields an I. F. degradation in every kind of MTI filter.
In the adaptive case the I. F. loss due to Doppler shift estimation adds to this degradation. But because of the presence of stagger the statistics of the Doppler shift between the two clutter samples in input to the estimator are changed and become time dependent.
So it is necessary to reevaluate the Improvement degradation due to adaptivity, in presence of stagger.
Simulation results for ACEC, double canceler, operating with the stagger sequence of pulse repetition periods (normalized to the average pulse repetition period): 0.89, 1.12, 0.82, 1.03, 0.94, 1.21, are given in table V together with the results obtained for a non adaptive double MTI with the same input data except for the zero mean Doppler shift. The difference between these data, which gives the Improvement loss for adaptivity, also appears in the table V. For purpose of comparison some of the data of last two columns of table IV are transcribed, together with their differences, in table VI.

$\rho(\bar{T})$	$\overline{I.F.}$ (dB) non adapt. staggered MTI	$\overline{I.F.}$ (dB) adapt. stagg. MTI	$\Delta\,\overline{I.F.}$ (dB) adaptivity I.F. loss
0.95	22.6	20.7	1.9
0.99	35.0	33.5	1.5
0.995	39.8	38.2	1.6

Table V - Improvement Factors for non adaptive (conventional) and adaptive (ACEC) double MTI in presence of a $\pm$ 20% prf stagger

$\rho(\bar{T})$	I.F. (dB) non adapt. MTI	I.F. (dB) adapt. MTI	Δ I.F. (dB) adapt.
0.95	23.2	21.4	1.8
0.99	37.0	35.6	1.4
0.995	43.0	41.2	1.8

Table VI - Improvement Factors for non adaptive (conventional) and adaptive (ACEC) double MTI, with no prf staggering

Being the adaptivity I. F. losses in the two cases almost equal it can be stated that, with usual prf staggering and for clutter spectrum width and $\overline{prf}$ such that $0.95 \leq \rho \leq 0.995$, I. F. degradations due to stagger and to adaptivity do not appreciably interact

4.3 ACEC insertion loss in clutter free environments

It is interesting to examine the variations of detection probability P_d when clutter level is under thermal noise level and ACEC is not bypassed.
With an heuristic although not rigorous approach we can say that the position of the stopband of the adaptive MTI is determined by noise samples and then randomly moves in the frequency interval (0,prf). Therefore, a non fluctuating but moving target with a Doppler f_D will be attenuated and amplified from pulse to pulse.
We are in a situation in which the target may be modeled in a similar way to the Swerling cases 2 and 4 (target fluctuation decorrelated from pulse to pulse).
The difference is in the target power (or cross-section) probability density. In our case the probability density of the clutter power is determined by the MTI power transfer function.
For the single canceler, if S_i is the input constant target power, the probability density of the output target power S_o was found to be:

$$p(S_o) = S_i / \left\{ 2\pi \sqrt{S_o(2 - S_o)} \right\}$$

But for evaluating P_d it is necessary to obtain the probability density of signal + noise amplitude at the output of the adaptive MTI. This is not done here but the ACEC behaviour in white noise was investigated by digital computer simulation.
In any way, it is intuitively obvious that like for the Swerling models, when signal to noise ratio is not too low there will be a reduction of P_d due to the signal fluctuations, or equivalently an increase of the SNR required to get a given P_d.
To obtain an indication on such losses a system consisting of an ACEC followed by a double threshold detector with a moving window of 10 pulses length was simulated.
The optimum value 6 was assigned to the digital threshold and, according to this and to a P_{fa} equal to 5.10^{-6}, the normalized analogic threshold was assumed equal to 2.45.
Simulation results are drawn in fig. 14 for single and double canceler.
For $P_d \leq 0.85$ the two curves are very close and SNR losses are less than 2 dB. For higher P_d the two curves are still close but differences in SNR losses become greater, owing to the little slope of the P_d versus SNR curves in that region.
It is noticeable that for the double canceler the ACEC

insertion loss does not exceed 2 dB for $P_d \leq 0.92$. So in many radar systems, for which the cancellation performances of an ACEC, double canceler, are satisfactory, if a SNR MTI insertion loss of 2 dB may be accepted, the ACEC must not necessarily be bypassed when clutter is not present.

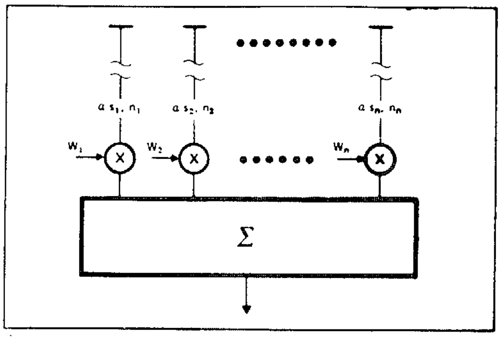

Fig. 1 Adaptive array

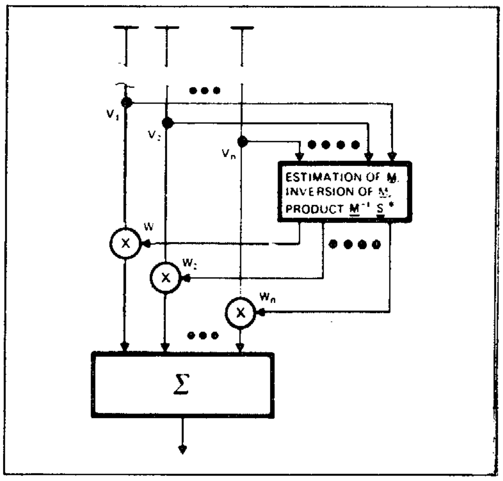

Fig. 2 Open loop adaptive array

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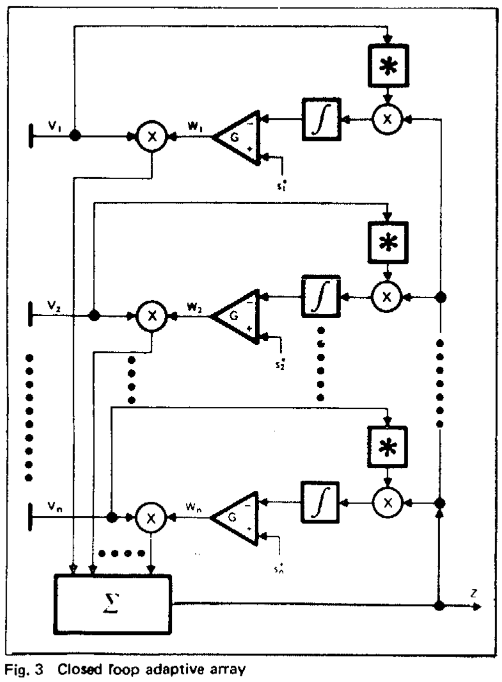

Fig. 3 Closed loop adaptive array

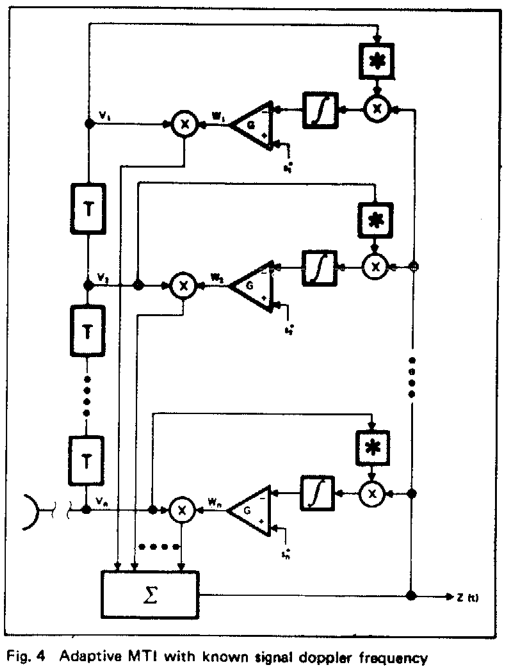

Fig. 4 Adaptive MTI with known signal doppler frequency

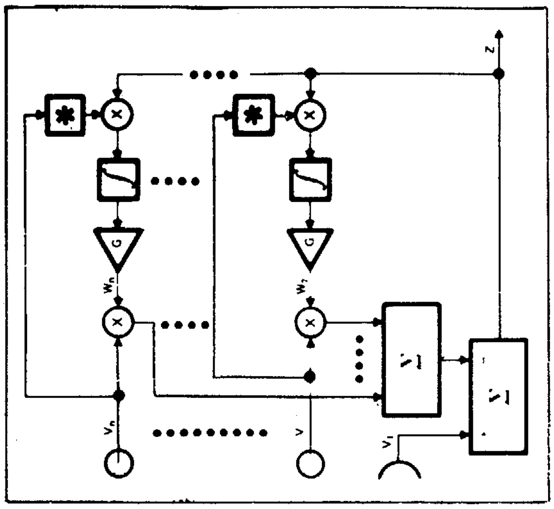

Fig. 5 SCS (Sidelobe Cancellation System)

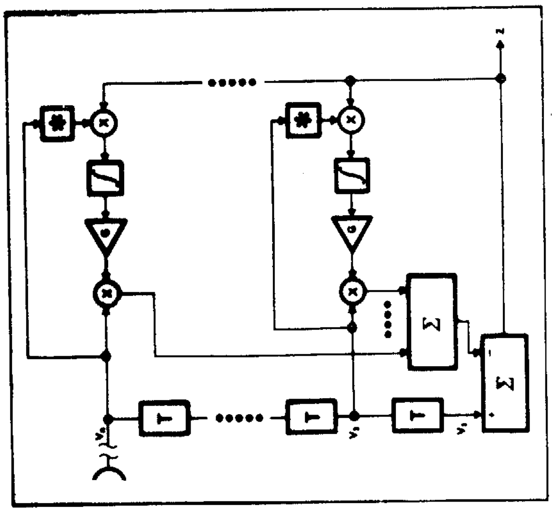

Fig. 6 Closed loop adaptive MTI with unknown signal doppler frequency

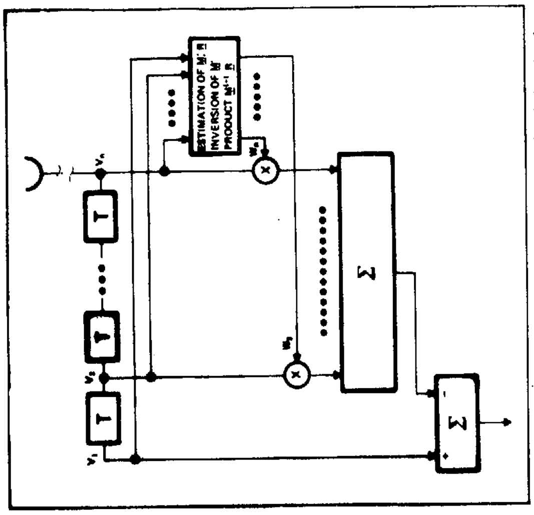

Fig. 7 Open loop adaptive MTI with unknown signal doppler frequency

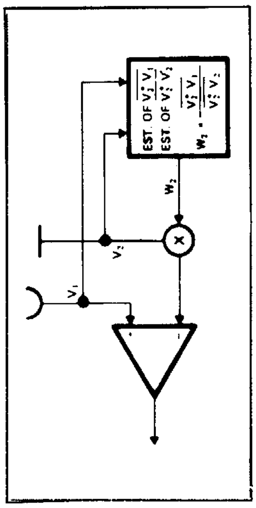

Fig. 8 Open loop single SCS

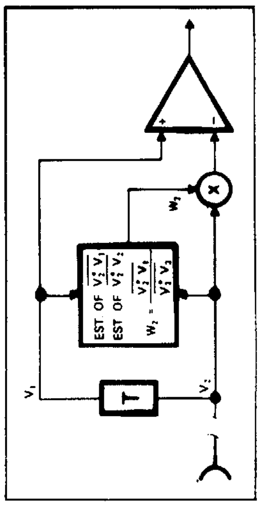

Fig. 9 Open loop single-delay adaptive MTI

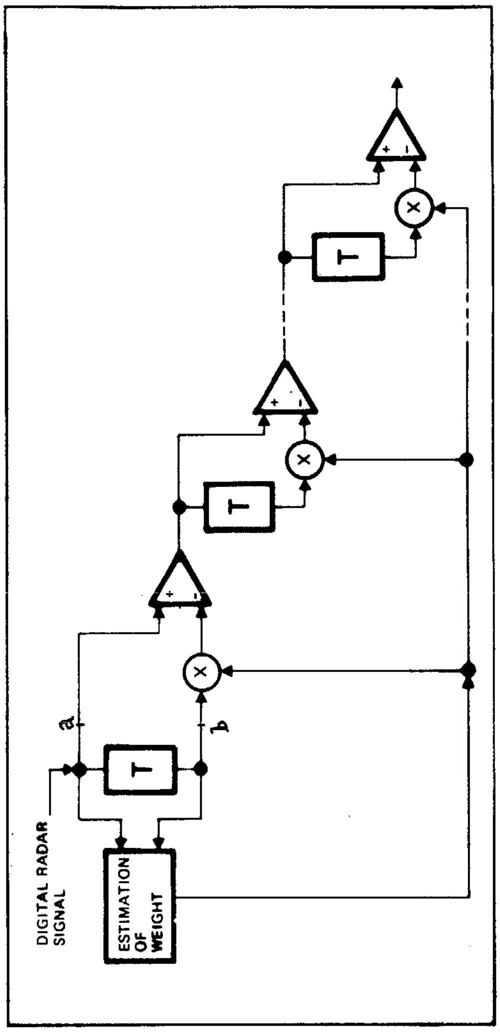

Fig. 10 Open loop multiple-delay adaptive MTI with coincident zeroes (suitable for the cancellation of a single clutter signal)

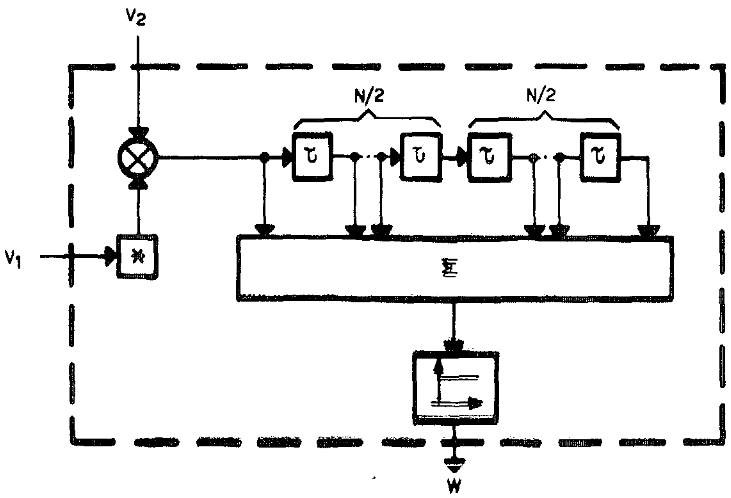

FIG.11 WEIGHT ESTIMATOR (N SAMPLES)

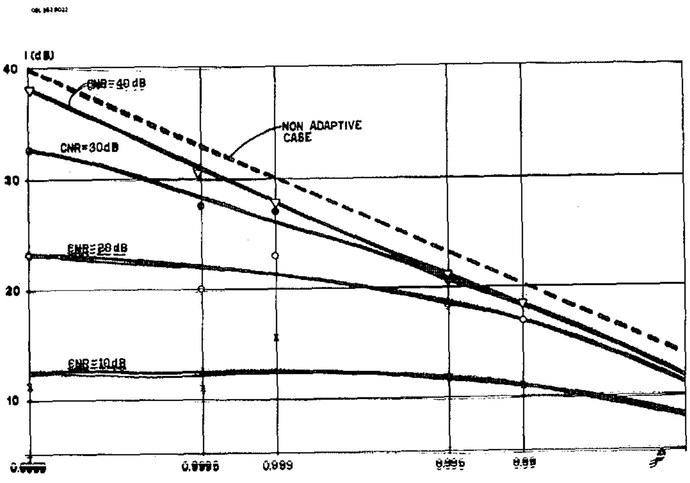

FIG.12 IMPROVEMENT FACTOR OF ACEC SINGLE CANCELER WITH 2 SAMPLES ESTIMATION IN PRESENCE OF CLUTTER AND NOISE

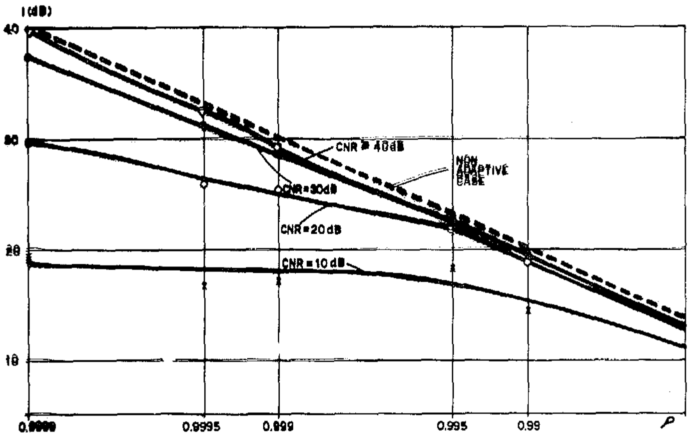

FIG.13.IMPROVEMENT FACTOR OF ACEC SINGLE CANCELER WITH 4 SAMPLES ESTIMATION IN PRESENCE OF CLUTTER AND NOISE

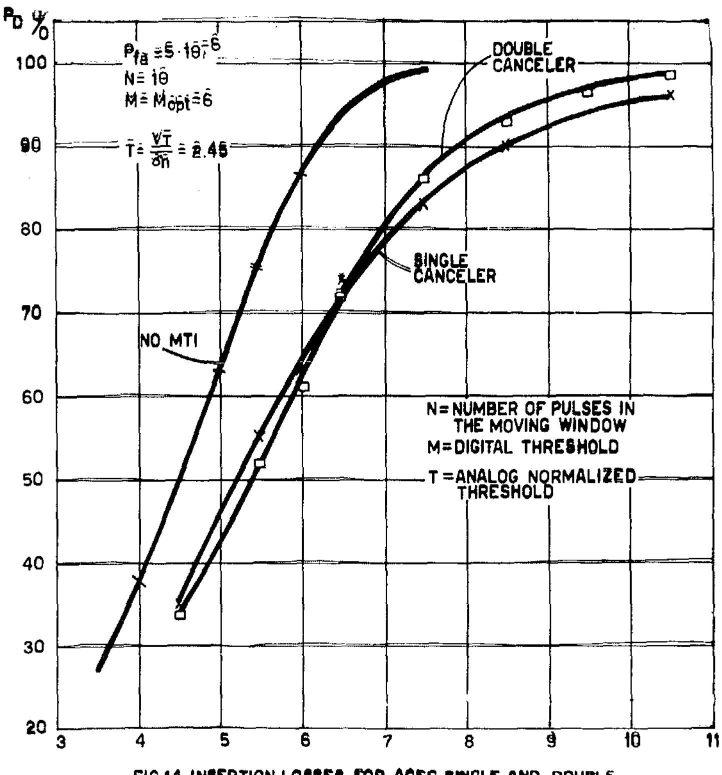

FIG.14-INSERTION LOSSES FOR ACEC,SINGLE AND DOUBLE CANCELER, IN CLUTTER FREE ENVIRONMENTS