CCD non-linearities in MTI receivers

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ABSTRACT

The influence of CCD non-linearities on the performances of an MTI (Moving Target Indicator) receiver is examined. Clutter Attenuation (CA) and Improvement Factor (IF) are evaluated using the non linear characteristic of a CCD delay line both for a single and a double canceller.

The analytical technique adopted can be used for more complex MTI's. Some experimental results are then shown.

INTRODUCTION

An MTI is a circuit able to detect useful moving targets in a ground (or sea or weather) clutter environment whose power is 40-60 dB above the signal power using the different doppler spectra of clutter and signal.

Later, no doppler carrier is thought present in clutter spectra; as a point of fact widely known techniques (f.i. adaptive MTI) are used to cancel this carrier, if present, and can be adopted also when CCD's are used.

It is useful to remember that nowadays MTI's, with 40-60 dB IF are implemented using digital devices in a transversal FIR configuration thus needing high resolution, high speed A/D converters (7-10 bits/200nsec).

In the early sixties MTI's were implemented analogically using quartz delay lines whose small dynamic range compelled the designer to cascade many stages with proper filters.

The same must be done with CCD's: dynamic range lower limits due to the device thermal noise, clock residues and residual transients, normally are declared by the factory;but upper limits due to non linearity are not known and they are the most important factor influencing the usable dynamic range in MTI's.

Abrupt non linearities has been studied (1) (2) to evaluate the best possible radar performances; the same has been done for an input hard limiter used to have clutter residues at noise level whichever the precessor.

These non linearities cause a spreading of clutter spectrum thus limiting the clutter attenuation. Recent papers have analyzed the influence of input quantization noise on the possible cancellation (3).

In this paper the influence of a typical CCD non linearity on the maximum allowable dynamic range is examined.

The parameter commonly used to describe the performance of an MTI is the Improvement Factor (IF) which is defined as the ratio between signal to clutter ratio at the output and the same ratio at the input of the MTI. The signal can have each possible speed so it is common to average signal

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power on all frequencies. If G is the averaged signal gain of the MTI IF is defined by the following equality

$$(1) \qquad IF = \frac{S/C|_o}{S/C|_i} = G \cdot CA$$

CA is the clutter attenuation (ratio between output and input clutter power). It is easy to realize that, in the above hypothesis, IF and CA represent a measure of the same characteristic. Later the CA will be used to estimate the performance of the MTI's; it is worth remind that it is useful to have an output clutter power smaller than noise power. Only single and double cancellers will be analyzed, but many cancellers can be examined using this tecnique. Moreover it can be observed that the implementation which uses cascaded cells allows a controll of dynamic ranges impossible for FIR implementation.

2 - Clutter Attenuation for a single and a double canceller

A single canceller, sketched in fig.1, makes a difference between the present signal and that one sweep delayed. The real delay line T (a CCD delay line) is thought a non linear device y=g(x) followed by a linear delay line. If input clutter is supposed gaussian in shape it is known its autocorrelation function

$$(2) \qquad R_{xx}(\tau) = \sigma^2 e^{-2\pi\sigma_c^2\tau^2} = \sigma^2\rho(\tau) \longleftrightarrow G_x(f) = \frac{\sigma^2}{\sigma_c\sqrt{2\pi}} e^{-\frac{f^2}{2\sigma_c^2}}$$

The output of the single canceller ($\varepsilon(t)$) has the autocorrelation function

$$(3) \qquad \begin{aligned} R_{\varepsilon\varepsilon}(\tau) &= \overline{\varepsilon(t)\cdot\varepsilon(t+\tau)} = \overline{[x(t)-y(t-T)][x(t+\tau)-y(t-T+\tau)]} \\ &= R_{xx}(\tau) + R_{yy}(\tau) - R_{yx}(\tau+T) - R_{xy}(\tau-T) \end{aligned}$$

So the CA for a single canceller $[CA^1]$ is

$$(4) \qquad CA^1 = R_{xx}(0)/R_{\varepsilon\varepsilon}(0) = \frac{1}{1 + \frac{R_{yy}(0)}{R_{xx}(0)} - \frac{R_{yx}(\tau)+R_{xy}(-\tau)}{R_{xx}(0)}}$$

If the delay line is linear $[y(t)=x(t)]$ then

$$(5) \qquad CA_L^1 = \frac{\sigma^2}{2\sigma^2 - 2\sigma^2\rho(T)} = \frac{1}{2[1-\rho(T)]}$$

as it is well known. A double canceller can be implemented cascading two single stages. The non linearity of the second stage increases the over all non-linearity effect only by a very small amount; the former stage handles a non cancelled clutter (a higher level), while the latter a cancelled one. Owing to the spectrum spreading due to the non linearity the filtering is small efficient. The output of the second canceller is

$$(6) \qquad z(t) = \varepsilon(t) - \varepsilon(t-T)$$

so

$$(7) \qquad R_{zz}(0) = 2R_{\varepsilon\varepsilon}(0) - 2R_{\varepsilon\varepsilon}(T)$$

and using eq. (3)

$$R_{zz}(0) = 2\left\{R_{xx}(0) + R_{yy}(0) - R_{yx}(T) - R_{xy}(-T) - R_{xx}(T) - R_{yy}(T) + R_{yx}(2T) + R_{xy}(0)\right\} \tag{8}$$

and it is easy to evaluate the CA^2 (clutter attenuation for a double canceller) which for linearity becomes

$$CA_L^2 = \frac{1}{2\left[3 - 4\rho(T) + \rho(2T)\right]} \tag{9}$$

3 - Non linearity analysis

The CCD delay lines non linearity is usually evaluated in two different ways: the former uses the second and third armonic distorsion coefficients (Reticon R5102/03); the latter uses the differential gain used in TV specifications. In the former case the non linearity, described by the general formula

$$g(x) = a_1 x + a_2 x^2 + a_3 x^3 + \cdots\cdots a_n x^n \tag{10}$$

where a_i's are constant (quasi-linearity), is completely described by a_1, a_2, a_3. Knowing the second and third armonic for a fixed input level it is easy to calculate [4] a_2 and a_3 (later a_1 will be considered always 1)

$$\begin{cases} a_2 \simeq 2\left(\dfrac{A_2}{A_1}\right)\dfrac{1}{E} \\ \\ a_3 \simeq 4\left(\dfrac{A_3}{A_1}\right)\dfrac{1}{E^2} \end{cases} \tag{11}$$

where $A_{2/A1}$ and $A_{3/A1}$ are the ratios of second and third armonic and E is the peak amplitude of the test signal; for the case $A_{2/A1}\big|_{dB}=43$ dB and $A_{3/A1}\big|_{dB}=50$ dB it is obtained

$$\begin{cases} a_2 = 0.025 \\ \\ a_3 = 0.037 \end{cases} \tag{12}$$

Such a description is non useful for a radar; each non linearity with its clipping effect is important. After an experimental analysis curves of differential gain (fig.2) have been obtained. So the curves of fig.3 and fig.4 [gain and I/O transfer function] have been obtained by a computer simulation for two different differential gains in useful dynamic range (1% $\rightarrow \sim$ .1 dB; 5% $\rightarrow \sim$ .5 dB). The curves of fig.3 have been obtained using a 16th degree polynomial whose even coefficients (odd are zero) are reported in Table 1. With the needed integration the a_i [s.(10)] have been evaluated (s. fig.4); it is obvious that the simmetry of curves determine the even coefficients to be null.

4 - CA evaluation in non linearity region

To evaluate the effects of non linearity on the clutter spectra the method proposed in [4] [5] to evaluate the autocorrelation for odd non linearities has been extended to even non linearities. So also cross correlation [eqs (3) and (8)] can be evaluated. As a point of fact using the characteristic function method

(13)
$$x = x(t) \longleftrightarrow F(u) = \int_{-\infty}^{\infty} x\, e^{-jux}\, dx$$
$$y = g[x(t-T)] = g(x_+) \longleftrightarrow G(q) = \int_{-\infty}^{\infty} g(x_+)\, e^{-jqx}\, dx_+$$

it is possible to write

(14)
$$R_{xy}(\tau) = \lim_{\theta \to \infty} \frac{1}{2T} \int_{-\theta}^{\theta} \left\{ \frac{1}{2\pi} F(u) e^{jux} du \cdot \frac{1}{2\pi} \int_{-\infty}^{\infty} G(q) e^{jqx_+} dq \right\} dt =$$
$$= \frac{1}{4\pi^2} \int_{-\infty}^{\infty} F(u) \int_{-\infty}^{\infty} G(q)\, C(u,q,\tau)\, dq\, du$$

that for a gaussian clutter becomes

(15)
$$C(u,q,\tau) = e^{\left\{ -\frac{1}{2}\sigma^2 u^2 - \frac{1}{2}\sigma^2 q^2 - R(\tau) uq \right\}}$$

After a few passages, with the notation

(16)
$$ju = w_1$$
$$jq = w_2$$

it is obtained

(17)
$$R_{xy}(\tau) = \sum_{k=0}^{\infty} \frac{R^k(\tau)}{k!} h_{KF} \cdot h_{KG}$$

when

(18)
$$h_{KF} = \frac{1}{2\pi j} \int_{-\infty}^{\infty} F(w_1)\, e^{\frac{1}{2}\sigma^2 w_1^2}\, w_1^k\, dw_1$$
$$h_{KG} = \frac{1}{2\pi j} \int_{-\infty}^{\infty} G(w_2)\, e^{\frac{1}{2}\sigma^2 w_2^2}\, w_2^k\, dw_2$$

If F(•)=G(•) then eq. (17) describes the output autocorrelation. So the coefficients h_{KG} must be computed to know $R_{\varepsilon\varepsilon}(0)$ and $R_{zz}(0)$. As g(x) is a sum of finite terms, G(•) itself is the linear combination of the characteristic functions of each addend of the polynomial (10) x^m (m=1÷n).

So

(19)
$$h_{KG} = \sum_{M=1}^{n} a_m h_{Km}$$

where

(20) $$h_{km} = \frac{1}{2\pi j}\int_{-\infty}^{\infty} G_m(W_2)\, e^{\frac{1}{2}\sigma^2 W_2^2}\, W_2^{\,k}\, dW_2$$

and

(21) $$G_m(W_2) = \int_{-\infty}^{\infty} x^m e^{-W_2 x}\, dx = G_{m+}(W_2) + G_{m-}(W_2)$$

Using the simmetry or antisimmetry of each addend it is obtained for m odd

(22) $$h_{km} = \begin{cases} 0 & k \text{ even} \\ 2\,\frac{1}{2\pi j}\int_{+} G_{m+}(W_2)\, e^{\sigma^2 W_2^2/2}\, W_2^{\,k}\, dW_2 & k \text{ odd} \end{cases}$$

for m even

(23) $$h_{km} = \begin{cases} 0 & k \text{ odd} \\ 2\,\frac{1}{2\pi j}\int_{+} G_{m+}(W_2)\, e^{\sigma^2 W_2^2/2}\, W_2^{\,k}\, dW_2 & k \text{ even} \end{cases}$$

Eq. (19) can be thought as a matrix product

(24) $$\tilde{h}_G = \begin{bmatrix} h_{1G} \\ h_{2G} \\ \vdots \\ h_{nG} \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} & & h_{1n} \\ h_{21} & h_{22} & & h_{2n} \\ \vdots & \vdots & & \vdots \\ h_{n1} & h_{n2} & & h_{nn} \end{bmatrix} = H \cdot \tilde{a}$$

So matrix $[H]$ must be evaluated while h_{KF} is obtained from the general case with n=1; for i and K odd eq. (22) can be written

(25) $$h_{km} = \frac{\Gamma(1+m)}{\left[\frac{\sigma^2}{2}\right]^{\frac{k-m}{2}} \Gamma\left[1-\frac{k-m}{2}\right]} = \frac{n!\ \sigma^{m-k}}{(\sqrt{2})^{m-k} \cdot \left(\frac{m-k}{2}\right)!}$$

The following recurrence formules can be used

(26) $$\frac{h_{(k+2)m}}{h_{km}} = \frac{m-k}{\sigma^2} \quad ; \quad \frac{h_{1(m+2)}}{h_{1m}} = \sigma^2(m+2)$$

where h_{1m} is obtained from eq. (25) with K=1. For m and k even the correspondent equations are

(27) $$h_{0m} = \frac{m!\ \sigma^m}{\sqrt{2}^{\,m}\ \frac{m}{2}!} \quad ; \quad \frac{h_{(k+2)m}}{h_{km}} = \frac{m-k}{\sigma^2} \quad ; \quad \frac{h_{0(m+2)}}{h_{0m}} = \sigma^2(m+1)$$

So matrix H can be evaluated allowing the computing of vectore $\tilde{h}_G = [h_{1G}, h_{2G} \ldots h_{nG}]$. So the addends needed to compute formula (4) can be evaluated

$$R_{yy}(0) = \sum_{k=0}^{\infty} \frac{\sigma^{2k}}{k!} h_{kG}^{2}$$

(28)

$$R_{yx}(\tau) = R_{xy}(-\tau) = \sigma^2 h_{11}^2 \rho(\tau)$$

thus obtaining

(29) $$CA^1 = \frac{1}{1 + \sum_{k=0}^{\infty} \frac{h_{0k}^2}{k!} \sigma^{2(k-1)} - 2 h_{11} \rho(\tau)}$$

for the single canceller and

(30) $$CA^2 = \frac{1}{2\left\{1 + h_{11} + \sum_{k=0}^{\infty} \frac{h_{kG}^2}{k!} \sigma^{2(k-1)} - (1+2h_{11})\rho(T) - \sum_{k=0}^{\infty} \frac{h_{kG}^2}{k!} \sigma^{2(k-1)} \rho^{k}(T) + h_{11}\rho(2T)\right\}}$$

for the double canceller. The behaviours of CA^1 and CA^2 are sketched in fig. 5 for two different ρ (T) values [.99,.999].
For the double canceller are shown the values obtained for the non linearity described by eq. (12) and table 1 for differential gain .1 dB and .5 dB. There is a small CA gain due to the better non linearity which can not be appreciated for a single canceller. The fact that CA^2 for .1 dB decreases sharply before CA^2 for .5 dB is due to the non linearity itself (as simulated).

5 - Experimental results

The Reticon delay line 5103 has bee used in order to implement the double canceller. Input sampling rate is 250 nsec, as the radar is thought to use 500 nsec pulses and two samples for pulse.

The most important constraints in the design are:

1) inefficiency compensation
2) system design to obtain linearity of both cancellers
3) fixed echoes cancellation
4) clock residues reduction

A solution is proposed to reduce the incomplete charge transfer (inefficiency) with the short-circuited lumped-parameter delay line (delay is half of sampling period). That delay line is connected to CCD input in parallel form (s.fig.6); the generator impedance matches the line one.

The experimental results demonstrated the validity of proposed solution. It has been shown that in order to satisfy the system requirements, referred to rain clutter level (with the mean speed-related to doppler frequency) and ground and sea clutter, the aplification (~20dB) the first canceller output signal is essential CCD input d.c. voltage adjustements and sampling and holding circuits have allowed to obtain optimum fixed echoes cancellation and clock residues reduction (s. fig.7-8). The input signal of double canceller (fig.7) is an optimum-speed target with -40,-50,-60 dB level related to $1V_{pp}$ (maximum input signal). Details of double canceller output are shown in fig.7, so the clock residues are below -60 dB level.

For the MTI described above, the cancellation residues are as shown in fig.8.

6 - Conclusions

The CCD non linearities are not very hard constraints to MTI filters design.

The MTI filter, which has been described, provides the clutter attenuation and the fixed clutter residues to a convenient level for normal military and civil radar applications.

The time and temperature stability appear to guarantee the system requirements.

CCD shift register allows the accurate delay control, typical of digital techniques, working on analogue input, with the obvious advantages of lower cost, lower power consumption, increased reliability, less size of whole filter than conventional digital approach (1 to 5÷7).

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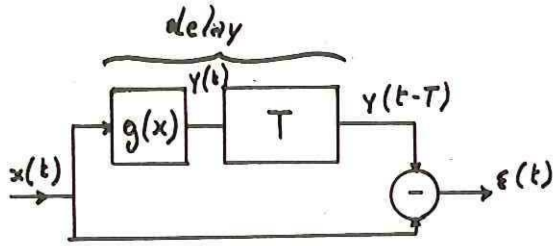


Fig.1 Single canceller block diagram

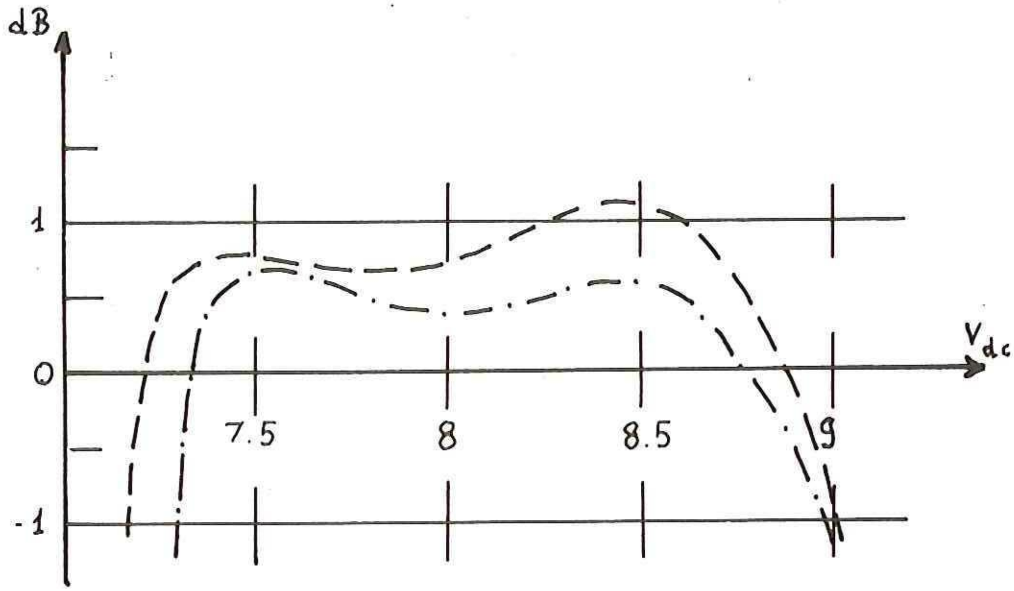


Fig.2 Differential gain versus d.c. level (P5103)

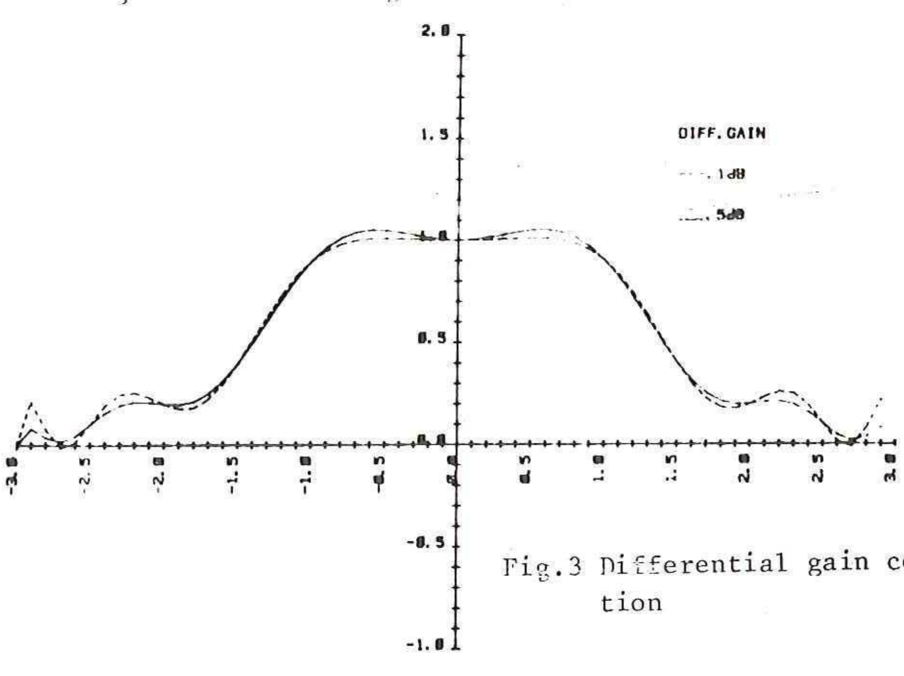


Fig.3 Differential gain computer simulation

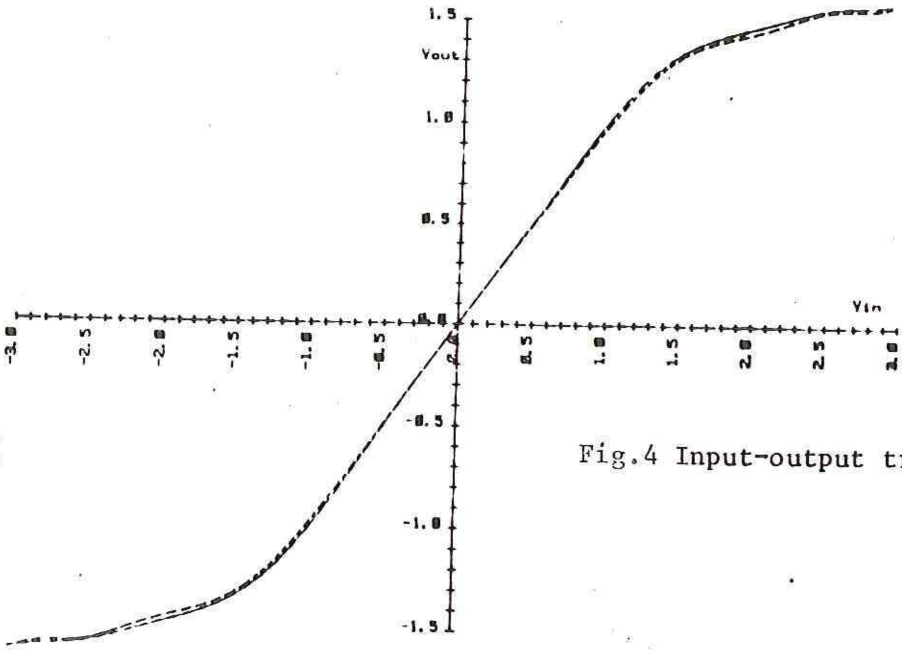


Fig.4 Input-output transfer function

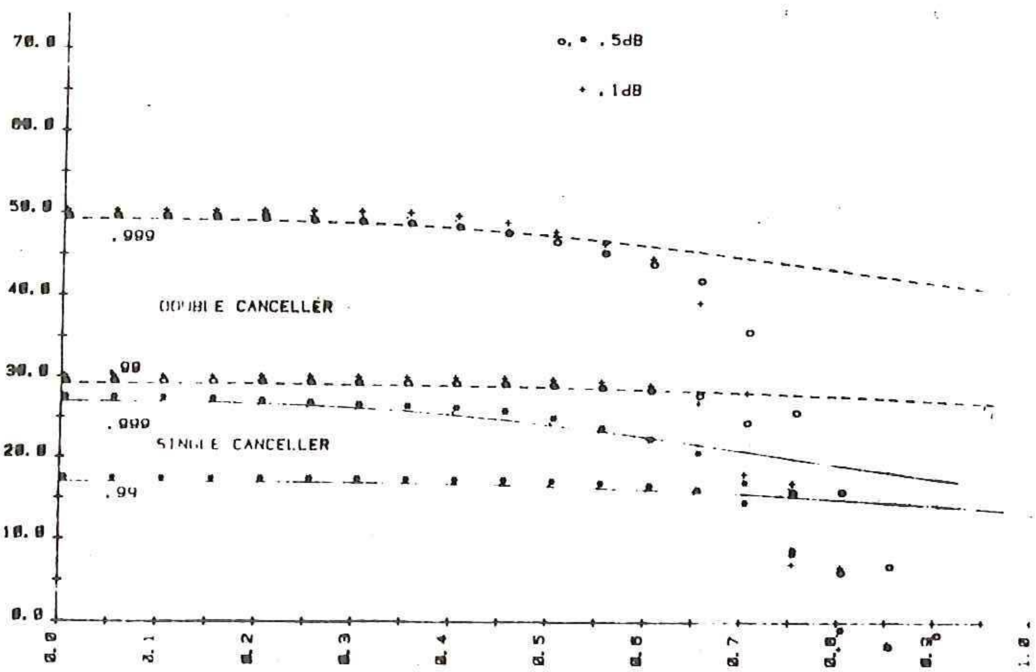


Fig.5 Clutter attenuation versus :

--- non linearity by (12)

" " " Tab.1- .5 dB

" " " " " .1 dB

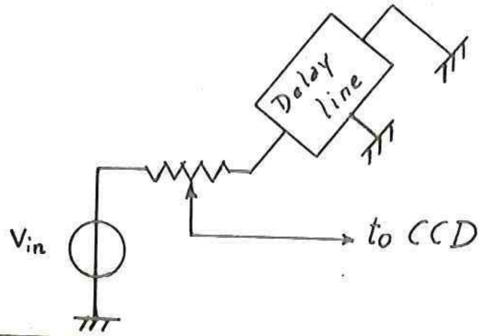
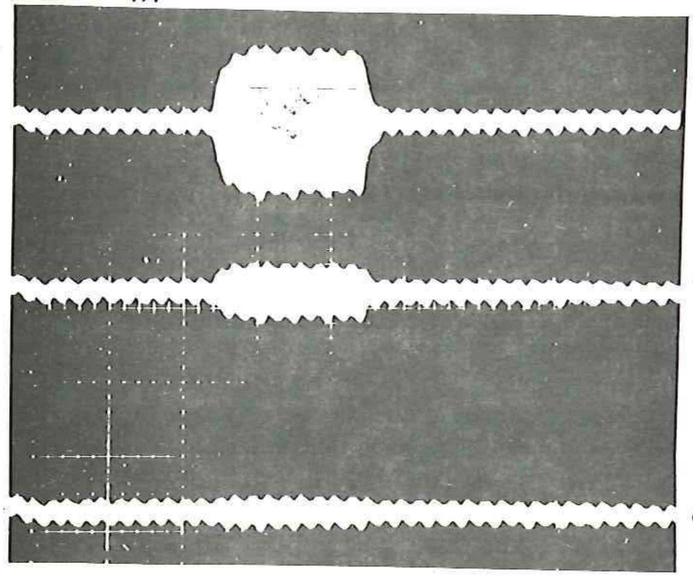


Fig.6 Solution proposed to reduce the inefficiency

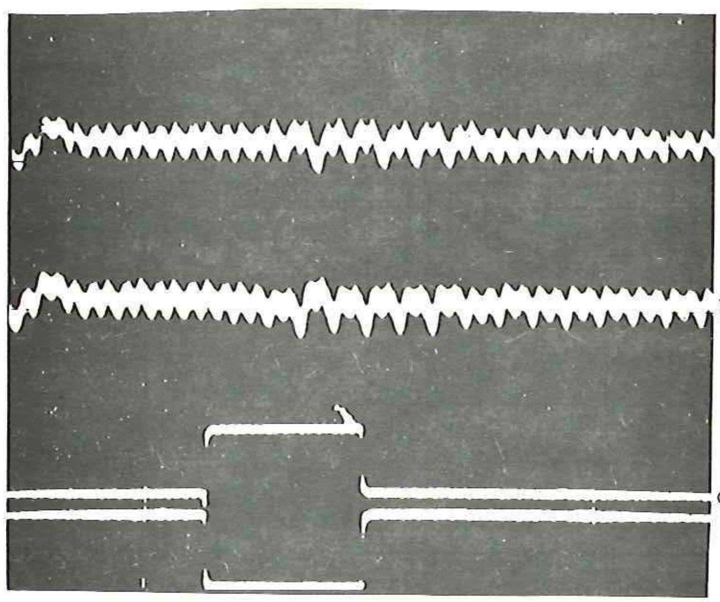


a) -40 dB (below $1V_{pp}$)

b) -50 dB

c) -60 dB

Fig.7 Output of double canceller to optimum speed target



a) positive input signal (.5V)

b) negative input signal (- .5V)

c) input signal

Fig.8 Cancellation residues to fixed echoes