

AN ACCURATE TIME-DOMAIN TECHNIQUE FOR DIRECT MEASUREMENTS OF THE EFFECTIVE TAP WEIGHTS IN CCD FILTERS

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ABSTRACT

A technique is introduced for accurate, direct measurements of the effective tap weights in transversal filters. Principles and obtainable results are discussed. An apparatus based on a multichannel analyzer is described and some representative experimental results, obtained on low pass split-electrode CCD filters, are discussed.

1. INTRODUCTION

The actual performance of CCD transversal filters is limited by the accuracy that can be achieved in the tap weights [1]. In split-electrode filters, a good experimental knowledge of the tap weight errors is required, in order to ascertain their relevance, the various possible causes and to individuate directions for technological improvement. The deviations of the effective weights with respect to the calculated values should therefore be measured down to very low levels, of the order of 0.1% of the maximum tap weight. The interest for such measurements increases in the case of programmable filters, where an accurate tap weight measurement technique is required to allow optimum setting of the weighting factors.

The available signal-to-noise ratio (S/N) at the output of CCD filters is quite low in comparison with the required resolution. Thus, suitable techniques must be used for enhancing the S/N ratio, of course at the expense of longer measurement times. Measurements are usually performed by means of spectrum analyzers (SA), that measure only the modulus $|G(f)|$ of the actual transfer function $G(f)$ of the filter. The required improvement in S/N ratio is obtained in this case by narrowing the analyzer's bandwidth. The deviations of the measured $|G(f)|$ from the designed transfer function module $|H(f)|$ are then analyzed on the basis of theoretical treatments, based on various models of tap weight errors. However, the information which can thus be obtained on the errors is quite indirect and not detailed. More complete information can be obtained by measuring also phase versus frequency. However, practical difficulties and inaccuracies in phase measurements are not to be overlooked, together with error propagation effects in the subsequent lengthy computation of the impulse response.

2. THE DIRECT APPROACH

Let us denote by g_k the measured tap weights and by h_k the designed values. The requirement of measuring every g_k directly leads to consider time-domain measurements of the impulse response $g(t)$ of the filter, that is, of the output corresponding to the injection of a single charge packet. The output waveform $g(t)$, if signal is detected by a charge integrator, is a sequence of rectangular pulses: the k -th pulse after the injection corresponds to the charge packet being under the k -th sensing electrode and its amplitude is proportional to the k -th effective weight.

A simple oscillographic recording is evidently not suitable, because of the insufficient S/N at the CCD output. It is possible, however, to obtain the required S/N enhancement by means of time domain techniques. This is better clarified by recalling how the CCD noise is seen in the time domain waveform $g(t)$; that is

a) within any single pulse in $g(t)$, noise originating from the output circuitry (detection noise) is seen superimposed on the top level;

b) from one repetition to the other of $g(t)$, the top level of any k -th pulse is not constant, but fluctuates (injection noise, transfer noise, etc.);

c) if the output circuitry includes facilities for resetting the integrator, the baseline level before the pulses will also fluctuate (reset noise).

The effect of a) can be reduced by suitable filtering of the pulses prior to the amplitude measurement. In particular, a further gated integration is a nearly optimum procedure in many cases [2].

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The effect of b) can be reduced by averaging over M repetitions the measurement of every k -th pulse; with uncorrelated fluctuations in the various repetitions the S/N is improved by a factor \sqrt{M} .

The effect of c) could also be reduced by averaging, but it is most effectively eliminated by using correlated double sampling (CDS). CDS has the additional advantage of eliminating also possible systematic reset errors due to poor characteristics of the circuitry.

The results that can be obtained with instrumentation based on these principles are illustrated by a typical example of measurement, performed with the apparatus described in the following section. The device is an n -channel 63 taps low-pass split-electrode CCD filter operated in the $1\frac{1}{2}$ phase mode: the d.c. voltage on the sensing gates is reset in each clock cycle. The filter is designed to give an equiripple transfer function with 0.05 dB ripple in the pass-band $0 \div 0.125 f_c$ and 63 dB attenuation in the stop-band $0.175 f_c \div 0.5 f_c$, where f_c is the sampling frequency. The calculated weights have been quantized to the discrete values allowed by the pattern generator used in photomask making, and the expected stop-band attenuation is correspondingly reduced to 54 dB.

Fig. 1 is a photograph of the oscillographic display of the apparatus: the top levels of the pulses and the reset levels before them, separately measured and averaged over $M = 10^4$ repetitions, are shown in linear scale. The effective weights are then calculated by performing the digital equivalent of CDS (subtraction of the reset level from the subsequent pulse top level).

Fig. 2 shows, in logarithmic scale, a plot of the measured weights, corrected for the effect of CTI (see section 4). The designed values interpolated by a solid line are also shown for comparison. The consistency of these measurements with those performed with a SA has been

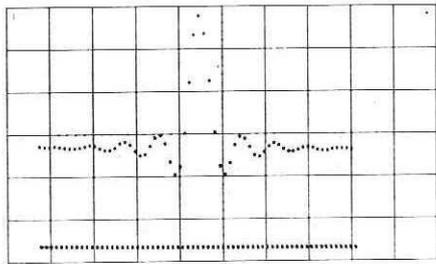


Fig. 1. Measurement of the impulse response (see text) of a low-pass CCD filter, averaged over 10^4 repetition (oscilloscope display of the apparatus).

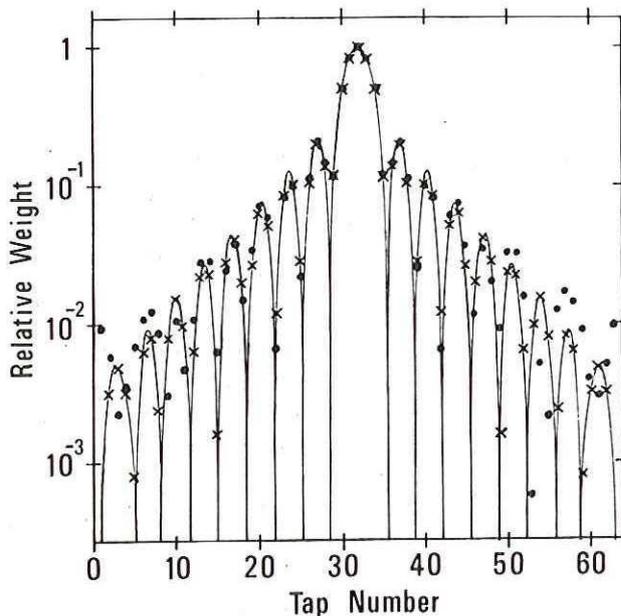


Fig. 2. Semilog plot of the weights of the filter. Dots = measured values, corrected for CTI; crosses = quantized calculated values, interpolated by a solid line.

verified, as depicted in Fig. 3. The modulus of the transfer function $|G(f)|$ has been computed by Fourier transform of the measured waveform $g(t)$ and the result (dots) is seen to be in excellent agreement with the recorded output of the SA (solid line) even in the high attenuation region.

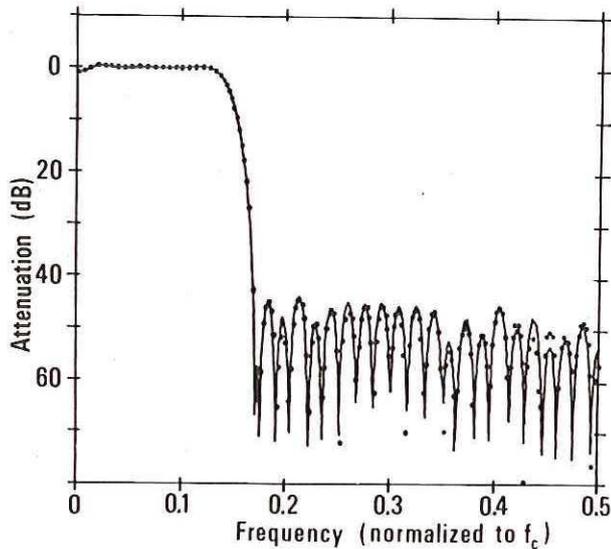


Fig. 3. Modulus of the transfer function $|G(f)|$ of the filter, as computed from the measured impulse response (dots) and measured by a Spectrum Analyzer (solid line).

3. THE MEASUREMENT SET-UP

The present implementation of the measurement system makes wide use of the facilities of an existing instrument, a multichannel pulse analyzer (MCA) Laben Modular 8000, complemented by a limited amount of specially designed external circuitry. The MCA is used in multiscaler mode; it behaves like a scaler associated with a multicell digital memory, so that the n -th cell stores the number of pulses counted by the scaler in the n -th time interval of a sequence of counting intervals.

The essential features of the system will be briefly illustrated by considering its main aspects, that is: a) how a single measurement on a given cell is done; b) how the various repetitions of such a measurement are averaged; c) how the complete set of measurements on the various cells is organized.

a) Single measurement on a cell

The output of the CCD filter, after further amplification by a conventional operational amplifier, is applied to a voltage-controlled-oscillator (VCO) Teledyne Philbrick 4707. The VCO makes a highly linear conversion of the input voltage V to an output frequency f_0 ; the input and output dynamic ranges are $0 \div 10$ V and $0 \div 5$ MHz respectively, the conversion constant is $k_f = df_0/dV = 0.5$ MHz/V.

As outlined in Fig. 4, a digital measurement of this frequency is made by counting the VCO pulses in a time interval with a precisely controlled duration T_G , slightly shorter than half the clock period T_c . This scheme implements at the same time a gated integration over T_G and an analog-to-digital conversion (ADC). For a given CCD cell two measurements are made and stored in separate memory cells, one for the reset level and one for the top level of the charge pulse. The quantization step in the frequency measurement is $q_f = 1/T_G$ and corresponds to quantizing the input voltage of the VCO with a step $q_v = q_f/k_f = 1/k_f T_G$.

b) Averaging measurements on a cell

In the M repetitions of the impulse response, the successive measurements made on a given cell are digitally summed in the corresponding memory cell. If in a single measurement

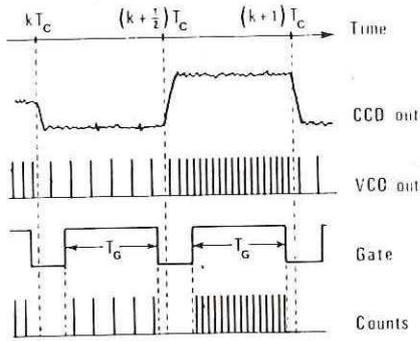


Fig. 4. Outline of a single measurement on the k -th cell of the CCD filter.

g and σ_g are the mean value and the r.m.s. deviation respectively, the corresponding values in the sum g are gM and $\sigma_g \sqrt{M}$, in case of uncorrelated fluctuations.

Errors due to the ADC stage should be kept within limits consistent with the required resolution. The level of the integral non linearities depends mainly on the VCO. It has been verified, by using an ultralinear voltage ramp as a test input to the apparatus, that nonlinearities did not exceed 0.01% of the maximum measured value, at least over a dynamic of 2.5 V. Additional noise is introduced by the ADC quantization, so that $\sigma_g > \sigma_i$, where σ_i is the r.m.s. deviation at the ADC input. This ADC technique has a triangular quantization profile [3,4]. In comparison to the usual rectangular profile, it has slightly higher quantization noise, but is free from systematic deviations in the average. The resolution in the averaging is thus inherently not limited by the number of ADC levels; it can be shown, for instance, that 0.01% resolution is obtained with 50 ADC levels in $M = 10^4$ repetitions.

c) The complete measurement set

The structure of the apparatus is outlined in Fig. 5. After a charge injection at the CCD input, measurements are performed in every clock cycle for a number of cycles greater than the number of CCD cells, that is 100 cycles with 63 cells. In this way the effect of CTI queues is negligible in the last channels, so that the reference level corresponding to zero signal charge can be accurately measured there and subtracted from the data.

The measurement system is synchronized to a high frequency clock. By suitable demultiplication the transfer clock $f_c = 1/T_c$ is generated. Every half period $T_c/2$, a channel advance signal increments the address of the MCA memory cell, and the gate signal with duration T_g (see Fig. 4) is generated. Every 100 transfer cycles the channel address is reset to zero by a scan start signal, and a charge packet is injected in the CCD. With $f_c = 10$ KHz, averaging over $M = 10^4$ repetitions requires 100 seconds. The data stored in the MCA can be continuously monitored on the oscilloscope analog display (see Fig. 1) during the averaging, and printed or recorded on punched tape at the end of it, for further off-line computer processing.

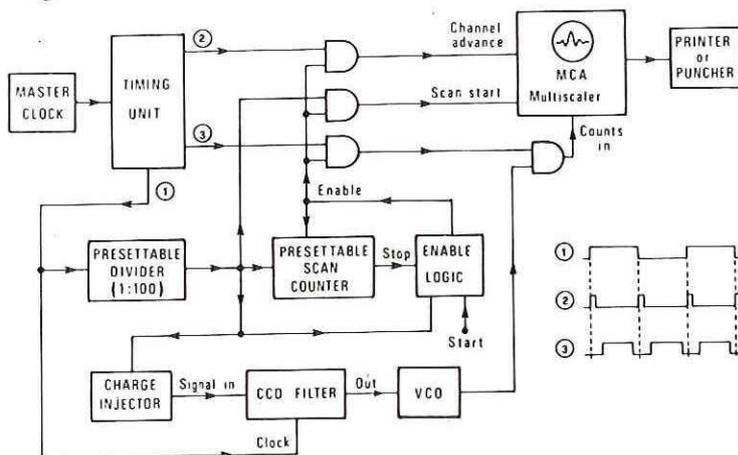


Fig. 5. Block diagram of the apparatus.

4. EXPERIMENTAL RESULTS

Measurements were performed on samples of low-pass CCD filters of the type described in section 2. They were designed with $600 \mu\text{m}$ effective channel width and a nominal $0.25 \mu\text{m}$ quantization step in the photomask, so that causes of errors other than quantization could be better studied. Signal is detected by a differential charge integrator followed by a gain stage. External operational amplifiers were used; the source follower stages required to buffer the sensing nodes, the reset transistors and the integrating capacitors are on-chip.

At the end of the transfer section, a differential floating diffusion amplifier allows to meter the charge transferred. Thus, by using the same apparatus, CTI can be accurately measured from the ratios of the queues to the main charge packet. For the device of Figs. 1-3, the measured CTI was $6 \cdot 10^{-4}$ per cell (4 electrodes). A first-order evaluation of the influence of CTI on the measured weights g_k gives the following expression in terms of the weights g'_k which would be measured in the absence of CTI, (ϵ):

$$g_k = g'_k + k \epsilon (g'_{k-1} - g'_k)$$

In fact, a correction computed by means of these equations removes a slight asymmetry of the experimental data with respect to the centre tap.

The errors in the corrected weights g'_k are individuated by comparing the g'_k with the quantized calculated weights h_k , scaled by a suitable factor Λ and shifted by a translation b (mainly due to mask misalignment). Errors are defined as $e_k = g'_k - (\Lambda h_k + b)$ and Λ and b are determined by the condition of minimizing the mean square error. It is worth noticing that this condition minimizes also the average quadratic difference between ideal and effective transfer function in the frequency domain, and corresponds to have a zero value of the zero shift cross-correlation between errors e_k and ideal weights h_k . The total errors ($e_k + b$) normalized to the maximum weight Λh_{max} are considered.

Fig. 6 reports the results obtained for two devices differing for the technology used to determine the split. In the first one, the two parts of the split electrode are defined by the active area mask in a Planox process, and the p^+ channel-stop island below the split does not allow any equilibration of surface potential between them. In the second one, the areas of the two parts of the split electrode critically depend on the alignment of the first poly-Si mask, where the split is defined, on the active area mask, which defines channel width. In this case, a n^+ connecting diffusion is automatically created in the position of the split and surface potential equilibration along the gate is allowed. As expected, the average error is almost negligible in the first device ($b/\Lambda h_{\text{max}} = 1.4 \cdot 10^{-3}$), while it is remarkable in the second one ($b/\Lambda h_{\text{max}} = 6.6 \cdot 10^{-3}$).

The behaviour of the two devices is quite different also from the standpoint of the autocorrelation function $K_{ee}(n)$ of the deviations e_k . As the number of weights is limited, the

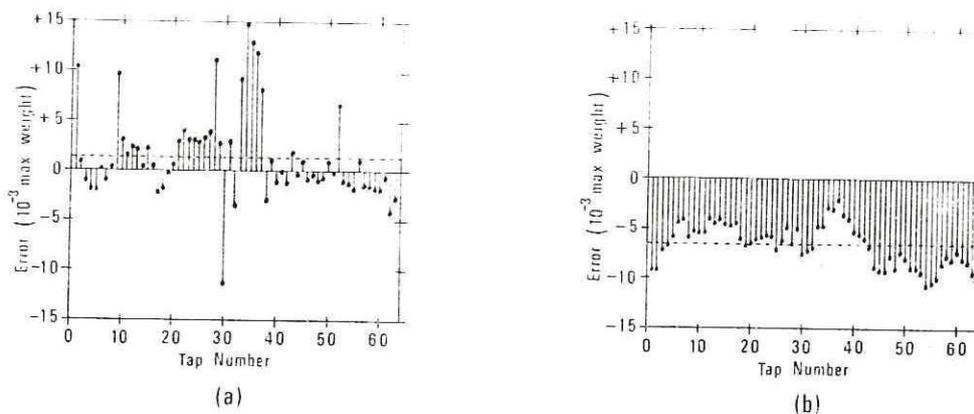


Fig. 6. Errors in the effective weights (corrected for CTI) in the filter samples described in the text: a) with p^+ channel-stop island b) with n^+ connecting diffusion. Dotted line = average error.

evaluation of $K_{ee}(n)$ is subject to considerable errors, that increase as n is increased [5]. $K_{ee}(n)$ was therefore evaluated up to $n = 10$; the normalised plots $K_{ee}(n)/K_{ee}(0) = K_{ee}(n)/e^2$ are reported in Fig. 7. Though measurements on a very large number of samples are required in order to draw more definite conclusions, there are indications of the existence of a wider correlation of the errors in devices of the second type.

Another difference is the sensitivity of the devices of the first type to the clock waveform. The influence of the slope of the trailing edge of the clock pulses can be accurately monitored; the results support an interpretation in terms of effects causing uneven charge transfer, surface potential equilibration along the gate being inhibited by the p^+ channel-stop island [6].

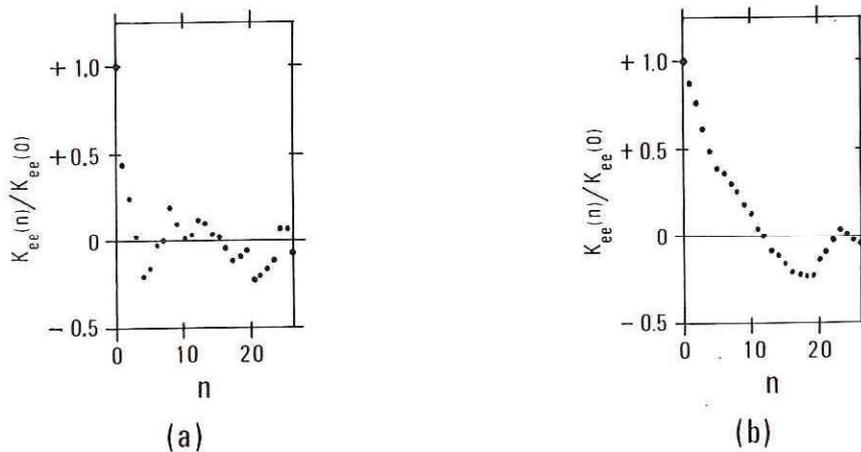


Fig. 7. Autocorrelation functions of the errors (referred to the average) for the samples of Fig. 6.

5. CONCLUSIONS

Accuracy and resolution of the described experimental technique together with its easy and fast operation make it very suitable for systematic analysis of fixed and programmable weights CCD filters.

Improvements are envisaged in order to enhance versatility and automatization of the measurements set up.

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