

CCD ADAPTIVE FILTERING TECHNIQUES WITH APPLICATIONS TO COMMUNICATION SYSTEMS

Vito Cappellini and Enrico Del Re

Istituto di Elettronica, Facoltà di Ingegneria

Florence, Italy

ABSTRACT

Some CCD variable or adaptive filtering techniques are described, in particular to perform multiple filtering, band-pass analysis, multifrequency-tone receiving and adaptive equalization. A special algorithm is presented in order to compensate for the charge transfer loss which is typical of CCD devices. A structure is in detail described, based on the use of a CCD tapped-delay-line and a microprocessor evaluating the filter coefficients and driving a set of MDAC converters. The implementation of a CCD multifrequency-tone receiver using this structure is shown and experimental results are reported.

INTRODUCTION

Variable or adaptive filtering techniques are of high interest in communication systems, in particular to perform multiple filtering, band-pass analysis, multifrequency-tone receiving and adaptive equalization. Charge-coupled-device (CCD) filtering structures have been proposed or used up to now for the above applications, mainly to perform fixed filtering operations.

In this paper we present some structures and procedures permitting to perform the required variable or adaptive filterings (in particular multiple filterings), combining the advantage of low-cost high-integration CCD implementation with the variability and adaptivity of the filtering operation. A structure is in detail presented, based on the use of a CCD tapped-delay-line and a microprocessor, evaluating the filter coefficients and driving a set of multiplying digital-to-analog converters (MDAC). Procedures increasing the adaptation speed and the filtering efficiency (compensating for the charge-transfer loss effects) are described. As a typical example, the project and implementation of a CCD multifrequency-tone receiver is presented with experimental results.

SOME CCD VARIABLE OR ADAPTIVE FILTERING STRUCTURES

A first approach to perform variable filtering with CCD structures corresponds to use suitable memories, in which the set of coefficients of the desired filterings are maintained for the necessary

time and switched subsequently to the analog multipliers at different CCD taps. The memories can be of analog type, using for instance precise potentiometers or CCD recirculating memories (in this last case a regenerative procedure must be provided to compensate for the charge transfer loss in CCD delay lines). Digital memories can be also used with the addition of digital-to-analog converters driving the analog multipliers or MDAC circuits (1). A general block-diagram for all the above considered structures is shown in Fig. 1.

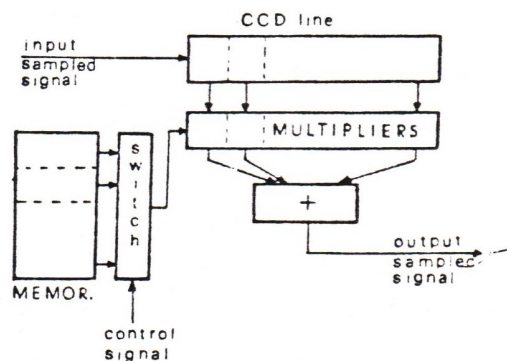
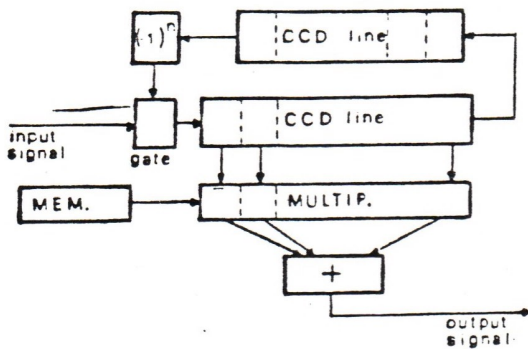


Fig. 1 - General block-diagram of variable CCD filtering structures using memories to store the set of coefficients for the desired filterings.

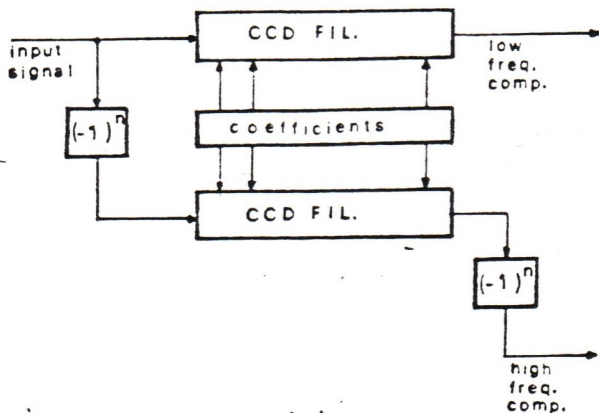
Another attractive solution to perform multiple filtering or band-pass analysis by using a CCD tapped-delay-line corresponds to make use of a single set of filter coefficients and of a suitable frequency shift procedure of the sampled signal spectrum. Indeed it is well known that a frequency shift of the sampled signal spectrum equal to half the sampling frequency can be obtained by multiplying the signal samples s_n by $(-1)^n$ (2) (3).

By assuming a sampling frequency $f_s = 2f_M$, f_M being the maximum signal frequency, the low-pass filtering of the original signal samples gives the signal components 0 to $f_M/2$, while the low-pass filtering of the $(-1)^n s_n$ samples gives the signal components $f_M/2$ to f_M (of course frequency-translated and inverted into the low frequency band 0 to $f_M/2$). The described operation can be implemen-

ted by two different structures. The first one is shown in Fig. 2 (a), where a unique hardware filtering unit processes the original signal and its frequency-translated version (after recirculation through the CCD delay line of sufficient length) in subsequent times. This structure does not allow to operate continuously on the input signal and can be suitable for applications where such an intermittent operation is accepted. One possible example is the multifrequency-tone communication, where at certain time instants the receiver has to determine whether a tone is present either in a low or in a high frequency band. A second solution, suitable for continuous operation on the input signal, is shown in Fig. 2 (b). In this case the same set of filter coefficients is supplied to two hardware filtering units (CCD tapped delay lines): the first one low-pass filters the input signal and gives at the output the low frequency signal components; the second one low-pass filters the $(-1)^n s_n$ and gives at the output, after another multiplication by $(-1)^n$ (frequency translation of $f_M/2$), the high frequency signal components.



(a)



(b)

Fig. 2 - CCD tapped-delay-line structures for multiple filtering, using a single set of filter coefficients: (a) with intermittent operation on the input signal; (b) with continuous operation.

As described for digital filtering structures and implemented in practice constructing hardware band-pass analyzers (3), the above-multiple filtering procedure shown in Fig. 2 can be easily repeated to obtain a band-pass analysis extending over all the signal spectrum.

Another solution to perform CCD variable or adaptive filtering in a flexible and efficient way, a part the cost, is represented by the use of a microprocessor evaluating the filter coefficients and driving a set of MDAC converters (Fig. 3). In this structure the filter transfer function can be easily digitally controlled by the microprocessor through the above coupling circuits and of course adaptation algorithms of the filter coefficients can be implemented more immediately (for instance the clipped Widrow LMS algorithm) (4).

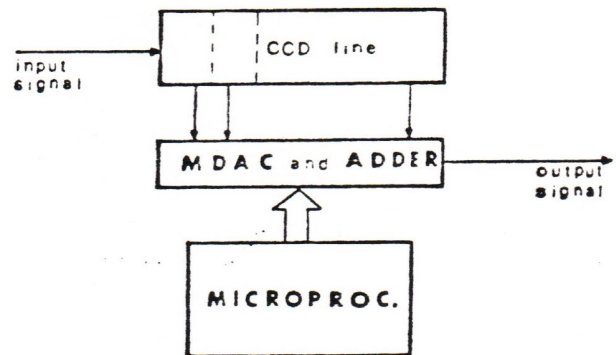


Fig. 3 - Block-diagram of a CCD variable or adaptive filtering structure, using a microprocessor evaluating the filter coefficients and driving a set of MDAC converters.

In all the above CCD structures, to increase the filter efficiency, a compensation of the charge transfer loss is highly desirable. A simple and efficient coefficient modification algorithm was recently defined to this purpose (5): if h_k are the usual filter coefficients, the modified coefficients h'_k are obtained from the nominal ones h_k through the recursive relation

$$h'_k = \frac{1}{1-ke} [h_k - \varepsilon(k-1)h'_{k-1}] \quad (1)$$

where ε is the value of the charge-transfer inefficiency. Fig. 4 shows an example of application of the above technique to the implementation of a low-pass filter (in particular Fig. 4 (b): the reduction of amplitude error from the uncompensated implementation (solid line) to compensated one (dotted line), which is practically reducing to zero).

An example of practical implementation of the last CCD structure with microprocessor, using the

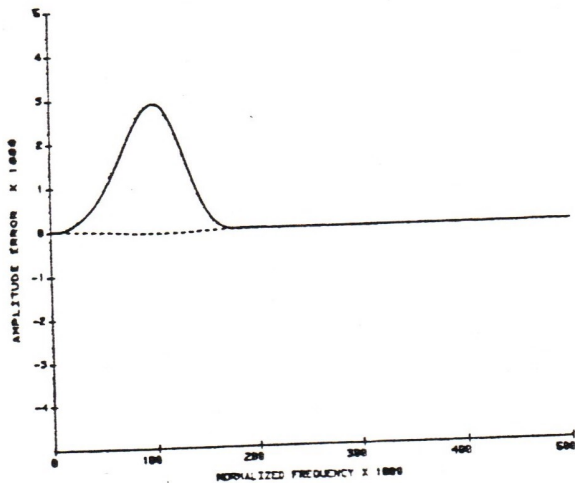
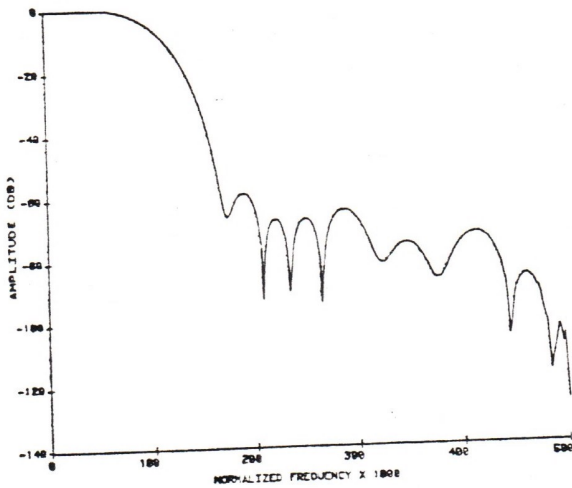


Fig. 4 - Example of CCD low-pass filter with compensation of the charge transfer loss: (a) amplitude response; (b) amplitude error for uncompensated implementation (solid line) and compensated one (dotted line).

above compensation algorithm; is shown in the following to perform a multifrequency-tone receiving.

APPLICATION TO MULTIFREQUENCY-TONE RECEIVING

Envelope Detection Through Quadrature Filters

The envelope of a real signal $x(t)$ can be defined as (6)

$$|c(t)| = |x(t) + j\hat{x}(t)| \quad (2)$$

denoting $\hat{\cdot}$ the Hilbert transform operation. If $x(t)$ is a pure tone of the form

$$x(t) = A \cos(\omega t + \varphi) \quad (3)$$

it follows easily from (2) that

$$|c(t)|^2 = x^2(t) + \hat{x}^2(t) = A^2 \quad (4)$$

Suppose now that we receive a signal $s(t)$ which may contain one of M possible tones having equal amplitude at the angular frequencies ω_i , $i = 1, \dots, M$. In other words let us suppose to receive a signal of the form

$$s(t) = \dots \cos(\omega_i t + \varphi) + n(t) \quad (5)$$

if the tone is present, or of the form $s(t) = n(t)$ if no tone is transmitted, being $n(t)$ the contribution of any type of noise, interference, disturbance, that may be received superimposed on the transmitted signal.

Let us consider now M couples of band-pass quadrature filters $H_i(\omega)$ and $\hat{H}_i(\omega)$, $i = 1, \dots, M$. The filters $H_i(\omega)$ are non overlapping band-pass filters having ω_i as respective center frequencies. The filters $\hat{H}_i(\omega)$ are defined as

$$\hat{H}_i(\omega) = -j \operatorname{sgn} \omega H_i(\omega) \quad i = 1, \dots, M \quad (6)$$

Therefore $H_i(\omega)$ and $\hat{H}_i(\omega)$ represent two band-pass quadrature filters and the output of $\hat{H}_i(\omega)$ is the Hilbert transform of the output of $H_i(\omega)$. Denoting these two outputs $x_i(t)$ and $\hat{x}_i(t)$ respectively and neglecting for the moment the contribution of the term $n(t)$ in (5) we have that either

$$x_i^2(t) + \hat{x}_i^2(t) = A^2 \quad (7)$$

if the i -th tone transmitted was ω_i , or

$$x_i^2(t) + \hat{x}_i^2(t) = 0 \quad (8)$$

if a different tone or no tone was transmitted.

Hence the receiver algorithm can be stated as follows:

a) determine the M envelope squares $|c_i|^2$

$$|c_i|^2 = x_i^2(t) + \hat{x}_i^2(t) \quad i = 1, \dots, M \quad (9)$$

b) choose the maximum of the M envelope squares

$$C = \max_i \{|c_i|^2\} \quad (10)$$

which will coincide with one of the M $|c_i|^2$ as

$$C = |c_j|^2 \quad (11)$$

c) if $C > T$, the j -th tone is detected, if $C < T$ no transmitted tone is assumed, where T is a specified threshold that depends on the allowable probability of false alarm P_F (i.e. the probability of detecting a tone when no tone was actually transmitted).

In the ref. (6) it is shown that, in presence of white Gaussian noise, the threshold T may be chosen to be

$$\frac{T}{N} = 21 \log \left(\frac{M}{F} \right) \quad (12)$$

where N is the noise power in the filter pass-bands (all supposed of equal bandwidth). In ref. (6) the probability of correct detection is also shown in terms of the receiver threshold T and the signal-to-noise ratio of the received signal.

A final consideration is to be done for the implementation of the filters $H_i(\omega)$ and $\hat{H}_i(\omega)$. If we choose to implement them in discrete form, it is more convenient to consider the FIR structure instead of the IIR one. In fact the first one requires to produce the output samples of $x_i(t)$ and $\hat{x}_i(t)$, $i = 1, \dots, M$, only at the time instants when they are necessary for the computation of the envelope squares. On the contrary the IIR structure would require the computation of all the output samples for very filter. Therefore the FIR solution allows the multiplexing of a definite hardware structure to the different filters to be implemented with considerable saving in system complexity and cost.

Implementation of the Receiver

The set of filters employed for frequency discrimination was implemented by means of a CTD tap-delay line. The resulting structure is a programmable filter in which the coefficient weights drive MDAC converters: this configuration makes possible to control the filter transfer function by means of a digital circuitry.

As it is shown in Fig. 5, a microprocessor has the task to control the filter coefficients. The final structure results little more complicated than a programmable filter, due to the fact that two sets of coefficients for the couple of band-pass quadrature filters of a given tone must be switched in a very short time (i.e. during the time interval between two sampling instants). In fact the output values of the two filters must be obtained starting from the same configuration of signal samples within the CTD delay line, as it is clear from eq. (9). Therefore it is necessary to employ the external set of latches so that the microprocessor can provide the coefficients of both filters and, subsequently, only a handshake signal is required for the switching of the two sets of coefficients.

The coefficient weights of the set of filters can further be computed in advance and stored in a suitable table so that the microprocessor can give the coefficients simply by transferring the values from the internal table to the output circuitry. Coefficient weights have been computed by means of the well known window method (3) and modified by the algorithm (1) in order to compensate for the charge transfer loss.

The receiver structure, implemented according to Fig. 5, uses a single 32 cell CTD delay line, time-multiplexed among all the filters. The sampling frequency is 8 KHz and $M = 4$ tones were considered in the tests.

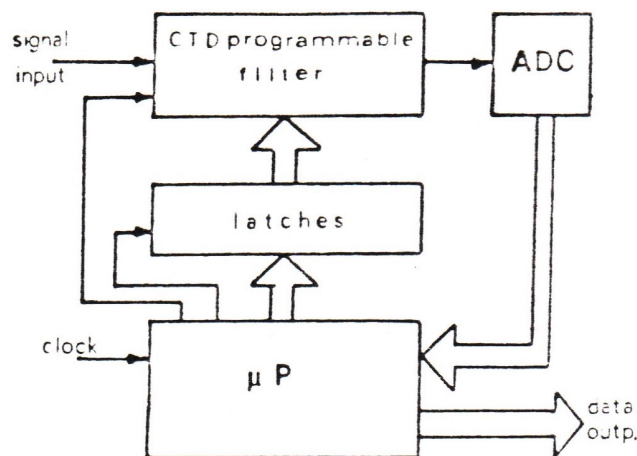


Fig. 5 - Block-diagram of the implemented multifrequency-tone receiver.

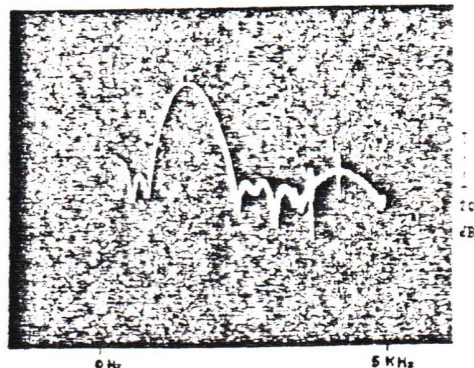


Fig. 6 - Example of an implemented band-pass Hilbert transformer ($f_c = 1.5$ KHz, $W = 0.5$ KHz).

An example of an implemented band-pass Hilbert transformer having a center frequency $f_c = 1.5$ KHz and a bandwidth (at -6 dB) $W = 0.5$ KHz is shown in Fig. 6.

REFERENCES

- (1) C.F.N. Cowan, and J. Mavor, "Analogue Signal Processing Using Charge-Coupled Devices", EU-SIPCO-80 Conf., Lausanne, pp. 669-674, Sept. 1980.
- (2) V. Cappellini, "Digital Filtering with Sampled Signal Spectrum Frequency Shift", PROC. IEEE, vol. 57, pp. 241-242, Feb. 1969.
- (3) V. Cappellini, and P.L. Emiliani, "A Special-Purpose On-Line Processor for Bandpass Analysis", IEEE Trans. Audio Electroac., vol. AU-18, pp. 188-194, June 1970.
- (4) B. Widrow, et al., "Adaptive Noise Cancelling Principles and Applications", PROC. IEEE, vol. 63, pp. 1692-1716, 1975.
- (5) E. Del Re, "Analysis and Compensation of CTI Effects on CCD Transversal Filter Response", IEEE Trans. ASSP, vol. 28, n. 6, pp. 756-768, 1980.
- (6) E. Del Re, "On the Performance Evaluation of a Multifrequency-tone envelope detector", Signal Processing, vol. 3, n. 1, Jan. 1981.