# SIGNAL PROCESSING APPLICATIONS OF CHARGE-COUPLED DEVICES

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## ABSTRACT

This paper reviews the application of charge coupled devices (CCD's) to analog delay, multiplexing, transversal filtering (fixed and variable weighting coefficients) and recursive filtering. It compares the various CCD approaches with regard to performance and cost for selected system applications. Results are given which reflect the current state of analog signal processing technology, and speculation is rendered regarding the competitiveness of CCD's in analog signal processing systems of the future.

## INTRODUCTION

Charge-coupled devices (ref 1,2) (CCD's) are silicon IC's which perform the functions of analog sampled-data delay, and as such, they are uniquely applicable to a wide range of analog signal processing functions. These functions can be divided into five major categories which are given below.

1) Analog Time Delay (ref 3). For time delays of less than one second CCD's are extremely promising. Applications include video frame storage for scan conversion, video line delay for PAL TV, audio delay, and sonar beam forming.

2) Multiplexing/Demultiplexing (ref 4). CCD's are attractive for multiplexing detector arrays due to their inherent low noise and their ability to operate at low temperatures. Demultiplexing has important application in radar range gating.

3) Transversal Filtering: Fixed Weighting Coefficients (ref 5). CCD transversal filters with fixed weighting coefficients offer the greatest cost advantages over digital techniques (ref 6) because this type of filter requires very little additional complexity over the basic CCD, and it replaces a complex digital processor. The CCD transversal filter will become a basic building block for integrated circuit designers. It will offer advantages in cost that make it attractive for many functions where transversal filtering is not now feasible.

4) Transversal Filtering: Electronically Variable Weighting Coefficients. Some analog signal processing functions require transversal filters with electronically variable weighting coefficients. Although several different approaches have been proposed, it is not yet clear which is best and how competitive these approaches are with digital techniques.

5) Recursive Filtering. Recursive filters are more flexible than transversal filters but require feedback amplifiers with very accurately controlled gain. Until such amplifiers can be integrated onto the chip,

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CCD recursive filters may not gain wide acceptance. Indeed, CCD transversal filters with their low cost advantages may be used instead of recursive filters for many filtering functions such as clutter rejection in moving target radar.

The primary motivation behind the development of CCD's for analog signal processing is their low cost. When compared with digital filters, CCD's are expected to be significantly cheaper in many applications for the following reasons (ref 7).

1) Signals are processed in an analog form thereby eliminating the need for A/D and D/A conversion.

2) A single analog filter replaces a multiplicity of digital filters.

3) The process for manufacturing CCD's is simple and compatible with standard MOS manufacturing technology, thereby permitting a high level of integration with other MOS circuitry.

In addition to their lower cost, CCD's offer advantages over digital techniques in smaller size, lighter weight, lower power and higher reliability due to the smaller number of packages.

In a sense, CCD's combine the best features of analog and digital techniques. Like digital filters, CCD's are controlled by a master clock, and as a result they have the same advantages of temperature stability and precise timing.

CCD vs SWD

CCD's are similar to acoustic surface wave devices (SWD's) in many of the functions which they perform. However, they should be viewed as complementary to SWD's and not competitive because their range of applicability does not significantly overlap that of SWD's. Practical, SWD's are limited in achievable time delay ($T_d$) to $T_d \lesssim$ 100 μsec whereas CCD's can achieve up to a second of time delay. SWD's, on the other hand can have bandwidth (W) of up to 1 GHz whereas CCD's are practically limited to $W \gtrsim$ 10 MHz (ref 8). CCD's have been operated at significantly higher frequencies (ref 9,10). However, two factors limit the potential of CCD's in the range $W \gtrsim$ 10 MHz, (1) competition from SWD's and (2) incompatibility with MOS circuitry for clock drivers, output amplifiers, etc.

CCD's are not without their drawbacks when compared with digital circuitry. For one thing they have performance limitations which do not limit digital approaches. These limitations are best discussed in conjunction with the function to be performed (Secs II through VI) but in general they limit bandwidth (W), time delay ($T_d$), time bandwidth product ($T_dW$), dynamic range, linearity within the dynamic range, minimum signal, etc.

Another drawback to CCD's is their limited availability to system designers at the present time. As CCD technology matures, a component family will emerge, and CCD's will become available as more or less standard functional IC components; delay lines, multiplexers, correlators, spectrum analyzers, etc. However, these components are not likely to become widely available for the next few years. In the near future CCD's will be used as custom IC's in systems where the volume is sufficiently high that a custom IC can be justified based on cost reduction and in systems where size, weight, power and reliability are the dominant considerations.

Bucket-brigade devices (ref 11,12) (BBD's) are functionally very similar to CCD's. They are made with standard MOS IC technology, but they have limited performance compared with CCD's. In this article, BBD's are not discussed specifically. However, much of the CCD discussion applies equally to BBD's, and some of the examples presented in this paper, are implemented with BBD technology.

In Secs. II through VI, the functional applications of CCD's are discussed from the point of view of system utilization. For device design and fabrication, reference is made to the published literature including this conference.

## II ANALOG TIME DELAY

The simplest application of CCD's is analog time delay in which the input voltage is sampled (ref 3,13) and a charge representative of that voltage is transferred serially down the CCD to the output. This is shown schematically in Fig. 1 (a) where each box represents one stage or clock period ($T_c$) of delay. For a p-phase device having fractional loss per transfer $\propto$, the fractional loss per stage is $\epsilon=p\propto$. This loss gives rise to dispersion characterized by the transfer function (ref 14)

$$H(z) = \left[\frac{1-\epsilon}{1-\epsilon Z^{-1}}\right]^{M} Z^{-M} \qquad (1)$$

$$\approx \exp\left[-M\epsilon(1-Z^{-1})\right] Z^{-M} \qquad (2)$$

where M is the number of delay stages in the device. Eq. (2) shows that the delay of Fig. 1 (a) behaves like an ideal delay ($Z^{-M}$) in series with a frequency dependent attenuator having magnitude

$$|H(f)| \approx \exp\left[-M\epsilon(1-\cos 2\pi f/f_s)\right] \qquad (3)$$

where $f_s$ is the sampling frequency which is equal to the clock frequency $f_c$ in this case.

The dispersion due to imperfect charge transfer can be reduced by using the multiplexed CCD (MCCD) configuration shown in Fig. 1 (b). In this configuration, q shift registers operate in parallel, each one having M/q delay stages and each one operating at a clock frequency of $f_c=f_s/q$. The dispersion of the MCCD is reduced considerably and is characterized by a transfer function whose magnitude is (ref 15)

$$|H(f)| \approx \exp\left[-\frac{M\epsilon}{2}(1-\cos 2\pi q f/f_s)\right] \qquad (4)$$

Two types of MCCD are commonly used; the series-parallel-series (SPS) structure in which the input steering and output multiplexing are accomplished with a CCD multiplexer and demultiplexer as discussed in Sec III and the phase multiplexed structure in which the input signal is sequentially clocked into shift register 1, 2....q when the 1,z,....,p clock phase is on. With the phase multiplexed structure q=p whereas with the SPS structure, q is arbitrary.

The drawbacks of the MCCD are threefold, 1) The demultiplexing and multiplexing circuitry require added complexity. 2) The leakage and the gain must be very nearly identical in each of the q parallel channels.

Otherwise, a periodic "fixed pattern noise" results. 3) The clock frequency $f_c = f_s/q$ occurs in the middle of the signal band $o < f < f_s/2$ and must be very accurately balanced to eliminate clock feedthrough.

The MCCD is taken to its logical extreme in the sequentially addressed memory (SAM) structure shown in Fig. 1(c). Here q=M, and each of the parallel channels contains only a single stage of delay. Obviously charge transfer loss is no problem in such a configuration, and many effective transfers can be realized without degradation due to charge transfer loss. In fact the SAM structure need not require charge coupling at all. Time delay can be achieved simply by sampling charge on a capacitor and reading that charge at a later time. The SAM structure has been extensively studied for delay applications requiring a large number of clock periods of delay (ref 16,17).

In the SAM structure, however, non-uniform thermal leakage, a common problem in CCD imagers, limits the total time delay to about 10 msec at room temperature (ref 18) as compared to about 1 sec for the serial structure. The very different effects which leakage has on the serial CCD and on the SAM can be seen by comparing Figs. 1(a) and 1(c). In the serial CCD, each charge packet spends an equal amount of time in each storage location, and so the leakage charge that appears at the output is the exact average of the leakage contribution from each cell. Thus the leakage charge as a function of time is uniform and can be eliminated by ac coupling at the output. Because of this, the amount of leakage charge can be quite large (up to $.1n_{max}$) before the dynamic range begins to be degraded. In the SAM, however, the leakage current is different for every storage location, and it is this variation that gives rise to fixed pattern noise. This "noise", which is proportional to the storage time ($T_d$), limits the dynamic range unless it is made small. Fig. 2 shows the approximate ranges of time delay ($T_d$) and signal bandwidth (W) which are practical with the serial CCD (solid lines) and the SAM (dashed lines). For the serial CCD the operating range is limited by $T_d \lesssim 1$ sec, $W \lesssim 10$ MHz and $T_d W \lesssim 10^3$. For the SAM, the operating range is limited by $T_d \lesssim 10$ msec, and $W \lesssim 10$ MHz. This comparison is further considered in another paper in this conference (ref 17).

It should be emphasized that the maximum time delay is limited by thermal leakage which increases exponentially with temperature. Fig. 2 is drawn for room temperature operation and slightly above. If 80°C operation is required, the limitations for the serial CCD and SAM drop by an order of magnitude.

Fig. 2 should not be construed as defining the performance limits of the devices but rather the range in which they are competitive on a cost/performance basis with alternate technologies. Higher frequency operation has been alluded to, and of course, longer delay can be achieved by cooling.

An obvious potential application for CCD delay is as a replacement for the ultrasonic delay line used in PAL TV receivers (ref 20). A 64 μsec delay is required for the chroma signal which has bandwidth of approximately 1 MHz. If the 4.4 MHz local oscillator is used to generate 2.2 MHz clock waveforms for the CCD, then 141 delay stages are required. This application has been considered by a number of companies (ref 11,21) but is not yet being implemented. The ultrasonic delay line may not be fashionable, but it is cheap (around $1.00) and performs adequately. If CCD's are to replace it they will probably have to be integrated with other chroma processing functions; synchronous demodulation, color difference matrixing, etc.

SCAN CONVERSION

Another important delay line application is scan conversion. In infrared imaging systems, for example the detectors scan the scene with a horizontal sweep, and the output of each detector is clocked into a CCD capable of storing an entire line. When the entire frame is stored in the band of CCD delay lines, the CCD's are read out sequentially and the video has the proper format to be displayed on a standard CRT. Two banks of CCD's are used: one is being loaded while the other supplies video information to the CRT (ref 22). CCD's are limited to those scan conversion applications which can be performed in 100 msec or less. This excludes many radar applications which require data retention for a period of seconds, but a number of picture processing applications are feasible within this constraint (ref 23).

Time axis equalization in video playback units (ref 24) appears to be an important potential application of CCD delay because a clock controlled variable delay is required and the only competitive approach is digital delay.

Beam forming in sonar or seismic data processing is accomplished by delaying the outputs from the acoustic sensors by different amounts. This function is now performed digitally and in many cases will continue to be done digitally. However, in cases where a small number of beams are to be formed in a disposable remote sensor where cost, size, weight, and power are all tightly constrained, CCD's are expected to be very important.

## III MULTIPLEXING

A CCD can be used as an analog multiplexer by loading the CCD from a number of parallel inputs, and then shifting the information out in serial form. Similarly, a demultiplexer can be constructed which operates like the multiplexer in reverse. Multiplexing is an important function, and CCD multiplexers will be prominent members of the family of CCD components. Their usefulness as components, however, is somewhat limited by the number of pins required for a device having a large number of outputs. CCD multiplexing is most advantageous where the multiplexing function is integrated with other functions. For example, the SPS delay line utilizes a demultiplexer and multiplexer on the same chip, thereby eliminating the pin-out problem.

CCD multiplexer/demultiplexer devices have two unique limitations. 1) Charge transfer loss is much more serious than it is in a CCD delay line because it introduces cross talk between adjacent input/output channels. 2) Non-uniformities in the gain or offset level of the input/output amplifiers gives rise to fixed pattern noise which limits the dynamic range. The first problem can be solved by inserting one or more isolation elements in between information storage elements (ref 4). Using this scheme cross talk is reduced to a second order or higher effect. The second problem is more bothersome for multiplexing low signal levels. Input techniques have been devised for this application which are independent of variations in the MOS threshold voltage, (ref 25) and these input techniques give stable, uniform inputs with low fixed pattern noise.

CCD multiplexers are especially useful for multiplexing the outputs from detector arrays because they can operate with very low signal levels and can be operated at low temperature.

CCD demultiplexers are important in radar range gating. Fig. 3 shows a schematic of a 10-stage range gate filter (RGF). The video return from a single transmitted pulse is sampled at $t_1$, $t_2$ ... $t_{10}$, and the samples are clocked into the CCD RGF. The clock is stopped when the $t_1$ sample reaches the leftmost CCD stage at which time all the samples are amplified and transferred to their respective outputs $v_1$, $v_2$, ...$V_{10}$. A 50-stage CCD RGF has been built which meets the electronic specifications of a field portable moving target radar.

The CCD range gate filter can be expanded to include doppler processing on the chip. This involves performing the discrete Fourier transform (DFT) on a pulse to pulse basis of each of the voltages $V_1$, $V_2$ ...$V_{10}$ in Fig. 3. The DFT is performed using the chirp z-transform (CZT) which is described in Sec. IV. A chip has been built which performs a 32-point DFT on each of 10 range bins, (ref 26) and a complete doppler processor for several thousand range bins can be constructed by cascading several hundred such chips.

## IV TRANSVERSAL FILTERING: FIXED WEIGHTING COEFFICIENTS

A CCD transversal filter can be made by sampling each node voltage in a CCD delay line, multiplying these samples by weighting coefficients, and summing the results. Such a filter, shown schematically in Fig. 4, performs the sampled-data convolution or correlation between the input signal and the filter impulse response which is determined by the $h_k$, k=1,M, of Fig. 4. Mathematically, the output during the $n^{th}$ clock period is given as a weighted sum of the M previous input samples.

$$v_{out}(nT_c) = \sum_{k=1}^{M} h_k \, v_{in}\left[(n-k)\, T_c\right] \tag{5}$$

Circuitry for performing the sampling weighting and summing functions is integrated with the CCD in a single IC, and different techniques for performing these functions are discussed elsewhere (ref 5, 27).

The CCD fixed weighting coefficient transversal filter is very attractive for a number of reasons. 1) The sampling, weighting, and summing are obtained with very little extra circuitry over the basic CCD shift register. 2) The amount of digital circuitry which it replaces is large. 3) This function is very important for a large number of important applications.

Variable weighting coefficient filters can be devised (ref 14,16,28) as discussed in the next section, and in many applications they are required. However, in all of the realizations proposed to date, (ref 14, 16,18) the inherent simplicity of the fixed weighting coefficient filter is sacrificed to a large extent.

Fixed weighting coefficient filters are factory programmable in the sense that the code or impulse response (the $h_k$ of Fig. 4) is determined by a single photomask which is used in the IC manufacture. Once the basic filter has been designed, a new code mask can be computer generated for as little as \$500.

The idea of using a general purpose correlator together with a ROM containing the impulse response is not practical for fixed-weighting-coefficient filtering. The CCD transversal filter itself is the most compact way of implementing an analog ROM.

CHARGE TRANSFER LOSS

The effect of charge transfer loss on transversal filters is very different from its effect on delay elements. If charge transfer loss were zero, a transversal filter having weighting coefficients $h_k$, k=1,M would have the ideal transfer function

$$H^I(z) - \sum_{k=1}^{M} h_k Z^{-k} \tag{6}$$

When non-zero charge transfer loss is present the transfer function is (ref 14)

$$H(z) = \sum_{k=1}^{M} h_k \left(\frac{1 - \epsilon}{1 - \epsilon Z^{-1}}\right)^k Z^{-k} \tag{7}$$

The effect of charge transfer loss can be determined by replacing Z in the ideal transfer function with

$$Z' = \left(\frac{1 - \epsilon Z^{-1}}{1 - \epsilon}\right) Z \tag{8}$$

$$\approx Z \exp\left[\epsilon(1-Z^{-1})\right] \tag{9}$$

In other words

$$H(Z) = H^I (Z') \tag{10}$$

Dispersion can in principle be inverted if it is predicted accurately. If it is desired to have effective weighting coefficients $h_k$, k=1,M, then the actual coefficients $h'_k$, k=1,∞ which are required, are obtained by iterating the relation (ref 14)

$$h'_k = \frac{\left[h_k - \sum_{j=1}^{k-1} h'_{k-j}\binom{k-1}{j} \epsilon^j (1-\epsilon)^{k-j}\right]}{(1-\epsilon)^k} \tag{11}$$

The principle of dispersion inversion has been demonstrated on an 11-stage Barker coded filter (ref 5), but this technique is not expected to be practical because of the difficulty in accurately predicting $\epsilon$.

MATCHED FILTERING

The matched filtering theorem of statistical communication theory states that when a particular signal is to be detected with optimum signal-to-noise ratio (SNR) in an environment of white, additive noise, then a matched filter should be used, i.e. a filter having impulse response equal to the time inverse of the signal to be detected (ref 29). CCD transversal filters are ideally suited to matched filtering especially in low-data-rate, spread-spectrum communication applications (ref 30).

An example of a matched filter detector for binary chirp signals is shown in Fig. 5. The RF input signal has $T_dW$ product of 100 and is of the form

$$v_i(t) = A \cos(\omega_o t \pm \mu t^2 + \phi) \tag{12}$$

$$-T_d/2 < t < + T_d/2$$

where the up-chirp (+) corresponds to a binary "1" and the down-chirp (-) corresponds to a binary "0". This signal is first mixed to base band so that it chirps from $-\omega/2$ through zero to $+\omega/2$ (or vice versa) and is then filtered in the 200-stage transversal matched filters marked SIN and COS. These filters were implemented using BBD technology, but several comparable CCD filters have also been built. The impulse responses of the two filters have the form

$$h_o(t) = \cos \mu t^2 \tag{13,a}$$

$$g_o(t) = \sin \mu t^2 \tag{13,b}$$

$$-\frac{T_d}{2} < t < + \frac{T_d}{2}$$

and are shown together with their correlation responses in Fig. 6. Both filters are packaged in a single, 28-pin, ceramic DIP.

The effects of charge transfer loss are clearly visible as an attenuation in the impulse responses. For these devices $\epsilon \approx 2 \times 10^{-3}$ making $M\epsilon \approx 0.4$. This amount of loss would be unacceptable for most delay line applications. However, in matched filtering applications, it can be shown that the degradation in filter sensitivity due to non-zero $\epsilon$ is a second order effect in $\epsilon$ (ref 14). In this example, the calculated sensitivity loss was less than .1 dB.

Fig. 7 shows the outputs of the up-chirp and down-chirp channels when the input signal is noncoherent, and chirps alternately up and down. Even when the input signal is masked by noise, the correlation peaks in the output are unambiguous. Measurements of the bit error probability were made from -28 dB to -8 dB input SNR with the result that over this entire range, the measured performance was within .5 dB of the ultimate sensitivity achievable with noncoherent FSK (ref 31).

CCD's will make possible the use of matched filtering techniques in applications where the cost, size,weight and power of digital filtering are not tolerable. In addition, M-ary communication, which requires a large number (M) of signaling waveforms may become feasible due to the simplicity with which arbitrary waveforms can be generated and detected using CCD's.

## RADAR PULSE COMPRESSION

Pulse compression is equivalent to matched filtering on a radar pulse. Acoustic SWD's are used extensively for compression of short radar pulses, but longer pulses continue to be processed digitally, and are amenable to CCD filtering at greatly reduced cost (ref 26). Radar pulses are usually chirp waveforms as in the previous communication example or phase shift keyed (PSK) waveforms: The latter can be processed, after mixing to baseband, in CCD filters which are matched to the pseudorandom pn sequence which modulates the PSK waveform. A typical pseudorandom code is given in Table I and a filter matched to that code is shown in Fig.8.

This filter was used to calculate the signal power, noise power and SNR at the output of the filter as a function of $M\epsilon$. The calculation shown in Fig. 9 illustrates that, although the signal power drops precipitously with increasing $\epsilon$, (due to mismatch between input signal and filter impulse response), the noise power also drops (due to correlation among noise charge packets) with the result that the output SNR is independent of $M\epsilon$ to first order.

CCD's will be most advantageous in compressing pulses having $W \lesssim 10$ MHz (range resolution greater than 15 m). However, by using multiplexing as discussed in Sec. III or by using "stretch" techniques common in synthetic aperture radar imaging, the bandwidth limitations of CCD's can be circumvented, and the use of CCD's for range correlation in high resolution imaging radar can be contemplated.

Weighting coefficient error does not have a significant effect on pulse compression or matched filtering functions. In both cases the input signal is contaminated with so much noise that the additional "noise" due to weighting coefficient error is negligible. In bandpass filtering or spectral analysis however, weighting coefficient error is a crucial limitation on sidelobe level or out-of-band rejection.

BANDPASS FILTERING

A linear phase bandpass filter can be constructed by selecting the impulse response of a transversal filter to be the Fourier transform of the desired frequency characteristic (ref 32). One of the principle advantages of such filters is that their frequency characteristic scales with the clock frequency $f_c$, and by varying $f_c$ the filter can be tuned.

The measured frequency characteristic of a 500-stage CCD bandpass filter is given in Fig. 10. This filter was designed using Hamming weighting to have a worst case sidelobe level of -42 dB. The measured sidelobe level of-38dB may be a result of weighting coefficient error which is estimated to be on the order of 1%.

The spectrum of Fig. 10 was obtained by clocking the device at $f_c$ = 1 MHz, and the device is designed to have a passband at $f_c/4$. Of course a spurious passband at 3 $f_c/4$ results from aliasing and must be suppressed by an anti-aliasing filter. Filters of this type have been clocked at frequencies of up to 4 MHz. The low frequency end is limited by leakage current to $f_c \gtrsim M/T_{d\,max} \approx 500$ Hz at room temperature.

SPECTRAL ANALYSIS

One approach to spectral analysis with CCD's is to construct a bank of bandpass filters. This is very wasteful, however, compared with the chirp z-transform (CZT) approach. The CZT is an algorithm for performing the discrete Fourier transform (DFT)(ref 33). This algorithm has no particular advantages over the conventional Cooley-Tukey FFT algorithm when it is implemented digitally, but it casts the DFT into a form where much of the computation is performed in a fixed weighting coefficient transversal filter which is amenable to CCD implementation. (ref 34,35)

The CZT algorithm can be derived in the following way. Starting with the definition of the DFT

$$X_k = \sum_{n=0}^{N-1} x_n \, e^{-i2\pi nk/N} \tag{14}$$

and using the substitution

$$2nk = n^2 + k^2 - (n-k)^2 \tag{15}$$

the following equation results.

$$X_k = e^{-i\pi k^2/N} \left[ \sum_{n=0}^{N-1} \left( x_n e^{-i\pi n^2/N} \right) e^{i\pi (k-n)^2/N} \right] \tag{16}$$

This equation has been factored to emphasize the three operations which make up the CZT algorithm: 1) Pre-multiplication by a complex chirp waveform, 2) filtering in a convolution filter having a complex chirp impulse response and 3) postmultiplication by a complex chirp waveform.

When only the power density spectrum is required, the postmultiplication by $e^{-\pi k^2/N}$ can be eliminated, and a block diagram of the simplified CZT system is shown in Fig. 11. This system has been implemented with 500-stage CCD filters whose impulse responses resemble those shown in Fig. 6. At the time of this writing, data are not available on the performance of this system, but the performance of an older system implemented with the 200-stage BBD filters is shown in Fig. 12. Here is shown the power density spectrum of a 100 kHz carrier amplitude modulated with a 1 kHz signal.

Spectral analysis via the CCD CZT is expected to be very important in radar doppler processing, speech recognition, target identification, sonar spectral analysis, video bandwidth compression and many other applications requiring electronic spectral analysis.

## V TRANSVERSAL FILTERING: ELECTRONICALLY VARIABLE WEIGHTING COEFFICIENTS

Many important analog signal processing functions require variable weighting coefficient filters, and several approaches have been proposed to meet this need (ref 14, 16, 28). Variable weighting is difficult to achieve in general, and sacrifices much of the inherent simplicity of the fixed weighting coefficient filter. For this reason, the advantages in cost, size, weight, power and reliability of the variable weight CCD filter over digital approaches are not as clear-cut as in the fixed weight case. There are, however, a number of applications where variable weight filters do appear to be cost effective.

For example it is relatively straight forward to design variable weight filters in which the weighting coefficients take on one of two values (ref 36) (1 and 0 or +1 and -1). For this reason, pn sequence filters of the type shown in Fig. 8 can be made electronically variable with a minimum of circuit complexity.

It is convenient to separate variable weighting coefficient filtering applications into two categories. 1) Convolution with an impulse response which is fixed for long periods of time but which must be varied slowly or changed infrequently. 2) Convolution or correlation of two arbitrary waveforms.

One important application in the former category is matched filtering on a waveform which changes intermittently. In this case a large number of codes must be stored in memory and supplied to a variable weight filter on request. If the codes are permanent and number of codes is not excessive, this problem is best solved by building a bank of fixed weight filters. The rationale for this choice becomes obvious when one considers that a fixed weight filter is a very efficient analog ROM.

Another important application in the former category is adaptive equalization in which the dispersion due to a changing transmission medium must be inverted in the receiver or MODEM (ref 37). In this case, the number of codes is not finite, and the solution requires a variable weight filter. Three approaches to this problem have been proposed; 1) Analog-binary serial CCD (ref. 14), 2) Analog-binary CCD SAM (charge sloshing) (ref. 16, 17) and 3) MNOS variable conductance CCD (ref. 28).

In the analog-binary approaches, use is made of the above-mentioned fact that binary weighted filters can be electronically programmed in a straight forward way.

The analog weighting coefficients are digitized to N-bit accuracy, and the analog signal is clocked through N filters whose binary weighting coefficients represent the desired impulse response. If the $h_k$ of Fig. 4 are written in binary form as

$$h_k \approx \sum_{m=1}^{N} h_k^m \; 2^{N-m} \tag{17}$$

then eq(5) for the filter output can be written as

$$v_{out}(nTc) \approx \sum_{m=1}^{N} \left[ \sum_{k=1}^{M} h_k^m \; v_{in}\left[(n-k)\; T_c\right] \right] 2^{N-m} \tag{18}$$

An adaptive filter implemented with the analog-binary serial CCD is shown in Fig. 13. The filtering is performed according to eq(18), and the coefficients are calculated in the usual way. The analog-binary scheme is ideally suited for this application because the weighting coefficients must be calculated and stored digitally anyway.

An alternative to the serial CCD for analog-binary filtering is the SAM (ref 16, 17) (See Sec. II) which is attractive when M is large and charge transfer loss is important.

MNOS transistors have also been proposed as a means of varying filter weighting coefficients (ref 28). The transistor's conductance is programmed by the analog voltage applied to the gate. At the time of this writing this technique has not been demonstrated in the literature and evaluation is premature. However, two serious drawbacks can be cited which jeopardize the cost effectiveness of this approach. 1) The MNOS circuitry introduces added processing complexity and hence added cost. and 2) The off-chip circuitry required to program the MNOS conductances is formidable. (The latter criticism applies to the analog-binary approach as well.)

The correlation of two arbitrary signals can be performed using any of the schemes discussed above. However, the cost effectiveness of these schemes in this application is doubtful compared with digital implementation. The full advantages of CCD's for correlation will be realized only if signals in two CCD shift registers can be multiplied in analog form, on chip, to provide an all-analog, self contained correlator IC. Schemes for doing this appear difficult but not impossible (ref 36).

## VI RECURSIVE FILTERING

Transversal or nonrecursive filters form a special class of sampled-data filters in which no feedback exists. In this sense, recursive filters form a more general class in that both feedback and feedforward can be used to shape the filter characteristic (ref 6).

The impulse response of a transversal filter is finite in time, whereas that of a recursive filter can be of infinite duration. As a result, bandpass filters having sharp skirts can generally be designed with fewer arithmetic operations using recursive filtering than with transversal filtering only. When filters are implemented digitally, the cost is roughly proportional to the number of arithmetic operations and therefore recursive filtering is used extensively in applications such as clutter rejection in MTI radar where sharp filter skirts are desired.

With CCD filtering, the economics of filter design are quite different. Within limits, the number of arithmetic operations is not a dominent consideration because the weighting and summing of a large number of node voltages can be performed easily and cheaply. It is somewhat more difficult, however, to include feedback in a CCD filter, and for this reason, the trade-off between recursive and nonrecursive techniques is weighted in favor of nonrecursive filtering. With CCD's it may be feasible to use a transversal filter having many delay stages rather than a recursive filter having fewer stages.

Recursive filters have been demonstrated for bandpass filtering (ref 38) and radar clutter rejection (ref 39) applications, and in both demonstrations, feedback was performed in off-chip amplifiers. Except in special cases, CCD recursive filters will not find wide acceptance until stable feedback amplifiers can be cheaply integrated on the CCD chip.

## VII CONCLUSIONS

CCD's are almost certain to have a large impact on analog signal processing systems of the future. Much of the CCD technology base which has been developed for optical imaging and digital memory applications is also applicable to analog signal processing, but there are differences in the manner in which analog signal processing technology needs to develop. These unique features are discussed below.

The CCD technology required to begin to make the impact discussed in this paper is here today. The remaining problems are largely system engineering. Compared to the application areas of optical imaging and digital memory, a low level of IC technology and complexity is required to make a major impact. Existing CCD technology needs to be pushed into engineering development of simple systems to determine whether the expected advantages of CCD analog signal processing justifies further device development.

The major advantage for CCD's in analog signal processing is low cost, and consequently the technology must be selected with manufacturability as the foremost consideration. For optical imaging, large CCD chips are generally required which have tight specifications on the level and uniformity of the leakage current. For analog signal processing, the chip size can be chosen for optimum manufacturability and most analog signal processing circuits are relatively insensitive to thermal leakage nonuniformity.

Standard processing is an important factor in low cost for two reasons. 1) Complex processes like MNOS cost more in terms of lower yield and 2) Compatibility with standard MOS circuitry permits a higher level of integration with other electronic functions and a consequent reduction in overall system cost. The ultimate goal should be self contained functional IC's (delay lines, filters, multiplexers, correlators, spectrum analyzers, Hilbert transformer's, etc) which are manufactured with standard high yield MOS processes and which require a minimum of off-chip circuitry (clock drivers, output amplifiers, etc., all integrated).

In the forseeable future, analog signal processing components will be developed for specific system applications as custom IC's and IC designers will utilize CCD building blocks to design integrated subsystems. Effective realization of the full potential of CCD's will require increasing cooperation between IC designers and system designers, both in industry and in government agencies. Equipment manufacturers who do not have IC manufacturing capability will be at a distinct disadvantage, because CCD analog signal processing components will probably not be commercially available except as custom items for a few years.

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## REFERENCES


1. W. S. Boyle and G. E. Smith, "Charge-Coupled Semi-conductor Devices," Bell Syst. Tech. J., 49, pp. 587-593, April 1970.

2. For a review of CCD's see M. F. Tompsett, "Charge Transfer Devices", J. Vac. Sci. & Tech., 9, pp. 1166-1181, July-Aug. 1972.

3. M. F. Tompsett and E. J. Zimany, Jr., "Use of Charge Coupled Devices for Delaying Analog Signals," IEEE J. Solid-State Circuits, SC-8, pp. 151-157, April 1973.

4. T. F. Cheek, Jr., A. F. Tasch, Jr., J. B. Barton, S. P. Emmons, and J. E. Schroeder, "Design and Performance of Charge-Coupled Device Time-Division Analog Multiplexers", Proc., CCD Applications Conf. San Diego, Calif., Sept. 1973, pp. 127-139.

5. D. D. Buss, D. R. Collins, W. H. Bailey, and C. R. Reeves, "Transfersal Filtering Using Charge Transfer Devices," IEEE J. Solid-State Circuits, SC-8, pp. 138-146, April 1973.

6. B. Gold and C. M. Rader, Digital Processing of Signals, McGraw-Hill, 1966.

7. D. D. Buss and W. H. Bailey, "Applications of Charge Transfer Devices to Communication," Proc. CCD Applications Conf., San Diego, Calif., Sept. 1973, pp. 83-93.

8. For a more detailed comparison between CCD's and SWD's see L. T. Claiborne and D. D. Buss, "Analog Signal Processing Using Surface Wave and Charge-Coupled Devices", International Microwave Symposium, June 12-14, 1974, Georgia Institute of Technology, Atlanta, Georgia.

9. L.J.M. Esser, M. G. Collet, and J. G. vanSanten, "The Peristaltic Charge Coupled Device", 1973 International Electron Devices Meeting, Tech. Dig., pp. 17-20.

10. M.J.J. Theunissen & L.J.M. Esser, "PCCD Technology and Performance", This conference.

11. F.L.J. Sangster, "Integrated MOS and Bipolar Analog Delay Lines Using Bucket-Brigade Capacitor Storage," 1970 IEEE Solid-State Circuits Conf., Dig. Tech. Papers, pp. 74-75, 185.

12. For a review of BBD's see L. Boonstra and F.L.J. Sangster, "Progress on Bucket-Brigade Charge-Transfer Devices", 1972 IEEE Solid-State Circuits Conf., Dig. Tech. Papers, pp. 140-141, 228.

13. H. Wallinga, "A Comparison of CCD Analog Input Circuit Characteristics", This conference.

14. D. D. Buss, W. H. Bailey, J. D. Holmes and L. R. Hite, "Charge Transfer Devices: A New Semiconductor Technology Applied to Communication Systems," Submitted to IEEE Trans. on Communication.

15. J. M. Caywood and D. D. Buss, "Frequency Response of a Multiplexed Charge Transfer Delay Line." To be published.

16. J. J. Tiemann, W. E. Engeler, R. D. Baertsch and D. M. Brown, "Intracell Charge-Transfer Structures for Signal Processing", IEEE Trans. Elec. Dev., ED-21, pp. 300-308, May 1974.

17. R. D. Baertsch, W. E. Engeler, J. J. Tiemann and H. S. Goldberg, "Application of Charge Transfer Devices to Signal Processing" This conference.

18. D. D. Buss & W. H. Bailey, "Applications of Charge Transfer Devices to Analog Signal Processing" 1974, IEEE Intercon Technical Papers, Session 9, paper 9/1.

19. M. H. White, D. R. Lampe, F. C. Blaha and I. A. Mack, "Characterization of Surface Channel CCD Image Arrays at Low Light Levels", IEEE J. Solid-State Circuits, SC-9, pp 1-12, Feb. 1974.

20. B. Townsend, PAL Colour Television, Cambridge University Press, 1970.

21. M. Kubo, M. Ashikawa, I. Takemoto and S. Ohba, "A Wideband Low-Noise Analog Delay Line", 1974 IEEE International Solid-State Circuits Conf. Dig. Tech. Papers, pp. 152-153,245.

22. Unpublished.

23. J. B. G. Roberts, "Video Integration with CCD Delay Lines", This conference.

24. W. J. Hannan, J. F. Schanne, and D. J. Waywood, "Automatic Correction of Timing Errors in Magnetic Tape Recorders," IEEE Trans. Military Electronics, MIL-9, pp. 246-254, July/Oct. 1965.

25. S. P. Emmons and D. D. Buss, "The Performance of CCD's in Signal Processing at Low Signal Levels," Proc. CCD Applications Conf., San Diego, Calif., Sept. 1973, pp. 189-205.

26. M. M. Whatley, C. J. Benning, W. H. Bailey and D. D. Buss, "Charge-Coupled Device Applications in Radar Signal Processing", Proc. of the IEEE 1974 National Aerospace and Electronics Conf., May 13-15, 1974, Atlanta, Georgia.

27. D. J. McLennan, J. Mavor, G. F. Vanstone & D. J. Windle, "Transversal Filtering Using CCD's," This conference.

28. M. H. White, D. R. Lampe & J. L. Fagan, "CCD MNOS Devices for Programmable Analog Signal Processing and Digital Nonvilatile Memory", 1973 International Electron Devices Meeting, Tech. Dig., pp. 130-133.

29. M. Schwartz, W. R. Bennett & S. Stein, Communication Systems and Techniques, McGraw Hill, 1966, pp. 63-68.

30. D. D. Buss, W. H. Bailey & D. R. Collins, "Spread Spectrum Communication Using Charge Transfer Devices," Proc. of 1973 Symposium on Spread Spectrum Communications, San Diego, Calif., March 1973.

31. M. Schwartz, W. R. Bennett & S. Stein, Op. Cit. pp-19, 81-82, 297-299.

32. D. D. Buss, C. R. Reeves, W. H. Bailey, and D. R. Collins, "Charge Transfer Devices in Frequency Filtering," Proc. 26th Annual Symposium

on Frequency Control, Atlantic City, N.J., June 1972, pp. 171-179.

33. L. R. Rabiner, R. W. Schafer, and C. M. Rader, "The Chirp Z-Transform Algorithm," IEEE Trans. on Audio and Electroacoustics, AU-17, pp. 86-92, June 1969.

34. H. J. Whitehouse, J. M. Speiser and R. W. Means, "High Speed Serial Access Linear Transform Implementations," presented at the All Applications Digital Computer Symposium, Orlando, Florida, 23-25 January 1973. NUC TN 1026.

35. R. W. Means, D. D. Buss, and H. J. Whitehouse, "Real Time Discrete Fourier Transforms Using Charge Transfer Devices," Proc. CCD Applications Conference., San Diego, pp. 127-139, September 1973.

36. C. S. Hartmann, L. T. Claiborne, D. D. Buss and E. J. Staples, "Programmable Transversal Filters Using Surface Wave Devices, Charge Transfer Devices, and Conventional Digital Approaches," Proc. International Specialist Seminar on "Component Performance and Systems Applications of Acoustic Surface Wave Devices" Aviemore, Scotland, Sept. 1973, pp. 102-114.

37. R. W. Lucky, J. Salz & E. J. Weldon, Jr., Principles of Data Communications, McGraw-Hill, 1968, Ch. 6.

38. D. A. Smith, C. M. Puckett and W. J. Butler, "Active Bandpass Filtering with Bucket-Brigade Delay Lines," IEEE J. Solid-State Circuits, SC-7, pp. 421-425, Oct. 1972.

39. J. B. G. Roberts, R. Eames and K. A. Roche, "Moving-Target-Indicator Recursive Radar Filter Using Bucket-Brigade Circuits," Electronics Letters, 9, pp. 89-90, Feb. 1973.

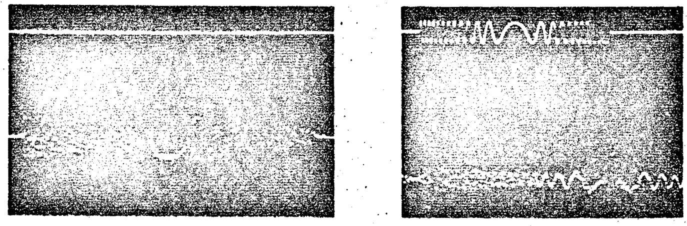
TABLE I

CODE FOR THE 100-BIT PN SEQUENCE FILTER

| | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|
| 1 - 10 | - | + | + | - | - | + | - | + | + | - |
| 11 - 20 | - | - | - | + | + | - | - | + | - | - |
| 21 - 30 | + | - | - | - | + | + | - | + | + | + |
| 31 - 40 | + | + | + | + | - | + | - | - | + | + |
| 41 - 50 | - | + | + | - | + | - | - | + | - | - |
| 51 - 60 | - | - | - | - | - | + | + | - | - | - |
| 61 - 70 | + | + | + | - | + | - | - | - | + | - |
| 71 - 80 | + | + | + | + | + | - | + | + | - | - |
| 81 - 90 | + | - | + | + | + | + | + | - | - | + |
| 91 - 100 | + | - | - | + | + | + | - | - | - | - |

The time signal consists of the elements of this table in the order given. The filter itself is coded with these values in reverse order.

| | $T_D^{Max}$ | $W^{Max}$ | $(T_D W)^{Max}$ |
|---|---|---|---|
| CCD | 1 s | 10 MHz | $10^3$ |
| SWD | 40 $\mu$s | 800 MHz | $10^3$ |

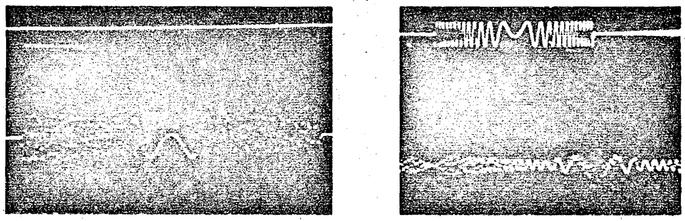


IMPULSE RESPONSE

CORRELATION RESPONSE

$$h(t) = \cos ut^2$$

$$-\frac{T_d}{2} < t < \frac{T_d}{2}$$



IMPULSE RESPONSE

CORRELATION RESPONSE

$$g(t) = \sin ut^2$$

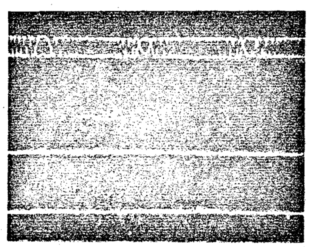
$$-\frac{T_d}{2} < t < \frac{T_d}{2}$$

Figure 6. The Impulse Response and Correlation Response of the Two Filters Required for the System of Figure 5. These filters are each 200 stages long and are both integrated in a single BBD IC.

INPUT

UP-CHIRP OUTPUT

DOWN-CHRIP OUTPUT

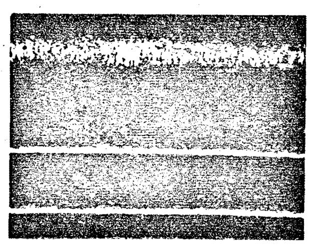


ORTHOGONAL MATCHED FILTER CORRELATION RESPONSE WITHOUT NOISE

INPUT

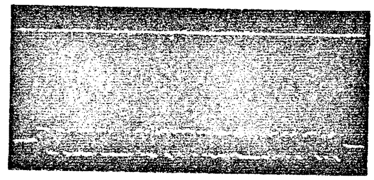
UP-CHIRP OUTPUT

DOWN-CHIRP OUTPUT



ORTHOGONAL MATCHED FILTER CORRELATION RESPONSE WITH 0-dB INPUT SIGNAL-TO-NOISE RATIO

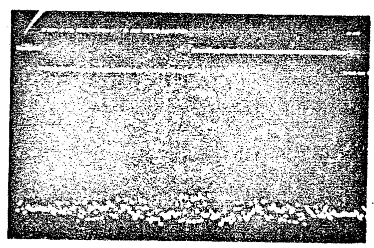
Figure 7. The Outputs of the Receiver of Figure 5 when Noncoherent Up-Chirp and Down-Chirp Signals are Sequentially Applied to the Input. Even when the input is obscured by noise, the output is unambiguous.



INPUT

OUTPUT

IMPULSE RESPONSE



INPUT

OUTPUT

CORRELATION RESPONSE

FIGURE 8. Operation of a 100-Bit, pn Sequence, CCD Filter Matched to a Pseudorandom Code. TOP: Response of the filter to a negative impulse. BOTTOM: Correlation response of the filter. Note the correlation peak in the output waveform.

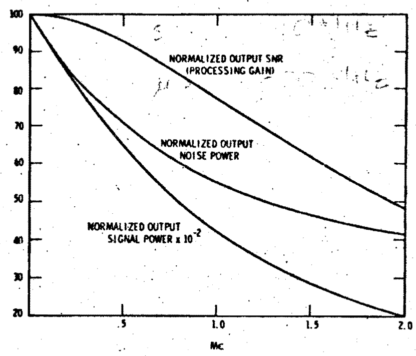
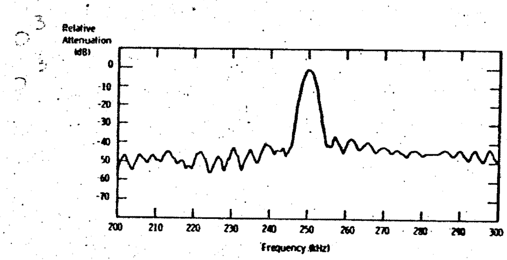



Figure 9. Signal Power, Noise Power, and SNR at the Output of the 100-Bit Filter Described in the Text. These results are normalized to their respective quantities at the input to the filter and plotted as a function of the loss parameter Mc.




Experimental Response of 500-stage CCD Bandpass Filter

Figure 10. The Measured Frequency Response of a 500-Stage Dolph-Chebyshev CCD Bandpass Filter. The filter was designed to have 42 dB out-of-band rejection and tunable passband at .25 $f_c$

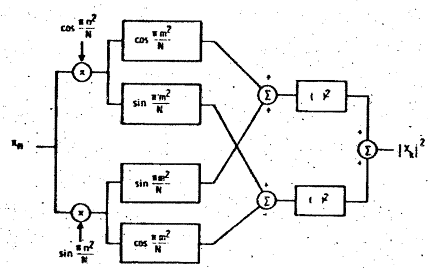



Figure 11. Circuit for Taking the Power Density Spectrum. The input signal is sampled to give the analog samples $x_n$, n=0, N-1.

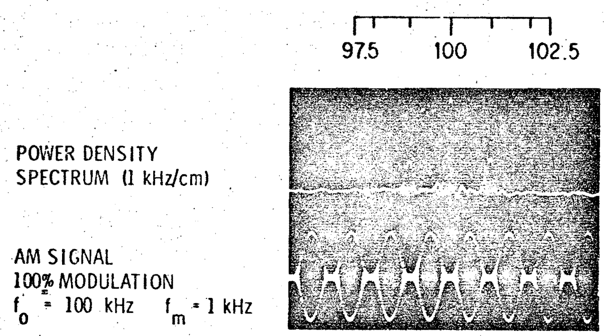



Figure 12. Power Density Spectrum of an AM Waveform. This spectrum was obtained using the filters whose characteristics are shown in Figure 6 in the CZT system of Figure 11.

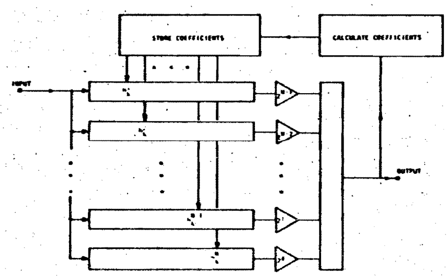



Figure 13. Adaptive Channel Equalization Using the Analog-Binary Filtering Scheme. The weighting coefficients of each filter are binary adjustable (one or zero) and are supplied from the coefficient store. The M coefficients supplied to the $i^{th}$ filter are the $i^{th}$ digits in the binary representation of the desired $h_k$.