THE USE OF CHARGE-COUPLED DEVICES FOR SINGLE-SIDEBAND MODULATION

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ABSTRACT

CCD transversal filters offer the possibility of splitting a varying frequency signal into two quadrature components over wide frequency ranges. Such a CCD filter, in effect realising the Hilbert transform, can therefore be used to generate the required quadrature signals and hence produce an SSB modulated wave by the well known phasing method. Experimental studies of the performance of such filters are presented. The useful frequency range about a given centre frequency is found to be restricted as a result of smearing within the device and by the accuracy with which the weighting coefficient can be realised. The dynamic range is limited by output problems and non-linearities.

INTRODUCTION

Single-sideband (SSB) modulation techniques (Ref. 1) are being increasingly used in a variety of communication systems because of their advantages over other types of linear modulation, particularly with respect to spectrum conservation, reduced power and weight requirements and improved propagation characteristics. The major drawback with these systems is the difficulty of achieving unwanted sideband suppressions by either conventional filtering or the phasing technique (Ref. 2). Charge-coupled devices (Ref. 3) offer novel ways of realising both filters and phasing systems for this purpose. Most existing systems employ filtering as the phasing technique is difficult to realise in practice with conventional circuitry. The aim of this paper is to discuss the application of charge transfer devices to the phasing technique because theoretically it offers infinite sideband suppression.

An SSB phasing system is shown in figure 1. The phasing elements are used to split the baseband and carrier signals into two components in phase quadrature. The two sets of signals are then fed to balanced modulators to produce two sets of double-sideband modulated waves. The phase relationships of these two waves are such that,say,the lower sidebands are in phase whilst the upper sidebands are 180 degrees out of phase. Thus if the two waves are summed at the output of the modulator an SSB modulated signal results with only the lower sideband being present. By reversing one of the phasing elements the lower sideband is eliminated and only the upper sideband remains. Infinite sideband suppression is only attained if each of the phasing elements satisfies the following criteria: (a) the two outputs are equal over all frequencies of interest, (b) the two outputs are exactly 90 degrees out of phase over all frequencies of interest. A one percent and one degree deviation from these requirements reduces the sideband suppression from infinity down to approximately 40dB. This is about the value that can be tolerated for a good practical system and it is evident that with

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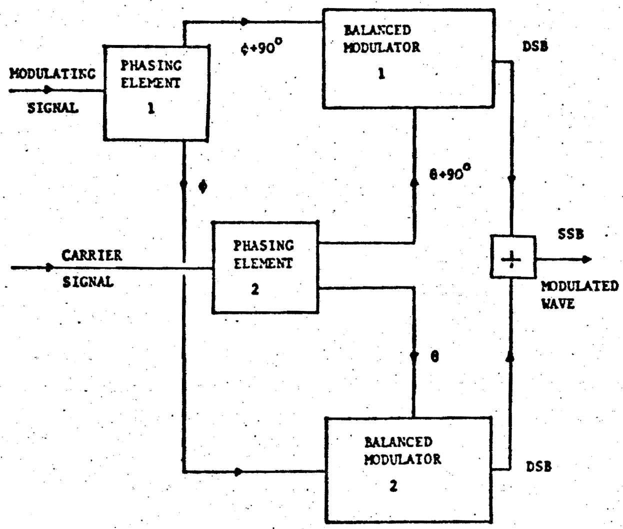

FIG. 1. AN SSB PHASING MODULATOR

conventional circuitry it is difficult to meet these specifications, particularly with baseband frequencies occupying several octaves.

As an alternative to conventional circuitry, the properties of transversal filters (Ref. 4) can be exploited to construct the phasing elements (Ref. 5). This approach shows considerable promise as with the advent of charge-transfer devices it has become possible to realise transversal filters for analogue signals with frequencies from d.c. up to several MHz (Ref. 6).

THE TRANSVERSAL FILTER AS A PHASING ELEMENT

A transversal filter (Ref. 7) as shown in figure 2(a) consists of a delay line which is tapped at a number of points along its length. The delay time between taps is T and the total delay of the line is 2nT. The output signal from the filter consists of the weighted sum of the signals at the taps and that at the input and output of the delay line. The amplitude characteristic is shown in Fig. 2(b). It consists of a series of passbands which repeat in frequency at intervals of 1/T. The shape of the passbands is governed by the number of taps and the weighting coefficients C_K. A further and unique property of such a

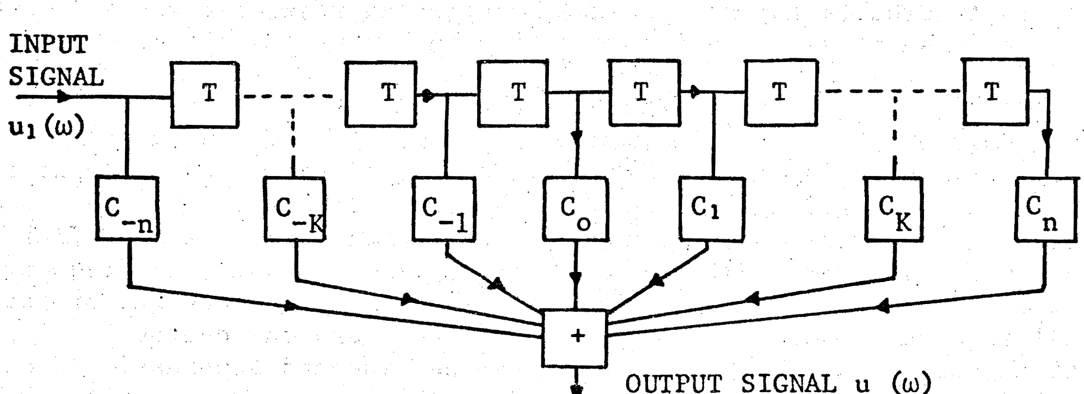

FIG. 2(a) A TRANSVERSAL FILTER WITH TOTAL DELAY = 2nT

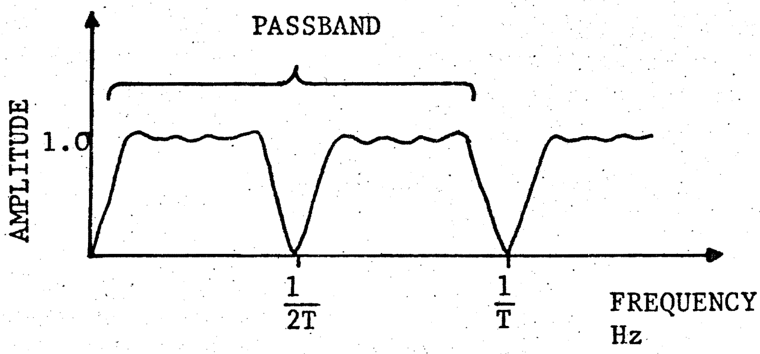

FIG. 2(b) AMPLITUDE CHARACTERISTIC OF TRANSVERSAL FILTER

filter is that its phase response is linear with frequency. This basically arises from the frequency independence of the delay, T. These properties can be illustrated by considering the simplest form of the filter which is that of a single delay line ($C_K = 0$, $k \neq 1$, $C_1 = 1$, $\tau = 2T$) as shown in figure 3(a). The output of the filter is simply the sum of the input and output signals of the delay line, with unity weighting. The characteristics of this filter are therefore described

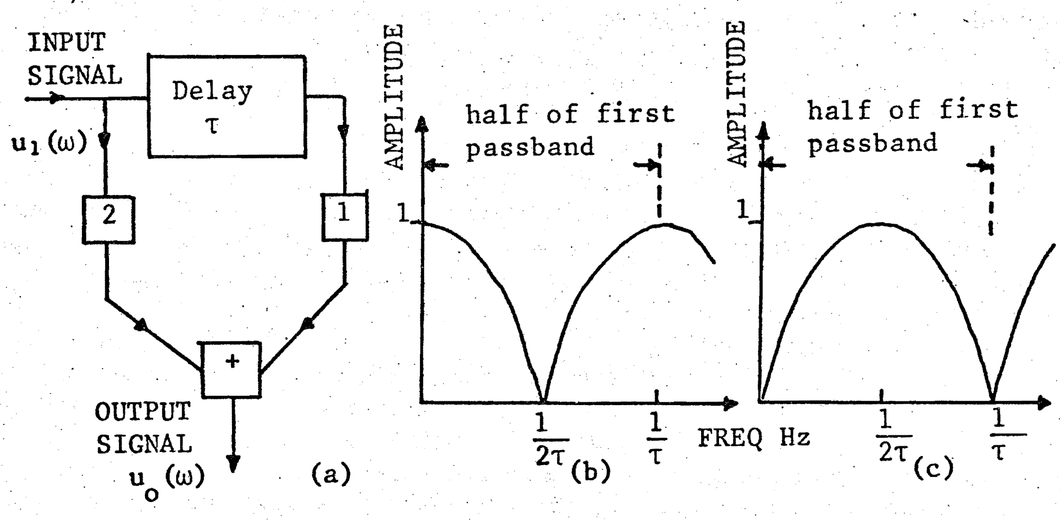

FIG. 3 (a) The simple transversal filter (b), (c) amplitude responses of simple filter for different weightings.

by the equation

$$
\begin{aligned}
u_o(\omega) &= u_1(\omega) + u_1(\omega)\exp(-j\omega\tau) \\
&= u_1(\omega)\left[1 + \exp(-j\omega\tau)\right] \\
&= 2u_1(\omega)\cos\left(\frac{\omega\tau}{2}\right)\exp\left(-\frac{j\omega\tau}{2}\right) \qquad \ldots(1)
\end{aligned}
$$

The first term in this equation is the amplitude response of the filter and corresponds to that illustrated in figure 3(b). The nulls occur at frequencies where the delay causes a phase shift of π or an odd integral multiple of π. The second term is the linear phase shift. By altering the weighting coefficients such that the two signals subtract at the output of the filter equation (1) becomes

$$u_o(\omega) = 2u_1(\omega)\left[1 - \exp(-j\omega\tau)\right]$$
$$= 2u_1\sin\left(\frac{\omega\tau}{2}\right)\exp\left[-\frac{j\omega\tau}{2} + \frac{j\pi}{2}\right] \qquad \ldots(2)$$

The amplitude response is illustrated in figure 3(c). The second term in the equation is again a linear phase shift but in addition, in this special case, the signal is always 90 degrees out of phase with equation (1). Thus by feeding the input signal simultaneously into two such transversal filters or into a single filter tapped twice at every point, two output signals which are in phase quadrature and linearly related in phase to the input signal, can be obtained. This forms the basis of a phase shifting element for SSB applications.

These simple filters, however, fulfil only one of the two requirements of a phasing element. They provide two outputs in phase quadrature at any frequency but the two outputs are not equal, one follows a cosine and the other a sine relationship. In order to overcome this problem additional stages must be added to the filters so that the shape of the passband can be modified. As shown in appendix I, the phase properties observed in the simple one stage filter can be retained with a multiple stage filter provided one filter is designed to have a passband that is an even function and the other an odd function of frequency. The weighting coefficients are then simply the fourier coefficients of the functions defining the passbands. From this it is evident that to define a given shape of passband precisely, an infinite number of taps is required and as in practice only a finite number can be used, the amplitude response can only be an approximation to the desired response. With ideal delay lines the two outputs are always exactly in phase quadrature independant of the number of stages.

In order to keep the output signals equal rectangular passband shapes are chosen. Figure 4 shows the calculated amplitude response of a ten stage filter with non-zero weighting coefficients at every second tap. Note that one passband is an odd and the other an even function. The passband shape is only a poor approximation to a rectangular response and for a constant input signal the output amplitudes can differ by almost 23% between the frequencies F_1 and F_2. These two points F_1 and F_2 define the useful frequency range within the first quarter of the pass-band. Even for an ideal rectangular response the frequency range is limited by the presence of nulls. It can be seen that in order to satisfy the amplitude requirements of a practical filter a larger number of taps is required or the amplitude response must be smoothed in some other way. The difference between the two amplitudes is given approximately by the ratio of the 1st to the last weighting coefficient used.

The transversal filter is realised using charge-transfer devices (Ref. 6). The analogue signal is sampled at the clock frequency and is reconstituted at the output of the filter by suitable band-pass filtering. The delay

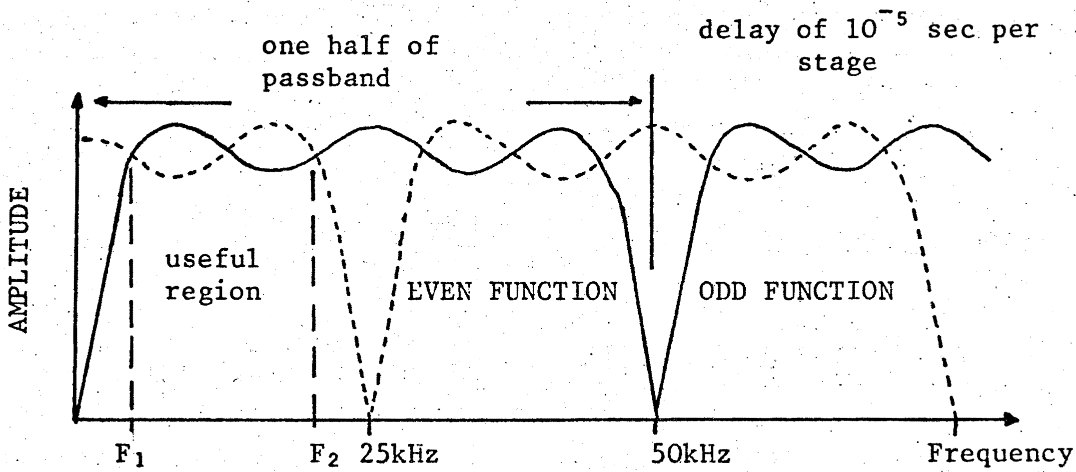

FIG. 4 CALCULATED AMPLITUDE RESPONSE OF 10 STAGE ODD AND EVEN TRANSVERSAL FILTERS.

of each stage of the filter is an integral number of clock cycles. The clock frequency therefore controls the position of the nulls of the filter and must also satisfy the sampling theorem.

PERFORMANCE OF SIMPLE PHASING ELEMENTS

The simple filters of figure 3 have been realised using single bucket-brigade devices giving amplitude characteristics with the first null at 12 kHz. The results show that the phase difference between the two outputs of such filters can be maintained within 2 degrees of 90 degrees between 2 and 8 kHz over a 30dB dynamic range although the errors do increase dramatically towards the nulls at 0 and 12 kHz.

An experimental CCD filter was realised using ten stages of a 14 stage tapped CCD, tapped at every stage and each stage having a delay equal to one clock period. The device used was constructed on 0.5Ωm <100> orientation n-type silicon with an oxide thickness of 1500 Å. The tapping arrangement is of the form described by McLennan et al (Ref. 8). Figure 5 shows the ten stages of the filter and the required weighting coefficients. Signal polarity inversion is achieved by sampling the signal at either the source or the drain load of the sensing mos device, (in this case a p channel FET) as shown in the inset of figure 5. The appropriate polarity signal is then fed via a fast operational amplifier source follower to the summing amplifier where the weighting coefficients are determined for each tap by the ratio of the input to feedback resistance. The source follower acts as a buffer and allows the coupling capacitor to be dispensed with. A capacitor would introduce excessive phase error and attenuation at low frequencies.

Input to the CCD is via a standard p-n junction/gate arrangement as shown in the other inset of figure 5. The diode is grounded and the analogue signal applied to the gate on a suitable dc level.

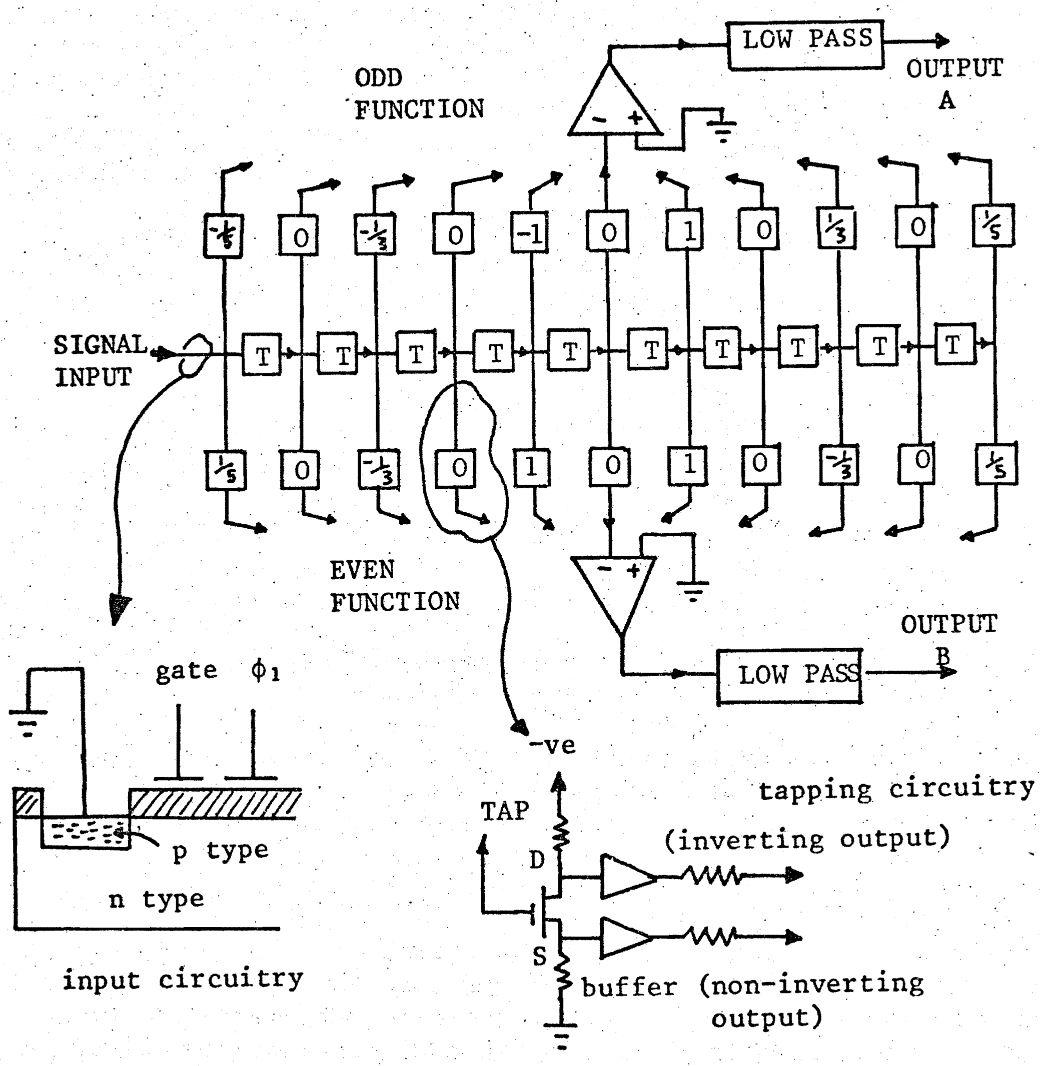

FIG. 5 CCD TRANSVERSAL FILTER SHOWING WEIGHTING COEFFICIENTS

Preliminary results were obtained using this configuration for a clock frequency at 100 kHz. The error in the phase response was found to be within ten degrees except towards the nulls where it again increases. The dynamic range was found to be very limited. The principle causes of error appear to be smearing and error in the setting of the weighting coefficients. The dynamic range is limited by the non-linear behaviour of the device which can partly be attributed to the smearing and partly to non-linearities associated with the input to the device and with the tapping technique. Phase error at the nulls increases due to the sensitivity of the filter to small deviations from the correct magnitudes of the signals at the summers (due to either smearing or weighting coefficient error). The amplitude response does not exhibit this sensitivity at the nulls, and was found to be within 15% of that predicted in figure 4.

CONCLUSION

Preliminary results for phasing elements constructed using charge-transfer device transversal filters indicate that such systems have considerable promise for applications in SSB modulation, although the requirements for a practical system cannot be met with the existing simple devices for two reasons:

(a) The amplitude response requires a large number of taps for a good approximation to the ideal rectangular response.

(b) The phase response is sensitive to errors in weighting coefficient, smearing and device non-linearities.

APPENDIX I

Consider a general transversal filter such as that shown in Fig. 2(a) with 2n delay stages each of which has a delay time T equal to an integral number of clock periods. The input signal is sampled at the input of the filter and after summing the output signal is reconstructed by suitable bandpass filtering. The output from this filter is given by

$$u_o(\omega) = u_1(\omega) \sum_{k=-n}^{n} C_K \exp\left[-j\omega(n+k)T\right] \qquad \text{...(A1)}$$

where C_K are the weighting coefficients. For a given amplitude response, $|u_o(\omega)|$, these can be calculated by noting that equation (1) is similar in form to a fourier series. Hence the coefficients C_K are given by

$$C_K = \frac{T}{2\pi} \int_{-\frac{\pi}{T}}^{\frac{\pi}{T}} u_o(\omega)\exp(jn\omega T)\,d\omega \qquad \text{...(A2)}$$

The above equation implies that the amplitude response of the filter is periodic with a period of $2\pi/T$. It consists of a series of passbands as shown in Fig. 2(b). In addition a fourier series has the following property:

For an odd function $C_{-K} = -C_K$ and for an even function $C_{-K} = C_K$.

Thus if the weighting coefficients are chosen such that the passbands are even functions of frequency equation (1) becomes

$$u_o(\omega) = u_1(\omega)\left[\sum_{K=0}^{n} C_K\exp(-j\omega\left[n+K\right]T) + C_K\exp(-j\omega\left[n-K\right]T)\right]$$

$$= 2u_1(\omega)\exp(-j\omega nT) \sum_{K=0}^{n} C_K\cos(\omega kT) \qquad \text{...(A3)}$$

and for an odd function

$$u_o(\omega) = u_1(\omega)\left[\sum_{K=0}^{n} C_K(\exp(-j\omega\left[n+K\right]T) - \exp(-j\omega\left[n-K\right]T)\right]$$

$$= 2u_1(\omega)\exp(-j\omega nT + \frac{j\pi}{2}) \sum_{K=0}^{n} C_K\sin(\omega kT) \qquad \text{...(A4)}$$

The resulting outputs from two such filters can be seen to be in phase quadrature. This is the basis of the CCD phasing element. By sampling a 2n stage CCD delay line twice after each delay element and weighting one set of samples so as to give an even and the other to produce an odd amplitude characteristic the summed outputs will be in phase quadrature. A further restriction on the choice of weighting coefficients arises from the need to keep the amplitude response flat in both cases. This requirement also implies the use of an infinite number of delay elements and the necessary truncation will introduce some error in a practical filter.

For the case of the baseband signal the operating region would be within the first passband whereas for a carrier frequency phasing element a higher passband would be used.

ACKNOWLEDGEMENT

The authors wish to thank the Royal Radar Establishment for supplying the charge-coupled devices used in the experiments.

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