A CCD PHASE-CANCELLATION SYSTEM FOR SINGLE-SIDEBAND MODULATION

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ABSTRACT

A prototype CCD implementation of a single-sideband modulation system for frequency-division multiplexing (FDM) in telephone systems is presented. The system is based on phase cancellation of the unwanted sideband. The prototype CCD system is built up of two split-electrode transversal filters in parallel. The clocking frequency is 8 kHz; the clock phase for the two filters differs by 90°. Some design considerations are discussed and performance data are presented. A sideband suppression of 48 dB has been achieved in the prototype devices.

INTRODUCTION

In frequency division multiplexing (FDM) telephone systems, each telephone channel requires one channel filter. Notwithstanding the substantial size reduction and the improved adjustment facilities over the last decades [1,2], these filters still consume a considerable share of the cost and size of modern FDM systems. In spite of the strong tendency nowadays towards digital implementation, there will be a continuing demand for small and low-cost single sideband (SSB) modulation filters, in the near future.

A number of methods for implementing SSB modulation filters with charge-coupled device transversal filters have been considered before by Tiemann and Sherrick [3]. One of these methods is the phase-cancellation system (PCS). Molo and Ricco [4] made a theoretical feasibility study on the system. To our knowledge no experimental evaluation of this principle has been reported.

In this contribution a prototype CCD implementation of the PCS is presented. General design aspects are reviewed and the performance data of the prototype device are discussed. Although the present prototype does not meet the CCITT requirements, it points out the possibility for CCD implementation of the phase-cancellation principle for SSB modulation.

THE PHASE-CANCELLATION SYSTEM

The purpose of the system is the transformation of the telephone voiceband (300 Hz until 3400 Hz) up to a single sideband of the carrier frequency fc. Hence the image band is from (fc - 300 Hz) until (fc - 3400 Hz) or from (fc + 300 Hz) until (fc + 3400 Hz). In the first modulation step of a SSB telephone system, usually a carrier frequency fc in the range of 20 kHz to 30 kHz is applied. The principle of the phase-cancellation system (PCS) has been reported in [4] and [5] and will be discussed briefly.

We assume the input signals to be bandlimited to the voiceband of 300 to 3400 Hz. The generation of the upper- and lower sidebands is performed by sampling the voiceband with sampling rate fc. The PCS is based on the phase differences between the output signals of two sampled data filters that are connected in parallel (fig. 1).

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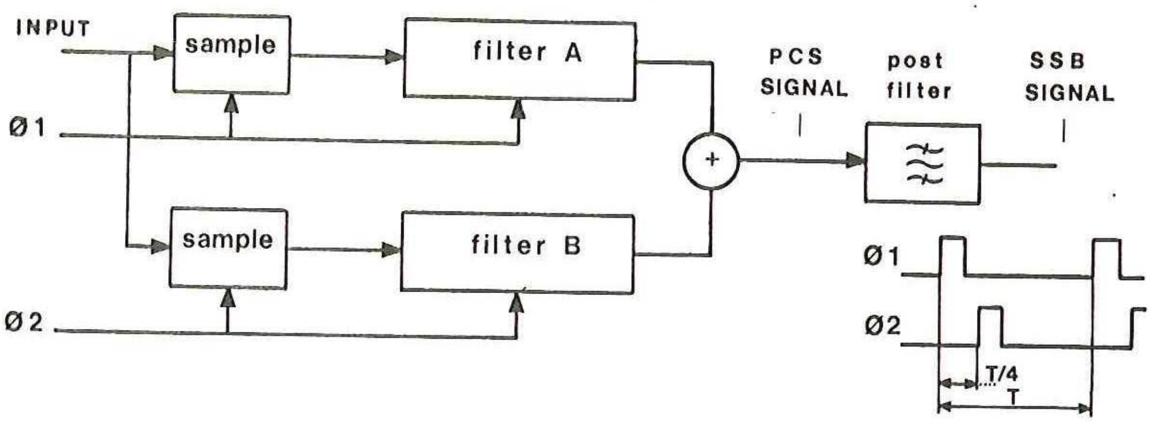

Figure 1. Block scheme of the PCS for SSB generation.
Filter A (HT) has an antisymmetrical impulse response,
filter B (DT) has a symmetrical impulse response. The clock $\phi 2$ is delayed by $\frac{1}{4}$T with respect to $\phi 1$.

In one of the sidebands of fc the filter output signals have to be in phase, while in the other sideband the difference in phase between the two filter outputs has to be 180°. By addition of the filter output signals, one sideband is enhanced, while the other sideband cancels.

In the PCS, a Hilbert transformer is used in branch A to provide for the 180° change in signal phase around fc. Actually, as will become clear from (2), the HT causes a phase shift of $+90^{\circ}$ in one of the sidebands and -90° in the other sideband. Within the desired frequency band, the HT has a flat amplitude response. The delay is independent on the signal frequency. In branch B a delay transformer (DT) is inserted, that has to introduce the same signal delay and within the desired frequency band, the same flat amplitude response. The phases for the upper- and lower sideband have to be equal.

It will be explained further down, that a HT transversal filter with an odd number of taps N is attractive for the implementation of a PCS. The HT requires an antisymmetrical impulse response. For the tap weight of the k^{th} tap (k=0, ..., N-1) it holds:

$$h_k = - h_{N-1-k} \tag{1}$$

The transfer function of the filter is [6]:

$$H(j\omega) = e^{-j\omega(N-1)T/2} \cdot 2j \sum_{k=0}^{(N-1)/2} h_k \cdot \sin \omega \left(\frac{N-1}{2} - k \right) T \tag{2}$$

T denotes the delay period, which is equal to the sample interval: T = 1/fc. The group delay of the linear phase HT filter is (N-1)T/2. If we want an image band of the carrier frequency fc = 24 kHz, the filter clock frequency should be fc = 24 kHz.

For the present purpose, the Hilbert transformation only has to be performed in the frequency band from 300 Hz to 3400 Hz, i.e. from 0.0125 fc to 0.142 fc. Hence it is allowed, just for the case of filter design, to use the unconstrained frequency band between 3400 Hz and the Nyquist frequency fc/2 for a periodic extension of the filter function. The filter design is also facilitated by a high symmetry in the transitionband [6].

Those considerations lead to the transfer function of fig. 2 , with a periodicity of fc/3 and symmetry around fc/12. Substitution of this symmetry requirement in (2) leads to the conclusion that only non-zero tap weights h_k are allowed for k values that satisfy:

$$\left(\tfrac{1}{2}(N-1)-k\right)/12 = (1+2n)/4 \quad \text{or: } k = \tfrac{1}{2}(N-7)-6n \tag{3}$$

$$\text{For } h_o \neq 0 \text{ it follows: } h_k = 0 \text{ for } k \neq 6n' \tag{4}$$

From (3) it follows that the number of delay elements N for the HT has to meet the requirement: $N = 7 + 12n''$ (5)

(n, n' and n'' denote integer values). The advantage of zero tap weights is that they don't have to be sensed and therefore don't contribute to any noise or distortion of the transfer function.

The same arguments may be applied to the design of the DT in the other filter branch. Here the desired phase response requires a symmetrical impulse response:

$$h'_k = h'_{N-1-k} \tag{6}$$

and the transfer function becomes:

$$H(j\omega) = e^{-j\omega(N-1)T/2} \cdot h'_{\frac{N-1}{2}} + 2 \sum_{k=0}^{(N-3)/2} h'_k \cos \omega \left(\frac{N-1}{2} - k\right) T \tag{7}$$

The most simple design that fulfils the requirements of constant group delay equal to $(N-1)T/2$, no extra phase shift and constant modulus of the transfer function, is a filter with only one tap weight: $h'_{(N-1)/2}$, while all $h'_k = 0$ for $k \neq (N-1)/2$. This is simply a delay line.

An improvement in the cancellation of the unwanted sideband can be obtained if the DT is designed as to approximate the ripple in the transfer function of the HT. Due to the periodicity in the HT design, the same periodicity is required for the ripple compensation in the DT design. The symmetry requirement has to be substituted in (7) and leads to non-zero tap weights h'_k for k values that satisfy:

$$\left(\tfrac{1}{2}(N-1)-k\right)/12 = \tfrac{1}{2}\, n \quad \text{or: } k = \tfrac{1}{2}(N-1) - 6n \tag{8}$$

It follows again: $h'_k = 0$ for $k \neq 6n'$ (9)
From (8) it follows that the number of delay elements N of the DT has to be equal to: $N = 1 + 12n''$ (10)

(n, n' and n'' denote integer values). The tap-weight values h'_k (k=6n) that provide for the matching of the ripple in the HT transfer function appeared to be very small. A further requirement for sufficient ripple matching is that the length of the DT exceeds the HT filter length. If no ripple matching would be applied, the ripple in the HT (and DT) transfer functions would determine the maximum sideband cancellation. In that case a maximum ripple of 0.1% would be required to meet theoretically the 60 dB sideband suppression. Those specifications, together with the requirement for a small transition bandwidth would lead to very long transversal filters ($\geq$ 145 taps).

Because the HT causes periodically an extra phase shift of $+90^o$ and -90^o, whereas the DT has no extra phase shift, the signal in one of the filter branches has to be subjected to an additional overall 90^o phase shift in order to enable phase cancellation.

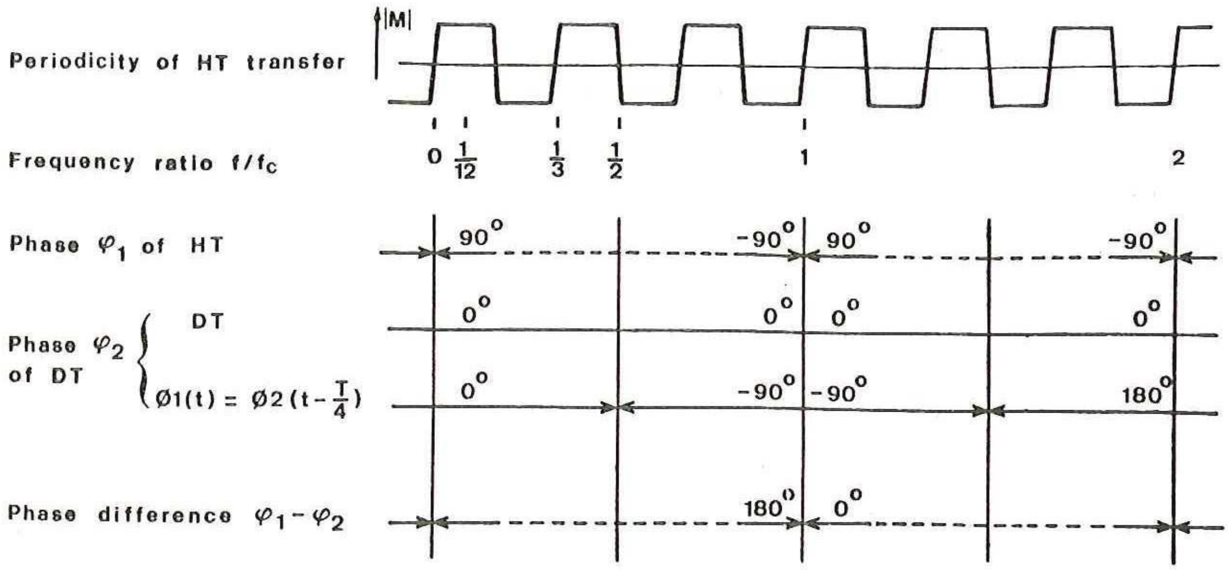

Figure 2. Periodicity and phase response diagram of the PCS.

This 90° phase shift is obtained by shifting the sampling moments and clock pulses of the DT over 1/4 of the clock period, with respect to the clock pulses and sampling moment of the HT. The 90° phase shift resulting from the delayed sampling may be explained at the hand of the Fourier series expansion of the sampling pulse sequence:

$$\sum_{k=-\infty}^{\infty} \delta\left(t - (k+\tfrac{1}{4})T\right) = \sum_{n=-\infty}^{\infty} e^{-jn\pi/2} \cdot e^{-jn2\pi t/T}$$

In the frequency domain, the sampled input signal $V_{in}(j\omega)$ becomes, with $\omega c = 2\pi fc$:

$$V_{in}(j\omega) = \sum_{n=-\infty}^{\infty} e^{-jn\pi/2}\, V_{in}(jn\omega c \pm j\omega) \qquad (11)$$

From (11) it is clear that a phase shift of -90° is added for the sidebands of the first clock harmonic (n=1). A graphical representation of the phase shift in HT and DT, as well as determined by (11) is given in fig. 2. With a filter length of 91 taps for the Hilbert transformer and 97 taps for the DT, a PCS with a passband region from 300 to 3700 Hz and a ripple in this passband less than 0.1 dB can be designed [7]. Owing to a good HT to DT ripple overlap, the theoretical unwanted sideband suppression is at least 60 dB. The system should be operated at 24 kHz clock frequency.

DESIGN AND PERFORMANCE OF THE PROTOTYPE PCS CCD FILTER

The purpose of the prototype is to study the feasibility of a CCD PCS implementation. The performance of the prototype should be comparable with the PCS filter design described above. Technological limitations of the in-house processing facilities forced us to limit the chip size to 2 x 2 mm^2. Therefore the filter length of the complete system has to be reduced to 25 delay elements or less. This is achieved by frequency scaling in the first place. The clock frequency is reduced to 8 kHz and the passband symmetry from 1/12 to 1/4 of the clock frequency. This leads to a design with only every second tap weight equal to zero, while the non-zero tap weights have exactly the same value as in the original design.

Frequency scaling only should reduce the filter length to 31 and 33 taps. To meet the chip size limitation, the useful band width of the filters has been reduced to 480 Hz - 3520 Hz.

The prototype PCS contains a 23 tap HT and a 25 tap DT on a single chip. To provide for the equal delay of the filter, the HT is preceeded by one delay element. The theoretical suppression of the unwanted sideband is 67 dB within the bandwidth of 500 Hz to 3500 Hz. The ripple in the passband is 0.5% ($\simeq$0.05 dB). The 10 μm technology used n-channel surface CCD's with overlapping 2-level polycrystalline silicon gates. Substrate resistivity was 10 Ωcm and the channel width of each CCD filter was 450 μm. The input signal is sampled by a fill-and-spill section. Charge transport is controlled by a $1\frac{1}{2}$ phase clock system. The peripheral circuitry consists of on-chip source-follower output buffers, reset MOST switches and feedback capacitors for the constant-voltage sensing. A microphotograph of the chip is shown in fig. 3. The prototype requires 4 external sense amplifiers, 1 differential amplifier and external clock generators.

For the specification of the PCS performance, three characteristics have been examined:

- In fig. 4 the modulus of the enhanced upper sideband (difference between HT and DT filter outputs) as well as the modulus of the suppressed upper sideband (addition of HT and DT signals) is shown. The sideband suppression of the PCS follows directly from fig. 4 without a correction for the modulation due to the sample-and-hold effects. Measurements on 6 devices turned out that 4 of them showed a sideband suppression of 45-48 dB within the bandwidth of 500 - 3500 Hz.
- Measurements of the second and third harmonic of the input signal showed a distortion level below -62 dB up to an input signal of 2V p-p.
- The noise level within a bandwidth of 30 Hz has been measured at -62 dB. Within the band of fc until fc + 4 kHz this appeared to have a white noise spectrum. The total signal/noise ratio within the telephone bandwidth is -40 dB.

Because the experimentally observed sideband suppression did not meet the theoretical level, the sources of errors in the system have been investigated. Therefore the magnitude and phase of the transfer functions of the HT and DT have been measured separately and the actual tap-weight values have been recalculated by inverse Fourier transformation. The accuracy of those measurements did not allow accurate conclusions. Tap weight errors in the range of 0.1 to 0.5% of maximum tap-weight value have been calculated.

The general indication is that one of the main sources of errors is a misalignment of the first poly mask, that defines the position of the splits of the sense gates. This is quite conceivable because 2 μm misalignment corresponds with a tap-weight deviation of 0.15% of the largest tap weight. Another source of errors may be induced by the summation of the four sense amplifier output signals. Particularly the large common mode signal may cause troubles. The use of double-split-electrode filters [8], may improve the filter performance. The signal to noise ratio in the prototype system was about 40 dB in the voiceband bandwidth. It should be noticed that the prototype design did not include any optimation for noise reduction.

CONCLUSIONS

With a design that is less sensitive for alignment errors and with the implementation of double-split-electrode transversal filters, the sideband cancellation may be improved considerable. Because the actual number of sensed taps is about equal to that of the proposed 97 tap design, it is expected that the presented performance data are also representative for the full filter design. The final conclusion is that with high quality CCD processing a PCS system for SSB seems feasible, although the CCITT specifications not yet have been achieved.

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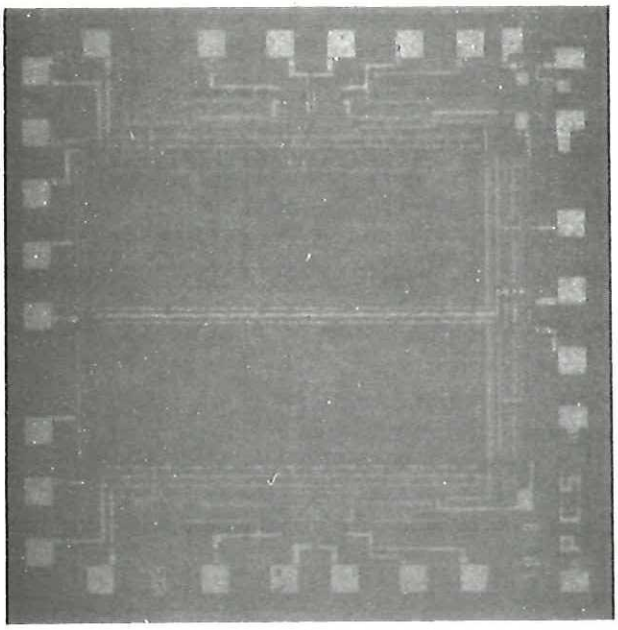

Figure 3. Photograph of the PCS prototype chip ($2 \times 2\ mm^2$).

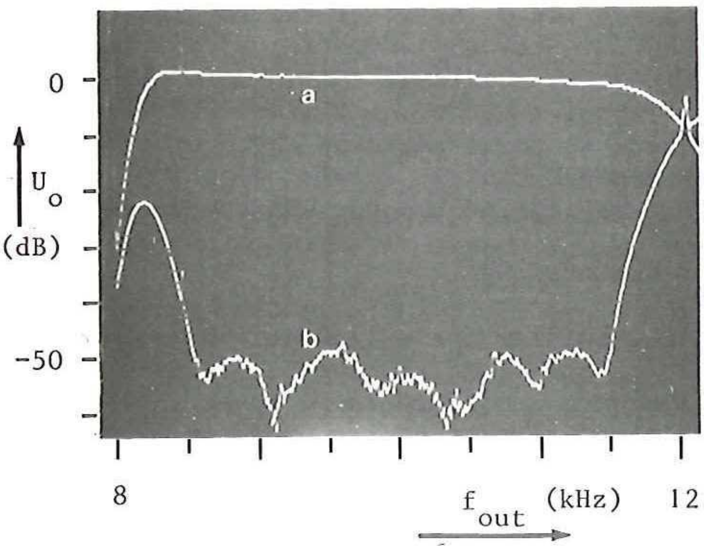

Figure 4. a: Enhanced upper sideband of fc (HT-DT).
b: Cancelled upper sideband of fc (HT+DT).

$f_{in} = f_{out} - 8$ kHz