FULLY-INTEGRATED CHARGE-COUPLED PCM LINE FILTERS.

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ABSTRACT

Two charge-coupled filters for use in a PCM system are described. Design considerations, noise calculations and experimental results are presented.

I. System considerations.

The integrated circuit contains essentially two charge-coupled transversal minimum-phase low-pass filters for use in a PCM system. The filter characteristics are designed to meet the CCITT recommendations G.712[1].

A PCM channel consists of a transmitter, containing an anti-aliasing filter and an A/D converter, and a receiver, containing a D/A converter and a signal reconstruction filter. Both filters are implemented as CCD's for which a clock frequency of 32 kHz is chosen. As the converters have to operate at a sample rate of 8 kHz only one sample is taken out of four at the output of the transmit filter whereas four identical samples are sent in one after another at the receive filter. The overall transfer function of the latter transformation combined with the sample and hold output operating at 32 kHz corresponds to the transfer function of a sample and hold operating at 8 kHz. The transmit filter has a prefiltering input which uses an oversampling at 64 kHz. The frequency responses of both charge-coupled filters are predistorted : the transmitting one to compensate for the rolloff of the prefilter input, the receiving one for the rolloff of a sample and hold operating at 8 kHz.

II. Circuit description.

The devices are processed in an n-channel two level polysilicon dynamic RAM technology adapted to yield enhancement and depletion transistors. A photograph of the chip is shown in fig. 1.

The charge-coupled devices are four phase structures. In both CCD's 47 weighting coefficients are implemented. The channel width is 600 μm. The length of one delay stage is 40μm. A parallel channel [2] is used to obtain an output DC level independent of the CCD input bias level and to achieve a better matching of the capacitances at the input nodes of the amplifier. The electrodes of the sense gates ($\emptyset$1) are double-split [3,4] resulting in smaller overlap capacitances, smaller common mode or bias charge and a lower sensitivity to capacitive mismatching.

With the used sensing method the outer parts and the center part of the split electrodes are not held at the same potential during sensing. Charge partitioning underneath such sense electrodes would cause a consi-

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derable error [2]. Our solution to that problem is to extend the islands of channel stop and thick oxide - upon which the sense electrode gaps are situated - underneath the preceding electrodes (Ø4) so that the charge packets are divided under unsplit electrodes.

The output circuit is shown in fig. 2. Only one operational amplifier is used as a differential integrator. The feedback capacitor value is chosen so that the total filter gain equals one. The amplifier circuit is based on Senderowicz's scheme [5]. The low frequency integrator noise is reduced and the sense gate reset noise totally suppressed by the correlated double sample and hold circuit [2,6]. The depletion source followers are designed to deliver no substantial level shift. V_{ref} is the reference voltage at which both the sense gates and the correlated double sampling structure are reset. V_{ref} is also applied to the center parts of the split electrodes.

The CCD charge injection method is a classical diode cutoff-technique with the second gate biased at V_{ref}. In this manner non-linearities due to the depletion capacitances under the sensing electrodes are largely reduced.

The capacitance at the amplifier input nodes is subjected to two opposite requirements : it has to be low to reduce the amplification of the equivalent input noise of the integrator and it has to be high if no other precautions are taken to restrict the voltage variation on the sense electrodes. With the use of $\overline{Ø4}$ and C_C (fig. 2) an excellent compromise is obtained. The voltage variation on the sense electrodes between the moments of resetting (before Ø4 falls off) and sensing is given by

$$V^{\pm} - V_{ref} = \frac{V_P(C_C - C_4) + Q^L_{bias} + Q^L_{signal}}{C_G + C_{FB}} \qquad (1)$$

$V^{\pm}$ are the voltages at the amplifier input nodes during the sensing. V_P is the phase pulse amplitude (14 V). C_4 is the overlap capacitance between sense gates and phase four. Q^L_{bias} and Q^L_{signal} are respectively the total bias and signal charge under the split electrode parts connected to the positive input node of the amplifier. C_G is the total capacitance at one amplifier input node, the feedback capacitor not included. By using $\overline{Ø4}$ and C_C a charge cancellation is realized. The compensation capacitor C_C is chosen so that formula (1) is reduced to

$$V^{\pm} - V_{ref} = \frac{Q^L_{signal}}{C_G + C_{FB}} \qquad (2)$$

A minimum value of 31 pF for C_G is now sufficient to limit the voltage variation on the sense electrodes to ± 0.5 V.

III. Noise calculations.

The output noise power in a bandwidth B with boundaries f_L=100 Hz and f_H=3.1 kHz is calculated for the transmit filter. The sample and hold at the filter output is operating at the CCD clock frequency f_C. The clock period is T_C. Depletion capacitances are neglected. The amplification of the correlated double sample and hold output structure is written as $A_{CDS\&H}$ and is equal to 0.86.

The formulae concerning the CCD noise sources can be found in ref. 7. The noise contributions due to dark current and to fast interface states are very low (Table 1). The noise power at the filter output due to the input noise of main and parallel channel is written as

$$v^2_{o,inp} = 2\int_{f_L}^{f_H} Q_i^2\, C_{FB}^{-2}\, A^2_{CDS\&H}\, f_C^{-1} \left| \frac{\sin \pi f T_C}{\pi f T_C} \sum_{n=1}^{N} h_n e^{-j2\pi f n T_C} \right|^2 df \qquad (3)$$

where σ_i^2 is the charge variance after the setting of the input gate oxide capacitance C_e and is taken equal to 4 kTC_e[7] with k the Boltzmann's constant and T the absolute temperature (300°K). N is the total number of the weighting coefficients h_n. For the main channel the modulus in formula (3) is taken equal for all passband frequencies to 4.44, which is the exact value at 800 Hz. For the parallel channel where all weighting coefficients are equal to one the modulus becomes $N\ |\sin\pi fNT_C/(\pi fNT_C)|$ which is zero at f_C/N. After calculating the integrals the results for the main and parallel channel are respectively

$$v^2_{o,inp,m} = \sigma^2_{i,m}\ C^{-2}_{FB}\ A^2_{CDS\&H}\ f^{-1}_C\ (4.44)^2\ 2\ B \tag{4}$$

$$v^2_{o,inp,p} = \sigma^2_{i,p}\ C^{-2}_{FB}\ A^2_{CDS\&H}\ N\ 0.67 \tag{5}$$

The charge integrator is the second important noise source. The white noise component, v_{MOA}, and the 1/f-noise component at 1 kHz, v^*_{MOA}, of the amplifier noise voltage referred to the input are respectively equal to 20 $nV/\sqrt{Hz}$ and 200 $nV/\sqrt{Hz}$. To find the integrator noise contributions we will first calculate the influence of the correlated double sample and hold.

A scheme of the correlated double sample and hold is given in fig. 3. A noise source r bandwidth limited to Ω is considered. Two samples taken respectively at $nT_C-\tau$ and nT_C are subtracted from each other and the resulting signal is held. The noise power is measured at the output of a bandpass filter. The equivalent scheme in fig. 4 is used for the calculation.

ϕ_r is the power spectrum of the noise source when the whole frequency axis is considered. When only positive frequencies are considered, we write v_r^2 with $v_r^2 = 2\phi_r$.

The power spectrum of signal s" (fig. 4) is written as

$$\phi_{s''}(f) = \phi_s(f)\left|e^{-j2\pi f\tau}-1\right|^2 = \phi_s(f)\ 2\ (1-\cos 2\pi f\tau) \tag{6}$$

or $$\phi_{s''}(f) = \phi_r(f)\ 2\ (1-\cos 2\pi f\tau) \quad \text{for } |f|<\Omega \tag{6a}$$

$$\phi_{s''}(f) = 0 \quad \text{for } |f|>\Omega \tag{6b}$$

The power spectrum of signal s^* is written as [8]

$$\phi_{s^*}(f) = f_C^2 \sum_n \phi_{s''}(f+nf_C) \tag{7}$$

The modulus of the transfer function of the hold operation is $T_C\ |\sin\pi fT_C/(\pi fT_C)|$. Neglecting the attenuation due to the $(\sin x)/x$ function the total output noise power is given by

$$v^2_{out} = \sum_n \int_{f_L}^{f_H} \phi_{s''}(f+nf_C)\,df + \sum_n \int_{-f_H}^{-f_L} \phi_{s''}(f+nf_C)\,df \tag{8}$$

Changing the arguments, using the symmetric property of a power spectrum and substituting $\phi_{s''}$ by $v_r^2\ (1-\cos 2\pi f\tau)$, expression (8) can be transformed

$$v^2_{out} = 2\sum_{n=o}^{n_m}\int_{f_L+nf_C}^{f_H+nf_C} v_r^2(1-\cos 2\pi f\tau)\,df + 2\sum_{n=1}^{n'_m}\int_{-f_H+nf_C}^{-f_L+nf_C} v_r^2(1-\cos 2\pi f\tau)\,df \tag{9}$$

As $\phi_{s''}(f) = 0$ for $|f|>\Omega$, n is restricted to n_m and n'_m.

Calculating expression (9) with an Ω-value of 0.45 MHz and a τ-value of 8 μs respectively with v_r^2 constant for white noise and with v_r^2 equal to (1000 Hz/f) v_r^{*2} for low frequency noise the formulae for the integrator contributions are obtained.

$$v_{o,int,w}^2 = v_{MOA}^2 \text{ Hz } 1.7\ 10^5\ (C_G+C_{FB})^2\ C_{FB}^{-2}\ A_{CDS\&H}^2 \tag{10}$$

$$v_{o,int,l}^2 = v_{MOA}^{*2} \text{ Hz } 1.2\ 10^3\ (C_G+C_{FB})^2\ C_{FB}^{-2}\ A_{CDS\&H}^2 \tag{11}$$

The results of the noise calculations for the transmit filter are summarized in table 1. The total output noise voltage in the specified band is 93 μV; with a signal of 3 V peak-to-peak the signal to noise ratio is 81 dB or 83 dB if CCITT psophometrical noise weighting is used. A similar performance is expected for the receive filter.

noise source	noise power at the filter output in a band between 100 Hz and 3.1 kHz 10^{-10} V^2
CCD-input	
main channel	21.5
parallel channel	17.1
dark current	
main channel	0.5
parallel channel	1.5
fast interface states	
main channel	2.2
parallel channel	1.2
integrator	
white noise	21.5
1/f-noise	15.2
CDS&H-structure	
transistors	3.4
capacitor resetting	1.9
Total	86.0

Table 1. Noise contributions at the filter output of the various noise sources.

IV. Experimental results.

Since the CCITT recommendation is only specified for a complete audio channel most of the figures below are given for a tandem configuration of transmit and receive filter except otherwise stated. The sample and hold at the transmit filter output is operated at 8 kHz.

The passband ripple is about $\pm$ 0.25 dB. The total stopband rejection is at least 72 dB while each filter separately gives more than 33 dB rejection in the stopband. The minimum phase design yields a group delay of 225 μs with superimposed a delay distortion of 170 μs between 500 Hz and 2800 Hz.

Total harmonic distortion is -45 dB at a signal level of 4 dBm. This yields a weighted noise to signal ratio of -76 dBmOp. The signal to noise ratio of the receive filter alone is 84 dBmOp which is in excellent agreement with the calculations made above. Intermodulation distortion is -40 dBmO. Gain deviation of 0.5 dB is met at 5 dBmO signal level.

No degradation of performance is observed by operating the devices at +70°C or at 150% of nominal ratings for extended periods. This indicates that the process yields reliable devices which do not suffer of dark current or threshold instabilities.

V. Conclusions.

Two filters suited for use in a PCM system are described. Noise calculations are presented and shown to be in excellent agreement with experiments. A signal to noise ratio of 84 dBmOp is obtained. With a tandem configuration of both filters a total distortion to signal ratio of -45 dBmO and a signal to noise ratio of 76 dBmOp are measured.

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References.

1. CCITT Sixth Plenary Assembly, Geneva, 27th September-8th October 1976, Orange Book, Vol. III, no. 2, pp. 415-423.
2. R.D. Baertsch, W.E. Engeler, H.S. Goldberg, C.M. Puckette, J.J. Tiemann: "The design and operation of practical charge-transfer transversal filters", IEEE Journal of Solid-State Circuits, Vol. SC-11, no. 1, pp. 65-73, February 1976.
3. T. Foxall, A. Ibrahim, G. Hupé : "Double-split-electrode transversal filter", Electronics Letters, Vol.13, no.11, pp. 323-324, 26th May 1977.
4. A.A. Ibrahim, G.J. Hupé, T.G. Foxall : "Double-split-electrode transversal filter for telecommunication applications", IEEE Journal of Solid-State Circuits, Vol. SC-14, no. 1, February 1979.
5. D. Senderowicz, D.A. Hodges, P.R. Gray : "High-performance NMOS operational amplifier", IEEE Journal of Solid-State Circuits, Vol. SC-13, no. 6, pp. 760-766, December 1978.
6. M.H. White, D.R. Lampe, F.C. Blaha, I.A. Mack : "Characterization of surface channel CCD image arrays at low light levels", IEEE Journal of Solid-State Circuits, Vol. SC-9, no. 1, pp. 1-13, February 1974.
7. S.P. Emmons, D.D. Buss : "The performance of charge-coupled devices in signal processing at low signal levels", in Proceedings of the CCD applications conference, (San Diego, Ca.), pp. 189-205, September 1973.
8. P.M.E.M. van der Grinten, J.M.H. Lenoir : Statistische procesbeheersing, PRISMA TECHNICA 50, Spectrum 1973, p. 164.

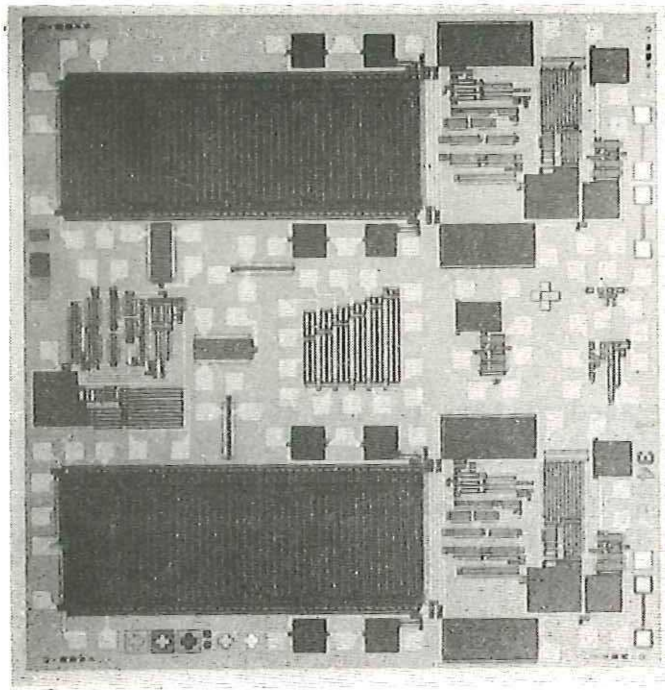

Figure 1: Photograph of the circuit.

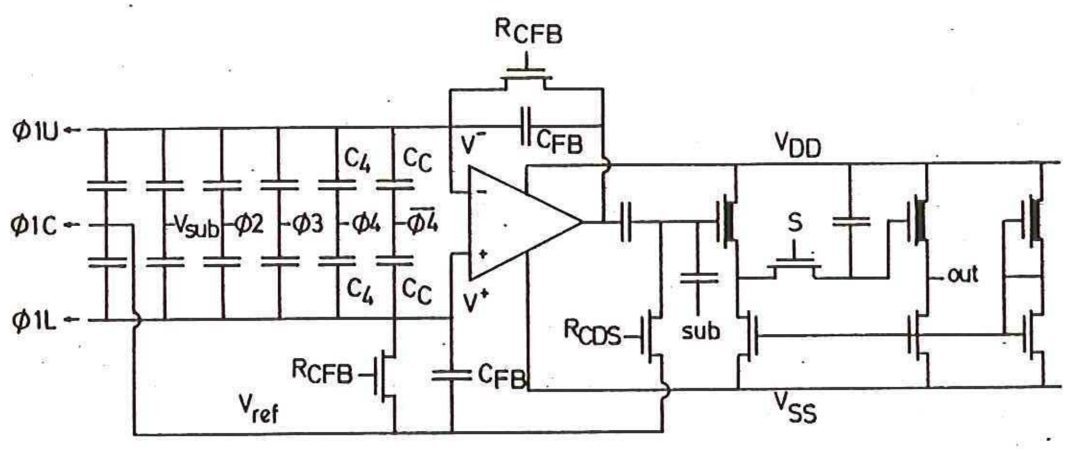

Figure 2: Output circuit

$\phi 1U$, $\phi 1C$ and $\phi 1L$ are resp. the upper, center and lower parts of the split electrodes.

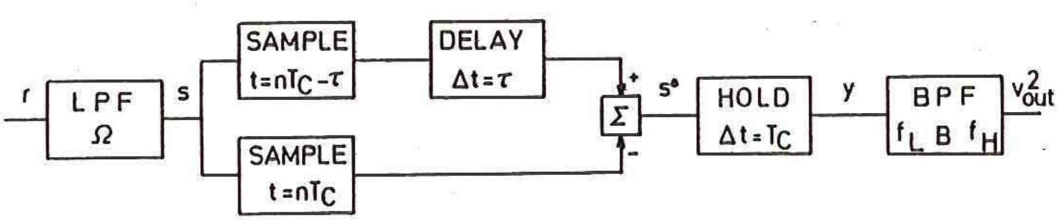

Figure 3: Scheme of the correlated double sample and hold.

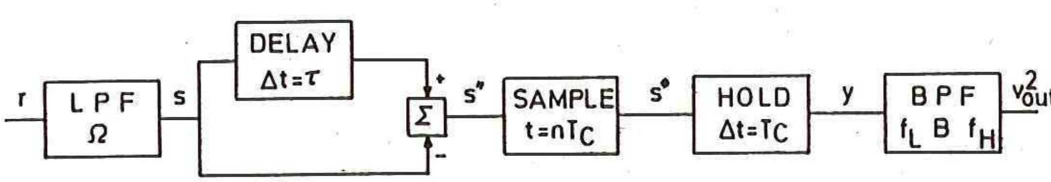

Figure 4: Scheme used for the calculation of the influence of a correlated double sample and hold on a noise signal.