$$-(\mathrm{Re}\{c_k\}+\mathrm{Im}\{c_k\})*\mathrm{Im}\{v(k)\}) \quad (9)$$
$$+j((\mathrm{Re}\{c_k\}+\mathrm{Im}\{c_k\})*\mathrm{Im}\{v(k)\}$$
$$+\mathrm{Im}\{c_k\}*(\mathrm{Re}\{v(k)\}-\mathrm{Im}\{v(k)\}))$$

Eqn. 9 and Fig. 2 indicate that, when compared with the conventional approach of Fig. 1, a complete filter section can be saved at the expense of a small increase in the number of external additions and subtractions from two to four. It should also be noted that the formation of each coefficient value for one of the three filter sections also requires an additional addition or subtraction operation during adaptation. Other three-filter variants, such as that derived from eqn. 6 above, will differ slightly in that they require only three external combining adders but need two coefficient adders per filter tap.

Furthermore, comparing a realisation based on three independent filter sections with one where a single complex equaliser is designed using three real multipliers per tap, the former requires only four combining adders (external to the filter sections) whereas the latter requires four per tap but requires fewer accumulation adders. Overall, the former saves $2N$ adders per equaliser at the expense of an extra N delay elements.

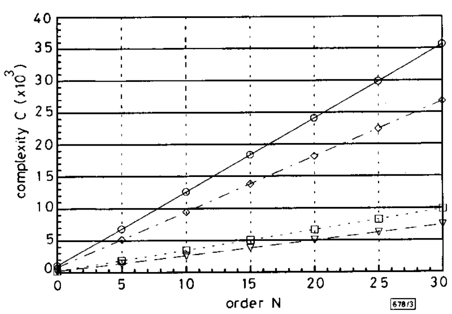

Fig. 3 *Complexity comparison between equalisers realised using three and four real filter sections for various order and wordlengths of 8 and 16 bit*

—▽— 3_sec: $B = 8$
—◇— 3_sec: $B = 16$
—□— 4_sec: $B = 8$
—○— 4_sec: $B = 16$

Complexity comparisons are presented in Fig. 3 for the case of complex filters realised with both three and four independent sections, for various values of N and with $B = 8$ and 16 bit. These results are based on the assumption that multiplication complexity is $O(B^2)$ and adder/subtractor (and delay element) complexity is $O(B)$ giving:

$$C_3 = 3(N+1)B^2+(6N+3)B$$
$$C_4 = 4(N+1)B^2+(8N+2)B \quad (10)$$

It can be observed that savings for the three-section case are asymptotic to 25% for large N. For mobile communications applications the combined number of taps in the feedforward and feedback parts of a DFE are typically less than 20, where the percentage saving is ~24%.

If the equaliser were to be implemented on a DSP microprocessor with a typical single-cycle multiply-accumulate-delay operation, then the conventional four-section solution would require $4N+6$ cycles per data item whereas the three-section variant will require only $3N+7$. For $N = 10$ this represents a saving of 20% whereas for $N = 20$ a saving of 22%. Again the saving is asymptotic to 25% as N increases. However, if the algorithm used were based on three multiplication operations per complex tap, this could increase to $8(N+1)$ cycles per data item, representing an overall reduction in throughput.

Conclusions: This Letter has shown that savings in implementation complexity for IQ filtering, approaching 25% are possible if filters are realised using the three multiplier approach. This can offer a hardware saving for VLSI, with associated power and cost reductions, or alternatively a throughput improvement for programmable DSP micoprocessor solutions. Although the method has been presented for the case of a linear transversal equaliser, it should be noted that the same principles may easily be extended to DFE and lattice structures.

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Kerwin-Huelsman-Newcomb circuit using current conveyors

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Indexing terms: Active filters, Current conveyors

Two equivalent circuits are generated using current conveyors to realise the same three transfer functions as in the Kerwin-Huelsman-Newcomb (KHN) biquad. Each circuit employs five current conveyors; two of them act as buffers. In both circuits, the two capacitors are grounded which is not the case in the KHN circuit.

Some years after Sedra and Smith [1] introduced the second generation current conveyor (CCII), it was used as the active building block in the realisation of filters [2, 3]. Recently several realisations were given for the Tow-Thomas biquad [4, 5] using CCIIs [6–8].

The purpose of this Letter is to introduce two equivalent realisations for the Kerwin-Huelsman-Newcomb (KHN) [9] biquad using CCIIs.

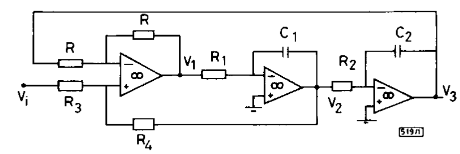

Fig. 1 *KHN circuit using three opamps*

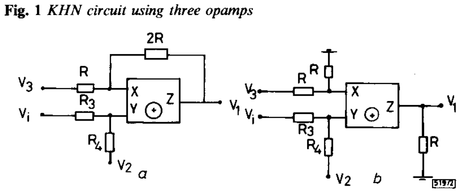

Fig. 2 *Two equivalent summer circuits using CCII*

Fig. 1 represents the KHN biquad using three opamps. The aim here is to have an exact equivalent circuit with three outputs, highpass, bandpass and lowpass, using CCIIs. The approach is based on realising the three basic building blocks, namely the summer and the two integrators [10], using CCIIs, and then connecting these blocks correctly.

Fig. 2 represents two new equivalent summer circuits having the output voltage V_1 given by

$$V_1 = 2V_i\left(\frac{R_4}{R_3+R_4}\right) + 2V_2\left(\frac{R_3}{R_3+R_4}\right) - V_3 \qquad (1)$$

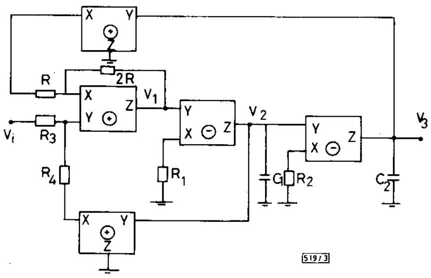

Fig. 3 *First KHN circuit using CCII*

The above expression is identical to that obtained using the opamp four resistor summer employed in the KHN biquad as can be seen from Fig. 1 (assuming an ideal opamp). The realisation of the inverting integrator using an inverting CCII, a grounded resistor and a grounded capacitor is well known [10]. Although the summer output and the two integrators can be easily cascaded, the integrator outputs must be buffered before being connected to the proper summer inputs. Here two more CCIIs are used acting as voltage followers [1] as seen in Figs. 3 and 4. The transfer functions of the circuits of Fig. 3 are given by

$$\frac{V_1}{V_i} = \frac{\frac{2R_4}{R_3+R_4}s^2}{D(s)} \qquad (2)$$

$$\frac{V_2}{V_i} = \frac{-\frac{2R_4}{R_3+R_4}\frac{s}{R_1C_1}}{D(s)} \qquad (3)$$

$$\frac{V_3}{V_i} = \frac{+\frac{2R_4}{R_3+R_4}\frac{1}{R_1R_2C_1C_2}}{D(s)} \qquad (4)$$

where

$$D(s) = s^2 + \frac{2R_3}{R_3+R_4}\frac{s}{R_1C_1} + \frac{1}{R_1R_2C_1C_2} \qquad (5)$$

The above equations are identical to those of the circuit of Fig. 1. The design equations are as in the opamp case and are given by

$$C_2 = C_1 = C \qquad R_1 = R_2 = \frac{1}{\omega_o C} \qquad (6)$$

$$R_4 = (2Q-1)R_3 \qquad (7)$$

The circuit of Fig. 4*b* will have the same equations if we take $R_5 = R_6 = R_7 = R$. Another design however is to take $R_7 = \infty$ (the circuit in this case will not be an exact equivalent to the KHN in Fig. 1). In this case

$$\frac{V_2}{V_i} = \frac{-\left(\frac{R_4}{R_3+R_4}\right)\left(\frac{R_5}{R_6}\right)\frac{s}{R_1C_1}}{s^2 + \left(\frac{R_3}{R_3+R_4}\right)\left(\frac{R_5}{R_6}\right)\frac{s}{R_1C_1} + \frac{1}{R_1R_2C_1C_2}\left(\frac{R_5}{R_6}\right)} \qquad (8)$$

Similar expressions can be obtained for the highpass and lowpass transfer functions. The design equations in this case are

$$\text{taking} \quad C_1 = C_2 = C \quad R_5 = R_6 \quad R_1 = R_2 = \frac{1}{\omega_o C} \qquad (9)$$

$$R_4 = (Q-1)R_3 \qquad (10)$$

In this design the resistor ratio spread is $(Q-1)$ and the magnitude gain at ω_o at each of the three outputs is $(Q-1)$, instead of $(2Q-1)$ in the previous design. Other designs are possible due to the available degrees of freedom but are not discussed here.

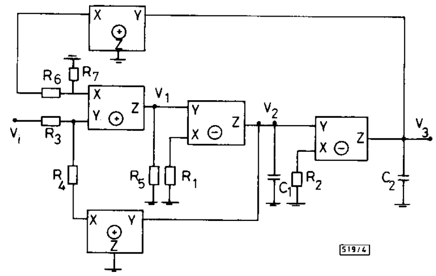

Fig. 4 *Second KHN circuit using CCII*

Conclusion: It is seen that with the current conveyor two equivalent circuits to the KHN biquad are generated, based on the equivalent building blocks . Each circuit requires five current conveyors, two of which however are used as buffers for the summer. Both circuits have the two capacitors grounded which is not the case in the opamp KHN biquad. One circuit requires one more resistor to be exactly equivalent to the KHN biquad. This resistor however can be removed resulting in an alternative design with a smaller spread in resistor ratio.

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