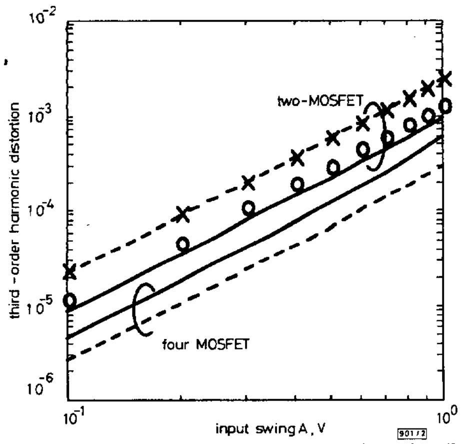

Fig. 2 *Third-order harmonic distortion calculated using charge-sheet (lines) and strong-inversion (symbols) models against input signal swing A*

C_{ox} = 1.5fF/µm², V_{FB} = –0.74V, ϕ_f = 0.37V, $\gamma = 0.6\sqrt{V}$, V_0 = 1V, f_0 = 1Hz, FFT sampling frequency = 16f_0;
two-MOSFET: V_C = 5V (—— and cir, r_{eq} = 0.34V^{-1}), V_C = 3.5V (– – – and ×, r_{eq} = 0.67V^{-1})
four-MOSFET: V_{C1} = 5V and V_{C2} = 4.25V (– – –, $_{req}$ = 1.36V^{-1}). V_{C1} = 5V and V_{C2} = 3.5V (——, r_{eq} = 0.68V^{-1})

The modified structure using two pairs of transistors with different gate voltages yields, when the differential input current is computed using the strong-inversion model, the ideal normalised linear resistance expression

$$r_{eq} = (V_{C1} - V_{C2})^{-1} \quad (5)$$

The complete cancellation of both even-order and odd-order terms is confirmed by the harmonic distortion analysis with the FFT. However, performing the analysis with the charge-sheet model reveals that the odd-order terms are far from being fully suppressed and may not be neglected under practical operating conditions. In a typical 5V implementation, the use of the four-MOSFET integrator instead of the two-MOSFET integrator may at best provide an attenuation of the third-order nonlinearity by a factor of 3 – 5 (Fig. 2). In the case of the control gate voltage limited to 3V as for low voltage implementations, the four-MOSFET integrator only provides an improvement over the two-MOSFET integrator if the DC input bias is scaled below 0.5V, thereby limiting the input swing to <1V peak-to-peak (Fig. 3).

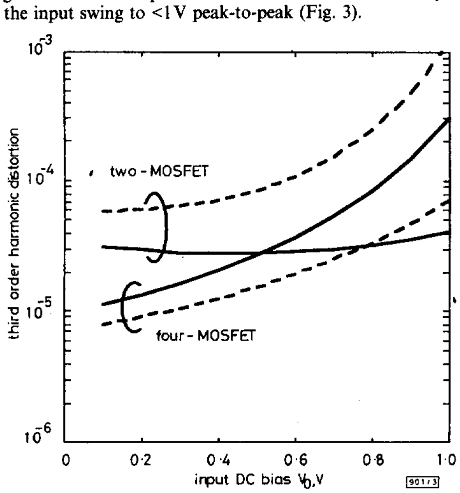

Fig. 3 *Third-order harmonic distortion calculated using charge-sheet model against DC input bias V_0*

A = 0.1 V;
two-MOSFET: V_C = 3V (——, r_{eq} = 0.62V^{-1}), V_C = 2.15V (– – –, r_{eq} = 1.26V^{-1})
four-MOSFET: V_{C1} = 3V and V_{C2} = 2.55V (– – –, r_{eq} = 2.3V^{-1}) V_{C1} = 3V and V_{C2}= 2.15V (——, r_{eq} = 1.2V^{-1})
Other parameters as in Fig. 2

The origin of the phenomenon lies in the physical slight variation with V_G of the surface potential in strong inversion. In the charge-sheet model, eqn. 3 accounts for this dependence of Ψ_s on V_G-V, whereas in eqn. 1, Ψ_s is fixed at $2\phi_f + V$. It is well-known that this dependence decreases for greater values of V_G-V, confirming the trends observed in Figs. 2 and 3. Since V_T is then a function of V_G and V_0, eqn. 5 should be rewritten as $r_{eq} = ((V_{C1} - V_{T1}) - (V_{C2} - V_{T2}))^{-1}$ with $V_{T1} \neq V_{T2}$. This underlies the existence, origin and impact of odd-order nonlinearities in the four-MOSFET structure.

Conclusions: Harmonic distortion analysis of the four-MOSFET linear resistance configuration using the charge-sheet MOS I-V model demonstrates the previously ignored existence of odd-order nonlinearities related to the body effect and clarifies their origin. These effects are not systematically nor significantly attenuated when compared to the two-MOSFET structure.

28 June 1995
Electronics Letters Online No: 19951016

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Current mode universal filter

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Indexing terms: Current-mode circuits, Active filters

A low sensitvity current mode filter circuit which realises highpass, bandpass and lowpass responses is proposed.

Very recently, a voltage mode universal filter using current conveyors (CCIIs) has been introduced [1]. In this Letter, a current mode bandpass-lowpass filter which employs three CCIIs, two grounded capacitors and three grounded resistors is given. To realise a highpass current response, a fourth CCII is added with a slight modification to the circuit. Both current-mode circuits have the following attractive features:

(i) very low input impedance
(ii) very high output impedance
(iii) independent control of Q without affecting ω_0
(iv) very low ω_0 and Q sensitivities to all circuit components.

Fig. 1 shows a new current-mode bandpass-lowpass filter which employs two grounded capacitors, three grounded resistors and three CCIIS. The two-output CCII is defined by the following matrix equation:

$$\begin{bmatrix} V_x \\ I_y \\ I_{z_1} \\ I_{z_2} \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ K & 0 & 0 & 0 \\ -K & 0 & 0 & 0 \end{bmatrix} = \begin{bmatrix} I_x \\ V_y \\ V_{z_1} \\ V_{z_2} \end{bmatrix} \quad (1)$$

where K can be either +1 or –1. If $K = 1$, Z_1 is defined as the noninverting output and Z_2 as the inverting output and vice versa. The two-output CCII has been used in other applications [2, 3].

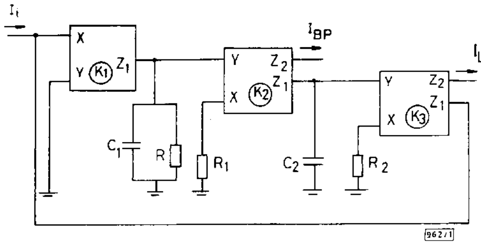

Fig. 1 *Bandpass-lowpass current-mode filter*

By direct analysis, the current transfer functions are obtained as

$$T_{BP} \triangleq \frac{I_{BP}}{I_i} = \frac{\frac{K_1K_2}{C_1R_1}s}{D(s)} \quad T_{LP} \triangleq \frac{I_{LP}}{I_i} = \frac{\frac{K_1K_2K_3}{C_1C_2R_1R_2}}{D(s)} \quad (2)$$

where

$$D(s) = s^2 + \frac{s}{C_1R} + \frac{K_1K_2K_3}{C_1C_2R_1R_2} \quad (3)$$

The product $K_1K_2K_3$ must be positive for stable transfer functions. The lowpass transfer function is always noninverting, whereas the bandpass transfer function can be either noninverting or inverting depending on the sign of K_1K_2 (which is the same as the sign of K_3). Table 1 includes the four possible sign combinations of K_1, K_2 and K_3 and the corresponding polarities of the transfer functions.

Table 1: Four possible sign combinations of K_1, K_2 and K_3 and corresponding polarities of transfer functions

Polarity CCII			Sign $T_i(s)$	
K_1	K_1	K_1	BP	LP
+	+	+	+	+
+	–	–	–	+
–	+	–	–	+
–	–	+	+	+

The ω_0 and the Q sensitivities to the circuit components are given by

$$S^{\omega_0}_{R_1} = S^{\omega_0}_{R_2} = S^{\omega_0}_{C_1} = S^{\omega_0}_{C_2} = -\frac{1}{2}$$
$$S^{\omega_0}_{R} = 0 \text{ and } S^{\omega_0}_{K_i} = \frac{1}{2} \quad (i = 1,2,3) \quad (4)$$

$$S^{Q}_{R_1} = S^{Q}_{R_2} = -\frac{1}{2}$$
$$S^{Q}_{R} = 1 \text{ and } S^{Q}_{K_i} = \frac{1}{2} \quad (i = 1,2,3) \quad (5)$$

$$S^{Q}_{C_1} = -S^{Q}_{C_2} = -\frac{1}{2} \quad (6)$$

From eqn. 3 and with $|K_i| = 1$ (i = 1,2,3), we obtain

$$\omega_0 = \frac{1}{\sqrt{R_1R_2C_1C_2}} \quad (7)$$

$$Q = R\sqrt{\frac{C_1}{R_1R_2C_2}} \quad (8)$$

For a specified ω_0 and Q, the design equations are given by

$$C_2 = C_1 = C \quad (9)$$

$$R_2 = R_1 = \frac{1}{\omega_0 C} \quad (10)$$

$$R = \frac{Q}{\omega_0 C} \quad (11)$$

The resistor R controls the Q of the filter response without affecting ω_0. The lowpass DC gain equals 1, whereas the magnitude of the bandpass gain at ω_0 equals Q.

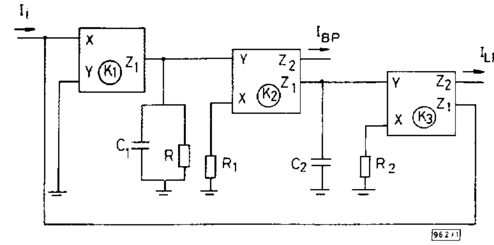

Fig. 2 *Universal current-mode filter*

Fig. 2 represents the universal current mode filter, with transfer functions given by

$$\frac{I_{HP}}{I_i} = \frac{K_1 s^2}{D(s)} \quad \frac{I_{BP}}{I_i} = \frac{\left(\frac{K_1K_2}{C_1R_1}\right)s}{D(s)} \quad \frac{I_{LP}}{I_i} = \frac{\left(\frac{K_1K_2K_3}{C_1C_2R_1R_2}\right)s}{D(s)} \quad (12)$$

where $D(s)$ is given by

$$D(s) = s^2 + \frac{K_1K_4}{C_1R}s + \frac{K_1K_2K_3}{C_1C_2R_1R_2} \quad (13)$$

The ω_0 and Q sensitivities are given by eqns. 4 – 6 with the following additional or modified equations:

$$S^{\omega_0}_{K_4} = 0 \quad (14)$$

$$S^{Q}_{K_1} = -\frac{1}{2} \text{ and } S^{Q}_{K_4} = -1 \quad (15)$$

Clearly, both K_1K_4 and $K_1K_2K_3$ must be positive; that is, K_4 must have the same sign as K_2K_3. The four possible sign combinations for the four CCIIs, and the signs of the current transfer functions, are given in Table 2. The current I represents another bandpass output with a unity gain at ω_0. The design equations for this circuit are the same as given by eqns. 9 – 11, and the highpass response has a high frequency gain magnitude of unity.

Table 2: Four possible sign combinations for four CCIIs, and signs of current transfer functions

Polarity CCII				Sign $T_i(s)$		
K_1	K_1	K_1	K_1	HP	BP	LP
+	+	+	+	+	+	+
+	–	–	+	+	–	+
–	+	–	–	–	–	+
–	–	+	–	–	+	+

A noninverting notch response can be realised by using a fifth CCII acting as a current summer for I_{HP} and I_{LP} with the CCII polarities as given by any of the first two combinations shown in Table 2. An all-pass current response can also be obtained by taking the current I to this summer after inverting it (using one more noninverting CCII).

 3 July 1995

Electronics Letters Online No: 19951018

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