Current controlled bandpass filter based on translinear conveyors

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Indexing terms: Bandpass filters, Current conveyors

A new concept to take advantage of the parasitic resistance that appears on port X of the second-generation current conveyors is introduced. This parasitic resistance, which is controllable in current, leads to the definition of the second generation current controlled conveyors (CCCII). A current controlled bandpass filter, operating in the current mode, is also described. It uses only two CCCII⁺s and two capacitors. Its central frequency can be adjusted by acting on the bias current of the conveyors. SPICE simulation results, in agreement with theory, are given for central frequencies around 30MHz.

Introduction: Second-generation current conveyors (CCII) have been introduced in 1970 [1]. They are now widely used for the implementation of high performance electronic functions operating either in the voltage-mode or current-mode [2, 3]. Thus, when they have been designed from translinear elements, mixed loops and complementary current mirrors, they allow implementation of electronic functions usable at frequencies in the megahertz range [2, 3]. However, at port X, these conveyors present a serial parasitic resistance that will be labelled R_X. This resistance is not negligible when the bias current I_0 is low (for example $R_X = 140\Omega$ at $I_0 = 100\mu A$). When this parasitic resistance is not taken into account, either conversion errors or incorrect frequency responses are experienced (in the case of filtering, for example when this resistance is found in series with a capacitor).

We will introduce the second generation current controlled conveyor (CCCII), deduced from the CCII. The expression of R_X as a function of the bias current I_0 is then explained. A bandpass filter, operating in the current mode, that uses ofIln two CCCIIs and two capacitors is then presented. The bias currents of the conveyors, that control the intrinsic value for their corresponding resistance R_X, allow us to modify the location of the central frequency of the filter. SPICE simulation results in agreement with theory will be given for the filter centred around 30MHz.

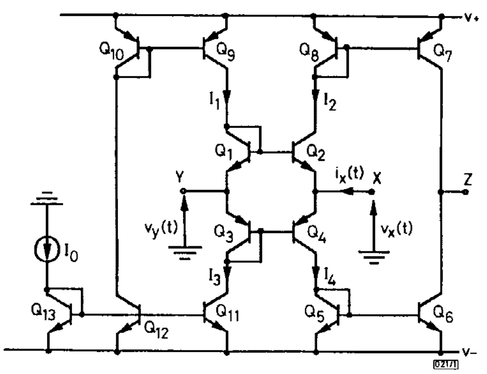

Fig. 1 *Schematic implementation for positive current controlled conveyor (CCCII⁺)*

Current controlled conveyors: Fig. 1 represents the schematic implementation of the positive current controlled conveyor (CCCII⁺). I_0 is the DC bias current of the circuit. The notations which are used for the input-output ports (X, Y, Z) are those usually used for the CCII⁺ [1, 4, 5]. Transistors Q_1 – Q_4 form the input translinear mixed loop that is characterised by the relationship $I_1 I_3 = I_2 I_4$ [6]. Assuming that the current gains β of the transistors are >1, the collector currents of Q_1 and Q_3 are equal to I_0. Thus Y is a high input impedance port. The previous relationship shows that the circuit, considered between points Y and X, is identical to a voltage follower (as $V_{xy} = 0$, in the absence of a load connected at X), [2]. When the current $i_x(t) \neq 0$, the voltage difference that exists between points Y and X, given by $V_{xy}(t) = -V_T \log(I_2(t)/I_0)$, allows us to determine the output resistance of the voltage follower. Therefore, after calculation of $I_2(t)$ [6] and by assuming that $|i_x(t)| < 2I_0$, the expression for this equivalent resistance is [5]

$$R_x = V_T/2I_0 \tag{1}$$

where V_T is the thermal voltage. This relationship shows that the value of this resistance can be modified by varying the DC bias current of the conveyor. A CCCII will also be obtained easily from the circuit in Fig. 1 by adding only two cross-coupled current mirrors, to reverse the current on output Z.

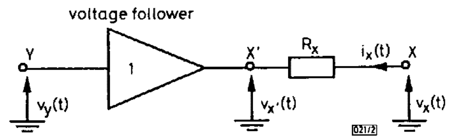

Fig. 2 *Equivalent circuit of translinear CCCIIs, seen between ports X and Y*

As a general rule, a current controlled conveyor (CCCII) is a CCII to which the possibility of modifying the value of the DC bias current I_0, will confer additional properties [5]. All the other characteristics of the CCIIs are preserved (and notably the matrix relation between its input and output variables [6]) provided that point X will be replaced, as shown in Fig. 2, by the intrinsic point X′.

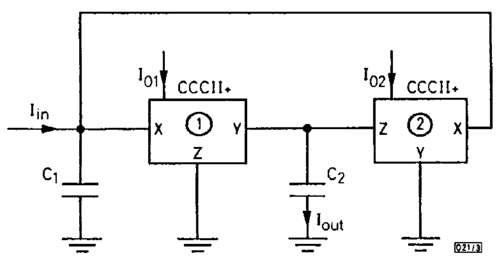

Fig. 3 *Current controlled bandpass filter implemented from CCCII⁺*

Current mode bandpass filter: Fig. 3 represents a second-order bandpass filter operating in current mode and using previously described conveyors of Fig. 1. The symbol of the CCCII⁺ is directly deduced from that of the CCII⁺ [3–5] by just adding the current I_0, which allows control of the value of the resistance R_X.

This filter contains only two capacitors and does not require any additional passive resistance Despite its relative simplicity, this implementation exhibits interesting properties and will be usable up to high frequencies. If I_{in} is the input current, the output current I_{out} which flows through C_1 will be carried out using a CCI⁺ with a virtual ground at the input, [2]. Furthermore, I_{out} can be directly obtained at high impedance, by simply adding a supplementary output Z to the second CCCII⁺. When both control currents I_{01} and I_{02} of the conveyors are equal to I_0, their intrinsic resistances have the same value R_X. Therefore, the transfer function of the circuit, whose gain at ω_0 is 1/2, will be characterised by

$$H(s) = \frac{I_{out}}{I_{in}}(s) = \frac{-R_x C_2 s}{1 + 2R_x C_2 s + R_x^2 C_1 C_2 s^2} \tag{2}$$

$$\omega_0 = (1/R_x)(C_1 C_2)^{-1/2} \tag{3}$$

$$Q = 1/2(C_1/C_2)^{1/2} \tag{4}$$

Eqn. 3 indicates that the value of ω_0 is adjustable by I_0 without affecting the quality factor and the gain.

Simulation results and discussion: To evaluate the performance of the circuit, several SPICE simulations have been performed using the typical parameters for the bipolar transistors of the complementary high performance HF3CMOS process from SGS Thomson. ±5V power supplies were used with a magnitude of 10μA for I_{in}. The simulated values for the controlled resistance of the CCCII were found to be in good agreement with those deduced from eqn. 1 (the deviations were <10% for I_0 = 1 – 150μA). For the filter in

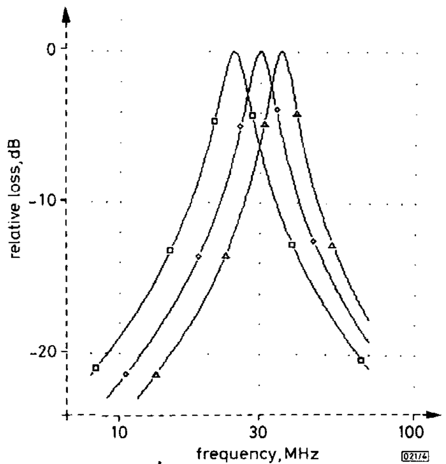

Fig. 4 *Frequency responses of filter*

□ I_0 = 40μA
◇ I_0 = 50μA
△ I_0 = 60μA

Fig. 3, low values for the bias currents I_0 allow us to obtain central frequencies beyond the megahertz range. For example, Fig. 4 represents the frequency responses obtained with the following values for passive components: C_1 = 130pF and C_2 = 2pF, using the successive values: 40, 50 and 60μA for I_0. Thus, the simulated values are 24.2, 29.6 and 35.0MHz, respectively, and correspond to the theoretical values (deduced from eqn. 3, taking into account the values of the parasitic capacitors on ports Y and Z of the CCCII⁺s, which appears in parallel with C_2; $C_{Y1} \simeq C_{Z2} \simeq 0.5$pF) 24.8, 31 and 37.2MHz, respectively. The corresponding values for Q vary from <6% around Q = 4.7, when I_0 changes from 40 to 60μA. Under the same conditions, variations of the gain are inferior to ±0.7dB around 6.7dB (theoretical value is –6dB).

Conclusion: Measurements performed with the above conditions for the CCCII fabricated by SGS Thomson confirmed the versatility of the controlled conveyor. Thus, R_X could easily be adjusted by changing I_0, from R_X = 1.2kΩ for I_0 = 10μA, to 28Ω for I_0 = 500μA. The circuit, which acts in the AB class [2], also benefits from low harmonic distortion. As an example, for any value of I_0, the total harmonic distortion of the voltage across R_X with Y grounded, was always found to be <1% for peak-to-peak magnitudes of the input current applied at X, as high as $2.54I_0$. Using two of these conveyors, a second-order bandpass filter operating in the current mode has been proposed. Its central frequency can be adjusted by modifying the control current I_0 of the conveyors. SPICE simulation results show that this implementation can be used at high frequencies, with the other advantage of low power consumption (4mW for f_0 = 30MHz, ±5V supplies and I_0 = 50μA).

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Minimum-phase filter design from linear-phase startpoint via balanced model truncation

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Indexing terms: Digital filters, Model order reduction

A new practical design approach for minimum-phase FIR or IIR filters, setting out from a high dimensionality FIR linear-phase prototype is described. The novelty of this technique lies in overcoming the inherent problem of finding the roots of a high order polynomial with repeated and/or very closely clustered roots.

The design of minimum-phase FIR or IIR filters through the indirect technique of starting from a high order linear-phase FIR prototype is attractive from several standpoints: flexibility, ease and speed of design, and abundant availability of design techniques and tools. The approach we employ here is simple and straightforward. Having designed the high order linear-phase FIR filter, we find its Z-plane roots, fold all zeros which lie outside the unit circle to their reciprocal-radius position inside (a process we refer to as 'mipizing') and generate the resulting minimum-phase FIR filter coefficients by multiplying the mipized roots back out into the polynomial form. This may turn out to be a lightweight task if the order of the FIR polynomial is low, or if it does not exhibit repeated or closely bunched roots. However, this is rarely the case. The problem of root finding is well established and off-the-shelf solutions such as found in [1, 2] are in existence and deliver acceptable results most of the time. We assume that our root finder has been successful in getting the roots of the desired polynomial and we have undertaken the necessary manipulations on these roots to mipize them. Our problem does not end here. As is the case in most digital filter design and implementation applications, the roots of the filter polynomial are of importance and of practical use in the design procedure, but most often they are the coefficients of the filter themselves which are employed in its realisation. To obtain to the coefficients, the final assembly procedure of multiplying out of the mipized roots is performed. This task can easily be performed in Matlab using the *poly* function [2]. This approach, although delivering satisfactory results most of the time, sometimes falls prey to numerical inaccuracies, and delivers results that are unacceptable.

In this Letter, we propose an alternative indirect approach based on balanced model truncation (BMT) [3, 4] to address the problem of arriving at a minimum-phase FIR or IIR filter from a linear-phase prototype, and demonstrate more favourable results compared to the conventional approach outlined above.

The rationale behind this new approach is to take the initial high order linear-phase FIR and convert it into a much lower order IIR equivalent [3, 4], while retaining its magnitude and phase characteristics faithful to (within acceptable error) the original prototype FIR. Now that a lower order IIR equivalent of the original FIR filter is available, the root-finding procedure is invoked on this smaller IIR, in the same manner as was done in the conventional direct approach for the FIR. However, the problems associated with the high order of the FIR polynomial are substantially reduced, by virtue of the reduced (IIR equivalent) we are now operating on. The procedure to follow now to obtain the minimum-phase filter is identical to that explained earlier. We are at this stage faced with two choices:

(i) implement this minimum-phase filter as an IIR filter, or

(ii) undertake an IIR to FIR transformation through long division, which is essentially identical to truncating the impulse