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Single-Amplifier Biquadratic MOSFET–C Filters

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circuit consisting of three CCII+ and two resistors with equal resistance. Such a circuit is shown in Fig. 2.5.

CCIIα_i: arbitrary current gain

In general, the CCII with positive or negative current gain α_i is described by the current equation $i_z = \alpha_i i_x$ [Schmid97, Schmid99b]. It is universal for any non-zero α_i. To prove this for positive α_i, it suffices to show that the circuit in Fig. 2.5 is a CCII− if the two resistances are chosen such that the overall current gain becomes one. The same circuit can also be used to prove universality for negative α_i: just use the Z terminal of the top right CCII as the output of the composite CCII−. (The bottom right CCII can then be omitted.)

CCI and CCIII: X input current mirrored to Y

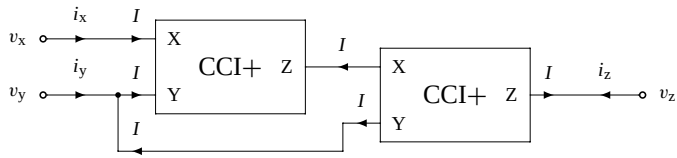
The CCII was originally derived from a device introduced as "the current conveyor," which is now called *first-generation current conveyor*, or CCI+. The CCI+ is described by the following three equations [Smith68]:

$$i_y = i_x,\ v_x = v_y,\ i_z = i_x\ . \tag{2.5}$$

To prove that it is universal, it is sufficient to show that a CCII− can be built using two instances of the CCI+. One way to do this is shown in Fig. 2.6. Defining $i_x = I$ and drawing this current I wherever it occurs makes it obvious that the circuit in Fig. 2.6 meets Eqs. (2.4) and thus is a CCII−. Other current conveyors similar to the CCI+ are the CCI− ($i_z = -i_x$), the CCIα_i ($i_z = \alpha_i i_x$), and the third-generation current conveyor, CCIII ($i_y = -i_x$, c.f. [Fabre95]), or, more generally, the CCIIIα_i. All first- and third-generation current conveyors are universal amplifiers, which can in every case be shown by a constructive proof, as for the CCII+ and the CCI+. Finally, it is also possible to choose a non-unity current gain from X to Y, i.e., to choose $i_y = \pm\alpha_j i_x$. The resulting amplifier is universal for any α_j.

VICC: voltage inverter instead of voltage buffer

A further idea is to use a voltage inverter instead of a voltage buffer at the input of any of these current conveyors, such that $v_x = -v_y$. It is not clear yet what kind of applications current conveyors containing a voltage inverter may have, we only include this case for the sake of completeness, and also because this functionality was used to build a filter (but not explicitly described) in [Chiu96, Fig. 10]. We propose the name *voltage-inverting current conveyor* (VICC) for such devices. Current conveyors of all three generations can be built



CCII− built using two CCI+. **Figure 2.6**

with a voltage inverter, thus there exist VICCIs, VICCIIs and VICCIIIs. All are universal, since two VICCs can be used to build one normal current conveyor, namely by using its voltage inverter to convert the inverting Y terminal to a non-inverting one. Note that using two VICCIs or two VICCIIIs gives a CCIII, whereas two VICCIIs give a CCII. Further research will show whether the VICCs are actually useful for network synthesis.

Twelve classes of current conveyors

It depends on the viewpoint how many different current conveyors our classification contains. If non-unity gains are just seen as a generalisation of a given current conveyor, then there exist twelve different current conveyors named according to the scheme *x*CC*yz*, where *x* is either "VI" or nothing to denote the polarity of the voltage buffer, *y* is either "I", "II", or "III" to denote the polarity or the absence of a Y-terminal current, and *z* is "+" or "−" to denote the polarity of the output current buffer.

Extended current conveyors

More universal amplifiers based on these twelve current conveyors can be derived by adding more current inputs and outputs (c.f. the balanced-signal CCII in [Schmid97, Schmid99b]) or more voltage inputs (c.f. the differential difference CCII in [Chiu96]). Like the extended operational amplifiers from Sec. 2.4, they are all trivially universal.

signals. The reasons for the difference are mainly the design preferences of the proponents of the current-mode approach.

3.4.3 Mixed-signal circuits

Advantages are open to debate

It has been pointed out that using mixed feedback (i.e. voltage to current or current to voltage) may result in speed advantages [Wilson92]. This is also open to debate. We will give two brief examples to illustrate the complexity of such comparisons:

Gm–C filters

Design advantage of Gm–C filters

Gm–C filters can reach higher frequencies than single-amplifier biquads (SABs), but they then also consume more power. From an overview of recently published Gm–C filters, it seems that it is easier to trade speed for power with Gm–C filters than with SABs (c.f. Chap. 8). However, there is still no fundamental reason for Gm–C filters to be faster than single-amplifier filters.

The current-feedback opamp

... has advantages and disadvantages ...

This device, which we already described in Chapter 2, was extensively discussed at the ISCAS 1993 [Bruun93, Bowers93, Franco93, Harvey93, Toumazou93]. The alleged advantages of the CFB opamp are that its bandwidth is very high and independent of the closed-loop gain, that it has no theoretical slew-rate limitation, and that its input-referred noise voltage is low compared to that of opamps. There are several applications in which the CFB opamp performs very well (c.f. Sec. 7.2). Its disadvantages are its inferior DC performance, the asymmetry of its inputs, the high input bias current necessary on the inverting input, and the dependence of its bandwidth on the feedback resistor [Bowers93]. Furthermore, the feedback cannot be capacitive, this would lead to stability problems [Franco93].

... but is not fundamentally better

There is always a trade-off between DC performance and bandwidth in opamps, and CFB opamps seem to be faster mainly because they are compared to voltage opamps having much better DC performance. Then, while the closed-loop gain is independent of the bandwidth, it is limited by the input resistance of the current-input terminal. Especially when the CFB opamp is set to its maximum bandwidth, the available

range of gain is surprisingly small. Furthermore, CFB opamps draw a considerable supply current under slewing conditions; thus, although the slew rate of the CFB opamp is indeed very high, it is set by the supply in practical applications. Many other problems were described in [Bowers93, Harvey93, Franco93].

Conclusion 3.5

The mode is not decisive

The notion of looking at circuits in terms of node impedances made it possible to derive a new, constructive proof of the circuit transposition theorem using signal-flow graphs. A discussion based on the same notion showed that there is no fundamental difference between current-mode and voltage-mode circuits. While it is true that many current-mode circuits live up to the reputed advantages of the current mode, the reason is not that current has been used as a signal, but that circuit simplicity, lower power consumption and speed are often achieved at the cost of higher distortion, higher gain variation, and so on.

On the difficulty of a valid comparison

What would happen if a designer set out to build a current-mode opamp that has approximately the same properties (CMRR, PSRR, linearity, chip area, etc.) as, e.g, the well-known opamp LM 741, but with the maximum possible speed? In the light of the above discussion, we believe that the speed would also be approximately the same, but until somebody tries this, which is not likely because the effort would be immense, the question will remain open.

Limit on the maximum pole frequency

The resistance of the low-impedance terminal therefore imposes fundamental limitations on the filter's pole frequency, and the highest achievable frequency for a given stopband attenuation is

$$\omega_{\mathrm{p}\max} \approx \frac{A_{\mathrm{pass}}}{\max(m,1/m)C_{\mathrm{o}} \cdot \max(n,1/n)R_{\mathrm{i}} \cdot A_{\mathrm{stop}}}\,, \tag{4.32}$$

which reaches a maximum at $m = n = 1$. Since the capacitors Y_4 and Y_2 must match well, Y_4 should not consist of parasitic capacitance only, and $\omega_{\mathrm{p}\max}$ should therefore not be approached too closely.

Bad high-frequency behaviour

Bandpass filter (BP1). Here the complex pair of zeros causes the TF to *rise 20 dB per decade* at frequencies above ω_{z}, until it flattens out again, at a gain of 1, because of a third high-frequency pole, which was cancelled from the Taylor series during the simplifications made above. Since $\omega_{\mathrm{z}}/\omega_{\mathrm{p}}$ is in the order of $\sqrt{\rho}$, the filter's gain reaches unity at a frequency of about $\rho\omega_{\mathrm{p}}$. *This may well make the filter useless for practical applications.*

Higher $\omega_{\mathrm{p}\max}$ than BP1

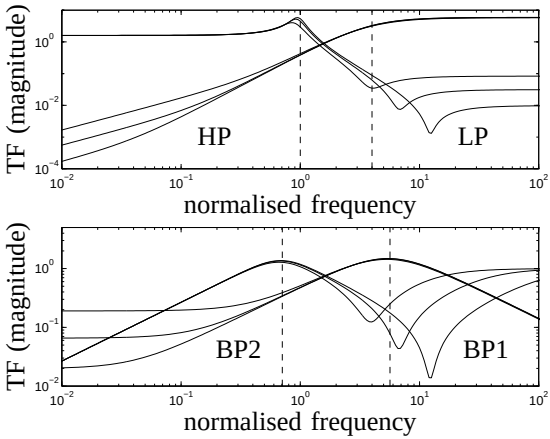
Bandpass filter (BP2). The single zero makes the TF constant for frequencies below ω_{z}, at a magnitude of approximately $\sqrt{2}\,\rho m$. Here it is a matter of convenience and interpretation to which level this should be referred, but the same fundamental frequency limitations occur as in the LP case.

No closed-form design expressions

High-pass filter (HP1). In this case, the single zero changes the slope of the TF from 40 dB per decade to 20 dB per decade for frequencies below ω_{z}. Again, the minimum capacitance to be used in the feedback network and the filter specifications impose frequency limitations, although in this case the dependence of the maximum frequency on the specifications is more complicated and is best evaluated graphically or numerically.

Examples

To clarify the above discussion, Fig. 4.12 shows the transfer functions of all four filters, where $m = 0.6$, $n = 1$, $\alpha_{\mathrm{I}} = -1.6$, $\kappa = 30$ and $\rho = 10, 30, 100$. The magnitudes of HP and BP2 have been multiplied by 4, and different pole frequencies have been chosen, both for graphical reasons only. The effects of the parasitic zeros can be seen clearly in all four cases. It is also evident that the LP filter has by far the highest q_{pi}, which already follows from (4.28_{LP})–(4.28_{HP}).



Transfer functions (TF) of the LP, BP1, BP2 and HP filters. The dashed lines indicate the different ω_{pi}. **Figure 4.12**

Practical example 4.4.4

Filter with medium q_p

As an example, consider a Sallen-and-Key low-pass filter biquad with $f_\mathrm{p} = 16.58\,\mathrm{MHz}$, $q_\mathrm{p} = 4$, and a stopband attenuation of at least 30 dB.[2]

A single-ended CMOS class AB second-generation current conveyor (CCII) is used as current amplifier. It is similar to the balanced CCII presented in [Schmid97], which is the balanced variant of the CCII shown in Fig. 2.10. Simulations show that the current input of the CCII has a resistance on the order of $100\,\Omega$, depending on the bias current, while the current output has a capacitance of $C_\mathrm{o} \approx 0.05\,\mathrm{pF}$.

Choose α_I, m and n

The choice of "optimum" values of m, n and α_I really depends on which sensitivity criterion should be optimised (c.f. Sec. 4.3). Here we choose reasonable values according to the criteria given in Section 4.4.2 without further explanation: neglecting the passband attenuation ($A_\mathrm{pass} \approx 0\,\mathrm{dB}$), and assuming $\max(m, 1/m) \approx 2$ and $\max(n, 1/n) \approx 1.25$, it follows that the input resistance of the CCII must be $R_\mathrm{i} = 240\,\Omega$. Then

[2] Although it is rather small, this attenuation already results in 60 dB stopband attenuation for a cascade of two biquads in a 4th-order filter.

through other methods that the main contribution to the total harmonic distortion (THD) is of odd order, then this method is a viable alternative to the much more time consuming method of using a spectrum analyser to measure the THD individually for every magnitude-frequency pair.

5.9 Conclusion

MOSFET–C SABs work

The discussion in this Chapter, together with the measurements provided in Chapter 7, show that MOSFET–C single-amplifier biquadratic filters work and are indeed a viable alternative to classical Gm–C filters. Three important questions remain open.

Suitability for mixed-signal integration

The filters we tested reject the substrate noise generated by the charge pump very well. This fact and the perfectly symmetrical structure of the filter gives rise to the assumption that they are well suited for use on mixed-signal circuits. It has yet to be shown that this is indeed the case by using our technique to integrate a true mixed-signal chip.

MOSFET–only filters

It was also shown above, by providing simulation results, that our technique can also be used to build MOSFET-only filters using a standard digital CMOS process (i.e., with one poly-silicon layer only). This possibility requires further investigations, since the THD will possibly have to be optimised using different criteria than the ones discussed in this chapter. It is also an open question whether *n*MOS or *p*MOS capacitors should be used, and if the latter are used, how the well should be polarised.

Higher-order MOSFET–C filters

Finally, it is also possible to build single-amplifier filters that generate three or more poles [Moschytz99a, Moschytz99b]. We think that this is feasible too and makes it possible to build even smaller filters with lower power consumption. However, the advance from Gm–C filters to MOSFET–C SABs is certainly much larger than the advance from MOSFET–C SABs to higher-order MOSFET–C single-amplifier filters.

Chapter 6

Implementation of the current amplifier

It is not likely that MAD circuit designers will be replaced by design tools in the foreseeable future.

(Yannis Tsividis)

This chapter is mostly descriptive, since the amplifier presented here is based on well-known concepts taken from the literature that were used to build symmetrical, balanced current amplifiers. Both the fixed-gain current amplifier and the variable-gain current amplifier presented in this Chapter can also be seen and used as second-generation current conveyors (CCIIs), as discussed in Chapter 2. Apart from a description of the amplifiers, this Chapter also discusses a few design criteria, and finally suggests improvements of the variable-gain current amplifier.

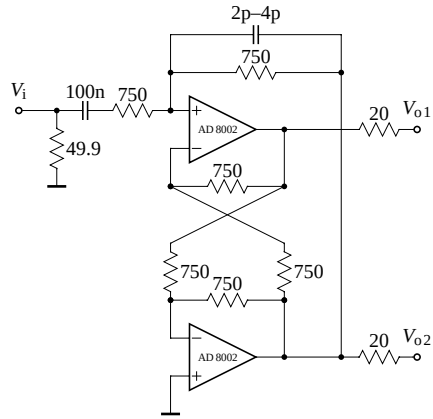


Figure 7.1 *Single-ended to balanced voltage converter using AD 8002 CFB opamps.*

enough chip pads were still free so that the inputs could be decoupled from each other and from the rest of the pads by placing grounded pads in between.

7.2.2 Output I–V converter

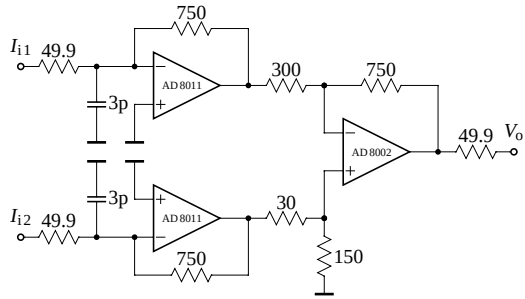
Two-stage transresistance amplifier

The current output of every test circuit was converted to a single voltage by the circuit shown in Fig. 7.2. It consists of tow independent I–V converters based on the AD 8011 (another CFB opamp) that has an $R_m = 750\,\Omega$. The following stage is a difference amplifier based on the AD 8002 with a voltage gain of 5. Together with the differencing, the overall R_m from a single current output to the converter output is $7500\,\Omega$. The reason that two different CFB opamps were used is that the AD 8011 is basically slower; because of the stability problems that often occur with high-speed amplifiers, it is not advisable to use amplifiers that are faster than necessary.

7.2.3 Measurement equipment

Brief description

All transfer functions and characteristics were measured with the 500-MHz spectrum analyser HP 8751 A; the noise and



Balanced-current to single-ended voltage converter using AD 8002 and AD 8011 CFB opamps. **Figure 7.2**

clock feed-through was measured with the 150-MHz spectrum analyser HP 3588 A. For the harmonic-distortion measurements, a 2-V_{pp} was generated with the Tektronix AFG 2020 function generator and then attenuated by a programmable attenuator, the Marconi MA 2186, in order to produce a harmonically clean signal for the measurements.

First test chip 7.3

V–I converter and signal inputs 7.3.1

Why an on-chip V–I converter is needed

Every circuit on the test chip that has a balanced current input is driven by an on-chip V–I converter that converts the balanced voltage input into a balanced current. The reason that such a converter is necessary is that otherwise the pad capacitance and the input resistance of the circuit would form a pole at unacceptably low frequencies (several MHz). This looks like a disadvantage of current-mode filters, but a voltage-mode filter that has an output impedance equal to the input impedance of the current-mode filter would simply have the same problems at its output, where a voltage buffer would have to be inserted.

Major mistakes on the first chip

On this first chip, two major mistakes were made. First, every circuit on the chip had its own pair of input pads. There were not enough pads remaining to isolate the input pads electromagnetically. As a result, there was a considerably

variance of the pole Q considerably compared to the pole-Q variance of an optimum design (c.f. 4.3).

8.3.6 Charge-pump or not?

Use a charge pump if possible

Finally, as has become apparent in Chap. 7, the advantages of having a charge-pump to drive the MOSFET resistor gates are so great that it should be done if possible. Also, the clock feed-through to the output of our filters is small enough for most applications. There are two things that could prevent the use of a charge pump.

Clock feed-through to other circuits

First, although our filters reject the substrate noise generated by the charge pump quite well, it must be made sure that the same is true for all other signal processing circuits on the chip. This may be a problem on purely analogue ICs, but is not really an issue on mixed-signal ICs, since there the substrate noise of the digital part dominates anyway.

Breakdown voltages

Second, the charge pump described in Sec. 5.5.2 is constructed so that although its output voltage can reach 5 V, no terminal voltage difference on any elements will exceed 3.3 V. Theoretically no break-down will occur even if the process used does not support 5 V as the 0.6-μm CMOS process by AMS does. The same is true for the MOSFET–C SABs. However, over-peaking during the transients might change this, and it must be made sure, by careful simulations, that the charge pump is compatible with the process at hand.

8.4 Conclusion

MOSFET–C filters are useful for video-frequency applications

We have shown in this dissertation that MOSFET–C SABs and filter cascades are a useful technique to build video-frequency filters. Their main benefit is that they require less chip area than conventional Gm–C filters having the same THD, SNR, and power consumption, typically the reduction is to 30…15 % of the size of the Gm–C filter. Since this PhD dissertation is, to our knowledge, the first comprehensive discussion of MOSFET–C SABs, many open questions still remain, which will be discussed in the following chapter.

Chapter 9

Ideas for future research

"Have you got an answer?"
"No, but I've got a different name for the problem."

(Douglas Adams)

Outline

As a conclusion of this thesis, we briefly discuss a few open questions and ideas for future research. They are mainly written for the benefit of the reader who wishes to apply MOSFET–C SABs, but also a list of directions in which the author's future research might go.

Suitability for MAD design

We have given plausible arguments, based on clock feed-through measurements, for the suitability of our filters for mixed analogue-digital (MAD) IC design. The reasons for this suitability is mainly the (theoretically) perfect balancing of the circuits. To be certain, however, this suitability has to be demonstrated by actually designing a MAD IC with a MOSFET–C SAB on it, or by a detailed discussion of the substrate-noise rejection and control-signal-noise rejection of our filters.

Amplifier Choice

In Chapter 2, we did not tell how a designer should actually choose the best amplifier for a certain application. This is always a difficult question, mainly because the definition of 'best' is very application specific. Also, the selection criterion for a designer is not which amplifier could be better from a theoretical point of view, but with which amplifier he personally can achieve better results. This is most probably the amplifier he is most familiar with. Thus, if the new and less well known amplifiers discussed in Chap. 2 should become viable candidates for applications, they must first be researched

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The author's biography

Hanspeter Schmid went to primary and secondary school in Seengen, Switzerland. In 1983 he lost both hands in an explosives accident, and had to repeat one school year to compensate for the long stay in hospital. From 1985–1988, he went to the Kantonsschule[1] Aarau, where he got his Maturität of type C [2].

He studied Electrical Engineering at the ETH[3] in Zürich, with a one-year break after the third year, during which he worked as a development engineer at Camille Bauer AG (Wohlen, Switzerland) and went to Edinburgh to study English. He received his diploma in electrical engineering in 1994 and the post-graduate degree in information technologies in 1999.

Hanspeter Schmid joined the Signal and Information Processing Laboratory of the ETH Zürich as a teaching assistant in 1994 and started his PhD studies late in 1995. After finishing this dissertation, he starts working as an analogue-IC designer for Bernafon, Switzerland, but continues being a senior lecturer in the field of analogue integrated filters at ETH Zürich. He also serves the IEEE Circuits and Systems Society as a member (currently the Secretary) of the Analog Signal Processing Committee.

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