On Computer-Aided Optimization of RC-Active Filter Designs

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Introduction

It is common practice to design RC-active filter circuits by treating operational amplifiers (OA's) as ideal devices. This simplification, however, can lead to significant deviations from the desired frequency response - primarily due to the finite gain-bandwidth product (G·B) of real amplifiers. In the past, some methods have been proposed to reduce the influence of amplifier imperfections, e.g. passive/active phase compensation, prewarping and composite amplifier configurations. However, these techniques are not very effective since they consider only a simplified single pole frequency response of the OA or require dual matched amplifiers tending to instabilities. Moreover, real input and output impedances of the amplifiers are not taken into account.

In this contribution, a novel technique of „pole tuning" is presented for more effectively compensating amplifier imperfections.Utilizing suitable EDA software, supported by amplifier macromodels, the most relevant non-ideal OA parameters can be taken

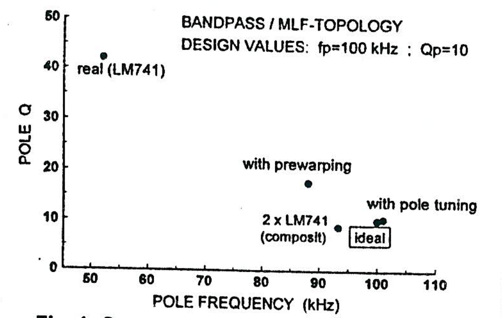

Fig. 1: Comparison of compensation methods

into consideration - as well as other error sources, e.g. deviations from calculated component values, load impedances or parasitic capacitances and inductances. By applying this technique the accuracy of the filter response is improved considerably. As illustrated in **Fig. 1** where the pole parameters Q_P resp. f_P are compared for different compensation methods it is possible to shift the dominant pole very close to the ideal position.

The method of „Pole Tuning"

The proposed technique is based on the „Substitution Theorem of Network Theory" [1] according to which an arbitrary branch *Z* of a time-invariant network can be replaced by an independent source without influencing node voltages or branch currents - provided the network matrix has a *singular* solution. Therefore, as far as voltage-to-current *ratios* (i.e. impedance values) are concerned, the theorem can be applied also to linear feedback systems which have for these ratios only *one* solution if the loop gain is *unity* (A_L=1). However, the idea behind the presented method applies the theorem in its *reverse* direction:

<u>Sentence</u>: *If an arbitrary element or branch Z of a circuit with feedback is substituted by a sinusoidal voltage source* V_Z *with a frequency* f_Z, *the voltage-to-current ratio* V_Z/I_Z *equals a complex impedance which can be used instead of the source in order to produce a loop gain* A_L=1 *at the frequency* f_Z

This property of feedback systems can be advantageously applied for correcting second-order filter stages, if the circuit to be „tuned" is part of an artificial loop as illustrated in **Fig. 2**. First, the circuit under concern (block NETWORK) is designed conventionally on the basis of *ideal* amplifiers using standard textbook routines or filter software.

Then, the parameters of the block TUNER are chosen in accordance with the Substitution Theorem (A_L=1 at the desired frequency f_Z).

Nevertheless, if the block NETWORK contains a circuit with *real* amplifiers, the loop gain will *not* be unity - unless a suitable part of the filter network is modified accordingly. Based on the Substitution Theorem, and with the aid of a program for analogue circuit analysis, the modifications necessary to correct this „error" are calculated within one simulation run only. As a result, the transfer function of the block NETWORK thereby is forced to assume its *ideal* value (in phase and magnitude) at frequency f_Z.

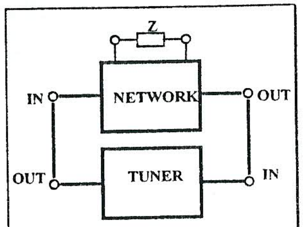

Fig. 2: Pole tuning loop

However, it is the primary aim of the procedure to shift the *pole* of the filter towards the ideal position ! Because, on the other hand, the pole location is directly related to the circuits *loop gain* at pole frequency f_P, the block NETWORK must contain, therefore, the filter circuit with an *open* feedback loop. Thus, the basic idea of the method is to set the filter's *loop gain* at f_Z=f_P to its proper value (in magnitude and phase), thereby correcting the pole parameters.

In the following, the „pole tuning" method is explained in detail by applying it to an example bandpass circuit. It is to be mentioned that an alternative tuning technique - applicable to *lowpass* filters only - is described in [2].

Example

The proposed optimization technique has been succesfully applied to several well-known filter topologies. As an example, **Fig. 3** shows a second-order bandpass stage with a bridged-T feedback network (multiloop-feedback, MLF). With a specified center frequency f_0=f_P=100 kHz and employing a 741-type OA (f_T≈1MHz) the pole-to-transit frequency ratio is f_P/f_T≈0.1. Note that this ratio is higher by a factor of 10 than usual recommendations [3].

Using standard design formulas or suitable filter software the following set of element values is calculated (OA ideal) for a relative bandwidth of 10% (Q=10):

R1=964.58 Ω R2=50.77 Ω

R5=19.29 kΩ C3=C4=1.65nF .

However, since the optimization process comprises also element deviations, the values as given in **Fig. 3** (out of the E24 series) are used as a starting point. As expected, a first simulation run based on a three-pole LM741-macromodel (National Semiconductor Inc.) reveals serious deviations from the specified frequency response (f_P≈52 kHz; Q_P≈42), see **Fig. 1.**

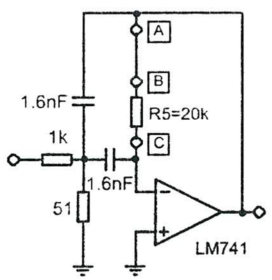

Fig. 3: Example circuit: MLF-Bandpass

As mentioned earlier, bandpass optimization requires adjusting the loop gain at f_P . Note that the MLF-topology is one of the exceptional cases where the OA is *not* used as a controlled source with local feedback. Therefore, the *ideal* loop gain of the filter approaches infinity, and only one sub-loop (in this case branch R5) may be considered.

(Hint: The same restriction applies to the popular GIC-block.)

Opening the sub-loop of R5, point A in **Fig. 3**, and simulating the circuit at f_P using an ideal OA (controlled voltage source with a gain as high as 1E12) yields *ideal* sub-loop gain values of A_L=1.0244 and ϕ_L=5.5725 deg.

To produce a closed loop gain of unity the TUNER, therefore, must have the two transfer parameters: $A_T=1/A_L=1/1.0244$ and $\phi_T=-\phi_L=-5.5725$ deg.

For simulation purposes, the following artificial transfer function is recommended: $H_{TUNER}=A_T\cdot\exp[\phi_T\cdot\pi\cdot s/(180\cdot|s|)]$ (s: complex frequency variable).

Additionally, to preserve proper loading of the OA output by the R5 branch the TUNER subcircuit should contain at its input a correct *mirror* of this branch.

A universal model can be used for this purpose in form of an artificial current source : $I_{LOAD}=I_T \cdot A_L \cdot \exp[\phi_L \cdot \pi \cdot s/(180 \cdot |s|)]$
with I_T being the current flowing through the TUNER output .

Next, one element of the feedback network must be replaced by a voltage stimulus V_Z to find at $f_Z = f_P$ the ratio I_Z/V_Z (or V_Z/I_Z), which gives the real and imaginary parts of a complex conductance (resp. impedance) to be used instead of the selected element. This is the most important and most critical design step of the procedure as it requires some understanding of the filtering function of the circuit! In this context, the following criterion has to be applied.

<u>Selection criterion</u>: *Any circuit modification with the aim of correcting the passband region of the filter must not influence the DC as well as the high-frequency behaviour. Therefore, even after substitution of an element by any parallel or series resistor-capacitor combination the filter function must still approach zero for very low and very high frequencies.*

This criterion can be fulfilled by replacing R5 by a parallel combination $R_P||C_P$ because, after that, there is still 100% feedback for low and high frequencies. Thus, running an AC analysis of **Fig. 3** at f_P=100 kHz, with the TUNER block inserted between points A and B and with a source V_Z between B and C, gives the elements: $R_P=V_Z/\mathrm{Re}(I_Z)=6.21\mathrm{k}\Omega$ $C_P=\mathrm{Img}(I_Z)/(V_Z \cdot 2\pi \cdot f_P)=20.53\mathrm{pF}$.
(Remark: In case of *negative* results another branch fulfilling the selection criterion is to be used. Otherwise, the deviations are too large to be compensated.)

Thereafter, replacing R5=20kΩ by $R_P||C_P$ (as calculated above) another AC analysis of the filter circuit confirms that the modified frequency response closely matches the ideal characteristic exhibiting the following parameters (see also **Fig. 1**): $f_0=f_P$=101.4 kHz and Q_P=10.4 .

Conclusion and summary

A simple computer-aided method of „tuning" the dominant pole pair of second-order bandpass stages has been introduced. Since all calculations are performed by a circuit analysis program, amplifier macromodels provided by manufacturers for a variety of active devices can be used to advantage (e.g. operational amplifiers, current-feedback amplifiers, transconductance amplifiers, current conveyors). If necessary, these macromodels can be enhanced by additional application-specific elements like load impedances and pin or socket capacitances.
Moreover, the proposed compensation technique can take into account also element deviations from their calculated values.
The usefulness of the method has been illustrated by means of an example circuit. All results obtained through computer simulations can be transferred to real circuits as long as the utilized amplifier macromodels reflect the actual device characteristics properly.
Finally, it should be mentioned that the desribed method is to be regarded not only as an *alternative* to some other „conventional" optimization techniques (see Introduction). For example, as can be seen from **Fig. 1**, it seems to be most promising to apply the „pole tuning" procedure to filter circuits containing either actively phase compensated or composite amplifiers.

References

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[2] v.Wangenheim, L.: „A simple method of incorporating amplifier imperfections in RC-active filter designs", Int. Journal of Electronics, 1995, pp. 653-665

[3] Herpy, M. & Berka, J.: Aktive RC-Filter, Franzis Verlag, 1984, Kap. 5.4.2