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ANALOGUE FILTERS WITH ELECTRONICALLY CONTROLLABLE ACTIVE ELEMENTS

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ANALOGUE FILTERS WITH ELECTRONICALLY CONTROLLABLE ACTIVE ELEMENTS

ANALOGOVÉ FILTRY S ELEKTRONICKY ŘIDITELNÝMI AKTIVNÍMI PRVKY

SHORT VERSION OF HABILITATION THESIS

VYSOKÉ UČENÍ TECHNICKÉ V BRNĚ
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Keywords

ACA, Adjustable Current Amplifier, analogue filter, active element, behavioral modeling, Balanced-Output Transconductance Amplifier, BOTA, Current-Gain Voltage Differencing Current Conveyor, CG-VDCC, CMOS, controllable filter, Digitally Adjustable Current Amplifier, DACA, Dual-Output Current Amplifier, DO-CA, fractional-order filter, Modified Current Differencing Unit, MCDU, Multiple-Output Current Follower, MO-CF, Operational Transconductance Amplifier, OTA, reconfigurable filter, transistor-level model, Voltage Differencing Transconductance Amplifier, VDTA, Z-Copy Current Gain Voltage Differencing Current Conveyor, ZC-CG-VDCC, Z-Copy Voltage Controlled Current Follower Differential Input Transconductance Amplifier, ZC-VCCFDITA.

Klíčová slova

ACA, Analogový filtr, aktivní prvek, behaviorální modelování, BOTA, CG-VDCC, CMOS, DACA, digitálně řiditelný proudový zesilovač, DO-CA, řiditelný filtr, řiditelný proudový zesilovač, filtr fraktálního řádu, proudový sledovač s více výstupy, proudový zesilovač s dvěma výstupy, MCDU, MO-CF, modifikovaná proudová diferenční jednotka, operační transkonduktanční zesilovač, OTA, rekonfigurovatelný filtr, model na tranzistorové úrovni, napěťově diferenční proudový konvejor s proudovým zesílením, napěťově diferenční transkonduktanční zesilovač, transkonduktanční zesilovač s dvěma výstupy, VDTA, ZC-CG-VDCC, Z-copy napěťově diferenční proudový konvejor s proudovým zesílením, Z-copy napěťově řiditelný transkonduktanční zesilovač s diferenčním vstupem a proudovým sledovačem, ZC-VCCFDITA.

THESIS IS AT DISPOSAL

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Author graduated in three-year Bachelors programme and two-year Masters programme both at Faculty of Electrical Engineering and Communication (FEEC), Brno University of Technology (BUT) and both with red diploma. He obtained rector's award for his Master's thesis. His studies were finished in 2011 by defending of his doctoral (Ph.D.) thesis titled as 'Frequency Filters with Current Active Elements' within four-year Doctoral programme Teleinformatics also at FEEC BUT. Author is member of academic staff at Department of Telecommunications BUT from 2007.

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Scientific and research activities of author are focused especially on area of fundamental research and most importantly on Czech Science Foundation projects. He very intensively deals with area of design of analogue frequency filters, design methods of these filters and design of new active elements suitable especially for filters in so-called current mode. Jan Jerabek participated in synthesis of four prototypes of subsequently manufactured active elements. Lately, he focuses especially on controllable and reconfigurable filtering structures and also on structures of so-called fractional order. Results of these research activities were published in many international journals or were presented at stage of many international conferences. He serves as a reviewer for many scientific journals and conferences and up to now he has evaluated 67 papers. Jan Jerabek is currently co-investigator of post-doctoral Czech Science Foundation project titled as 'Research for electronically adjustable advanced active elements for circuit synthesis' and he is currently also team member of two other Czech Science Foundation projects titled as 'Synthesis and Analysis of Fractional-Order Systems Using Non-Conventional Active Elements' and 'Active devices with differencing terminals for novel single-ended and pseudo-differential function block design'. He as well participated in three successfully defended projects of this category in the past. Jan Jerabek is junior scientist of SIX research centre at FEEC BUT and member of team of recently awarded COST EU project titled as 'Fractional-order systems: analysis, synthesis and their importance for future design'. To sum up, Jan Jerabek published 37 papers in journals with citation index according to the Web of Science (WoS), in case of 9 of them as main author. He also presented or co-authored 47 contributions at international conferences. There are 231 citations of his works according to the WoS database and h-index in WoS is currently 9. Note: all numbers are valid for March, 21st 2016.

1 INTRODUCTION

Filtering of a part of the frequency spectrum of the signal is a basic operation of suppressing or sometimes also amplifying of some spectral components or parts of processed bandwidth. Filters are essential in a huge variety of electronic systems: noise reduction systems, demodulators, signal detectors, multiplexors, anti-aliasing circuits, sound and speech processing systems, transmission lines equalizators, image processing systems etc. Some kind of a filter is a part of almost every electronic system.

Digital techniques dominate the signal processing area and therefore also play significant role in the domain of filtering of undesired spectral components. Although very old and well known, analogue filtering techniques are still being researched and are still very important for interaction of modern electronic systems with real environment and natural phenomena. Natural environment is essentially analogue and it is usually mandatory or at least much easier for electronic system to interact with it via analogue input and/or output circuits, including frequently frequency filters.

Domain of analogue filters splits in to two fundamental areas: passive filters and active filters. Area of active filters still evolves in several directions. One of the most significant paths is the usage of newly introduced active elements instead of classic Operational Amplifiers (op-amps) or simple Operational Transconductance Amplifiers (OTAs). With respect to the previously published solutions with op-amps and OTAs, this way usually extends variability, controllability or adaptability, brings possibility to operate up to higher or down to lower frequencies, decreases power consumption or harmonic distortion, extends dynamic range or simply fits better for particular application.

Approximations represent methods for finding the ideal positions of poles and zeros of a transfer function of the filter in order to poses particular features in magnitude, phase and time domain. Mathematical basics of these calculations were introduced decades and sometimes also hundreds years ago. Traditionally, integer-order approximations are used for filter design, with order starting from one and most commonly having the second order. With introduction of fractional-order circuits and thus also fractional-order filters, novel approximations were introduced, usually following properties of integer-order prototype functions in certain part of the frequency band. Note that probably the most frequently used approximation in both the integer-order and fractional-order filtering responses is Butterworth approximation, for its maximally flat magnitude response in the pass band and therefore it is also used all over this thesis.

2 THESIS OVERVIEW

2.1 Motivation

Active elements are very important parts of communication systems for analogue signal processing, amplification, filtering, generating, mixing and shaping of signals in almost all electronic devices [2], [23]. The analogue way is still used in many cases and the search for new active elements and approaches is an up-to-date topic for many research teams worldwide. The classical op-amp has been very important and practically available active element in synthesis of analogue circuits for decades [8]. Lack of electronic and external adjustability of the parameters of circuit using op-amps is a fundamental drawback. Therefore, other active elements were defined and began to be used in circuit synthesis and design. These novel active elements are still being developed and interesting modifications or elements based on non-traditional principles are still appearing. Active elements with controllable features have received an increasing attention in recent years [3]. However, there is still space for further research or significant improvements.

Novel approaches in active element conceptions could bring interesting results in the field of standard circuit synthesis or dynamical non-linear circuits unconventional design and modeling of real physical systems [47]. Analogue solutions represent an interesting topic for current research because: requirements for variability and adaptability of complex electronics systems increase; electronic adjustability of parameters in an analogue solution allows to compensate the influences of temperature, fluctuances of supply voltage, etc. simply and dynamically; modern approaches in design of active elements allow targeting all requirements of control in frame of active elements; analogue solutions are able to offer lower costs, higher efficiency in many cases (e.g. in case of high frequencies) and lower power consumption (in case of on-chip implementation) in comparison with digital solutions; a digital solution has a certain delay of processing and therefore these systems are not able to response immediately.

Electronic control has been a topic of interest for many years. We can divide electronic control in analogue circuits into two groups. The first group includes solutions where control (referred to as indirect) employs digital potentiometers, field effect transistors (FETs) as controllable resistor in linear region or other controllable active elements (transconductor for example), which replace grounded or floating resistors. Although researched deeply, several problems remain. Approaches, where the parameters of an controllable active element are directly controlled by an external parameter (bias or control DC current or voltage), form the second group. Many of these elements are still only hypothetical, they are not commercially available or have not been investigated yet nor used for circuit synthesis [3]. The presented results are sometimes based on computer simulations with trivial models only. There are several special types of commercially available active parts (e.g. [14], [56]) together with several active elements developed at our workplace (e.g. [51], [66]†) that are suitable for behavior modeling of hypothetical complex elements and most importantly for experimental verification. A correctly and sufficiently operating behavioral model promises and predetermines usually even better features of future on-chip implementation.

2.2 Goals

This thesis has the ambition to provide basic overview of the progress in analogue filtering with particular active elements. The main part of this text focuses on the presentation of several novel active elements and direct electronic control of parameters in filtering systems with help of tuning of internal parameters of these active devices. The aims can be identified as follows:

- to *provide* readers with resources covering the *fundamentals and state of the art* of analogue filtering.
- to *present* and *analyse* parameters of several recently introduced active elements with possibilities of current and voltage gain electronic control or control of other parameters within the frame of active element.
- to *synthetize* and *design* novel and suitable structures of frequency filters with possibilities of control of its parameters. The first group of filters features electronic controllability, i.e. one or more of basic parameters of the structure should be dependent on electronic tuning. The second group (reconfigurable circuits) represents solutions that provide the possibility of the change of transfer function without physical/galvanic reconnection of input or output terminal by appropriate electronic control of the particular parameter(s) of active elements. The third group of filters is of fractional-order type and these structures also have some of its parameters controllable by external parameter.
- to *verify* features of designed structures by appropriate analyzes, simulations or in particular cases also by laboratory measurements. The obtained results are *discussed*, *evaluated* and *compared* with another solutions. This aim is completely covered in full version of this habilitation thesis. In this short version, unfortunately, it has to be partly omitted because of limited space.

2.3 Contribution

The text is written to have not only scientific but also pedagogical contribution. Therefore, it should serve not only to experts in the field of analogue frequency filtering, but also to students interested in this topic in order to gain not only basic knowledge but also an overview of current state of the art and to get familiar with some of the latest achievements.

Relation to Author's Other Publications

The goal is to base this thesis on the most recent achievements of the author. Therefore, some original parts of this thesis are currently at the state of submission for publication, some of them are expected to be most likely accepted soon and then published, some of them are accepted for publication but are not published yet. However, the most of the results are already published. None of the results presented in this thesis were included in the author's Ph.D. thesis or any past author's theses.

Research described in the following chapters is partially subject of the recent Czech Science Foundation research project (No. GP-14-24186P, from 2014 to 2016). This research, however, builds on results of three previous fundamental research grants, in which Jan Jerabek also actively participated.

2.4 Structure

The Chapter 3 describes the state of the art and explains the latest achievements in the area of analogue filtering focused on Controllable Filters (Section 3.1), Reconfigurable Filters (Section 3.2) and Fractional-Order Filters (Section 3.3). Mutual comparison of selected filtering solutions published in literature and of particular solutions included in this thesis is provided in each of these sections. Chapter 4 introduces basic active elements together with advanced active elements based frequently on modular approach, i.e. consisting of several basic active elements. It is not possible to include detailed parameters of particular solution of each of advanced active elements especially in this short version of the thesis, therefore appropriate references where models or transistor-level structures and results were published are provided. Chapter 5, 6 and 7 respectively presents briefly 5, 2 and 2 designed frequency filters respectively, based on active elements introduced in the previous chapter. These filters are capable of electronic control of at least one of its parameters and most of them are universal (Chapter 5), reconfigurable (Chapter 6) or of fractional-order (Chapter 7). Chapter 8 concludes this thesis a summarizes the main achievements.

3 STATE OF THE ART

Many scientific works focus on filters referred to as multifunction or universal [37], [61], [11], [62], [59], for instance. These filters have several transfer functions available between different nodes of the network. These circuits are referred to as a triple input - single output (TISO) [53] or single input - triple output (SITO) [35], or more generally as a single input - multiple output (SIMO) [28] or a multiple input - single output (MISO) [24]. The most general form is multiple input - multiple output (MIMO) type [10], usually with many input and output terminals. These structures are sometimes too much general and its parameters are limited by the complexity of the structure. When a filter is universal, each of five standard transfer functions, i.e. low pass (LP), high pass (HP), band pass (BP), band stop (BS) and all pass (AP), is available from the same topology by proper selection of position of input, output or by reconfiguration in case of reconfigurable filters. Universal and reconfigurable filters are also useful for design libraries of analogue circuits and designer can then implement only one required transfer function in particular solution. Note that many other relevant references are included in Table 3.1 introduced in the following section.

3.1 Controllable Filters

Possibility to change or adjust an analogue circuit (filter, oscillator or generator) is always very important requirement in the area of signal processing. A filter is adjustable or tunable if one or more of its parameters (angular pole frequency, quality factor, bandwidth, pass-band or stop-band gain) are controllable [39], [29], [46]. In order to be useful in practice, their control should be mutually independent. Controllability of the parameter of the filter can be achieved by driving of parameter of active element (transconductance g_{m} [52], transresistance R_{T} [19], intrinsic resistance R_{X} [63], current inter-terminal gain B [66]† or voltage inter-terminal gain referred to as A [32]). These parameters are most frequently controlled by external DC voltage or DC current. Sometimes more than one parameter of active device can be controlled in case of advanced active elements. The second possibility of control of filter's parameters that is usually not so beneficial, is to change value(s) of passive component(s), i.e. mostly resistors [11] usually replaced by their controllable substitutes (digital potenciometers, controlled FETs or sometimes by other active elements such as OTAs).

Selected filtering solutions published in literature and their basic parameters are summarized in Table 3.1. References included in this table prove the variability of published solutions in type and a number of active and passive elements in filter's topology, in type of (non)electronic control of one or more parameters of the filter and in type of standard filtering transfer functions that are provided directly from the proposed circuitry. Three solutions included in the table provide so-called dual-parameter control, i.e. one parameter of the filter can be controlled by control of two parameters, which can be also beneficial. Note that there are also other possibilities how to extend the tuning range of the filter – circuits referred to as shadow or agile filters were published in [30], for instance. It is obvious that a number of required active elements is frequently around two, however, one active element (usually more complex) is sufficient in many cases. The most popular type of control is by g_m of OTA active element, forming also frequently input or output part of advanced active elements. The most popular topology is SITO type having one input terminal and three output terminals providing LP, BP and HP responses – as well-known from Kerwin-Huelsman-Newcomb (KHN) filters [31]. Note that the selected active elements that are important for purposes of this thesis are described in Chapter 4 in more detail.

Tab. 3.1: Overview of selected controllable filtering solutions.

Reference	Year	Active elements (number of devices)	Passive components (number)	Controlled by	Type of topology	Available function(s)
[27]	2009	FD-CF (1)	C(2), R(3)	R	MIMO	HP, BP, LP
[42]	2010	OTA (2)	C(2)	g_m	SIMO	LP, BP, HP, BS, AP
[21]	2010	CCCFTA (1)	C(2)	g_m, R_X	SITO	LP, BP, HP, BS, AP
[48]	2010	MO-CCCCTA (3)	C(2)	R_X, g_m*	SIMO	LP, BP, HP, BS, AP
[5]	2012	VDTA (2)	C(2)	g_m	SITO	HP, BP, LP
[9]	2012	FDCCII (1)	C(2), R(2)	R	TISO	LP, BP, HP, BS, AP
[65]†	2012	CF (3), DACA (1)	C(2), R(2)	R, B	SITO	LP, BP, HP, BS, AP
[68]‡	2014	CG-VDCC (1)	C(2), R(1)	B, g_m, R	TISO	LP, BP, HP, BS, AP
[67]‡	2014	VDTA (2)	C(2), R(2)	g_m	TISO	LP, BP, HP, BS, AP
[28]	2014	CCCII (3)	C(2)	R_X, B	SIMO	LP, BP, HP, BS, AP
[57]	2015	CFTA (3)	C(2)	g_m	SIMO	LP, BP, HP, BS, AP
[6]	2015	CCCDTA (3)	C(2)	g_m, R_X	MISO	LP, BP, HP, BS, AP
[70]‡	2015	MO-CF (1), MCDU (2)	C(2)	g_m, B*	TISO	LP, BP, HP, BS, AP
[86]‡	2016	ZC-VCCFDITA (1)	C(2)	A, g_m, R_X	SISO	BP
[77]‡	2016	BOTA (2), DACA (2)	C(2)	B, g_m*	SITO	LP, BP, HP, BS, AP

Abbreviations of active elements from Table: BOTA – Balanced Output Operation Transconductance Amplifier, CFTA – Current Follower Transconductance Amplifier, CCCFTA - Current Controlled Current Follower Transconductance Amplifier, CF – Current Follower, FDCCII – Fully Differential Current Conveyor of the second generation, FD-CF – Fully-Differential Current Follower, MO-CCCCTA – Multiple-Output Current Controlled Current Conveyor Transconductance Amplifier, VDTA – Voltage Differencing Transconductance Amplifier, DACA – Digitally Adjustable Current Amplifier, CG-VDCC – Modified Controlled-Gain Voltage Differencing Current Conveyor, CCCII – Current-Controlled Current Conveyor of the second generation, CCCDTA – Current-Controlled Current Differencing Transconductance Amplifier, MO-CF – Multiple-Output Current Follower, MCDU – Modified Current Differencing Unit.
Previously unexplained symbol: * – dual-control of one parameter.

The solutions identified in Table by double dagger (‡) symbol represent filters presented in own papers and moreover discussed in more detail in this thesis. This identification is valid only in frame of this chapter. These solutions are identified and included in order to be comparable with other solutions in their basic parameters. Note that more details about their features, including advantages and disadvantages of these solutions are discussed at the end of the respective sections of this thesis.

3.2 Reconfigurable Filters

It is very interesting concept to use two-port structure having some special features without changing internal topology or a position of input or output port. This concept is used also in microwave techniques, for instance [7]. These topologies are referred to as single input - single output (SISO). These filtering solutions allow reconnection-less and switch-less control of the type of the transfer function. This control is usually carried out electronically by the parameter(s) of advanced active element(s), or by digital bus provided to control parameters of advanced active element [12]. Frequently, also tuning of the stop-band and/or pass-band attenuation that actually represents continuous change between particular transfer functions is available.

Many papers present filters referred to as reconfigurable [25], [58] but from the bandwidth or quality factor adjustability point-of-view (in the case of BP response most frequently). From more strict point of view, reconfigurable filter [36] represents the type of transfer function, where reconfigurability is available without any reconnection in the structure.

Controllability of the type of the transfer function of the filter can be achieved by driving the same parameters of active elements as mentioned in Section 3.1 for the case of controllable filters.

The basic parameters of selected solutions of reconfigurable filters from non-microwave domain are compared in the following Table 3.2. Note that there are both the first- and the second-order reconfigurable filters that are the most common types. Solutions identified by ‡ symbol represent again filters discussed in more detail in this thesis.

Tab. 3.2: Overview of selected reconfigurable filtering solutions.

Reference	Year	Active elements (number of devices)	Passive components (number)	Change of function by control of	Order	Reconfi-guration between functions
[78]†	2011	ECCII (1), VB (1)	C (2)	B	2	AP, BS
[79]†	2014	OTA (5), VB (1)	C (2), R (1)	B	2	AP, BS
[40]	2014	OTA (2)	C (1), R (1)	g_{m}	1	AP, HP, LP
[80]†	2014	OTA (2), ECCII (1)	C (1)	g_{m}, R_{X}, B	1	AP, LPZ, iDT, AZ
[81]†	2014	ZC-CG-VDCC (1)	C (1)	R_{X}	1	iAP, HP, DT
[83]†	2015	MCDU (1)	C (1)	B	1	AP, iHP, LP
[82]†	2015	MCDU (1)	C (1)	B	1	AP, HP, LP
[85]†	2015	OTA (4)	C (2)	g_{m}	2	AP, iBP, BS, LPZ, HPZ
[84]†	2015	OTA (4)	C (2)	g_{m}	2	AP, iBP, BS, HP, LPZ, HPZ
[75]‡	2016	VDCC (1), CA (1)	C (2), R (1)	B	2	LP, BS, HP, LPZ, HPZ
[76]‡	2016	ZC-CG-VDCC (1)	C (2)	B	2	iAP, iBS, iDT

Previously unexplained abbreviations from Table: ECCII – Electronically Controllable Current Conveyor of Second Generation, VB – Voltage Buffer, ZC-CG-VDCC – Z-Copy Controlled-Gain Voltage Differencing Current Conveyor, VDCC – Voltage Differencing Current Conveyor, CA – Current Amplifier. LPZ – Low Pass with Zero, iDT - inverting Direct Transfer, DT – Direct Transfer, AZ – Adjustable Zero, iHP - inverting High Pass, iBP – inverting Band Pass, HPZ – High Pass with Zero.

3.3 Fractional-Order Filters

Traditionally, integer-order analogue filters are studied. Transition from pass band to stop band of the filter is based on the integer-order approximation of the filter. Therefore, there is a step change of the order of the filter and their attenuation is equal to $20n$ dB/decade theoretically (n stands for order of the filter).

In case of the fractional-order filters, the change of the order of the filter is more precise (fractional) or even continuous, represented by $20 \cdot (n + a)$ dB/decade slope of the attenuation in case of the LP or HP responses, where n is non-zero unsigned integer and parameter a is defined by $0 < a < 1$. Note that some sources refer a as α [44], [45], but the meaning is exactly the same, and in the case of integer-order filters $a = 0$.

Design of fractional-order electronic systems, for instance passive components, frequency filters or oscillators is of great interest of many research groups, e.g. [15]. Fractional-order calculations are useful not only for electrical engineering but also for other scientific areas such as biology [18] or even agriculture [17].

There are two standard ways how to obtain fractional-order filter. The first one is to use fractional-order element (FOE) in the filtering structure, usually nowadays approximated by passive RC network [26], [13], for instance. Some filters are designed with respect to the passive prototype [1], some of them are of the KHN topology [43], [50], [1] or Sallen-Key topology [43], [50], [1]. These structures usually include two FOEs [50], [1]. FOE is usually designed in order to behave as fractional-order element of particular order in certain frequency band of interest, hardly ever exceeding two decades.

The second way how to obtain fractional-order filter is to approximate fractional-order transfer function by integer-order function with order higher than $(n + a)$ [55], [33]. The term $\boldsymbol{s}^a$ in the transfer function is approximated by a ratio of two second-order rational polynomials. This leads to common approximation of the transfer function of the fractional filter with order between 1 and 2 by a third-order function. It can be then implemented by various topologies, e.g. by the 3rd-order Follow the Leader Feedback (FLF) topology [55], [33], [73]‡, [74]‡, which is probably the most popular. Note that double dagger symbol (‡) has exactly the same meaning as in Tables above, i.e. it identifies the solutions discussed in more detail within this thesis. Before real fractional-order elements (not only emulators) are fabricated and commercially available, the second method is really interesting in order to prove the concept of the designed filter.

4 DESCRIPTION OF SELECTED ACTIVE ELEMENTS USEFUL FOR THE DESIGN

There have been reported many novel and advanced active elements in recent works, a detailed summarization is available in [3]. It is well-known fact that novel and advanced active elements are based only on several basic building blocks. Most of all, they are current followers (CFs) or current amplifiers (CAs), many times with differential or multiple input/output, voltage buffers (VBs) or voltage amplifiers (VAs) including inverting variants, voltage or current differencing units, transconductance stages, transresistance stages, current conveyors (CCs) or voltage conveyors (VCs). At least some of its parameters are usually controllable and in many cases there are also some auxiliary ports that widen possibilities of usage in the final application [3]. The following sections were prepared in order to describe briefly the most important active elements that are used as part of the structures presented in Chapters 5, 6 and 7. Note that behavioral models and/or transistor-level implementations of all presented active elements were omitted in this short version of thesis.

4.1 Current Conveyors

This section discuss briefly the particular types of Current Conveyors that are important for the purposes of this thesis. Note that detailed overview of all generations and types of Current Conveoyors is given in [47].

Current Conveyors are well-known active elements existing in three generations and having many particular variants. The most common type is basic Current Conveyor of the second generation (CCII) [49] having voltage input Y, current input X and one or more current output(s) designated as Z, see Fig. 4.1(a). Outer behavior of this active element (in case of CCII+/− variant) is described by the following matrix:

$$\begin{bmatrix} i_{\mathrm{Y}} \\ v_{\mathrm{X}} \\ i_{\mathrm{Z+}} \\ i_{\mathrm{Z-}} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & -1 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{Y}} \\ i_{\mathrm{X}} \\ v_{\mathrm{Z+}} \\ v_{\mathrm{Z-}} \end{bmatrix}. \tag{4.1}$$

Particular meaning of variables in matrix above is obvious from Fig. 4.1(a). Note that in frame of CCII+/− device there are no controllable parameters. However, this element forms important subpart of other active elements, such as Voltage Differencing Current Conveyor (VDCC) introduced in Section 4.5.

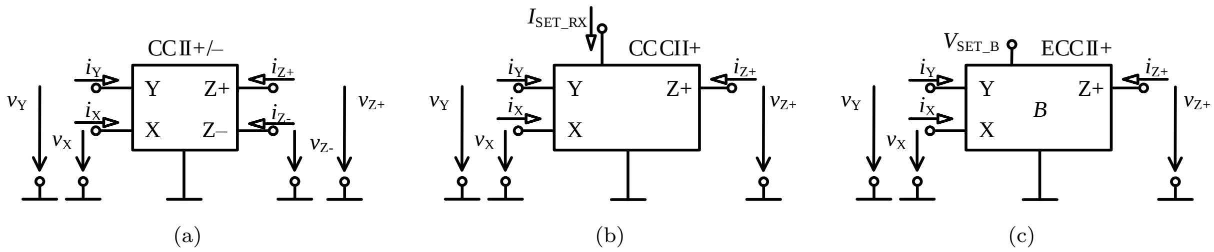


Fig. 4.1: Schematic symbols of the active elements: (a) CCII+/−, (b) CCCII+, (c) ECCII+.

Real implementation of CCII suffers from some parasitic features, especially the current input X has always non-zero input resistance in realistic scenarios. There are many works [48], [6], [28] (including references cited therein), that take this parasitic parameter as a part of the design of the circuit. It is referred to as intrinsic resistance R_{X} and its value is controlled electronically by bias DC current ($I_{\mathrm{SET\_RX}}$ in Fig. 4.1(b), where Current Controlled Current Conveyor of the second generation in CCCII+ variant is shown).

Electronically Controllable Current Conveyor of Second Generation (ECCII) [34], Fig. 4.1(c), differs from basic CCII by the possibility of electronic control of current transfer (gain, B) from current input X terminal to current output Z. Note that this concept can also be combined with R_{X} control [28]. B is electronically controlled usually by DC bias current or ratio of two currents [34], [28], or sometimes also by DC voltage ($V_{\mathrm{SET\_B}}$ in Fig. 4.1(c)) as in case of EL2082 IC from Intersil/Elantec [14], where numerically $B \approx V_{\mathrm{SET\_B}}$ is valid. However, note that EL2082 represents ECCII− type.

CCCII+ and ECCII+ are described respectively by:

$$\begin{bmatrix} i_{\mathrm{Y}} \\ v_{\mathrm{X}} \\ i_{\mathrm{Z+}} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & R_{\mathrm{X}} & 0 \\ 0 & 1 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{Y}} \\ i_{\mathrm{X}} \\ v_{\mathrm{Z+}} \end{bmatrix}, \qquad \begin{bmatrix} i_{\mathrm{Y}} \\ v_{\mathrm{X}} \\ i_{\mathrm{Z+}} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & B & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{Y}} \\ i_{\mathrm{X}} \\ v_{\mathrm{Z+}} \end{bmatrix}. \tag{4.2}$$

Control of current-input intrinsic resistance R_{X} is used also within the frame of other active elements, namely: Current Controlled Current Conveyor Transconductance Amplifiers (CCCCTA) [48], Modified Current Differencing Unit (MCDU) [70]†, described in more detail in Section 4.7, Current Controlled Current Follower Transconductance Amplifier (CCCFTA) [21] and Current-Controlled Current Differencing Transconductance Amplifier (CCCDTA) [6], for instance.

Concept of electronic control of current gain B is used similarly in the case of the following active elements: Adjustable Current Amplifiers (ACAs), discussed in Section 4.3, Current-Gain Voltage Differencing Current Conveyor (CG-VDCC), see Section 4.5 and also MCDU (Section 4.7).

4.2 Operational Transconductance Amplifier

Operational Transconductance amplifier (OTA) (Fig. 4.2) belongs to the group of well-known active elements that are widely used in analogue circuits [3]. It is characterized by the relation:

$$i_{\mathrm{OUT}} = g_{\mathrm{m}}(v_{+} - v_{-}). \tag{4.3}$$

Transconductance g_{m} can be controlled electronically by DC bias current ($I_{\mathrm{SET\_gm}}$ in Fig. 4.2). OTA is available also in variants with multiple outputs [20].

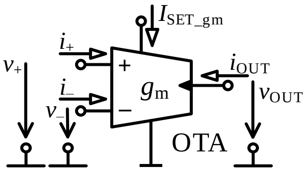


Fig. 4.2: Schematic symbol of OTA active element

Since OTA is really simple and popular active element, is is also used as a part of other active elements. The most important examples are: Voltage Differencing Transconductance Amplifier (VDTA), see Section 4.4, VDCC (Section 4.5), Voltage Controlled Current Follower Differential Input Transconductance Amplifier (VC-CFDITA), discussed in Section 4.6, CCTA [48] and many other derived variants of these active elements.

4.3 Adjustable Current Amplifiers

Adjustable Current Amplifiers (ACAs) represent special case of Current Followers with adjustable current gain. There are two common and well-known ways how to control its gain: analogue [66]† with theoretically unlimited number of available gain values in particular range, Fig. 4.3(a), and digital [41], [66]† with limited set of discrete values of gain, Fig. 4.3(b).

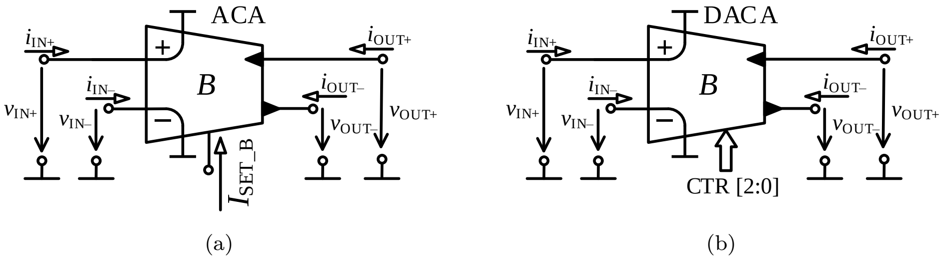


Fig. 4.3: Schematic symbol of the (a) Adjustable Current Amplifier (ACA) (b) Digitally Adjustable Current Amplifier (DACA). Both shown in fully differential variant.

Current transfers of both the ACA and DACA elements are given by the following relations:

$$i_{\mathrm{OUT+}} = B(i_{\mathrm{IN+}} - i_{\mathrm{IN-}}), \quad i_{\mathrm{OUT-}} = -B(i_{\mathrm{IN+}} - i_{\mathrm{IN-}}). \tag{4.4}$$

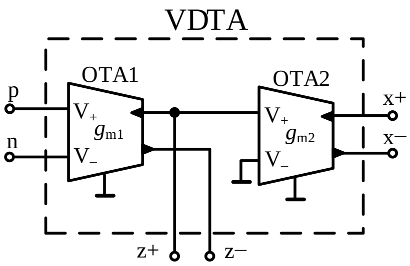
Note that the above described variants of Adjustable Current Amplifiers are presented as differential from both the input and the output side. These circuits are very much suitable for fully-differential (F-D) signal processing [66]†. However, also single-ended input and/or single-ended output variants are sufficient for many applications of ACAs [66]†, [68]†, [14]. Another possibility is to allow setting of independent gain for each of outputs, e.g. in case of Dual-Output Current Amplifier (DO-CA) [75]†. Note that this element can be internally based on interconnection of one MO-CF and two SISO ACAs [75]†.

ACAs are used in filtering applications not only as standalone electronically controlled active elements, but also as a part of advanced active elements. To mention some of them, Current-Gain Voltage Differencing Current Conveyor (CG-VDCC) (Section 4.5), Z Copy-Controlled Gain-Current Differencing Buffered Amplifier (ZC-CG-CDBA) [4] or MCDU (Section 4.7) should be listed as the examples.

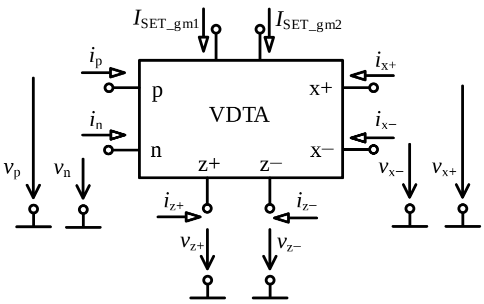
4.4 Voltage Differencing Transconductance Amplifier

The voltage differencing transconductance amplifier (VDTA) [60], [5] is another active block that was recently introduced. The VDTA is actually composed of two voltage-controlled current sources (represented by OTAs with electronically and independently controllable transconductances g_{m1}, g_{m2}) with the required number of outputs and these sub-circuits are interconnected internally.

The basic structure of one possible variant of VDTA active element is shown in Fig. 4.4(a). Block structure is depicted in simplified form not showing controlled currents that are actually shown in case of the schematic symbol that is placed as Fig. 4.4(b). Note that similar rule is going to be used in case of other active elements discussed later because of better readability of figures.



(a)



(b)

Fig. 4.4: Voltage Differencing Transconductance Amplifier (VDTA) with two independently electronically adjustable transconductances (a) principle block structure (b) schematic symbol.

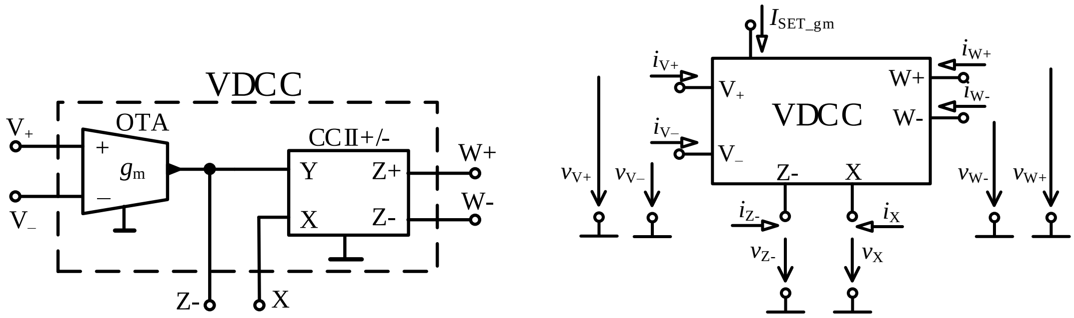
Outer behavior of the VDTA element is described by the following matrix:

$$\begin{bmatrix} i_p \\ i_n \\ i_{z+} \\ i_{z-} \\ i_{x+} \\ i_{x-} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ g_{m1} & -g_{m1} & 0 & 0 \\ -g_{m1} & g_{m1} & 0 & 0 \\ 0 & 0 & g_{m2} & 0 \\ 0 & 0 & -g_{m2} & 0 \end{bmatrix} \cdot \begin{bmatrix} v_p \\ v_n \\ v_{z+} \\ v_{z-} \end{bmatrix}. \tag{4.5}$$

Behavioral model can be prepared easily with help of two OTAs, as represented by block structure shown in Fig. 4.4(a). This was also the initial point for the design of the CMOS implementation of VDTA [67]† that is however omitted in this short version of thesis.

4.5 Voltage Differencing Current Conveyors

Simple Voltage Differencing Current Conveyor (VDCC) was initially reported in [3]. It consists of a transconductance amplifier (Section 4.2) connected to the Y terminal of standard current conveyor of second generation (Section 4.1). Its block structure and schematic symbol are depicted in Fig. 4.5.



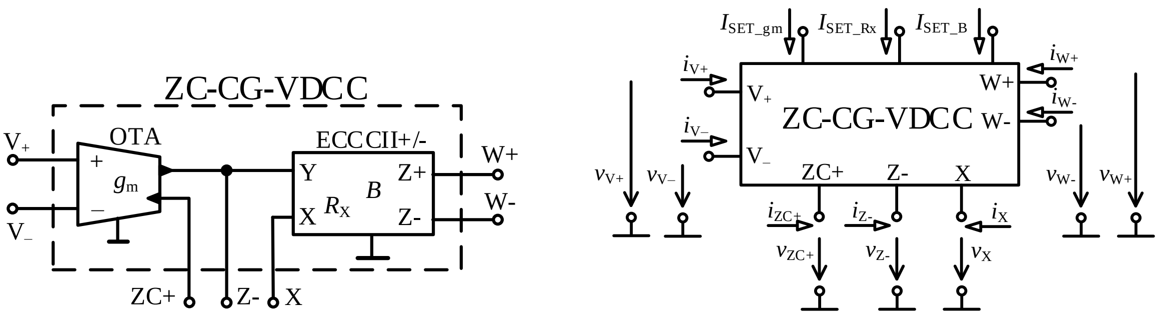
(a) (b)

Fig. 4.5: Voltage Differencing Current Conveyor (VDCC) (a) principle block structure with only single-output OTA section (b) schematic symbol.

Note that there is only one controllable parameter in frame of conventional VDCC (g_m of OTA section controlled by $I_{\mathrm{SET\_gm}}$ in this particular case). Simple VDCC current and voltage inter-terminal transfers are described by the following matrix:

$$\begin{bmatrix} i_{\mathrm{V+}} \\ i_{\mathrm{V-}} \\ i_{\mathrm{Z-}} \\ v_{\mathrm{X}} \\ i_{\mathrm{W+}} \\ i_{\mathrm{W-}} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ -g_{\mathrm{m1}} & g_{\mathrm{m1}} & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & -1 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{V+}} \\ v_{\mathrm{V-}} \\ v_{\mathrm{Z-}} \\ i_{\mathrm{X}} \\ v_{\mathrm{W+}} \\ v_{\mathrm{W-}} \end{bmatrix}. \tag{4.6}$$

Active element referred to as Z-Copy Controlled-Gain Voltage Differencing Current Conveyor (ZC-CG-VDCC) [81]† was derived from basic theoretical concept of VDCC. The main design goal was to allow simultaneous and mutually independent control of three parameters. In this particular case it is g_{m} of OTA section, R_{X} of current input terminal X in frame of the CCII section, and B between X and Z terminals also in case of the CCII section, as obvious from Fig. 4.6. Note that variant shown includes only one ZC terminal and its phase is opposite with respect to phase of Z terminal.



(a) (b)

Fig. 4.6: Active element with three tunable parameters: Z-Copy Controlled-Gain Voltage Differencing Current Conveyor (ZC-CG-VDCC) (a) principle block structure (b) schematic symbol including all three control currents.

When we consider features of ZC-CG-VDCC (meaning of all variables is obvious from Fig. 4.6):

$$\begin{bmatrix} i_{\mathrm{V+}} \\ i_{\mathrm{V-}} \\ i_{\mathrm{Z-}} \\ i_{\mathrm{ZC+}} \\ v_{\mathrm{X}} \\ i_{\mathrm{W+}} \\ i_{\mathrm{W-}} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ -g_{\mathrm{m1}} & g_{\mathrm{m1}} & 0 & 0 & 0 & 0 & 0 \\ g_{\mathrm{m1}} & -g_{\mathrm{m1}} & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & R_{\mathrm{X}} & 0 & 0 \\ 0 & 0 & 0 & 0 & B & 0 & 0 \\ 0 & 0 & 0 & 0 & -B & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{V+}} \\ v_{\mathrm{V-}} \\ v_{\mathrm{Z-}} \\ v_{\mathrm{ZC+}} \\ i_{\mathrm{X}} \\ v_{\mathrm{W+}} \\ v_{\mathrm{W-}} \end{bmatrix}. \tag{4.7}$$

ZC-CG-VDCC active element shown in Fig. 4.6 is quite complex and it is therefore useful to have possibility to verify the concept of active element in stage of behavioral model [71]†. After successful verification of the concept on the level of behavioral model, CMOS model can be designed [71]†.

Simple VDCC and ZC-CG-VDCC active elements presented can be further enhanced [68]†. The original ZC-CG-VDCC has three independently tunable parameters and one of them is B standing for current gain that is common for transfers from X to both W+ and W- outputs. The key feature of Modified CG-VDCC solution is that each of current gains can be controlled independently (B_1, B_2, B_3). Particular variant with three independently controlled current outputs (designated as $\mathrm{W_{B1}}$, $\mathrm{W_{B2}}$ and $\mathrm{W_{B3}}$) is shown in Fig. 4.7.

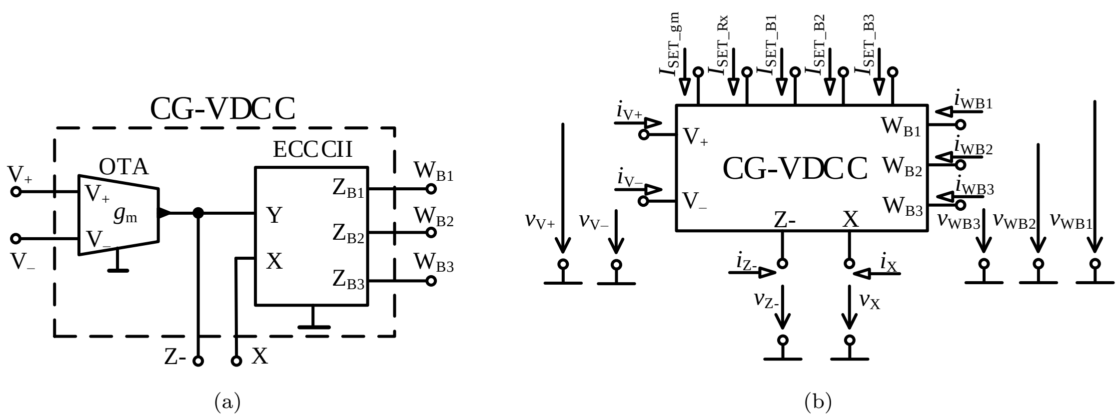


Fig. 4.7: Modified Controlled-Gain Voltage Differencing Current Conveyor (CG-VDCC) having three independently tunable current gains (a) principle block structure (b) schematic symbol including control currents.

Features of Modified CG-VDCC are presented in form of the following matrix:

$$\begin{bmatrix} i_{\mathrm{V+}} \\ i_{\mathrm{V-}} \\ i_{\mathrm{Z-}} \\ v_{\mathrm{X}} \\ i_{\mathrm{W_{B1}}} \\ i_{\mathrm{W_{B2}}} \\ i_{\mathrm{W_{B3}}} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ -g_{\mathrm{m1}} & g_{\mathrm{m1}} & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & R_{\mathrm{X}} & 0 & 0 & 0 \\ 0 & 0 & 0 & B_1 & 0 & 0 & 0 \\ 0 & 0 & 0 & B_2 & 0 & 0 & 0 \\ 0 & 0 & 0 & B_3 & 0 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{V+}} \\ v_{\mathrm{V-}} \\ v_{\mathrm{Z-}} \\ i_{\mathrm{X}} \\ v_{\mathrm{W_{B1}}} \\ v_{\mathrm{W_{B2}}} \\ v_{\mathrm{W_{B3}}} \end{bmatrix}. \tag{4.8}$$

Behavioral model of Modified CG-VDCC element can be easily derived from model of ZC-CG-VDCC. CMOS structure and its features can be studied in [68]†. Again, details have to be omitted because of limited space in this short version of habilitation thesis.

4.6 Z-Copy Voltage Controlled Current Follower Differential Input Transconductance Amplifier

The active element referred to as Z-copy Voltage Controlled Current Follower Differential Input Transconductance Amplifier (ZC-VCCFDITA) [69]†, [86]† is a more versatile derivative of the Generalized Current Follower Differential Input Transconductance Amplifier (GCFDITA) [22] and of other similar devices.

The main benefits of the ZC-VCCFDITA are controllable intrinsic resistance of the current input terminal (R_X), referred to as (R_F) in this particular case [69]†, controllable voltage gain (A) in frame of Variable Gain Amplifier (VGA) stage [56] and controllable transconductance (g_m) of the transconductance section. The explanation of the ZC-VCCFDITA element concept is given in Fig. 4.8, where variant with VGA having both voltage inputs non-grounded is presented. In other words, this ZC-VCCFDITA element consists of three blocks: current follower with two outputs while one of them is connected to OTA that represents second section and second input of OTA is sourced by VGA output that has differential input.

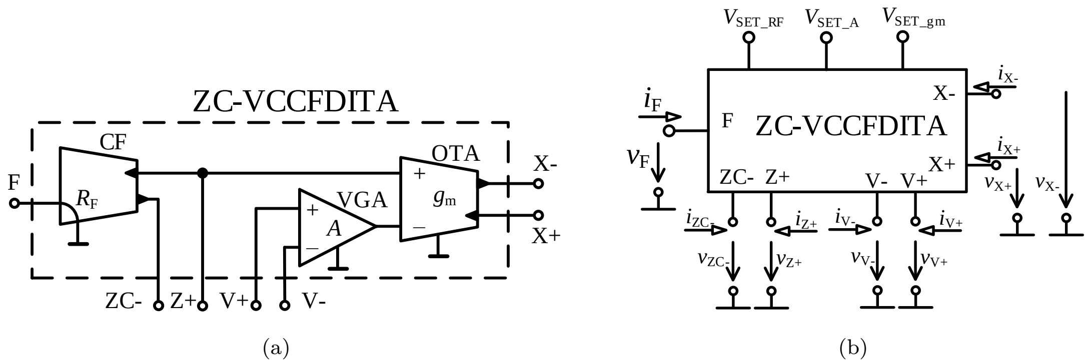


Fig. 4.8: Z-Copy Voltage Controlled Current Follower Differential Input Transconductance Amplifier (ZC-VCCFDITA) in variant with two auxiliary voltage inputs (a) principle block structure (b) schematic symbol.

Inter-terminal relations of ZC-VCCFDITA active element are described by the following matrix:

$$\begin{bmatrix} i_{\mathrm{Z}+} \\ i_{\mathrm{ZC}-} \\ v_\mathrm{F} \\ i_{\mathrm{V}+} \\ i_{\mathrm{V}-} \\ i_{\mathrm{X}-} \\ i_{\mathrm{X}+} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & -1 & 0 & 0 & 0 & 0 \\ 0 & 0 & R_\mathrm{F} & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ -g_\mathrm{m} & 0 & 0 & g_\mathrm{m}A & -g_\mathrm{m}A & 0 & 0 \\ g_\mathrm{m} & 0 & 0 & -g_\mathrm{m}A & g_\mathrm{m}A & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} v_{\mathrm{Z}+} \\ v_{\mathrm{ZC}-} \\ i_\mathrm{F} \\ v_{\mathrm{V}_+} \\ v_{\mathrm{V}_-} \\ v_{\mathrm{X}-} \\ v_{\mathrm{X}+} \end{bmatrix}. \tag{4.9}$$

Again, meaning of used the variables is obvious from Fig. 4.8. Behavioral model and its features verified both by the simulation and measurement are published in [69]†, [86]†.

4.7 Modified Current Differencing Unit

Modified Current Differencing Unit (MCDU) [83]†, [70]†, [72]† enhances the concept of Current Differencing Unit (CDU) [3] shown in Fig. 4.9(a). Block structure of MCDU is shown in Fig. 4.9(b). Key feature of MCDU, i.e. subtraction, is done after independent amplification of currents in this particular case.

If output current i_x is considered as flowing out of x terminal, as is common in case of CDUs, transfer is given by the following relation:

$$i_\mathrm{x} = i_\mathrm{p} - i_\mathrm{n}, \tag{4.10}$$

where i_p and i_n stand for currents flowing into p and n terminals respectively.

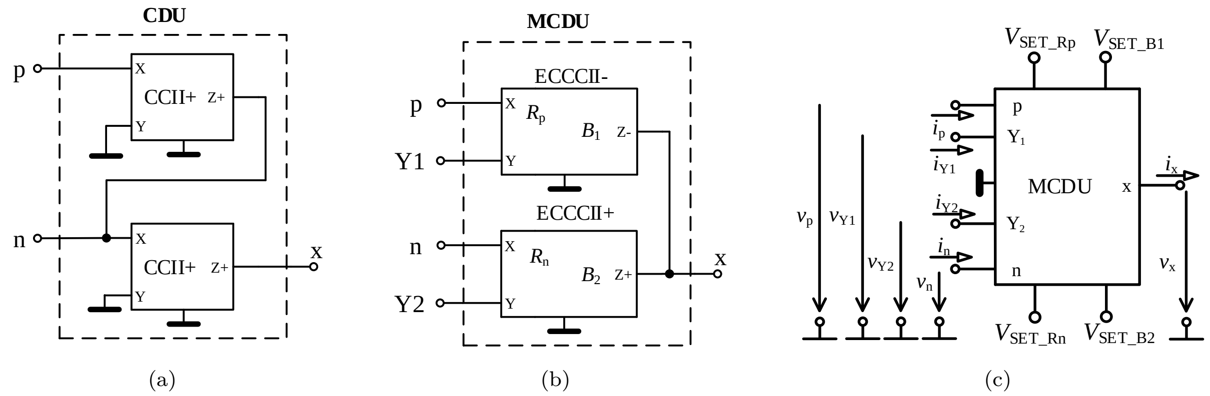


Fig. 4.9: Basic block structure of (a) Current Differencing Unit (CDU) (b) Modified Current Differencing Unit (MCDU) (c) schematic symbol of MCDU.

Schematic symbol of MCDU is shown in Fig 4.9(c), while its outer behavior is described by:

$$\begin{bmatrix} v_\mathrm{p} \\ v_\mathrm{n} \\ i_\mathrm{x} \\ i_\mathrm{Y1} \\ i_\mathrm{Y2} \end{bmatrix} = \begin{bmatrix} R_\mathrm{p} & 0 & 0 & 1 & 0 \\ 0 & R_\mathrm{n} & 0 & 0 & 1 \\ B_1 & -B_2 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} i_\mathrm{p} \\ i_\mathrm{n} \\ v_\mathrm{x} \\ v_\mathrm{Y1} \\ v_\mathrm{Y2} \end{bmatrix}. \tag{4.11}$$

Possible behavioral emulator of MCDU element can be found in [82]† and [83]†, [70]†, respectively. After successful verification of the concept in stage of behavioral model, CMOS transistor-level model can be designed and verified [72]†.

4.8 Concluding Remarks of This Chapter

Commercially available active device able to be used as representation of simple active element is sometimes available, e.g. EL2082 [14] as ECCII–, OPA860 [38] as OTA or also CCII+ element and VCA810 [56] as VB or VGA. However, number of inputs or outputs is usually limited in case of these elements and obviously cannot be changed. Nevertheless, it is useful to consider these devices also because of their general availability.

There are also some prototypes developed by our research group in the past. These devices are usually more universal in order to allow experiments in wider application field. This applies for Universal Current Conveyor (UCC) [64]† and Universal Voltage Conveyor (UVC) [51] and also for DACA [66]†, for instance.

Advanced active elements can be formed of devices from each of these two groups. Advantages in particular application or solution are taken into account when deciding what particular device is going to be used.

Preparation of transistor-level implementation is reasonable when it is proven that particular advanced active element is usable in several applications. Manufacturing of every newly designed device is not economical during research. When developing new device, it is usually useful to design it with help of simpler blocks that are interconnected externally out of a chip. This concept was used also in case of our recent Czech Science Foundation research project No. GP-14-24186P.

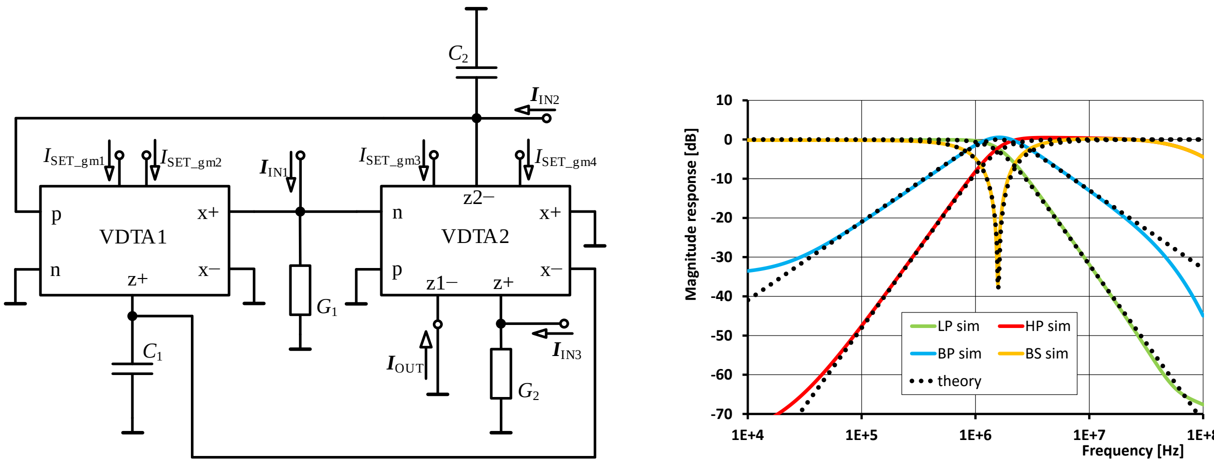
5 DESIGNED CONTROLLABLE FREQUENCY FILTERS

This chapter includes an overview of several designed second-order frequency filters, which are capable of electronic control of at least one of its key parameters (angular pole frequency, quality factor, bandwidth, pass-band gain). As an example, each of the following sections includes filter with various active elements presented in chapter 4, while most of the structures are universal. Most of these own results have been already published in international journals with citation index or presented during recognized international conferences, or were submitted lately and are expected to be published soon. Note that many details were omitted because of limited space, however, the key idea should be always visible and some details should be easily obtainable from preserved text, equations or graphs.

Note that in case of all presented graphs there are several rules that apply through the rest of this thesis. Colored curves stand for the results of bigger significance and credibility, while dotted lines stand for less significant results. The simulation results (colored) are usually compared with theoretical expectations (dotted) or sometimes the measured results (colored) are compared with the simulation results (dotted).

5.1 Filtering Solution Based on Two VDTAs

This section presents electronically adjustable TISO filter with two VDTA active elements presented in [67][†] whose structure is shown in Fig. 5.1(a).



(a) (b)

Fig. 5.1: (a) Electronically adjustable TISO filter with two VDTA active elements having 4 electronically adjustable g_m parameters. (b) Comparison of theory and simulation results in case of magnitude (and phase) responses for starting parameters of the filter, i.e. $f_\mathrm{p} = 1.6$ MHz and $Q = 0.707$. LP, iBP, HP and BS functions.

The ideal transfer functions (LP, HP, inverting band pass (iBP), BS, AP) of the filter can be easily derived from AP response transfer function (that is obtained by combination of LP, HP and iBP responses):

$$K_\mathrm{AP}(\boldsymbol{s}) = \frac{\boldsymbol{I}_\mathrm{OUT}}{\boldsymbol{I}_\mathrm{IN1} + \boldsymbol{I}_\mathrm{IN3} + \boldsymbol{I}_\mathrm{IN2}} = \frac{\boldsymbol{s}^2 C_1 C_2 G_2 g_\mathrm{m3} - \boldsymbol{s} C_2 g_\mathrm{m2} g_\mathrm{m3} g_\mathrm{m4} + G_2 g_\mathrm{m1} g_\mathrm{m2} g_\mathrm{m3}}{\boldsymbol{s}^2 C_1 C_2 G_2 G_1 + \boldsymbol{s} C_2 g_\mathrm{m2} g_\mathrm{m3} g_\mathrm{m4} + G_2 g_\mathrm{m1} g_\mathrm{m2} g_\mathrm{m3}}\,, \tag{5.1}$$

where terms in nominator are given by input currents in order exactly shown in denominator of the first fraction when taken from the left. Transconductances g_m1 and g_m2 are parameters of VDTA1, and transconductances g_m3, g_m4 are parts of the VDTA2 as designated in Fig. 5.1(a) by control currents. If simple matching condition is fulfilled ($G_1 = g_\mathrm{m3}$), unity gain (i.e. 0 dB) is obtained in the pass band of each of the transfer functions. If this is ensured, the angular pole frequency ω_p and quality factor Q are

$$\omega_\mathrm{p} = \sqrt{\frac{g_\mathrm{m1} g_\mathrm{m2}}{C_1 C_2}}\,, \quad Q = \frac{G_2}{g_\mathrm{m4}} \sqrt{\frac{g_\mathrm{m1} C_1}{g_\mathrm{m2} C_2}}\,. \tag{5.2}$$

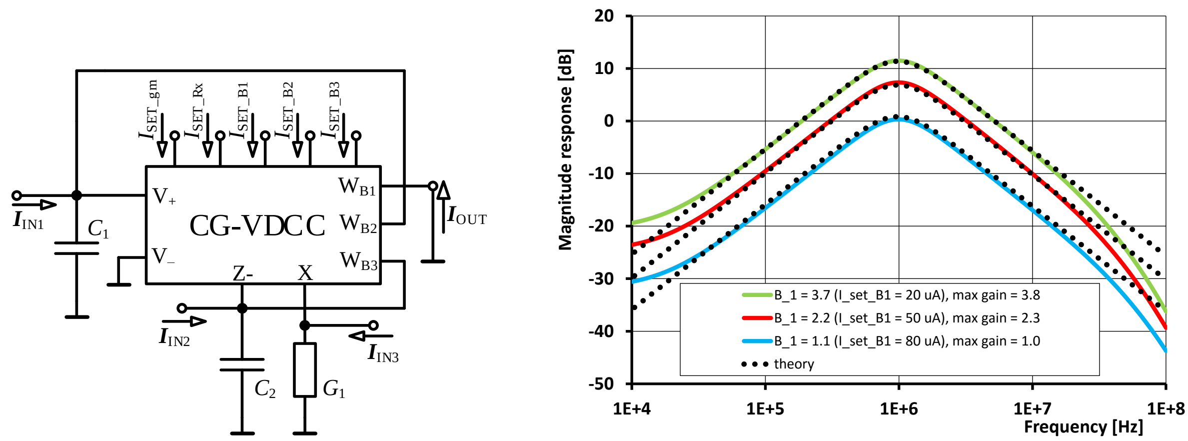
The angular pole frequency could be tuned independently of the quality factor by simultaneously changing the g_m1 and g_m2, while keeping $g_\mathrm{m1} = g_\mathrm{m2}$, i.e. both transconductances are controlled simultaneously. The ratio of G_2 and g_m4 could be used for independent adjustment of the quality factor. If only electronic control is required, only g_{m4} parameter should be used for Q control, nevertheless, particular value of G_2 selected during the design calculations moves the tuning range up or down significantly as presented in [67]†.

The simulation results were prepared with VDTA modeled by CMOS structure [67]† and only several examplary characteristics are shown in Fig. 5.1(b). The parameters of the filter and passive components are provided in [67]† and are omitted in this thesis.

The disadvantage of the presented version of filter from the complexity point of view is the presence of two resistors, nevertheless all passive components including these resistors are grounded. The filter consists of two active elements and four passive elements, but some of them could be omitted if the control of pole frequency independently of the quality factor and/or some of the filtering functions are not required in a particular application. The control possibilities were satisfactory demonstrated in [67]†.

5.2 Filter with One Modified CG-VDCC

This section presents an example of current-mode TISO universal controllable filter (Fig 5.2(a)) with only one Modified CG-VDCC element that was published in [68]†. Note that this active element was described in more detail in Section 4.5.



(a) (b)

Fig. 5.2: (a) Adjustable TISO filter with one Modified CG-VDCC having 3 electronically adjustable parameters of the filter. (b) Comparison of theoretical and simulation results in case of magnitude responses of iBP (f_p = 1.0 MHz, $B_3 = -1$) when controlled pass-band gain by B_1 with constant $B_2 = -1$.

The ideal transfer functions (LP, iBP, HP, BS, AP) can be again easily derived from the following equation valid for AP response (obtained when $\boldsymbol{I}_{IN1} = -\boldsymbol{I}_{IN}$, $\boldsymbol{I}_{IN3} = \boldsymbol{I}_{IN}$ and $\boldsymbol{I}_{IN2} = \boldsymbol{I}_{IN}$):

$$K_{AP}(\boldsymbol{s}) = \frac{\boldsymbol{I}_{OUT}}{\boldsymbol{I}_{IN3} + \boldsymbol{I}_{IN2} + \boldsymbol{I}_{IN1}} = \frac{\boldsymbol{s}^2 C_1 C_2 B_1 - B_1 \boldsymbol{s} C_1 G_1 + B_1 G_1 g_m}{\boldsymbol{s}^2 C_1 C_2 - B_3 \boldsymbol{s} C_1 G_1 - B_2 G_1 g_m}\,, \tag{5.3}$$

where individual terms in nominator are again obtained by input currents in order exactly shown in denominator of the first fraction when taken from the left. It is obvious that filter is stable only for $B_2 < 0$ and $B_3 < 0$, therefore outputs W_{B2} and W_{B3} have to be of negative type. The angular pole frequency and quality factor are equal to

$$\omega_p = \sqrt{\frac{-B_2 G_1 g_m}{C_1 C_2}}\,, \quad Q = \frac{1}{B_3}\sqrt{\frac{-B_2 g_m C_2}{G_1 C_1}}\,. \tag{5.4}$$

From the equations (5.3) to (5.4) it is obvious that:

- B_1 can be used for electronic control of gain (and also phase) of each of filtering functions,

- negative B_2 can be used for electronic control of pole frequency of iBP function with constant bandwidth (BW),
- negative B_3 can be used for electronic control of quality factor in case of all filtering functions,
- controlling simultaneously $g_\mathrm{m} = G_1$ can be used for control of pole frequency while keeping Q value constant.

Some of the most important simulation results are summarized in [68]† together with particular numerical design of all coefficients of the filter. The simulation results were prepared with CG-VDCC modeled by CMOS structure shown in [68]†. Just one example of the simulation results is shown in Fig. 5.2(b).

The issue of this particular solution of the filter is the presence of current divider in case of HP response causing pass-band attenuation as discussed in [68]†. However, the advantage is a possibility to control three parameters of the filter. Filter consists of one active element and three passive elements, but G_1 can be omitted if control of pole frequency independently on quality factor and some of filtering functions are not required.

5.3 ZC-VCCFDITA Used in Filtering Solution

Designed resistor-less filtering structure with controllable features with just one ZC-VCCFDITA active element, presented in Section 4.6, two grounded capacitors and one additional voltage buffer that is not necessary if the following section has high input impedance, is going to be published in [86]† and is shown in Fig. 5.3(a).

This structure provides only voltage-mode BP response in the following form:

$$K_\mathrm{BP}(\boldsymbol{s}) = \frac{\boldsymbol{V}_\mathrm{OUT}}{\boldsymbol{V}_\mathrm{IN}} = \frac{\boldsymbol{s}C_2 g_\mathrm{m} A}{\boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s}C_2 g_\mathrm{m} A + \frac{g_\mathrm{m}}{R_\mathrm{F}}}\,. \tag{5.5}$$

Angular pole frequency ω_p and Q of the filter are expressed as follows:

$$\omega_\mathrm{p} = \sqrt{\frac{g_\mathrm{m}}{R_\mathrm{F} C_1 C_2}}\,, \quad Q = \frac{1}{A}\sqrt{\frac{C_1}{R_\mathrm{F} g_\mathrm{m} C_2}}\,. \tag{5.6}$$

As obvious from the equations above, Q can be controlled independently on ω_p electronically by A and angular pole frequency ω_p can be controlled independently on Q of the filter by voltage-controlled $g_\mathrm{m} = 1/R_\mathrm{F} = 1/R$. Moreover, it can be easily calculated that ω_p of the filter can be controlled without disturbing the BW of the BP filter by R_F.

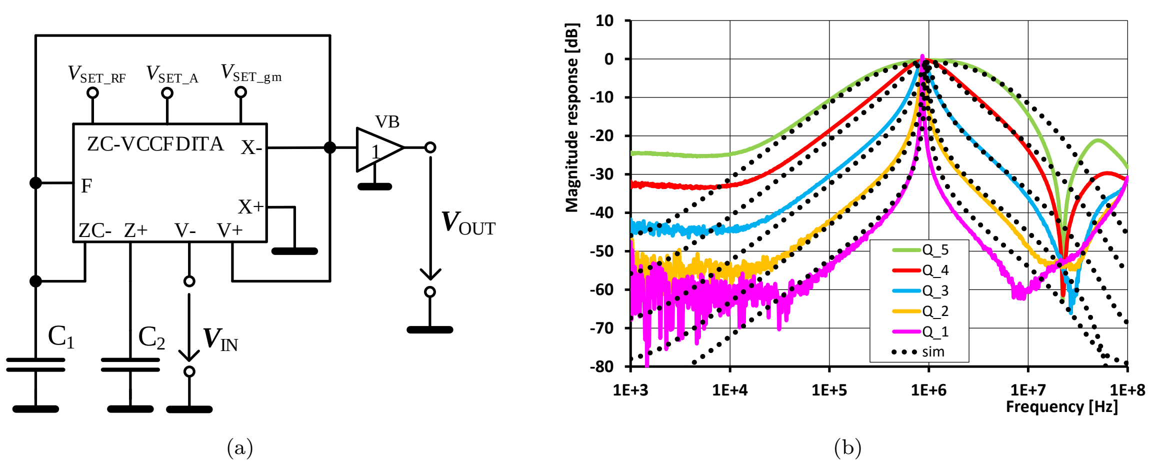


Fig. 5.3: (a) Controllable resistor-less band-pass filter operating in the voltage mode with one ZC-VCCFDITA element and two grounded capacitors. (b) Example results of electronically tuned parameter of the BP filter (measured vs. simulated magnitude responses): Q control while keeping f_p constant.

Paper [86]† includes detailed non-idealities analyses together with measured results compared with simulation results, both obtained with behavioral model presented also in [86]† including all numerical parameters. In this thesis, one example of obtained results is shown, see Fig. 5.3(b).

Based on all the results presented in [86]†, in final application it is required to tune not only one parameter as expected from the theoretical calculations, but also fine-tune other tunable parameter(s) in order to reach the expected results more precisely. Disadvantage of this concept of control is that control DC voltages adjusting R_F and g_m are in different ranges as discussed in detail in [86]† and that it provides only BP response, but the tunability possibilities are enormous.

5.4 Filter with two BOTAs and two DACAs

The structure of the filter with dual-parameter (extended) type of control is shown in Fig. 5.4(a). It is dual-parameter controlled because pole frequency is dependent on two electronic parameters. In this particular case it is dependent on the current gain and also transconductance parameters. Filter consists of two Balanced Output Operational Transconductance Amplifiers (BOTAs) – described in Section 4.2, and two Digitally Adjustable Current Amplifiers (DACAs) – described in Section 4.3, serving as the active elements of the filter, and the filter is of the SITO type. Note that this filtering solution was submitted to be published in [77]†.

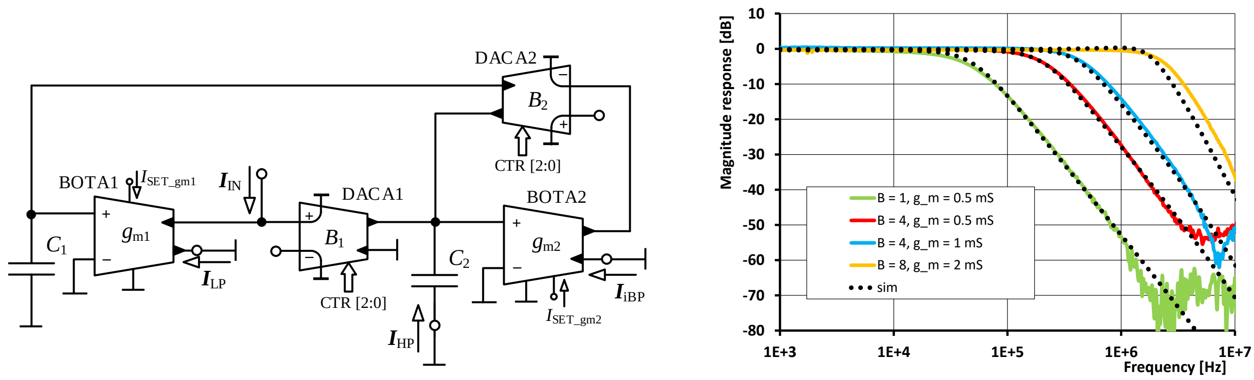
The ideal transfer functions of the filter (LP, iBP, HP, BS and AP) can be derived from the transfer function of AP response:

$$K_\mathrm{AP}(\boldsymbol{s}) = \frac{\boldsymbol{I}_\mathrm{HP} + \boldsymbol{I}_\mathrm{iBP} + \boldsymbol{I}_\mathrm{LP}}{\boldsymbol{I}_\mathrm{IN}} = \frac{B_1 \boldsymbol{s}^2 C_1 C_2 - \boldsymbol{s} C_1 g_\mathrm{m2} B_1 + g_\mathrm{m1} g_\mathrm{m2} B_1 B_2}{\boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s} C_1 g_\mathrm{m2} B_2 + g_\mathrm{m1} g_\mathrm{m2} B_1 B_2}\,. \tag{5.7}$$

As can be seen from the following equations, its f_p can be controlled independently on Q if $g_\mathrm{m1} = g_\mathrm{m2}$ and also by $B_1 = B_2$:

$$\omega_\mathrm{p} = \sqrt{\frac{g_\mathrm{m1} g_\mathrm{m2} B_1 B_2}{C_1 C_2}}\,, \quad Q = \sqrt{\frac{C_2 g_\mathrm{m1} B_1}{C_1 g_\mathrm{m2} B_2}}\,. \tag{5.8}$$

Detailed non-ideal analysis focused on impact of parasitic resistances, capacitances, evaluation of one-pole models of active transfers and examination of stability criteria can be found in [77]†. Moreover, simulation results are supported also by measurement results. In this thesis, example of dual-parameter tuning in case of LP response is shown in Fig. 5.4(b).



(a) (b)

Fig. 5.4: (a) Universal filter with the dual-parameter control of the pole frequency with two BOTAs, two DACAs and two grounded capacitors. (b) Example of measurement results (solid lines) of the filter in comparison with simulation results (dotted lines) - magnitude responses: control of the f_p in the case of LP response by tuning of both g_m and B values (dual-parameter control).

Dual-parameter type of control of some of the filter's parameters can be very useful and can help to extend the useful range of tuning of one of the filter's parameters (f_{p} in this particular case). The disadvantages of this solution are that one of outputs is not taken from high-impedance output of active element and that dual-parameter type of control is not available in case of all the transfer functions. However the measurement results match the simulation results very much and prove the workability of the concept as obvious from [77]†.

5.5 Filter Based on Two MCDUs and One MOCF

The target feature of the filter [70]†, [72]† presented in this section is the widest pole-frequency tunability range. Similarly to the previous section, it is obtained by possibility to dual-control the pole frequency. Topology of the filter is shown in Fig. 5.5(a) and it contains two MCDU active elements introduced in Section 4.7 and one auxiliary MO-CF element [64]†.

The ideal transfer functions of this MISO type filter (LP, iBP, HP, BS and AP) can be similarly to previous cases derived from the following formula for AP response (when $\boldsymbol{I}_{\mathrm{IN1}} = -\boldsymbol{I}_{\mathrm{IN}}$, $\boldsymbol{I}_{\mathrm{IN4}} = \boldsymbol{I}_{\mathrm{IN}}$, $\boldsymbol{I}_{\mathrm{IN2}} = -\boldsymbol{I}_{\mathrm{IN}}$ and $\boldsymbol{I}_{\mathrm{IN3}} = 0$):

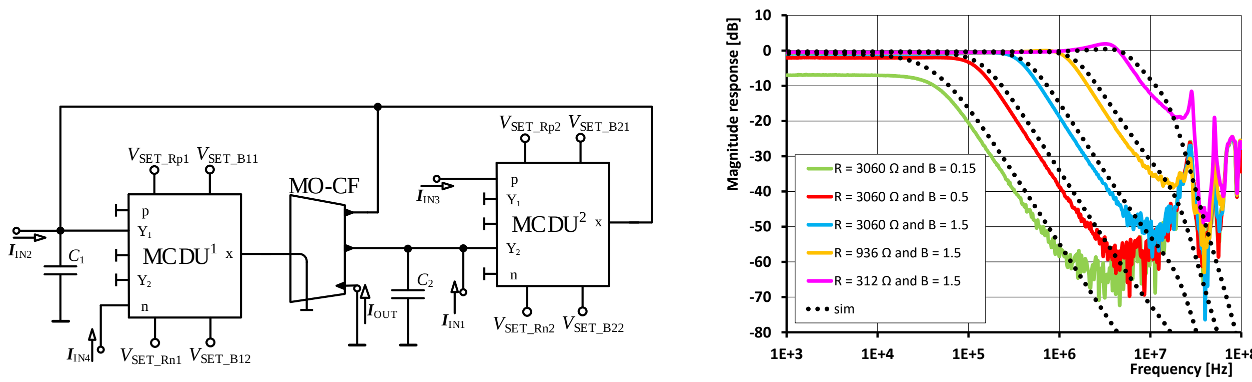
$$K_{\mathrm{AP}}(\boldsymbol{s}) = \frac{\boldsymbol{I}_{\mathrm{OUT}}}{\boldsymbol{I}_{\mathrm{IN4}} + \boldsymbol{I}_{\mathrm{IN2}} + \boldsymbol{I}_{\mathrm{IN1}}} = \frac{\boldsymbol{s}^2 C_1 C_2 B_{12} - \boldsymbol{s} C_2 G_{\mathrm{p1}} B_{11} + G_{\mathrm{p1}} G_{\mathrm{n2}} B_{11} B_{22}}{\boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s} C_2 G_{\mathrm{p1}} B_{11} + G_{\mathrm{p1}} G_{\mathrm{n2}} B_{11} B_{22}} \,. \tag{5.9}$$

The meaning of symbols is obvious from Fig. 5.5(a) and $G_{\mathrm{p1}} = 1/R_{\mathrm{p1}}$, $G_{\mathrm{n2}} = 1/R_{\mathrm{n2}}$. Angular pole frequency ω_{p} and quality factor Q of the filter are expressed as follows:

$$\omega_{\mathrm{p}} = \sqrt{\frac{G_{\mathrm{p1}} G_{\mathrm{n2}} B_{11} B_{22}}{C_1 C_2}}\,, \quad Q = \sqrt{\frac{C_1 G_{\mathrm{n2}} B_{22}}{C_2 G_{\mathrm{p1}} B_{11}}}\,. \tag{5.10}$$

From (5.10) it can be derived that if $G_{\mathrm{p1}} = G_{\mathrm{n2}} = G = 1/R$ is controlled electronically, ω_{p} is adjusted without disturbing Q. Let us assume that this is the first tuning parameter. Same is valid for $B_{11} = B_{22} = B$ (the second tuning parameter). Therefore, there are two ways how to control ω_{p} that can be used mutually independently or can be combined in order to obtain extended control range. Note that B_{12} can be used to adjust independently the pass-band gain of HP response and B_{21} can independently control the pass-band gain in case of iBP2 transfer function (but it is not presented within this short version of thesis).

Presentation of this filter in [72]† is accompanied by detailed tuning suitability analyses providing answers in what kind of range should individual parameters be tuned in order to obtain the widest possible tuning range. Performance of the filter was analysed with help of behavioral model (simulation and also measurement) and moreover, transistor-level model was designed and used in order to verify workability of possible on-chip



(a) (b)

Fig. 5.5: (a) Universal filter with the dual-parameter control of the pole frequency with two MCDUs, one MOCF and two grounded capacitors. (b) Examples of measurement vs. simulation results of the filter when behavioral models were used (magnitude responses): dual-parameter control of f_{p} in case of LP response.

implementation. More details can be found in [72]†. In this thesis, only one example of results is shown in Fig. 5.5(b).

Dual-parameter type of control of f_p was proven to be very useful as obvious from [72]†. The disadvantage of this solution is the complexity of the structure. In case of CMOS implementation, each MCDU consists of 124 transistors as shown in [72]†, when considering only main circuitry. However, this complexity brings wide tuning possibilities as an advantage. Moreover, because of continuous character of control quantities, required value (e.g. f_p) can be fine-tuned in particular application. Transistor-level simulations provided significantly better results than simulations or measurements with behavioral models consisting of many commercially available ICs and several passive elements [72]†.

5.6 Concluding Remarks of This Chapter

Chapter 5 presented briefly five designed solutions of controllable filters with different tuning possibilities. Achieved results proved that advanced and controllable active elements are useful for the design of the controllable filters. All filtering solutions were presented in their variants as published in literature. In each case there is an essential trade-off between the complexity and usability and all conceivable parameters of the filters are not always controlled electronically. However, if it is useful for particular application, it is usually an easy task to enhance presented structure in order to provide additional freedom of tunability.

As obvious from Table 3.1 presented in Chapter 3, designed filtering solutions are based on various active elements and various types of topology. Not all published filters are universal, however almost all own solutions of controllable filters presented in this chapter have this capability.

6 DESIGNED RECONFIGURABLE FREQUENCY FILTERS

This chapter presents several frequency filters, which are capable of electronic switching of type of transfer function without any reconnection of input or output in the structure. Sometimes also one or more of its key parameters (pass-band or stop-band gain, for instance) can be controlled. Structures take advantage of the same type of active elements as controllable filters presented in the Chapter 5. Note that in case of all the presented graphs the same rules are applied regarding colors and type of curves as used in the previous chapter.

6.1 Reconfigurable filter with DO-CA and VDCC

Reconfigurable filter presented in this section is of the SISO type and consists of one DO-CA together with VDCC active element. In this particular case, basic VDCC, introduced in Section 4.5, in variant with three outputs is the core of the filter. Filtering structure is shown in Fig. 6.1(a), and is going to be published in [75]†.

Ideal current-mode transfer function of the filter is:

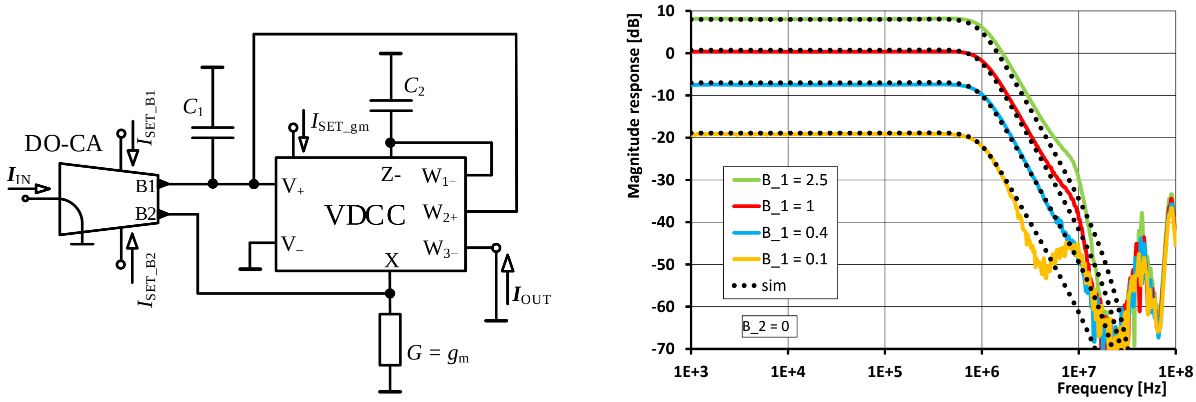
$$K(\boldsymbol{s}) = \frac{\boldsymbol{I}_\mathrm{OUT}}{\boldsymbol{I}_\mathrm{IN}} = \frac{\boldsymbol{s}^2 C_1 C_2 B_2 + G g_\mathrm{m} B_1}{\boldsymbol{s}^2 C_1 C_2 + \boldsymbol{s} C_1 G + g_\mathrm{m} G}\,. \tag{6.1}$$

Routine calculations yield formulas for angular pole frequency and quality factor:

$$\omega_\mathrm{p} = \sqrt{\frac{g_\mathrm{m} G}{C_1 C_2}}\,, \quad Q = \sqrt{\frac{C_2 g_\mathrm{m}}{C_1 G}}\,. \tag{6.2}$$

The following transfer functions are obtained by electronic control of adjustable current gains (B_1, B_2):

1. LP response, if $B_2 = 0$ and $B_1 \neq 0$. Pass-band gain of the filter is adjusted by B_1. If $B_1 > 0$, non-inverting LP response is obtained and pass-band gain is directly proportional to B_1. If $B_1 = 1$, unity gain in low-frequency pass band is achieved. Obviously, if $B_1 < 0$, inverting LP response is obtained.



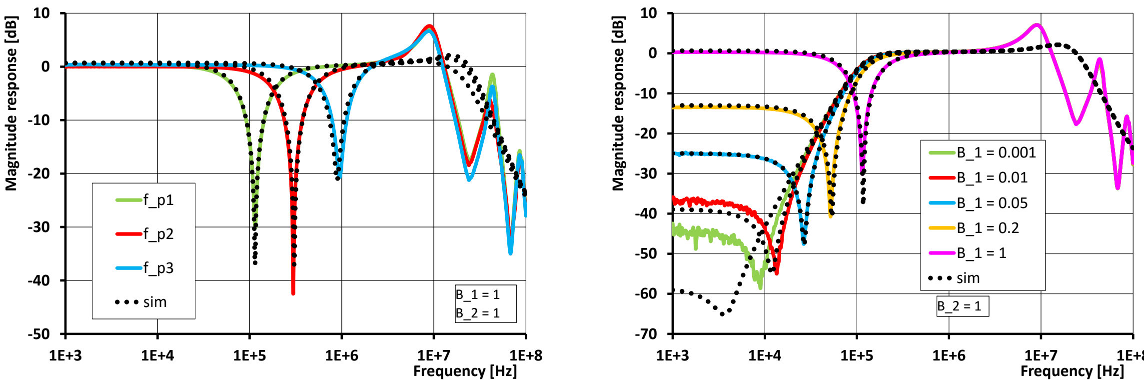
(a) (b)

Fig. 6.1: (a) Reconnection-less reconfigurable filter of the SISO type with one DO-CA and one VDCC element. (b) Examples of measurement vs. simulation results: pass-band gain adjusting of LP response by B_1.

2. HP response, if $B_1 = 0$ and $B_2 \neq 0$. Pass-band gain of the filter is adjusted by B_2. If $B_2 > 0$, non-inverting HP response is obtained and pass-band gain is directly proportional to B_2. If $B_2 = 1$, unity gain in high-frequency pass band is achieved. Obviously, if $B_2 < 0$, inverting HP response is obtained.
3. BS response, if $(B_1 = B_2) \neq 0$. Pass-band gain of the filter is adjusted by $B_1 = B_2$. Other features can be easily derived.
4. LPZ response, if $B_1 > B_2 > 0$ or $B_1 < B_2 < 0$. Also in this case, pass-band gain of the filter is adjusted by B_1, stop-band gain is dependent on B_2. Other features are similar to basic LP response.
5. HPZ response, if $B_2 > B_1 > 0$ or $B_2 < B_1 < 0$. Also in this case, pass-band gain of the filter is adjusted by B_2, stop-band gain is dependent on B_1. Other features are similar to basic HP response.

Note that $B_2 < 0$ and/or $B_1 < 0$ can be obtained with help of current mode multiplier or by using different type of DO-CA element, but this concept was not used in this case. Moreover, if $g_m = G$ is varied, this parameter controls the pole frequency without disturbing Q value.

Details about simulation and also measurement results can be found in [75]† together with details about numerical design. In order to provide basic preview of real behavior of the filter, Fig. 6.1(b) and Fig. 6.2 are included in this thesis.



(a) (b)

Fig. 6.2: (a) Overall measurement results of BS transfer function with unity pass-band gain while tuning pole frequency in comparison with simulation results. (b) Examples of measurement vs. simulation results in case of stop-band gain adjusting: HP response with zero and reconfiguration between HP and BS.

Disadvantage of this solution is that one R is required and it is necessary to change it simultaneously with g_m when pole frequency is tuned. However, this is not the goal function of the presented solution. The key feature was to provide electronic control of change of the transfer function type and additionally to be able to control gain in both the pass and stop band of particular transfer function.

6.2 ZC-CG-VDCC-based Reconfigurable Filter

This section presents interesting reconfigurable and also controllable filtering solution [76]† that is shown in Fig. 6.3(a). It consists of only one ZC-CG-VDCC active element (presented in Section 4.5) and two capacitors and therefore it is obvious that it is resistor-less filtering solution. All controllable features of the ZC-CG-VDCC are fully utilized in order to control parameters of the filtering structure as will be shown.

Transfer function of this filter has the following form:

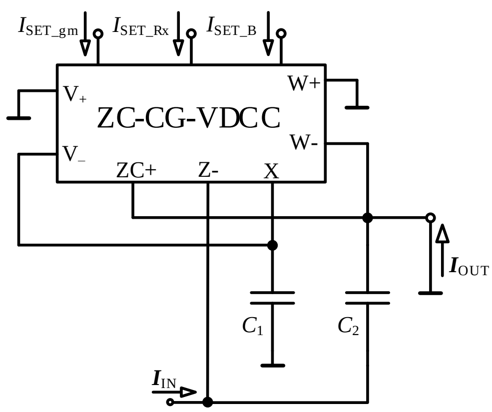
$$K(\boldsymbol{s}) = \frac{\boldsymbol{I}_\mathrm{OUT}}{\boldsymbol{I}_\mathrm{IN}} = -\frac{\boldsymbol{s}^2 C_1 C_2 R_\mathrm{X} + \boldsymbol{s}(C_2 - C_1 B) + g_\mathrm{m}}{\boldsymbol{s}^2 C_1 C_2 R_\mathrm{X} + \boldsymbol{s} C_2 + g_\mathrm{m}} \,. \tag{6.3}$$

Angular pole frequency, quality factor and bandwidth of the filter are expressed by these equations:

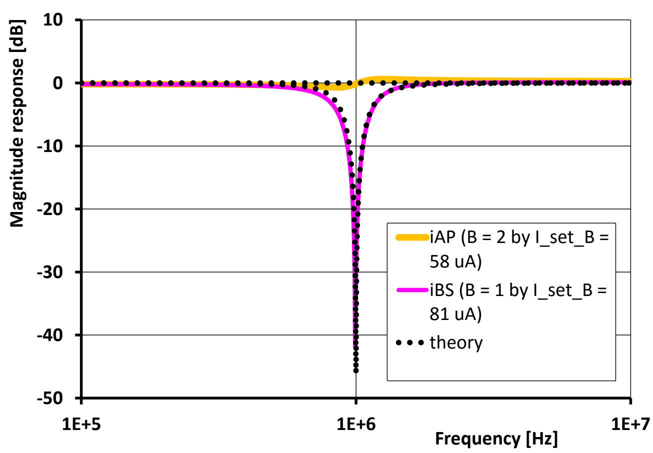
$$\omega_\mathrm{p} = \sqrt{\frac{g_\mathrm{m}}{C_1 C_2 R_\mathrm{X}}}\,, \quad Q = \sqrt{\frac{g_\mathrm{m} R_\mathrm{X} C_1}{C_2}}\,, \quad BW = \frac{1}{C_1 R_\mathrm{X}}\,. \tag{6.4}$$

If we suppose $C_1 = C_2$, inverting AP response is available for $B = 2$ and inverting BS for $B = 1$. As obvious from the text and equations above, B is used for control of transfer function type and it also controls the stop-band gain in case of iBS response. Afterwards, g_m is capable of controlling of the angular pole frequency while keeping constant BW. Moreover, angular pole frequency can be controlled by $1/R_\mathrm{X} = g_\mathrm{m}$ without disturbing the value of quality factor. Of course, it is also possible to change value of Q by control of both R_X and g_m while keeping constant value of angular pole frequency. Example of simulation results obtained with transistor-level models [76]† is shown in Fig. 6.3(b).

The disadvantage of this solution is floating capacitor C_2 that is moreover connected between input and output terminal of the filter. Obviously, in case of current-mode filter, high-impedance outputs should be preferred. This could be easily achieved by adding one auxiliary current follower connected as output buffer. Reconfiguration between just two types of transfer function is possible, however, electronic controllability of their parameters is very good thanks to the features of ZC-CG-VDCC element as is presented in [76]†.



(a)



(b)

Fig. 6.3: (a) Filter with electronic reconfiguration between iAP and iBS response using one ZC-CG-VDCC and two capacitors. (b) Examples of simulation results vs. theory in case of starting values with $f_\mathrm{p} = 1$ MHz and $Q = 3$: magnitude responses.

6.3 Concluding Remarks of This Chapter

The obtained results in both cases confirm workability of the concept of reconfigurable filters with certain controllability also available. Since advanced active elements with controllable parameters are used, it is possible to finetune particular parameter or moreover, to design system that will be able to self-adjust itself dynamically, usually in coordination with digital signal processing component of the more complex system (e.g. analysis of frequency spectrum in order to identify level of undesired spectral components).

7 DESIGNED FRACTIONAL-ORDER FREQUENCY FILTERS

This chapter presents two solutions from the promising area of fractional-order analogue filters. Similar to controllable and reconfigurable filters, controllability of the parameters of the filter is expected. In case of fractional-order, it is usually more difficult to electronically control any of its parameters. In these cases, not only traditional parameters (pole frequency, quality factor, etc.) are evaluated. Very important value is also non-integer order of the filter. Unfortunately, without significant change of topology, only one type of transfer function is usually provided.

7.1 Fractional-order LP Filter with OTAs and ACAs

Design procedure of the fractional-order low-pass filter (FLPF) having order of $(1 + a)$ was precisely described in [55], for instance. Starting transfer function should be in the following form:

$$K_{1+\mathrm{a}}^{LP}(\boldsymbol{s}) = \frac{K_1}{\boldsymbol{s}^{1+a} + \boldsymbol{s}K_3 + K_2}\,. \tag{7.1}$$

Coefficients K_{X}, when $X = 1, 2, 3$ influence the shape of the transfer characteristic similarly to the kind of approximation in integer-order filters. The values of K_{X} can be computed for Butterworth approximation according to the equations shown in [55], depending on the parameter a standing for fraction from 0 to 1. They are expressed as follows:

$$K_1 = 1\,, \quad K_2 = 0.2937a + 0.71216\,, \quad K_3 = 1.068a^2 + 0.161a + 0.3324\,. \tag{7.2}$$

The biggest issue is how to obtain $\boldsymbol{s}^a$ term. In order to overcome this problem, there are some approximating integer-order functions of this term. The second-order approximation of this term is given by [55]:

$$\boldsymbol{s}^a \cong \frac{a_0\boldsymbol{s}^2 + a_1\boldsymbol{s} + a_2}{a_2\boldsymbol{s}^2 + a_1\boldsymbol{s} + a_0}\,, \quad \text{where} \quad a_0 = 2(1+a)\,, \quad a_1 = 5 - a^2\,, \quad a_2 = 2(1-a), \tag{7.3}$$

and they are a part of approximation originally published in [33]. The following 3rd-order transfer function was used to emulate FLPF with Butterworth approximation of order $(1 + a)$:

$$K_{1+\mathrm{a}}^{LP}(\boldsymbol{s}) \cong \frac{K_1}{a_0}\frac{\boldsymbol{s}^2a_2 + \boldsymbol{s}a_1 + a_0}{\boldsymbol{s}^3 + \boldsymbol{s}^2b_2 + \boldsymbol{s}b_1 + b_0}\,, \tag{7.4}$$

where coefficients a_0, a_1 and a_2 are given in this case by (7.3) and

$$b_0 = \frac{a_0K_2 + a_2K_3}{a_0}\,, \quad b_1 = \frac{a_1(K_2 + K_3) + a_2}{a_0}\,, \quad b_2 = \frac{a_1 + a_0K_3 + a_2K_2}{a_0}\,. \tag{7.5}$$

It is worth mentioning that all these coefficients are dependent on initial value of a that determines the fractional-order of the filter.

One of the possible ways how to implement 3rd-order transfer function is to use Follow the Leader Feedback (FLF) filtering topology. Fig. 7.1(a) shows the possible block diagram implementing this type of general topology, consisting of inverting integrators in this case, three feedback loops and three forward signal paths. Its general transfer function is

$$K(\boldsymbol{s}) = \frac{\boldsymbol{I}_{\text{OUT}}}{\boldsymbol{I}_{\text{IN}}} = \frac{\boldsymbol{s}^2 \frac{G_1}{\tau_1} + \boldsymbol{s}\frac{G_2}{\tau_1\tau_2} + \frac{G_3}{\tau_1\tau_2\tau_3}}{\boldsymbol{s}^3 + \boldsymbol{s}^2\frac{1}{\tau_1} + \boldsymbol{s}\frac{1}{\tau_1\tau_2} + \frac{1}{\tau_1\tau_2\tau_3}}\,. \tag{7.6}$$

The particular equations how to obtain τ_{X} and G_{X} (when $X = 1, 2, 3$) from a_{X} and b_{X} coefficients are obvious when comparing (7.4) and (7.6).

Full topology of designed FLPF (3rd-order LP filter actually) respecting block FLF structure shown in Fig. 7.1(a) is depicted in Fig. 7.1(b). Note that this solution together with key steps of its design procedure and the simulation results was published in [73]†.

Particular transfer function of the inverting FLPF from Fig. 7.1(b) is as follows:

$$K(\boldsymbol{s}) = \frac{\boldsymbol{I}_{\text{OUT}}}{\boldsymbol{I}_{\text{IN}}} = -\frac{\boldsymbol{s}^2 \frac{g_{\text{m1}}B_1}{C_1} + \boldsymbol{s}\frac{g_{\text{m1}}g_{\text{m2}}B_1B_2}{C_1C_2} + \frac{g_{\text{m1}}g_{\text{m2}}g_{\text{m3}}B_1B_3}{C_1C_2C_3}}{\boldsymbol{s}^3 + \boldsymbol{s}^2\frac{g_{\text{m1}}}{C_1} + \boldsymbol{s}\frac{g_{\text{m1}}g_{\text{m2}}B_1}{C_1C_2} + \frac{g_{\text{m1}}g_{\text{m2}}g_{\text{m3}}B_1}{C_1C_2C_3}}\,. \tag{7.7}$$

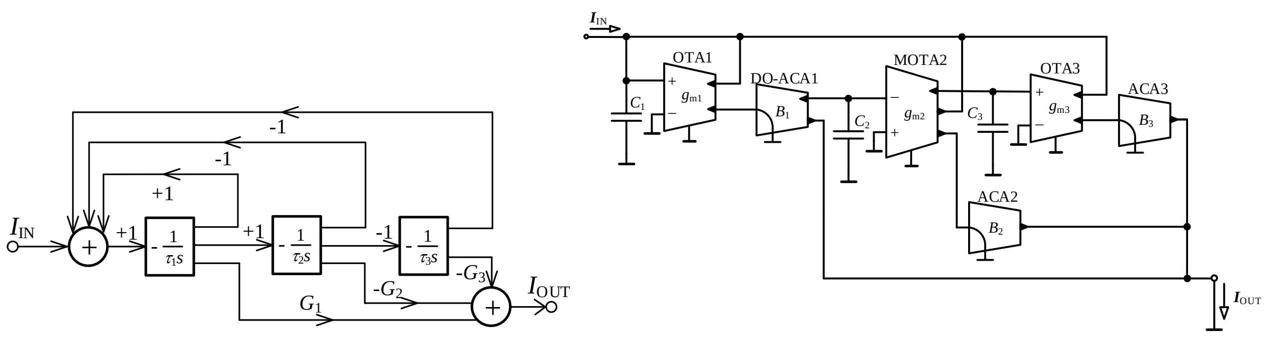
Meaning of quantities and variables used in these equation is obvious from Fig. 7.1(b). Note that many of these parameters can be controlled electronically, which is a beneficial feature for change of the order of the filter and also for control of the pole frequency of the filter in this particular case.

Similarly to previously presented filters, details about particular design and simulations can be found in [73]†. As an example of achieved results, only Fig. 7.2 is included in this thesis.

Disadvantages of not only this particular solution but of most of the fractional-order filters is that particular approximation is usually valid (i.e. corresponds to fractional-order character) only in range shorter than two decades (second-order approximation of $\boldsymbol{s}^a$ as used also in this case). Another disadvantage is that really precise values of resistors and also of current gains are required. Of course, in case of real implementation, the values of resistors have to be selected from at least E48 row, while precise values of current gains are usually not an issue. This step was omitted in [73]† in order to suppress impact of inaccurate resistors and to focus more on features of approximation.

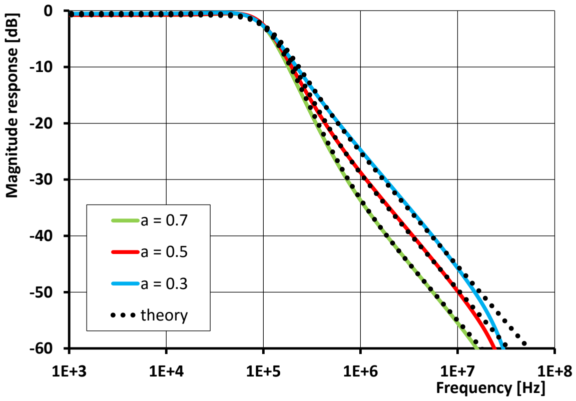
7.2 Solution of Fractional-Order HP Response

LP fractional-order filter, presented in the previous section, can be easily transformed to a solution providing HP response, i.e. Fractional-order High-Pass Filter (FHPF). When standard transformation of LP to HP

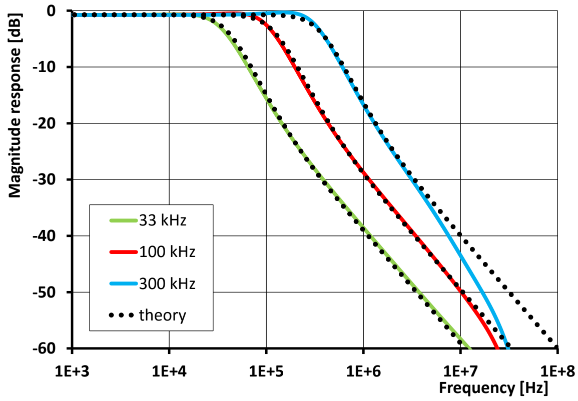


(a) (b)

Fig. 7.1: (a) Block diagram of a FLF (follow-the-leader feedback) topology used for approximation of the FLPF (fractional-order low-pass filter) of (1 + a) order. (b) Designed 3rd-order LP filter approximating (1+a)-order FLPF with OTAs and ACAs.



(a)



(b)

Fig. 7.2: (a) Simulation results vs. theory in case of inverting FLPF response with $f_\mathrm{p} = 100$ kHz and for 3 different values of parameter a: magnitude responses. (b) Simulation results vs. theory in case of tuning f_p of inverting FLPF response while $a = 0.5$: magnitude responses.

transfer function by $\boldsymbol{s} \rightarrow 1/\boldsymbol{s}$ is applied, eq. (7.1) changes to:

$$K^{HP}_{1+\mathrm{a}}(\boldsymbol{s}) = \frac{K_1 \boldsymbol{s}^{1+a}}{K_2 \boldsymbol{s}^{1+a} + K_3 \boldsymbol{s} + 1}\,. \tag{7.8}$$

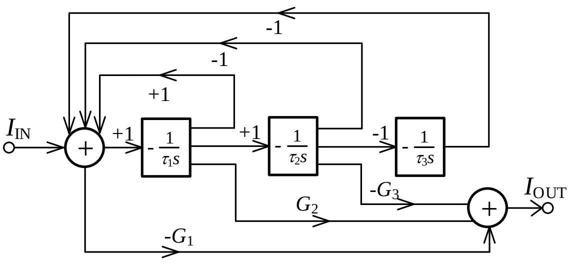
Note that coefficients K_X, where $X = 1, 2, 3$, are exactly the same as already shown in (7.2), so as second-order approximation of $\boldsymbol{s}^a$ as shown by (7.3). In this particular case, two variants of approximation of coefficients a_0, a_1 and a_2 are going to be used and mutually compared. The first set is already defined by (7.3). The second approximation of these coefficients is as follows [16]:

$$a_0 = a^2 + 3a + 2\,, \quad a_1 = 8 - 2a^2\,, \quad a_2 = a^2 - 3a + 2\,. \tag{7.9}$$

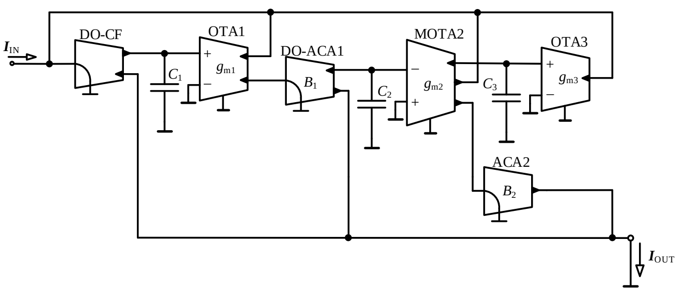
The following 3rd-order transfer function was used to emulate FHPF with Butterworth approximation of order $(1 + a)$, with b_0, b_1 and b_2 already determined by eq. (7.5):

$$K^{HP}_{1+\mathrm{a}}(\boldsymbol{s}) \cong \frac{K_1}{a_0} \frac{\boldsymbol{s}^3 a_0 + \boldsymbol{s}^2 a_1 + \boldsymbol{s} a_2}{\boldsymbol{s}^3 + \boldsymbol{s}^2 \frac{b_1}{b_0} + \boldsymbol{s} \frac{b_2}{b_0} + \frac{1}{b_0}}\,, \tag{7.10}$$

Similarly to FLPF, this particular transfer function of the 3-rd order can be implemented by FLF topology that is slightly modified for the case of FHPF as shown in Fig. 7.3(a) with transfer function:



(a)



(b)

Fig. 7.3: (a) Block diagram of a FLF topology used for approximation of the FHPF (fractional-order high-pass filter) of $(1 + a)$ order. (b) Designed 3rd-order HP filter approximating $(1{+}a)$-order FHPF with OTAs and ACAs.

$$K(s) = \frac{I_{\text{OUT}}}{I_{\text{IN}}} = \frac{s^3 G_1 + s^2 \frac{G_2}{\tau_1} + s \frac{G_3}{\tau_1 \tau_2}}{s^3 + s^2 \frac{1}{\tau_1} + s \frac{1}{\tau_1 \tau_2} + \frac{1}{\tau_1 \tau_2 \tau_3}} . \tag{7.11}$$

Full topology of designed FHPF (3rd-order HP filter actually) respecting block FLF structure shown is depicted in Fig. 7.3(b). Note that this solution is going to be published as [74]†.

Particular inverting transfer function of the filter from Fig. 7.3(b) is in the following form:

$$K(s) = \frac{I_{\text{OUT}}}{I_{\text{IN}}} = -\frac{s^3 + s^2 \frac{g_{\text{m}1} B_1}{C_1} + s \frac{g_{\text{m}1} g_{\text{m}2} B_1 B_2}{C_1 C_2}}{s^3 + s^2 \frac{g_{\text{m}1}}{C_1} + s \frac{g_{\text{m}1} g_{\text{m}2} B_1}{C_1 C_2} + \frac{g_{\text{m}1} g_{\text{m}2} g_{\text{m}3} B_1}{C_1 C_2 C_3}} . \tag{7.12}$$

Fig. 7.3(b) explains particular meaning of quantities and variables used in this equation. Note that except values of C_1, C_2 and C_3, all these parameters can be controlled electronically, which is beneficial feature not only for control of the pole frequency of the filter but also for electronic change of the order of the HP filter.

Similarly to previously presented filters, details about particular design and simulations can be found in [74]†. As an example of achieved results, only Fig. 7.4 is included in this thesis.

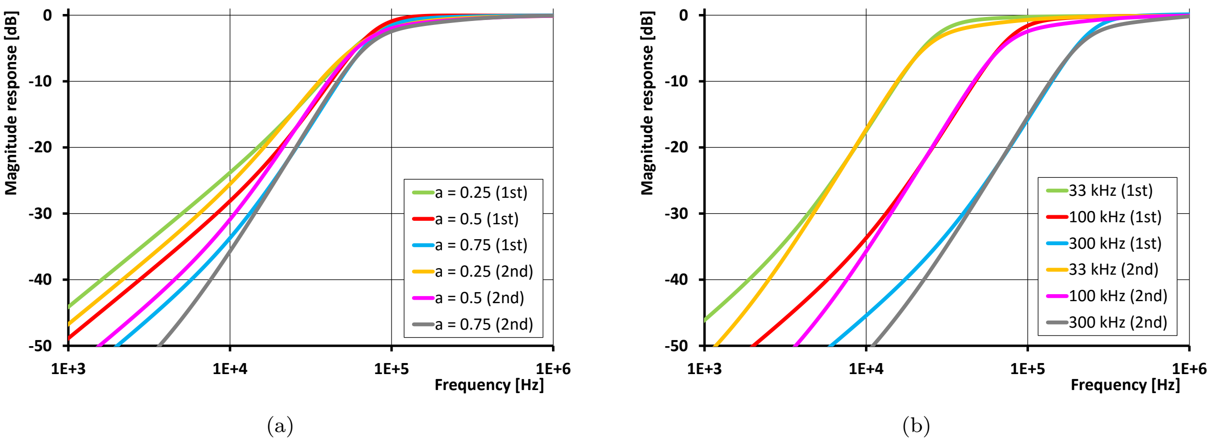


Fig. 7.4: Mutual comparison of both approximations (detail of 1st vs. 2nd version) in case of tuning (a) a parameter: magnitude responses. (b) f_{p}: magnitude responses for fixed $a = 0.75$.

Based on key differences, proper type of approximation should be selected for particular application in order to obtain required character of response. Note that order of the FHPF is controlled electronically in both cases by simultaneous change of five parameters: $g_{\text{m}1}$, $g_{\text{m}2}$, $g_{\text{m}3}$, B_1 and B_2. Moreover, if only $g_{\text{m}1}$, $g_{\text{m}2}$, $g_{\text{m}3}$ parameters are tuned, pole frequency can be controlled without disturbing the order of the filter.

This topology has advantages and disadvantages similar to solution of LP fractional-order filter presented in the previous section.

7.3 Concluding Remarks of This Chapter

The simulation results of both fractional-order filters were obtained with help of behavioral models and the design is prepared in order to allow also future verification by laboratory measurements. Most of the published solutions of fractional-order filters are designed to operate in low-frequency range below 10 kHz. The designed FLPF and FHPF topologies introduced in [73]† and [74]† confirmed possibility to use concept of fractional-order filters also at higher frequencies. Of course, obtained features are always limited by features of active elements that represent the significant components of the whole circuit. The biggest issue is the real-world and really beneficial application of any fractional-order filter that is being constantly searched worldwide.

8 CONCLUSIONS

This thesis had three main topics going through whole text that are controllable filters, reconfigurable filters and fractional-order filters. The thesis had 4 main goals described at the beginning of the text (Section 2.2). Basic overview of the fundamentals and progress in analogue filtering domain was provided in Chapter 3. This chapter included also basic comparison of the chosen solutions from literature with the solutions presented within this thesis and it is divided in order to not mix three main topics of the text. Chapter 4 covers several recently introduced advanced active elements together with basic active elements that usually form a part of these advanced active elements. A description of these elements is essential for the following chapters.

An electronic controllability of parameter(s) of the filter and also reconfiguration of transfer function type is an interesting research topic. As obvious from Table 3.2 and Chapter 6 it is difficult to obtain higher number of possible transfer functions and full controllability of all functions when using reasonable complex circuitry. However, it is usually possible to obtain special types of transfer functions such as LPZ and HPZ response as in case of first solution presented in Chapter 6. Area of the fractional-order filters is not fully researched as discussed in Chapter 3. Breakthrough is expected after discovery of simple or at least favorable implementation of Fractional-Order Element (FOE) on chip. Note that currently, fractional-order elements are also very useful for design and tuning of oscillators. However, this area was not the topic of this thesis. Key advantages and also disadvantages of each of presented solutions are briefly discussed at the end of each of respective sections. Based on text above, it can be concluded that goals of the thesis were fulfilled satisfactory.

It is worth mentioning at this place that this thesis covers only several selected topics of the author's scientific and pedagogical work in years after Ph.D. defence. The same is valid for own references that are included in the Bibliography – only the references that are relevant and cited in this thesis are included. As stated before, Jan Jerabek is author or co-author of 37 journal papers with non-zero impact factor and 47 international conference contributions (numbers were updated in March 2016).

Our research team closely cooperates with several highly-recognized experts in the field. There are several joint works with assoc. prof. Winai Jaikla from King Mongkut's Institute of Technology Ladkrabang, Bangkok, Thailand. Our cooperator is also assoc. prof. Jiun-Wei Horng from Chung Yuan Christian University, Taoyuan, Taiwan. Very appreciated colleague in the field of CMOS integrated circuits design is Dr. Abhirup Lahiri from ST Microelectronics, India. We also cooperate with assoc. prof. Bilgin Metin from Department of Management Information Systems, Bogazici University, Turkey in area of electronically controlled circuits. In the promising area of fractional-order systems we appreciate the cooperation with prof. Costas Psychalinos from University of Patras, Greece.

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Brno, March 24th, 2016

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ABSTRACT

This habilitation thesis focuses on research in area of analogue frequency filters. Particularly and most importantly it focuses on such solutions that are capable of electronic controllability of some of filter's parameter or parameters with help of active elements providing controllable parameter or parameters. The thesis is divided into seven chapters that focus on motivation of origin of this text and goals of the thesis, brief description of current state of the art, basic introduction of important and relevant active elements, their features and possible models of various types and subsequently on controllable, reconfigurable and fractional-order solutions of frequency filters. All presented filtering solutions are accompanied by verification of its features with help of simulations or laboratory measurements and achieved results are evaluated, but only in full version of this thesis. This short version of habilitation thesis includes only examples of achieved results for each of presented filters in order to provide at least basic preview about obtained characteristics because of limited space. This text is written to have not only scientific but also pedagogical contribution. The thesis consists mostly of original research of its author in years after Ph.D. thesis defence. The most of presented solutions was already published in journals with impact factor or presented at recognized international conferences.

ABSTRAKT

Tato habilitační práce je zaměřena na výzkum v oblasti analogových kmitočtových filtrů. Konkrétně pak především na taková řešení, která umožňují elektronickou řiditelnost některého nebo některých parametrů filtru za pomoci aktivních prvků s jedním nebo více řiditelnými parametry. Práce je členěna do sedmi hlavních kapitol, které jsou zaměřeny na motivaci vzniku práce a cíle práce, dále pak na stručný popis současného stavu dané problematiky, na základní představení důležitých a relevantních aktivních prvků, jejich vlastností a možných modelů různého typu a následně pak na řiditelné, rekonfigurovatelné a fraktální řešení kmitočtových filtrů. U všech prezentovaných filtračních řešení je provedeno ověření vlastností za pomocí simulací nebo laboratorního měření a dosažené vlastnosti jsou vyhodnoceny, avšak pouze v plné verzi habilitační práce. Tato zkrácená verze habilitační práce (teze) obsahuje pouze ukázky dosažených výsledků pro jednotlivé prezentované filtry za účelem poskytnutí alespoň základního náhledu na dosažené charakteristiky kvůli omezenému rozsahu. Práce je psána takovým způsobem, aby měla nejen vědecký, ale i pedagogický přínos. Práce je sestavena především z originálního výzkum jejího autora v době po obhajobě doktorské práce. Většina z prezentovaných řešení již byla publikována v impaktovaných časopisech nebo prezentována na uznávaných mezinárodních konferencích.