VYSOKÉ UČENÍ TECHNICKÉ V BRNĚ

FAKULTA ELEKTROTECHNIKY A KOMUNIKAČNÍCH TECHNOLOGIÍ

ÚSTAV TELEKOMUNIKACÍ

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REAL-TIME DIGITAL SIMULATION OF GUITAR AMPLIFIERS AS AUDIO EFFECTS

ČÍSLICOVÁ SIMULACE KYTAROVÝCH ZESILOVAČŮ JAKO ZVUKOVÝCH EFEKTŮ V REÁLNÉM ČASE

ZKRÁCENÁ VERZE PH.D. THESIS

Obor: Teleinformatika

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Datum obhajoby: x. 12. 2012

KLÍČOVÁ SLOVA

Nelineární setrvačné systémy, zpracování číslicových signálů v reálném čase, digitální simulace, hudební efekt, kytarový zesilovač.

KEYWORDS

Nonlinear dynamic system, real-time digital signal processing, digital simulation, audio effect, guitar amplifier.

Disertační práce je k dispozici na Vědeckém oddělení děkanátu FEKT VUT v Brně, Technická 10, Brno, 616 00

ISBN 80-214-
ISSN 1213-4198

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1 Introduction

Digital signal processing influences many fields of human interests these days and together with the progress of computer science it addresses many of new applications. One of these applications is digital audio signal processing in computers. Original analog records on magnetic tapes have been replaced with digital records. Furthermore, an additional processing of audio signals by audio effects is replaced by digital signal processing using real-time digital audio effects. Basically, the digital audio effects work similarly as their analog prototypes, but the output audio signal usually differs. Some musicians claim that the digital audio effects sound too accurate and thin, and therefore they prefer the analog audio effects. Because of this, there is a big effort to simulate these analog audio effects digitally, including their imperfections, such as nonlinear distortion of audio signals, nowadays. A special category of these audio effects is built by simulations of guitar amplifiers and effects where nonlinear distortion is required and the aim is to get the same nonlinear distortion like the analog prototypes have. To analyze and simulate this type of systems, more complex techniques, which often have enormous computational demands, must be used. Therefore, there is effort to find such algorithms that will lead to the same audio perception as the simulated nonlinear system, while the computational demands will stay relatively low to be workable in real-time.

Development of algorithms for simulation of analog electronic circuits has begun with computer programs for analysis of the electronic circuits. Program Simulation Program with Integrated Circuit Emphasis (SPICE) [8] is a typical example of such programs. The time-domain simulation of nonlinear audio effects here is related to the transient analysis of given circuits. These programs nowadays allow simulation of majority of electronic circuits, but this generality of algorithm has enormous computational demands and thus, they are not suitable for real-time simulations.

In the case of audio effects, the generality of the simulation algorithm is not necessary. In spite of generality, attention is rather paid to several typical circuits which can be simplified, approximated and optimized without loosing the same audio perception. Therefore, it is important to understand how the analog audio effects work in order to design proper simplifications (e.g. division into several separate blocks according to their function). The simplified blocks can be then simulated using different algorithms exploiting benefits which different algorithms provide.

2 State of the Art

Generally, the algorithms for simulation of analog audio effects can be classified into several categories – black box approach or white box approach, linear and nonlinear transfer function, etc. Good overview can be found e.g. in literature [9]. Black box approach or white box approach differ in the fact whether the inner structure of simulated systems is known.

2.1 Volterra Series

Volterra Series represent the type of the black box approach for the modeling of nonlinear dynamic systems. The inner structure is unknown and all the necessary information for designing the simulating system is extracted only from relation between the input and output signal [10]. The implementation of the algorithm is very straightforward according to its definition but on the other hand, obtaining of N-dimensional Volterra kernels is a very difficult task. The dimension of kernels can be further reduced using their factorization [10] and the new model can be treated as the Wiener nonlinear model. There are also other simplified models, namely the Wiener-Hammerstein model and the Hammerstein model, which is called the nonlinear convolution and was further investigated in [11]. Results showed that it is possible to obtain the accurate simulation of audio effects. However, all the simulated effects contained one nonlinearity surrounded by linear circuit components, which is in correspondence with the Wiener-Hammerstein model of the nonlinear system. The system containing more nonlinear blocks connected with linear dynamic parts would probably require usage of Volterra series in full form. The simulation of especially highly nonlinear systems, which guitar amplifiers are, requires use of high order Volterra and therefore, it is impractical for the real-time implementation.

2.2 Numerical Integration of Nonlinear Ordinary Differential Equations

These algorithms represent the white box methods. They know the inner structure of the simulated system. Talking about the analog audio effects, the inner structure of the system is directly the circuit schematic of the system. To simulate these circuits, circuit equation have to be obtained firstly using e.g. Modified Nodal Analysis (MNA) method. A set of nonlinear Ordinary

Differential Equations ODE is result of the circuit analysis and they can be solved using the numerical integration. The usage of numerical integration for real-time simulation of audio effects was studied intensively in literature [12] with focus on different numerical integration formulas: implicit integration formulas, which require a numerical algorithm to solve the nonlinear equation, further semi-implicit methods which require one iteration of the numerical algorithm and finally explicit methods which do not require numerical solving. Results showed that the implicit formulas must be generally used in order to ensure the stability of the solution because the explicit formulas are not stiff stable [12].

Computational costs and accuracy of the solution are thus closely related to the stability of the solution. Although explicit formulas offer constant and low computational cost, the requirement for enormous sampling frequencies makes them unpractical for usage in simulations of electronic circuits in real-time. Computational cost of implicit formulas is related to the number of iterations of a numerical nonlinear solver (e.g. Newton-Raphson method). The semi-implicit methods seem to be trade-off providing constant computational cost and stability, however according to [12] they introduce artificial sound artifacts at high frequencies.

Based on the comparison of integration formulas made in [12], it is possible to state that Backward Euler (BE) integration and Trapezoidal Rule (TR) integration are applicable for real-time simulations if sufficient computational power is available but only for simple circuits.

2.3 Simulation by Static Waveshaping and Digital Filter Design

Algorithms based on static waveshaping on the contrary to the numerical integration minimize overall computation costs. This type of algorithms was designed in correspondence with low computational power of computational systems and even today they can find their employment in Digital Signal Processing (DSP) applications.

A nonlinear behavior of algorithms is specified by a simple static mapping between input and output signals given by $y[n] = f(x[n])$ where $f(x)$ is the nonlinear transfer function. The transfer function can be obtained by a numerical solution of the given circuit using MNA and state derivatives set to zeros , from output data of programs for electronic circuits simulation [13] or by a measurement of the real circuit. The transfer function is very often approximated [1, 2] or stored in look-up tables [10, 14, 9, 15, 12].

Nevertheless, simulated circuits often contain frequency dependent circuit components which influence the frequency response of the simulated circuit. Because the static transfer function is not capable to change the frequency response, these algorithms must be supplemented with additional digital filters connected before and behind the nonlinear function.

Although these algorithms are not as accurate as the numerical integration, they are useful for simulation of complex systems where the numerical integration would require extreme computation power to be workable in real-time. The complex circuit of the audio effect is divided into linear and nonlinear parts and the whole system is simulated using a series of digital filters and memoryless nonlinear functions [9, 14, 12]. Results showed validity of this approach although it is not the exact emulation of circuits. Further accuracy can be achieved by using better division into blocks.

2.4 State-Space Based Approach

The main problem of direct numerical integration of MNA ODEs is, except the necessity of use of a nonlinear iterative solver, total number of equations and thus number of unknown variables to be solved numerically. The nonlinear equations are spread over the unknowns forming non-computable loops – delay free loops. However, as it was showed in prior work [16] and then reviewed in [12], the number of unknown variables to be solved numerically can be significantly reduced by using transformation of linear parts of the circuit equations which highlights nonlinear mappings and decompose the system into a nonlinear multiple-input multiple-output mapping part and a linear dynamic part. This method has a few variants, the most suitable is the DK-method with incidence matrices introduced in [17] given by the equations

$$\mathbf{x}[n] = \mathbf{A}\mathbf{x}[n-1] + \mathbf{B}\mathbf{u}[n] + \mathbf{C}\mathbf{i}_{\mathrm{n}}(\mathbf{v}[n]), \tag{1}$$

$$\mathbf{y}[n] = \mathbf{D}\mathbf{x}[n-1] + \mathbf{E}\mathbf{u}[n] + \mathbf{F}\mathbf{i}_{\mathrm{n}}(\mathbf{v}[n]), \tag{2}$$

$$\mathbf{v}[n] = \mathbf{G}\mathbf{x}[n-1] + \mathbf{H}\mathbf{u}[n] + \mathbf{K}\mathbf{i}_{\mathrm{n}}(\mathbf{v}[n]) \tag{3}$$

where matrices $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$, $\mathbf{D}$, $\mathbf{E}$, $\mathbf{F}$, $\mathbf{G}$, $\mathbf{H}$ and $\mathbf{K}$ define the nonlinear state-space description, $\mathbf{x}[n]$ is the state vector, $\mathbf{y}[n]$ is the output vector, $\mathbf{u}[n]$ is the input vector, $\mathbf{i}_{\mathrm{n}}(\mathbf{v}[n])$ is the nonlinear circuit component model and the equation (3) is the implicit nonlinear equation to be solved numerically. Although the number of unknown variables in the nonlinear equation was reduced, direct use of this method for the real-time simulation of more complex is still not possible.

2.5 Nonlinear Wave Digital Filters

Wave digital filters are another approach to elimination or reduction of delay free loops. The basic principle of wave digital filters is substitution of classic port variables – Kirchhoff variables (voltage v and current i) by wave variables – incident and reflected waves A and B and a port resistance R, which is however not related with the physical resistance, all representing one port element (e.g. resistor). As a result, transformed circuit components in term of wave variables are introduced [15, 12]. These component are being connected using series and parallel adapters forming Binary Connection Tree (BCT). A big advantage of using these adapters is the linear computational complexity in contrast to the general scattering matrices providing quadratic computational complexity [12]. The biggest disadvantage is that the nonlinear components have to be connected to the root of BCT and therefore, this method can efficiently handle only one nonlinear circuit component, until the waves of nonlinear component are delayed, which is insufficient.

3 Goals of Thesis

Despite of the recent significant progress in the field of real-time simulation of the analog audio effect, few problems still remain. Foremost, all mentioned algorithms were tested on relatively simple audio circuits. Therefore, the overall goal of this thesis is to investigate simulation of the complex nonlinear audio effect circuits in real-time – guitar amplifier in this case. The block diagram of the guitar amplifier is given in Figure 1 [10, 14, 18]. The preamp consists of several triode amplifier stages or limiter units, built by operational amplifiers and diodes, and the passive gain and the tone stack circuit. The power amplifier consists of the phase splitter and the push-pull amplifier built by two penthodes and the output transformer.

The DK-method with incidence matrices was chosen as the perspective method for the simulation of more complex circuits, because the second candidate, nonlinear wave digital filters, is not very convenient for the simulation of circuits with more nonlinear functions although it is possible. The main objectives are:

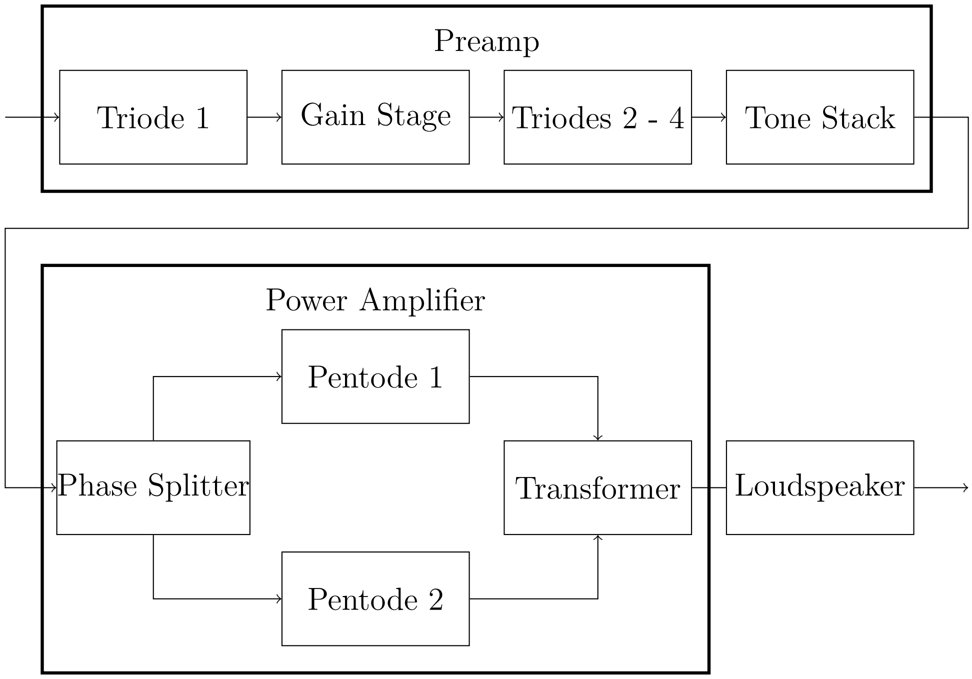

Figure 1: Block diagram of guitar amplifier

- The extension of DK-method with incidence matrices by introducing the nonlinear model of the audio transformer and also the nonlinear model of the operational amplifier. This will serve as a base for further more complex systems discussed in later parts. The extension will be tested on some audio circuits.
- The efficient approximation of system nonlinearities. Firstly, the offline precomputation of the system nonlinearities will be discussed as a way of reduction of the computational complexity. The approximation algorithms will be investigated with regards to computational complexity as well as to the amount of data stored in look-up tables.
- Simulation of complex systems. Because the simulation of the complex system can be computationally demanding, techniques of decomposition into simpler parts preserving mutual interaction between blocks will be designed.
- Simulation quality evaluation. The validity of designed algorithm for simulation of complex systems will be tested using listening tests. Finally, the presence of aliasing distortion in output signals will be investigated. On the basis of these findings, the minimal required sampling frequency will be stated.

4 Circuit Analysis of Audio Effects

The derivation of circuit equations is always the first step in simulation of the circuit. It can be done manually or this process can be automated, which is very convenient especially for complex systems. This work continues in work of the automated derivation of circuit equations proposed by Holters and Zölzer [17] whose method of incidence component matrices for derivation of the state-space model of the simulated system will be used. It deals with circuits with basic circuit components. However, several other components, foremost the model of the audio transformer and the model of Operational Amplifier (OPA), should be considered to be able to simulate audio effects circuits more generally.

5 Simulation of Circuits with Audio Transformer

Transformers can be found in many audio effect or audio amplifier circuits. However, they are often a part of power supply circuits and thus, they can be mostly neglected in simulations of these electronic devices. However, transformers are also often connected directly in the audio path in many types of audio circuits, e.g. input and output stages of mixing consoles, professional studio compressors, etc. Similar types of the transformers can be also found in tube power amplifiers where their main function is to match the high output impedance of used tubes to the low input impedance of the loudspeaker. Although designers of the audio transformers try to minimize the distortion effect of transformers in order to get a transparent signal path, the presence of the transformer can be often heard and sometimes it is even required by musicians and audio engineers. The main reason for this is the characteristic nonlinear distortion caused by the soft saturation of the transformer core when a large magnetization occurs. However, the saturation effect is not the only one which manifests in the output signal, the hysteresis effect and frequency properties also have to be included in the simulation. The model of the transformer, which will be used in later simulation only with three windings instead of two, derived from [19] is shown in Figure 2.

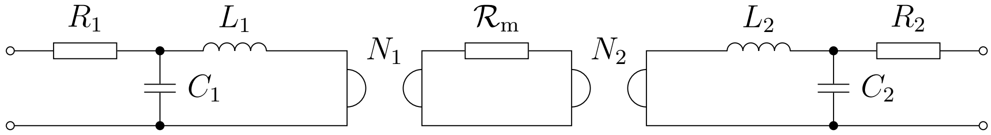

Figure 2: Model of a transformer with two windings.

The transformer model can be split into three parts – the electric part modeling parasitic capacitances C_1, C_2 of windings as well as leakage inductances L_1, L_2 and windings resistances R_1, R_2, the electromagnetic part responsible for interaction of electric and magnetic field, represented by two gyrators with resistances corresponding to numbers of turns of the windings N_1, N_2, and the magnetic part which models the core saturation, the hysteresis and eddy currents. Because the electric part of transformer can be easily incorporated directly into the whole circuit, if necessary, these electric components are no further considered as a part of the transformer model but rather as a part of the whole circuit. The electro-magnetic interaction is described using the Farraday's induction law and Amper's law.

The magnetic part of the model can be either linear or nonlinear. In case of the linear magnetic core, the relation between magneto-motive force $\mathcal{F}$ and flux Φ is given by Hopkinson's law $\mathcal{F} = \mathcal{R}\Phi$ where $\mathcal{R}$ is the magnetic reluctance. In case of the nonlinear core model, the reluctance $\mathcal{R}$ is nonlinear because of nonlinear permeability μ. To simulate the nonlinear core behavior, the following transformer core model were used:

- **The Frohlich equation** – provides a simple approximation of the transformer core permeability but does not support simulation of the hysteresis loop [20].
- **Modified Jiles-Atherton (JA) model** – with improved modeling of minor hysteresis loops [21], which provides the simulation of the hysteresis.
- **Gyrator-Capacitor (GC) model** – proposed in [22] and used in [19]. It also provides the simulation of the hysteresis.

All transformer core models were tested in simulation of the input stage circuit used as asymmetrical-to-symmetrical signal converter. The numerical Newton-Raphson method was used for solving the circuit equations. The comparison of the computational efficiency can be found in Table 1 where the maximum number and the average number of iterations is given for each transformer model.

Table 1: Number of iterations of Newton-Raphson method.

Model	Frohlich	GC-model	JA-model
Max	5	4	53
Average	2.93	2.96	12.58

The conclusion is that the JA-model provides good modeling of the hysteresis loop but on the other hand, the computational cost is very high and problems with the numerical stability can also occur foremost in the area where Φ and $\mathcal{F}$ are close to zero. The Frohlich model and GC model have almost the same computational cost (computational cost of GC-model is higher due to the more complex model equation) but the advantage is modeling of the hysteresis loop.

The next step is incorporation of the transformer model into the DK-method. The DK-method requires prior component-wise discretization of energy storing circuit components. The Farraday's induction law can be discretized by either TR or BE discretization formula. Using TR, it is possible to derive companion circuit consisting of conductance $G = \frac{2N}{T}$ where N is the number of turns and T is the sampling frequency and current source x according to

$$v[n] = G\Phi[n] - x[n-1] \tag{4}$$

where $x[n-1] = G\Phi[n-1] + v[n-1]$. The state variable x can be updated by the equation

$$x[n] = 2G\Phi[n] - x[n-1]. \tag{5}$$

Subsequently the conductance matrix used for the derivation of state-space matrices using connection matrices have to be build. Firstly, the linear model of the transformer will be considered. The conductance matrix can be divided into a regular part without the transformer model and a part with the connected transformer model. This division is illustrated in Figure 3 left. The regular part without the transformer model is denoted by symbol $\mathbf{S}$ and it can be obtained in standard way according to [17]. The rest of the matrix contains parameters of the transformer. Several parts can be there further identified, namely the connection matrix $\mathbf{N}_\mathrm{t}$ which specify which nodes is the transformer connected to, the vector containing numbers of turns of each winding $\mathbf{N}_\mathrm{w}$, the transformer core parameter and several zero matrices. Thus, the whole conductance matrix can be built up from submatrices $\mathbf{S}$, $\mathbf{N}_\mathrm{t}$, $\mathbf{N}_\mathrm{w}$, zero matrices and the magnetic reluctance parameter $\mathcal{R}$ according to Figure 3 right.

The matrix $\mathbf{N}_\mathrm{t}$ is the connection matrix with the number of rows given by the number of windings and the number of columns given by the number of circuit nodes.

Once the whole matrix, further denoted as $\mathbf{S}_\mathrm{t}$, is built up, DK-method matrices $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$, $\mathbf{D}$, $\mathbf{E}$, $\mathbf{F}$, $\mathbf{G}$, $\mathbf{H}$ and $\mathbf{K}$ can be obtained according to [17]. However, because of the changed size of the conductance matrix, equations are not consistent anymore. Therefore, incidence matrices must be extended

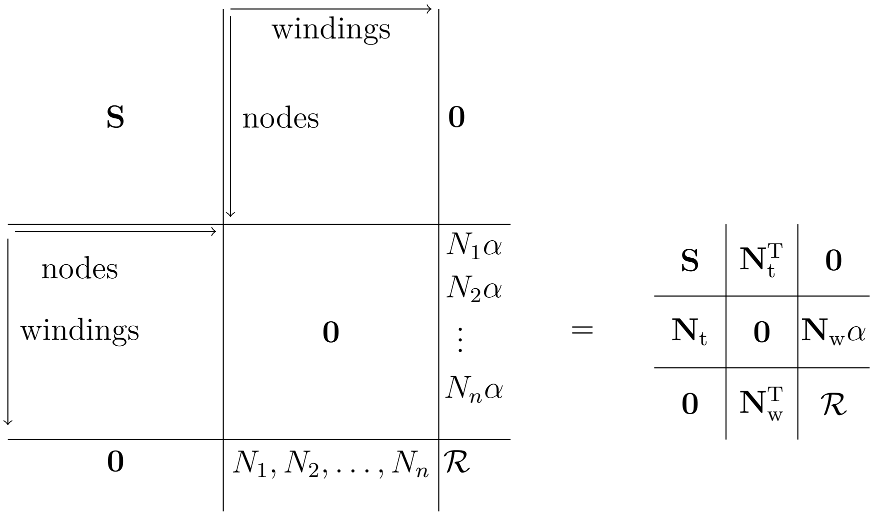

Figure 3: Stamp of linear transformer model for conductance matrix

by zero vectors $\mathbf{0}$ with the number of columns given by number of windings increased by one.

Similarly, the incorporation of the nonlinear transformer model into the automated DK-method can be derived. The magnetic reluctance $\mathcal{R}$ can not be isolated from transformer core models and therefore the transformer model will be included in nonlinear models vector. The conductance matrix $\mathbf{S}_{\mathrm{t}}$ can be again decomposed into submatrices according to Figure 4.

$\mathbf{S}$	$\mathbf{N}_{\mathrm{t}}^{\mathrm{T}}$	$\mathbf{0}$
$\mathbf{N}_{\mathrm{t}}$	$\mathbf{0}$	$\mathbf{N}_{\mathrm{w}}\alpha$
$\mathbf{0}$	$\mathbf{N}_{\mathrm{w}}^{\mathrm{T}}$	0

Figure 4: Stamp of nonlinear transformer model for the conductance matrix

The only change in the conductance matrix is the zero value instead of the magnetic reluctance $\mathcal{R}$. Another change is in the nonlinear model vector where one nonlinear function has been added. The incidence vector $\mathbf{N}_{\mathrm{n}}$ is now

$$\mathbf{N}_{\mathrm{nt}} = \begin{bmatrix} \mathbf{N}_{\mathrm{n}} & \mathbf{0} \\ \mathbf{0} & \begin{bmatrix} 0 & 0 & \dots & 0 & 1 \end{bmatrix} \end{bmatrix}. \tag{6}$$

One new row was added to introduce one more nonlinear function in the DK-method core function. This vector consists of zero values and 1 at the end of the vector because the equation describing the magnetic circuit is the last one in the nonlinear vector.

5.1 Simulation of Circuits with Operational Amplifier

Another commonly used circuit component which has not been discussed well in real-time audio effect simulations is the operational amplifier (OPA). The operational amplifier is usually used as the active component of analog filters which can often be found in audio effects circuits. The operational amplifiers are usually simulated as macro models [23]. For the case of the real-time audio effect modeling, a linear or a behavioral nonlinear model can be used instead of complex OPA macro models. These models can be used because of the limited frequency bandwidth ($f < 20$ kHz) which allows to use frequency independent models of operational amplifiers. The linear model can be described by

$$v_{\mathrm{out}} = A\left(v_{+} - v_{-}\right) \tag{7}$$

The nonlinear function can be implemented according to

$$v_{\mathrm{out}} = v_{\mathrm{EE}} + \left(\frac{1}{2}\tanh\left(a\left(v_{+} - v_{-}\right)\right) + 0.5\right)\left(v_{\mathrm{CC}} - v_{\mathrm{EE}}\right) \tag{8}$$

where v_{EE} is the negative power supply voltage and v_{CC} is the positive power supply voltage and the parameter a can be used for the modeling of the different amplification factor A.

Further, incorporation into the DK-method is designed. Firstly, the whole circuit without operational amplifiers can be analyzed using the standard automated DK-method to get the conductance matrix $\mathbf{S}$ consisting of resistors and discretized models of inductors and capacitors.

Subsequently, the operational amplifier is modeled as the nonlinear voltage controlled voltage source. Therefore, the model of the operational amplifier can be added into the system similarly as any other voltage source resulting in extension of the conductance matrix $\mathbf{S}$ according to

$$\mathbf{S}_{\mathrm{opa}} = \begin{pmatrix} \mathbf{S} & \mathbf{N}_{\mathrm{opa-O}}^{\mathrm{T}} \\ \mathbf{N}_{\mathrm{opa-O}} & \mathbf{0} \end{pmatrix} \tag{9}$$

where the incidence matrix $\mathbf{N}_{\mathrm{opa-O}}$ holds information where the output of the operational amplifier is connected. The matrix $\mathbf{N}_{\mathrm{opa-O}}$ can be for one operational amplifier in form

$$\mathbf{N}_{\mathrm{opa-O}} = \begin{bmatrix} \cdots & 0 & 1 & -1 & 0 & \cdots \end{bmatrix} \tag{10}$$

where values 1 and -1 determine which nodes the output of the operational amplifier is connected to and other values are zeros. The total number of columns of the $\mathbf{N}_{\mathrm{opa-O}}$ matrix is given by the number of columns of the matrix $\mathbf{S}$. One terminal is very often connected to the reference node and therefore the -1 element is often missing.

Subsequently, equations for deriving DK-method matrices must be extended. Firstly, the control voltage must be obtained. The control voltage is the voltage between the terminals of the operational amplifier and these terminals are specified by the matrix

$$\mathbf{N}_{\mathrm{opa-I}} = \begin{bmatrix} \cdots & 0 & -1 & 1 & 0 & \cdots \end{bmatrix} \tag{11}$$

for one operational amplifier.

The construction of the $\mathbf{K}$ matrix differs from original method [17] because the nonlinear function is not connected to the control voltage nodes but is connected as the additional voltage source. The $\mathbf{K}$ matrix can be obtained from

$$\mathbf{K} = \begin{pmatrix} \mathbf{N}_{\mathrm{n}} & \mathbf{0} \\ \mathbf{N}_{\mathrm{opa-I}} & \mathbf{0} \end{pmatrix} \mathbf{S}_{\mathrm{opa}}^{-1} \left(\begin{pmatrix} \mathbf{N}_{\mathrm{n}} \\ \mathbf{0} \end{pmatrix} + \begin{pmatrix} \mathbf{0} \\ \mathbf{I} \end{pmatrix} \right) \tag{12}$$

where the identity matrix is used to express the connection of the nonlinear voltage controlled voltage sources of the operational amplifiers and $\mathbf{N}_{\mathrm{n}}$ is the incidence matrix of nonlinear components without considering the operational amplifiers. The matrix $\mathbf{C}$ is derived using

$$\mathbf{C} = 2\mathbf{G}_{\mathrm{x}}\mathbf{Z} \begin{pmatrix} \mathbf{N}_{\mathrm{x}} & \mathbf{0} \end{pmatrix} \mathbf{S}_{\mathrm{opa}}^{-1} \left(\begin{pmatrix} \mathbf{N}_{\mathrm{n}} \\ \mathbf{0} \end{pmatrix} + \begin{pmatrix} \mathbf{0} \\ \mathbf{I} \end{pmatrix} \right) \tag{13}$$

and the matrix $\mathbf{F}$ from

$$\mathbf{F} = \begin{pmatrix} \mathbf{N}_{\mathrm{o}} & \mathbf{0} \end{pmatrix} \mathbf{S}_{\mathrm{opa}}^{-1} \left(\begin{pmatrix} \mathbf{N}_{\mathrm{n}} \\ \mathbf{0} \end{pmatrix} + \begin{pmatrix} \mathbf{0} \\ \mathbf{I} \end{pmatrix} \right). \tag{14}$$

The other matrices are obtained in standard way.

If the linear OPA model is used, the conductance matrix $\mathbf{S}_{\mathrm{opa}}$ is constructed according to

$$\mathbf{S}_{\mathrm{opa}} = \begin{pmatrix} \mathbf{S} & \mathbf{N}_{\mathrm{opa-O}}^{\mathrm{T}} \\ \mathbf{N}_{\mathrm{opa-O}} + \mathbf{A}\mathbf{N}_{\mathrm{opa-I}} & \mathbf{0} \end{pmatrix} \tag{15}$$

where $\mathbf{A}$ is the diagonal matrix with the amplification factor A for particular operational amplifiers on the diagonal. All DK-method matrices are computed in standard way according to [17]

6 Approximation of Implicit Nonlinear Circuit Equations

Direct usage of numerical algorithms is not very convenient for real-time applications. It is foremost due to two properties – high computational complexity and variable computational complexity, which does not allow efficient utilization of the DSP system computational power. Moreover, the numerical solution can also diverge if suitable initial conditions are not used. Therefore, usage of approximation of nonlinear functions is essential for real-time applications. It ensures constant and low computational complexity and unambiguous solutions.

6.1 Precomputation for Approximation of Nonlinear ODEs

For some circuits, the number of inputs is lower than number of nonlinear equations and therefore the direct approximation of the equations requires lower dimension and also lower computational cost than if the DK-method is used for the simulation. The nonlinear equations to be solved are in form of

$$\mathbf{0} = \mathbf{f}(v_{\mathrm{in}}, v_{\mathrm{x1}}, v_{\mathrm{x2}}, \cdots, v_{\mathrm{xM}}) \tag{16}$$

where v_{in} is the input voltage or the input signal value and v_{x1}, v_{x2},..., v_{xM} are voltages on energy-storing components. The system has $N = M + 1$ inputs and a particular solution for a combination of the inputs. Further, the system thus have N outputs and if the functions to be approximated are denoted as f_{out}, f_{x1}, f_{x2},...,f_{xM}, the all system may be rewritten into

$$\begin{aligned}
&v_{\mathrm{out}}[n] = f_{\mathrm{out}}(v_{\mathrm{in}}[n], v_{\mathrm{x1}}[n], v_{\mathrm{x2}}[n], \cdots, v_{\mathrm{xM}}[n])\\
&v_{\mathrm{x1}}[n+1] = v_{\mathrm{x1}}[n] + T f_{\mathrm{x1}}(v_{\mathrm{in}}[n], v_{\mathrm{x1}}[n], v_{\mathrm{x2}}[n], \cdots, v_{\mathrm{xM}}[n])\\
&v_{\mathrm{x2}}[n+1] = v[n]_{\mathrm{x2}} + T f_{\mathrm{x2}}(v_{\mathrm{in}}[n], v_{\mathrm{x1}}[n], v_{\mathrm{x2}}[n], \cdots, v_{\mathrm{xM}}[n])\\
&\vdots\\
&v_{\mathrm{xM}}^{n+1} = v_{\mathrm{xM}}[n] + T f_{\mathrm{xN}}(v_{\mathrm{in}}[n], v_{\mathrm{x1}}[n], v_{\mathrm{x2}}[n], \cdots, v_{\mathrm{xM}}[n])
\end{aligned} \tag{17}$$

where T is the sample period, and the superscript n denotes time index. Computational complexity depends on the number of circuit inputs N (input signal and states). The computation of one output signal sample requires M add operations, M multiply operations and N computations of approximating functions. Therefore, the total computational complexity markedly depends on the chosen approximation type.

6.2 Approximation Techniques Comparison

The several approximation techniques were compared, all of them were based on interpolations between stored data in look-up tables. The computational cost was investigated foremost. Table 2 shows the maximal number of interpolations which can be computed in real-time. The best performance is provided by cubic spline and linear interpolation.

Table 2: Maximal number of interpolations per sample for $f_s = 48$ kHz.

Dim.	Spline 1	Spline 2	Spline 3	Newton	Hermite	Linear
1-D	5210	3470	2980	833	1740	4170
2-D	801	386	372	167	282	992
3-D	342	107	101	43	77	563
4-D	109	30	27	11	20	311
5-D	26	6	7	2	4	96

The amount of data stored in look-up tables can be reduced by using non-uniformly gridded interpolations.

7 Complex System Simulation

The typical guitar preamp or the power amplifier can consist up to ten nonlinear equations, which is not feasible to compute in real-time numerically and the approximation of the N-dimensional nonlinearity would require huge data sets of coefficients. The reasonable number of dimensions of the approximation function is 4 as was shown in the previous chapter and this number is not sufficient for simulation of more complex systems. Therefore, the main aim is to discuss techniques of decomposition into block which however preserve mutual interactions between separate blocks.

7.1 Modified Block-Wise Method

This method was designed as the efficient and simple method enabling simulation of the mutual interaction between adjoined blocks [3, 4, 5, 6]. Audio effect circuits are often built by a cascade of amplifiers, filters, nonlinear waveshapers etc. There is therefore often coupling only between directly adjacent blocks. The simulation of the first block has to take account of the current, typically nonlinear, flowing into the second block and one has to simulate both the first two blocks together. As a result of this, the output impedance of the first block is matched with the input resistance of the adjoined second block and the output voltage or the output signal of the first block is correct.

Up to this point, only the simulation of the first block was considered although the second block had been used and there is also available the output signal of the second block. This output signal is not, however, matched with the third block and cannot be therefore used. The simulation of the second block requires the simulation of the second block together with the third one. The output signal between the first and second block is then used as the input signal for the connected second and the third block. If there are more blocks connected in series, we can continue with the decomposition in similar way. The whole process of such as decomposition is illustrated in Figure 5.

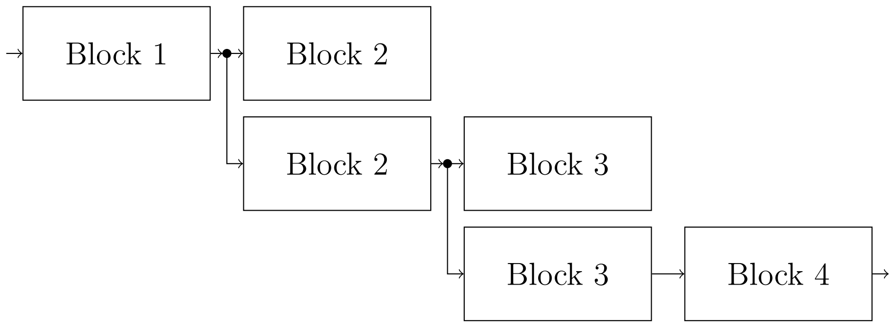

Figure 5: Example of decomposition into separate blocks using the modified block-wise method.

This approach is especially suitable for guitar preamp simulations, which was proved in the simulations. The implementation of individual blocks can be based on the direct approximation of nonlinear ODEs resulting in very efficient simulation algorithm. It was tested in simulation of the guitar preamp consisting of four triode stages.

The algorithm with the spline interpolation and the block-wise method was implemented in C++ language as the Virtual Studio Technology (VST) plug-in effect to test it in real-time. It was tested with a 2.66 GHz i7 Intel Mac with 4 GB RAM at the sampling frequency of 48 kHz using the external audio interface M-Audio Fast Track Pro. The CPU load was around 3 %. The total size of all tables required for the simulation of the whole preamp is 10 MB.

7.2 Decomposition of the DK-method nonlinear core

The block decomposition discussed in the previous section provides very efficient simulation of circuits where the number of circuit inputs represented by real inputs and energy-storing components is lower than the number of nonlinear equations. Otherwise, the DK-method is more appropriate and it can be of course used together with the modified block-wise method. The advantage of this approach is the simulation of the circuit with more energy-storing components. This approach will be studied on the simulation of the guitar preamp with topology typical for Fender preamps [24] published in [7]. The circuit schematic of the studied preamp is in Figure 6. It consists of two triodes 12AX7 and the tone stack connected between them. Components R_6, R_7, R_8 and R_9 are potentiometers of variable circuit parameters treble θ, middle μ, bass β and volume λ. The nonlinear equation has 4 independent inputs (according

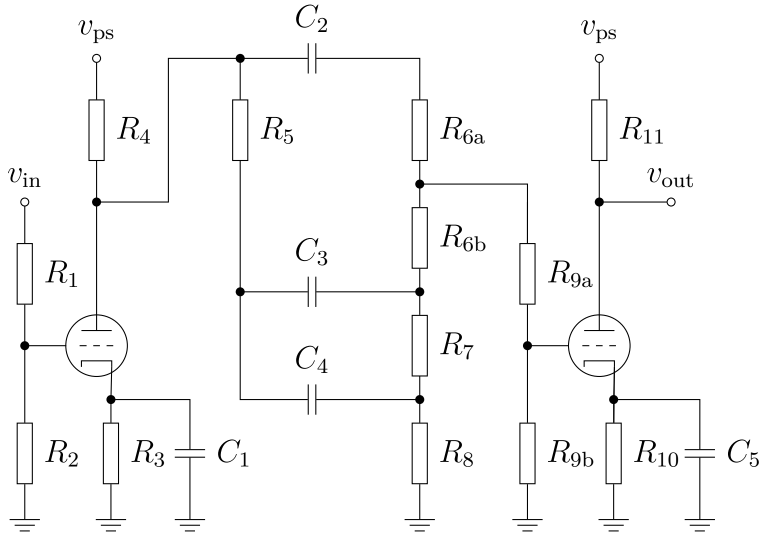

Figure 6: Circuit schematic of the Fender type guitar preamp.

to the dimension of the matrix $\mathbf{K}$) for which it has to be precomputed. First of all, it is necessary to find ranges for input variables p_1, p_2, p_3 and p_4. This can be done using $\mathbf{p}[n] = \mathbf{Gx}[n] + \mathbf{Hu}[n]$ and it will depend on input signal values and circuit state values as well. As the result of the precomputation, there are four 4D look-up tables for circuit currents or voltages for constant parameters or 7D look-up tables for the parametric circuit.

Reduction of data can be done using the correlation analysis of precomputed data. Covariance matrices for input variables and function values were in the form (expressed only for one row of input data)

$$C = \operatorname{cov}\left(\left[p_1^{i_1}\, p_2^{i_2}\, p_3^{i_3}\, p_4^{i_4}\, f(p_1^{i_1}, p_2^{i_2}, p_3^{i_3}, p_4^{i_4})\right]\right) \tag{18}$$

where the superscripts denote indexes $i_1 \in [1, N_1]$, $i_2 \in [1, N_2]$, $i_3 \in [1, N_3]$ and $i_4 \in [1, N_4]$ for lengths of vectors N_1, N_2, N_3 and N_4 and function $f(p_1, p_2, p_3, p_4)$

was substituted with precomputed nonlinear functions $i_{\mathrm{g1}}(\mathbf{p})$, $i_{\mathrm{p1}}(\mathbf{p})$, $i_{\mathrm{g2}}(\mathbf{p})$ and $i_{\mathrm{p2}}(\mathbf{p})$. The resulting covariance between nonlinear functions and inputs p_1, p_2, p_3 and p_4 is stated in Table 3.

Table 3: Covariance between precomputed functions and inputs.

	$i_{\mathrm{g1}}(\mathbf{p})$	$i_{\mathrm{p1}}(\mathbf{p})$	$i_{\mathrm{g2}}(\mathbf{p})$	$i_{\mathrm{p2}}(\mathbf{p})$
p_1	$-5.23 \cdot 10^{-4}$	$-6.98 \cdot 10^{-3}$	$5.55 \cdot 10^{-4}$	$8.61 \cdot 10^{-4}$
p_2	$1.12 \cdot 10^{-6}$	$-1.02 \cdot 10^{-1}$	$6.65 \cdot 10^{-3}$	$1.17 \cdot 10^{-2}$
p_3	$-1.17 \cdot 10^{-8}$	$1.11 \cdot 10^{-3}$	$-1.40 \cdot 10^{-2}$	$-2.47 \cdot 10^{-2}$
p_4	$5.49 \cdot 10^{-14}$	$5.02 \cdot 10^{-9}$	$5.40 \cdot 10^{-8}$	$-8.81 \cdot 10^{-3}$

As it can be seen from the table, some nonlinear functions are almost independent of some input variables, which enables omitting some inputs.

The second important point is that due to missing connections between nonlinear functions, the further simplification and decomposition can be done. The nonlinear DK-method equation can be split in this case into

$$\begin{aligned} v_{\mathrm{gk1}} &= k_{11} i_{\mathrm{g}}(v_{\mathrm{gk1}}) + k_{12} i_{\mathrm{p}}(v_{\mathrm{gk1}}, v_{\mathrm{pk1}}) + p_1 \\ v_{\mathrm{pk1}} &= k_{21} i_{\mathrm{g}}(v_{\mathrm{gk1}}) + k_{22} i_{\mathrm{p}}(v_{\mathrm{gk1}}, v_{\mathrm{pk1}}) + \underbrace{k_{23} i_{\mathrm{g}}(v_{\mathrm{gk2}}) + p_2}_{\overline{p_2}} \end{aligned} \tag{19}$$

and

$$\begin{aligned} v_{\mathrm{gk2}} &= \underbrace{p_3 + k_{32} i_{\mathrm{p}}(v_{\mathrm{gk1}}, v_{\mathrm{pk1}})}_{\overline{p_3}} + k_{33} i_{\mathrm{g}}(v_{\mathrm{gk2}}) + k_{34} i_{\mathrm{p}}(v_{\mathrm{gk2}}, v_{\mathrm{pk2}}) \\ v_{\mathrm{pk2}} &= k_{43} i_{\mathrm{g}}(v_{\mathrm{gk2}}) + k_{44} i_{\mathrm{p}}(v_{\mathrm{gk2}}, v_{\mathrm{pk2}}) + p_4 \end{aligned} \tag{20}$$

where terms $k_{23} i_{\mathrm{g}}(v_{\mathrm{gk2}})$ and $k_{32} i_{\mathrm{p}}(v_{\mathrm{gk1}}, v_{\mathrm{pk1}})$ are mutual impacts of adjacent triodes. Equations (19) and (20) can be computed separately for input variables p_1, $\overline{p_2}$ and $\overline{p_3}$, p_4 and further functions

$$\overline{f_{\mathrm{ip1}}}(p_1, \overline{p_2}) = i_{\mathrm{p}}(v_{\mathrm{gk1}}(p_1, \overline{p_2}), v_{\mathrm{pk1}}(p_1, \overline{p_2})) \tag{21}$$

$$\overline{f_{\mathrm{ig2}}}(\overline{p_3}, p_4) = i_{\mathrm{g}}(v_{\mathrm{gk1}}(\overline{p_3}, p_4), v_{\mathrm{pk1}}(\overline{p_3}, p_4)) \tag{22}$$

can be introduced. Considering this decomposition together with the reduction of look-up tables based on the correlation analysis, the whole model is split into two independent parts, each containing one tube. Grid and plate currents of the first tube are computed from the approximated functions

$$i_{\mathrm{g1}} = i_{\mathrm{g1app}}(p_1, p_2) \tag{23}$$

$$i_{\mathrm{p1}} = i_{\mathrm{p1app}}(p_1, p_2, p_3, \theta, \mu, \lambda) \tag{24}$$

$$\tag{25}$$

where the redundant inputs were neglected. Then, the mutual interaction of both tubes is known and it has been already included in the plate current i_{p1}. This current is subsequently used as the additional contribution to the input p_3 as it can be seen in (20) and currents of the second tube are obtained from

$$i_{\mathrm{g2}} = i_{\mathrm{g2app}}(p_3 + k_{32} i_{\mathrm{p1}}, k_{33}) \tag{26}$$

$$i_{\mathrm{p2}} = i_{\mathrm{g2app}}(p_3 + k_{32} i_{\mathrm{p1}}, p_4, k_{33}). \tag{27}$$

Inputs θ, β, μ, λ and k_{33} are used for modeling of the parametric part of the circuit.

The algorithm for the simulation of the preamp was written as the Matlab mex function in C language. The algorithm consists of nonuniform spline interpolations up to third order for p variable interpolation and linear interpolations for the parameter variable interpolation. The parameter interpolation is performed before the main processing loop. The computational complexity of the algorithm was around 10 % on the 2.66 GHz Intel processor but more than half of it was spent on matrix operations. The original state-space model without approximations consumes around 76 %.

7.3 DK-model Decomposition Using Connection Components

The decomposition described in the previous section allows the efficient decomposition of feed-forward circuits with triodes or transistors without any feedbacks. This is due to the form of the **K** matrix, which is tri-diagonal, and the model of the triodes or transistors which are built by two nonlinear current functions controlled by two control voltages. Nevertheless, the previous model of the preamp did not take the Miller effect into account which in case of circuit with tubes can manifest in the audible area [25]. This effect is caused by the parasitic capacitance between the grid and plate node of the tube. This capacitance causes the local feedback of each tube amplifier stage between the plate and grid circuit and as a result, the **K** matrix does not have the tri-diagonal form anymore. The full model with Miller capacitances will be tested in the Marshall JCM 800 preamp simulation from Figure 7 [18].

Nevertheless, the circuit still has the feed-forward topology where there can be identified some blocks which are connected in one node with current flowing between them and thus the decomposition into blocks should be still possible. This decomposition of the circuit is illustrated in Figure 7.

It consists of three blocks – tube amplifier stages, each containing one triode. The mutual interaction between the blocks is caused by connection currents

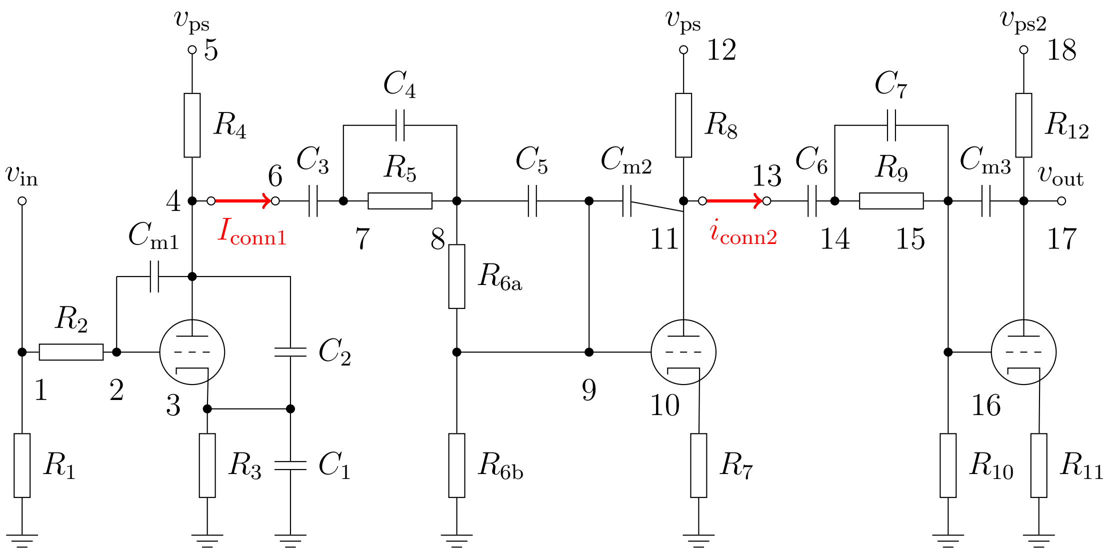

Figure 7: Circuit schematic of the Marshall JCM 800 guitar preamp with the decomposition into blocks.

i_{conn1} and i_{conn2} which are marked in Figure 7 with the color. If we consider the simulation of each tube amplifier stage separately, then these connection currents are actually additional current inputs of given tube amplifier stages. Further, we can define a nonlinear connection component with the connection current flowing through but zero voltage across the component terminals. The analysis of the whole circuit can be then done in the same way using incidence matrices as the analysis of the separate parts or any other circuit without these connection components. The only change is in the matrix $\mathbf{N_n}$ which defines positions of nonlinear components and in which the connection components have to be added – each connection component as one new row in the matrix with 1 for the nodes where the component is connected and 0 for the others nodes. As a result, tube amplifier stages are clearly separated in the $\mathbf{K}$ matrix.

However, connection components are not standard circuit components but as was stated above, the voltage across the component must be zero. Therefore, the standard DK-method nonlinearity has to be transformed into

$$\mathbf{0} = \mathbf{p} + \mathbf{Ki} - \mathbf{vZ}_{\text{n}} \tag{28}$$

where $\mathbf{Z}_{\text{n}} = \text{diag}\left(\begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 & 0 & 0 \end{bmatrix}\right)$ where zeros are at positions of rows with connection components in the $\mathbf{N}_{\text{n}}$ matrix and vectors $\mathbf{v}$ and $\mathbf{i}$ are extended with unknown connection currents and voltages. Further, the nonlinearity can be decomposed into

$$\mathbf{0} = \begin{bmatrix} \overline{p_1} \\ \overline{p_2} \end{bmatrix} + \begin{bmatrix} k_{11} & k_{12} \\ k_{21} & k_{22} \end{bmatrix} \begin{bmatrix} i_{\mathrm{g}}\left(v_1, v_2\right) \\ i_{\mathrm{p}}\left(v_1, v_2\right) \end{bmatrix} - \begin{bmatrix} v_1 \\ v_2 \end{bmatrix}, \tag{29}$$

$$\mathbf{0} = \begin{bmatrix} \overline{p_3} \\ \overline{p_4} \end{bmatrix} + \begin{bmatrix} k_{33} & k_{34} \\ k_{43} & k_{44} \end{bmatrix} \begin{bmatrix} i_{\mathrm{g}}\left(v_3, v_4\right) \\ i_{\mathrm{p}}\left(v_3, v_4\right) \end{bmatrix} - \begin{bmatrix} v_3 \\ v_4 \end{bmatrix}, \tag{30}$$

$$\mathbf{0} = \begin{bmatrix} \overline{p_5} \\ \overline{p_6} \end{bmatrix} + \begin{bmatrix} k_{55} & k_{56} \\ k_{65} & k_{66} \end{bmatrix} \begin{bmatrix} i_{\mathrm{g}}\left(v_5, v_6\right) \\ i_{\mathrm{p}}\left(v_5, v_6\right) \end{bmatrix} - \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \tag{31}$$

for arbitrary inputs $\overline{p_1}$, $\overline{p_2}$, $\overline{p_3}$, $\overline{p_4}$, $\overline{p_5}$, $\overline{p_6}$ and nonlinear triode currents i_{g1}, i_{p1}, i_{g2}, i_{p2}, i_{g3} and i_{p3} can be approximated for these inputs. Finally, the nonlinear connection function is given by

$$\begin{bmatrix} k_{71} & 0 \\ k_{72} & 0 \\ k_{73} & k_{83} \\ k_{74} & k_{84} \\ 0 & k_{85} \\ 0 & k_{86} \\ k_{77} & k_{87} \\ k_{78} & k_{88} \end{bmatrix}^{\mathrm{T}} \begin{bmatrix} i_{\mathrm{g1}}\left(p_1 + k_{17} i_{\mathrm{conn1}}, p_2 + k_{27} i_{\mathrm{conn1}}\right) \\ i_{\mathrm{p1}}\left(p_1 + k_{17} i_{\mathrm{conn1}}, p_2 + k_{27} i_{\mathrm{conn1}}\right) \\ i_{\mathrm{g2}}\left(p_3 + k_{37} i_{\mathrm{conn1}} + k_{38} i_{\mathrm{conn2}}, p_4 + k_{47} i_{\mathrm{conn1}} + k_{48} i_{\mathrm{conn2}}\right) \\ i_{\mathrm{p2}}\left(p_3 + k_{37} i_{\mathrm{conn1}} + k_{38} i_{\mathrm{conn1}}, p_4 + k_{47} i_{\mathrm{conn1}} + k_{48} i_{\mathrm{conn2}}\right) \\ i_{\mathrm{g3}}\left(p_5 + k_{58} i_{\mathrm{conn2}}, p_6 + k_{68} i_{\mathrm{conn2}}\right) \\ i_{\mathrm{p3}}\left(p_5 + k_{58} i_{\mathrm{conn2}}, p_6 + k_{68} i_{\mathrm{conn2}}\right) \\ i_{\mathrm{conn1}} \\ i_{\mathrm{conn2}} \end{bmatrix} = - \begin{bmatrix} p_7 \\ p_8 \end{bmatrix} \tag{32}$$

with two unknown variables i_{conn1} and i_{conn2} to be solved numerically.

The validity of the block decomposition was successfully tested by the comparison of the output signals of the simulation without the block decomposition and the simulation with this block decomposition. The computational cost of the algorithm implemented using C language was between 10 and 15 % oc CPU on the Intel 2.6 GHz processor.

8 Quality of Simulation

The quality of the simulation of audio effects can be determined using the subjective evaluation of quality, which requires listening tests with several people. Results of the listening test must be then processed statistically to find final results of the test. The quality was tested on the simulated preamp ENGL E530. The ABX Double Blind Test was chosen [26]. Listeners have two reference samples A and B and they have to determine whether the unknown sample X is the sample A or B. The statistical hypothesis H_0 was introduced:

differences between the measured and the simulated signal are not audible and listeners have to guess the answer with the probability $p_0 = 0.5$ The alternative hypothesis H_1 says that the differences are audible.

The listening test had 15 participants – guitarists, professional musicians, sound and audio engineers. Each listener had to identify 10 sound samples for each channel. The results are in Table 4.

Table 4: Results of subjective evaluation of preamp simulation.

Channel	Correct identifications	Difficulty to judge
Clean	83	4.41
Crunch	90	4.60

Results showed very good quality of the simulation. The number of correct identifications of the clean channel of the preamp was not sufficient to reject the null hypothesis and therefore we can say that measured and simulated signals are not distinguishable. The number of correct identifications of the crunch channel was sufficient to reject the null hypothesis. It means that signals are distinguishable but it requires big effort due to the high value of difficulty.

9 Conclusion

The main aim of the thesis was to design algorithms for the digital signal processing which emulate analog audio circuit with sufficient accuracy but still are able to work in real-time. The main contributions of this work are:

- **Incorporation of transformer model into the automated DK-method.** It allows automated derivation of DK-method matrices from incidence matrices for circuits containing the transformer (e.g. input stages of some audio effects or the tube push-pull power amplifier). Further, four types of transformer models were compared with regards to the computational complexity, the stability of solution, the accuracy and audio perception. Modeling the hysteresis effect significantly increases the computational cost and can introduce the numerical instability but the impact of the hysteresis effect to the output signal is very subtle. It is therefore sufficient to use the GC-model without the hysteresis or the linear model of transformer core whose use is still sufficient, without introducing audible difference in the output signal, in some circuits.

- **Incorporation of the operational amplifier model into the automated DK-method.** It allows the automated derivation of DK-method matrices from incidence matrices for circuits containing operational amplifiers using the linear and the nonlinear model of OPA.
- **Method of direct approximation of nonlinear ODEs.** This algorithm provides very efficient simulation of dynamic nonlinear systems which have order of the system (number of states) lower than the number of nonlinear components (nonlinear equations) in the system. The big advantage is that stored data in look-up tables are independent on the sampling frequency value.
- **Modified block-wise method.** This method was designed for the decomposition into simpler blocks in such way that the mutual interaction between adjoined blocks is preserved. This method however assumes that there is no interaction between the first block and third block connected to the second block as the nonlinear load.
- **Block decomposition of the DK-method nonlinear core.** This method enables the simulation of the whole circuit without a prior decomposition into blocks. Only the DK-method nonlinear equation can be decomposed into separate equations in case that the simulated circuit has feed-forward topology. This method also assumes very small dependence between the first and the third nonlinear circuit component (connected in series). This dependence can be revealed using the correlation analysis.
- **Block decomposition using connection currents.** This decomposition also enables the simulation of the whole circuit at once. But comparing to the previous one, it is suitable also for feedback systems. DK-method nonlinear equations are split using the unknown connection current. Each simple block, typically of dimension of two, can be easily approximated. Then only the unknown connection currents are computed. The disadvantage is that connections currents have to be computed numerically. On the other hand, results of this method are comparable to the numerical solving of the whole circuit while the number of unknown variables to be solved is much lower and thus the computational cost is also much lower.

All proposed algorithms can find their employment in several applications for musicians and some often have already been used in software for guitar real-time processing [1]. Listening test showed that very high level of quality of emulation was achieved.

[1]http://www.audiffex.com/EN/amplion.html

Author's Publications

[1] J. Mačák, "Simulation of the diode limiter using linear time variable filter," in *Proceedings of the 15th Conference STUDENT EEICT*, Brno, Czech Republic, 2009, pp. 264–268.

[2] ——, "Návrh algoritmů číslicového zpracování zvukových signálů pro simulaci kytarového komba," in *Proceedings of the 14th Conference STUDENT EEICT*, Brno, Czech Republic, 2008, pp. 73–75.

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Curriculum Vitae

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Personal information

- Born on 4th October, 1983.
- Czech nationality

Education

- Brno University of Technology, Faculty of Electrical Engineering and Communication, Czech Republic, doctoral degree programme, 09/2008 – 10/2012
- Helmut Schmidt University - University of the Federal Armed Forces Hamburg, Department of Electrical Engineering, Hamburg, Germany, 02/2012 – 05/2012
  – research stay
- Brno University of Technology, Faculty of Electrical Engineering and Communication, Czech Republic, master's degree programme, 09/2006 – 06/2008
  – final state exam with honours
- Brno University of Technology, Faculty of Electrical Engineering and Communication, Czech Republic, bachelor's degree programme, 09/2003 – 09/2006
  – final state exam with honours
- Gymnázium Lipník nad Bečvou, 09/1995 – 06/2003
  – final exam with honours

Employment

- Brno University of Technology, Department of Telecommunication, Brno, 08/2009 – 05/2012
  – assistant on project CZ.1.07/2.3.00/09.0222

- Janáček Academy of Music and Performing Arts in Brno (JAMU in Brno), Brno, 01/2010 – 08/2012
  - specialized assistant and lecturer in field of audio effects and instruments design
- Disk Multimedia, spol. s r.o., Boskovice, 01/2006 – 10/2012, software developer
  - DSP algorithm design
    - digital simulation of Gallien-Krueger bass amplifiers `http://www.audiffex.com/EN/gallien.html`,
    - digital simulation of TC Electronic vintage guitar pedals `http://www.tcelectronic.com/vintageguitarpedalbundle.asp`,
    - digital simulation of analog guitar amplifiers and effects – ampLion `http://www.audiffex.com/EN/amplion.html`,
    - audio effects for TC Powercore system – inValve `http://www.audiffex.com/EN/invalve.html`,
  - audio plug-in development
  - team-leader of audio effects development
  - GUI development

Participation in Projects

- CZ.1.07/2.3.00/09.0222 – Educational Center for Increasing the Interest of Young People in Research into Information and Communication Technologies. Holder: Ing. Kubánek. 2009–2012
- FR-TI1/495 – Manifold System for Multimedia Digital Signal Processing. Holder: Ing. Schimmel. 2009–2012
- 2704/G1/2011 – Extension of Laboratory Assignments Addressing Audio Effect Measurement. Holder: Ing. Mačák. 2011
- FEKT-S-11-17 – Research of Sophisticated Methods for Digital Audio and Image Signal Processing. Holder: Prof. Z. Smékal. 2011
- MSM21630513 – Electronic Communication Systems and Technologies of Novel Generations (ELKOM). Holders: Prof. Z. Raida, Prof. K. Vrba, Prof. J. Jan. 2008–2011
- 2912/G1/2010 – Innovation of Laboratory Assignments of Subject Electroacoustics. Holder: Ing. Mačák. 2010
- FEKT-S-10-16 – Research on Electronic Communication Systems. Holder: Prof. K. Vrba. 2010

- FT-TA3/010 – Spatial Effects for Multichannel Digital Signal Processing Systems. Holder: Ing. Schimmel. 2009.

Invited Talks

- Simulation of analog audio devices in real-time. Audio Engineering Society, Prague 1/12/2010.
- Real-time digital simulation of analog audio effects, SPLab Workshop 2011, Brno 28/10/2011.

Results

- Publications: 13
  - In international journals with Impact Factor: 1
  - In proceedings of international conferences: 6
  - In other journals: 4
  - In other conferences: 2
- Software/Products: 1
- Citations (without self-citations): 8
- h-index according to Web of Science: 1
- responses to publications from foreign experts: 3

Awards

- EEICT 2008 student conference and competition – 3rd prize
- EEICT 2090 student conference and competition – 1sd prize

Abstract
The work deals with the real-time digital simulation of guitar amplifiers considered as nonlinear analog audio effects. The main aim is to design algorithms which are able to simulate complex systems in real-time. These algorithms are mainly based on the automated DK-method and the approximation of nonlinear functions. Quality of the designed algorithms is evaluated using listening tests.
Abstrakt
Práce se zabývá číslicovou simulací kytarových zesilovačů, jakož to nelineárních analogových hudebních efektů, v reálném čase. Hlavním cílem práce je návrh algoritmů, které by umožnily simulaci složitých systémů v reálném čase. Tyto algoritmy jsou převážně založeny na automatizované DK-metodě a aproximaci nelineárních funkcí. Kvalita navržených algoritmů je stanovana pomocí poslechových testů.