Abstract: - Unfortunately, complex transformer design problems cause that transformers are not common in audio circuits design except professional devices. This happens partly for their physical construction disadvantages as bulky dimensions and relatively high price and partly because of the fact that designers tend to overlook them even if their employment is reasonable because their design is not trivial. However, in some cases using the transformer can simplify the circuit and improve its noise parameters. In order to make the audio transformer design easier the authors of this paper are developing progressive methods to enable the computer-aided transformer design. One of these methods is based on employing the evolutionary algorithms. In this paper the transformer optimisation by means of Differential evolution is described.

Key-Words: - audio transformers, artificial intelligence, Differential evolution, computer aided design, electrical circuits

1 Introduction
Evolutionary algorithms are computational methods inspired by natural processes generally described as Darwin’s theory. The similarity of those algorithms with processes which occur in nature is fairly close. Both of these are aimed on optimization. The optimization which proceeds in the nature can be considered as a way how to select those best of individuals (organisms) whose abilities to survive and adapt to surrounding environment are the highest. Thanks to organic variability ensured by mutation and cross-over it is possible to expect even improved abilities of descendants of them in the next generation. The abilities of individuals can be imagined as objects of optimization. In the case of evolutionary algorithms the individuals have a form of vectors. The parameters of the vectors represent the “abilities” which are subsequently optimized. Mutation and cross-over processes are realized by vectors transformations and mathematic computations (see below). The results of optimization by evolutionary algorithms can be unexpectedly (and positively) surprising because of ability of evolutionary algorithms to exceed local extremes especially. In this paper, the optimisation of an audio transformer design by means of the Differential evolution [6-8] is described in order to show how the complex design of the transformer can be simplified by using the methods of artificial intelligence.

Considering optimization, evolutionary algorithms can be used for other purposes too (e.g. for function approximation and pattern recognition).

Except Differential evolution, there are also other evolutionary algorithms (e.g. SOMA [9] etc.). But there is chosen as one of them Differential evolution in this research. A short description of Differential evolution and its parameters can be found below.

2 Problem Formulation
Any audio small-signal transformer can be modeled with an equivalent circuit (see [5]) and a set of appropriate equations. This set of equations is not easy to solve manually, because there are many mutual dependences. On the other hand, methods of artificial intelligence seem to be ideal for such evaluation. For the purpose of this paper, the transformer described in [4-5] has been considered to be optimized. A basic method of determining whether such transformer can be manufactured or not is to be found in [4] as well as the tables of standardized dimensions of the transformer cores and usually employed wires. In this paper the solution of the set of the equations by means of the differential evolution is described.
2.1 Basic transformer equations

In low-power audio transformers the number of primary winding turns is not determined primarily by the power of the transformer and its core induction but by the frequency response of the transformer at low frequencies. This is because the power and the core induction is usually low enough to be neglected but the inductance of the primary winding together with the resistances at the primary and the secondary part of the transformer create a frequency dependent divider (see the equivalent circuit of the transformer in [5]). By recalculating the parameters to the primary side of the transformer one can for the range of low frequencies obtain the following equation:

\[ Z_{PRI} = \frac{RP + \frac{1}{1 + \frac{1}{\frac{1}{RX}}} \frac{1}{\frac{1}{N^2(\frac{1}{RS + RL})}}}{} \]  \[ (\Omega) \]  \[ (1) \]

Where \( Z_{PRI} \) is the impedance at the primary winding of the transformer, \( RP \) is the resistance of the primary winding, \( RS \) is the resistance of the secondary winding, \( RL \) is the load resistance, \( N \) is the transformer conversion ratio, \( LP \) is the primary winding inductance and \( RX \) represents losses in the core as an additive resistor. Usually \( RX \) can be neglected while \( RG \) as the internal source resistance must be considered to create a voltage divider together with \( Z_{PRI} \). Then the minimum frequency \( f_{\text{min}} \) is determined as follows:

\[ f_{\text{min}} = \frac{(RG + RP)(RS + RL)\frac{1}{N^2}}{2\pi LP \cdot \left(\frac{1}{N^2} + \frac{1}{RS + RL} + \frac{1}{(RS + RL)(RG + RP)}\right)} \]  \[ [Hz] \]  \[ (2) \]

Therefore there is a need to express the minimum \( LP \) that is necessary to achieve the required \( f_{\text{min}} \):

\[ LP = \frac{(RG + RP)\frac{1}{N^2}}{2\pi f_{\text{min}} \cdot \left(\frac{1}{N^2} + \frac{1}{RS + RL} + \frac{1}{RG + RP}\right)} \]  \[ [H] \]  \[ (3) \]

The number of the primary winding turns \( n_1 \) is then determined by the required \( LP \) according to [3] as follows:

\[ n_1 = \sqrt{LP \cdot \left(\frac{l_{\text{air}}}{l_{\text{air}} + l_{\text{air}}}\right) \cdot 10^5} \]  \[ (4) \]

Where \( S \) is the core mass cross-section in \([\text{cm}^2]\) as can be found in the appropriate table (see [4]), \( k \) is a constant considering the inaccuracy and losses (usually \( k = 0.9 \)), \( l \) is the average length of a magnetic line of force inside the core mass, \( \mu_r \) is the relative permeability of the core sheets and \( l_{\text{air}} \) is the width of the air gap between the sheets E and I.

caused by the inaccuracy of the sheet cutting (usually \(10^{-4}\)). However it is not easy to determine the number of the secondary wiring turns. In order to consider the attenuation caused by \( RP \) and \( RS \) the correction factor \( m \) representing the relative winding resistance to the source and load resistances shall be employed. Then the following equations can be applied:

\[ m = \left(\frac{RP + \frac{RS}{N^2}}{RG + \frac{RL}{N^2}}\right) \cdot 100 \]  \[ [%] \]  \[ (5) \]

\[ n_2 = \left(1 + \frac{m}{100}\right) \cdot n_1 \cdot N \]  \[ (6) \]

At this point, according to [3], the following equation for the core induction can be deduced:

\[ B = \frac{3243.24 \cdot U_{i \text{max}}}{f_{p \text{max}} \cdot S \cdot k \cdot n_1} \]  \[ [T] \]  \[ (7) \]

For ordinary EI sheets the \( B \) should not be higher than approximately 0.5 T. \( U_{i \text{max}} \) stands for maximum input voltage of the transformer and \( f_{p \text{max}} \) is the lowest frequency at which the distortion caused by the non-linearity of the core is observed.

Another problem to be solved consists in determining the primary and secondary winding resistances that must be found in order the equations (1) to (7) could be evaluated. The resistances depend on the cross-sections of the appropriate wires and their length which is determined by the number of the turns and the average length of one turn for the predefined core sheets. Assuming the specific electrical resistivity of a copper wire is 0.0178 \( \Omega \text{mm}^2/\text{m} \), following expressions can be used:

\[ RP = \frac{0.0178 \cdot n_1 \cdot o}{S_1} \]  \[ [\Omega] \]  \[ (8) \]

\[ RS = \frac{0.0178 \cdot n_2 \cdot o}{S_2} \]  \[ [\Omega] \]  \[ (9) \]

Where: \( S_1 \) is the primary winding wire cross-section in \([\text{mm}^2]\), \( S_2 \) is the primary winding wire cross-section in \([\text{mm}^2]\), \( n_1 \) is the number of primary winding turns, \( n_2 \) is the number of secondary winding turns and \( o \) is the average length of a single current turn for the predefined core.

It is also usual to define primary to secondary winding resistance ratio (WRR) that can be in specific cases employed for the purpose of the noise optimisation:

\[ r = \left(\frac{N}{1 + \frac{m}{100}}\right)^2 \]  \[ RS \]  \[ (10) \]

The attenuation caused by the transformer due to the resistances and the load can be expressed as follows:
At high frequencies the performance of the transformer is dependent on the leakage inductance of all windings and parasitic capacities that occur on all windings, among them and between the windings and the transformer core. These parameters are highly dependent on the internal winding arrangement. If the winding arrangement is made in the way described in [4-5], the following equations can be used to express the parameters:

$$n_{l1} = \frac{[n_1]}{[h_w - 2t_f] \cdot d_{out 1}}$$  \hspace{1cm} (12)  

$$n_{l2} = \frac{[n_2]}{[h_w - 2t_f] \cdot d_{out 2}}$$  \hspace{1cm} (13)  

$$t_w = n_{l1} \cdot d_{out 1} + 2 \cdot n_{l2} \cdot d_{out 1} + t_f + 2 \cdot t_{iz}$$  \hspace{1cm} (14)

The meanings of the variables from (12), (13) and (14) are as follows: $n_1$ is the number of layers of the primary winding, $n_2$ is the number of layers of one section of the secondary winding (symmetrical two-section winding is considered!), $t_f$ is coil former mass thickness, $t_{iz}$ is the isolation between the sections thickness, $d_{out 1}$ is the outer diameter of the primary winding wire and $d_{out 2}$ is the outer diameter of the secondary winding wire (considering the isolating lacquer). The total thickness of the coil $t_w$ should be slightly lower than the maximum sheet window height $t_0$ but not much in order all the above mentioned equations were valid. See (15):

$$0.8 \cdot t_0 < t_w < 0.95 \cdot t_0$$  \hspace{1cm} (15)

According to [3] the leakage inductance $LL$ can be expressed as follows:

$$LL = \frac{1.68 \cdot n_1^2 \cdot \varepsilon \cdot (4 \cdot t_{iz} + t_w)}{h_w} \cdot 10^{-6} \ [H]$$  \hspace{1cm} (16)

More complicated situation occurs at expressing the parasitic capacities. The prevailing resonant frequency of the transformer can be estimated using the consideration that according to the internal transformer arrangement (see [4-5]) only those capacities cannot be neglected:

- Primary winding capacity $C_p$,
- Prim. to sec. winding capacity $C_{ps}$ (see [4]),
- Secondary winding capacity $C_s$ that will at the primary part of the transformer seem as $C_s'$,
- Secondary winding to the core $C_{SC}$ capacity that will at the primary part of the transformer seem as $C_{SC'}$.

The capacity is spread across the whole winding and to be evaluated generally, it shall be calculated on one side of the transformer. Following approximations may be used to estimate the total capacity and the prevailing resonant frequency of the transformer. First of all a capacity between a two of layer may be estimated according to (17)

$$C_{2l} = \frac{885 \cdot a \cdot h_w \cdot \varepsilon_r \cdot 10^{-14}}{t} \ [F]$$  \hspace{1cm} (17)

Where: $\varepsilon_r$ is a relative permittivity (usually $\varepsilon_r = 3$) and $t$ is the distance between the two layers that is determined by the thickness of their isolation (see (18)).

$$t = d_{out} - d_{in}$$  \hspace{1cm} (18)

If the winding consists of more turns, the voltage is spread across the whole winding and the capacity decreases according to the number of layer which is described by (19).

$$C = \frac{1.333}{n_l} \left(1 - \frac{1}{n_l}\right) C_{2l} \ [F]$$  \hspace{1cm} (19)

The $n_l$ parameter stands for the number of the layers (see (12), (13)). According to (17) the capacities $C_{PS}$ and $C_{SC}$ can be evaluated as well. The $t$ parameter then depends on the thickness of the isolation layer $t_{iz}$ or the core former mass thickness $t_f$:

$$t_{PS} = \frac{(d_{out 1} - d_{in 1})}{2} + \frac{(d_{out 2} - d_{in 2})}{2} + t_{iz}$$  \hspace{1cm} (20)

$$t_{SC} = \frac{(d_{out 2} - d_{in 2})}{2} + t_f$$  \hspace{1cm} (21)

When recalculating the $C_S$ and $C_{SC}$ to the primary side, the following equations shall be employed:

$$C_{SC}^{\prime} = \left(1 + \frac{m}{100}\right) \cdot C_{SC}$$  \hspace{1cm} (22)

$$C_S^{\prime} = 2 \cdot \left(1 + \frac{m}{100}\right) \cdot C_S$$  \hspace{1cm} (23)

In (23) the multiplying by 2 refers to the two parallel secondary windings of the transformer. The total wiring capacity recalculated to the primary side of the transformer can be then expressed as follows:

$$C_{tot} = C_p + C_{PS} + C_s' + C_{SC}'$$  \hspace{1cm} (24)

According to [2] the resonant frequency of the transformer may be expressed according to (25):
\[ f_r \approx \frac{1}{\pi \sqrt{\frac{2}{3} C_{tot} L L}} \quad [Hz] \quad (25) \]

Usually the prevailing resonant frequency is lower than the calculated \( f_r \). The following approximation may be employed:

\[ f_r' = 0.71 f_r \quad [Hz] \quad (26) \]

All the above mentioned equations have been employed in the optimisation task driven by DE.

2.2 Brief insight into Differential evolution

Differential evolution (DE) (described e.g. in [7, 8]) strictly comes out of above mentioned. In the case of DE, it is considered optimization as origination of new descendants with improved parameters in the dependence on their predecessors. The basis of optimization is a definition of a specimen which represents a general description of individuals thus number of parameters and their range. The specimen is a vector containing four parameters:

- Minimal transferred frequency \( f_{min} \),
- Index of wire parameters set for the primary winding \( i_{primary} \),
- Index of wire parameters set for the secondary winding \( i_{secondary} \),
- Index of isolation layer \( i_{tps} \).

The ranges for the optimization have been predefined as follows:

\[ f_{min} \in (0; 100) \quad [Hz] \quad (27) \]
\[ i_{primary} \in \{1; \text{number of rows in } T_{wires}\} \quad (28) \]
\[ i_{secondary} \in \{1; \text{number of rows in } T_{wires}\} \quad (29) \]
\[ i_{tps} \in \{1; \text{number of items in } v_{isolation}\} \quad (30) \]

Where: \( T_{wires} \) is a table with parameters of available wires (see [4]) and \( v_{isolation} \) is a vector containing available thicknesses of the isolation layer.

\[ v_{isolation} = \{100, 200, 300, 400\} \quad [\mu m] \quad (31) \]

When initial generation of individuals is created (or generated because the parameters of each individual are set randomly in accordance with ranges stated in specimen) the optimization can start.

The optimization is consisted of individual evaluations. It means that all individuals occurred in the current generation are evaluated thus their cost values are computed. The vector transformations and mathematic computations which ensure mutation and cross-over proceed as follows as well as the mathematic expression of DE (see [8]).

Within optimization, individuals of the current generation are subsequently selected. The selected individual is known as a current individual \( x_r \).

Additionally, other three individuals \( x_{r_1}, x_{r_2}, x_{r_3} \) are randomly selected. The first two of randomly-selected individuals are subtracted. The result is a new vector which is known as the differential vector. As the next step there is a mutation. The mutation has a form of multiplication of the mutation constant and the differential vector. The result of multiplication is also new vector. This one is known as the weighted differential vector.

\[ x_{r_j}^w = x_{r_j}^d + F \cdot (x_{r_j}^d - x_{r_i}^d) \quad (32) \]

\[ j = \{1, \ldots, \text{Dimension of problem}\} \]

As the final operation there is an origination of the test vector \( x_{test} \). The test vector is originated as a cross-over of the noise vector and the current individual selected in the beginning.

\[ x_{r_{ij}}^{test} = \left\{ \begin{array}{ll}
[&x_{r_{ij}}^d + F \cdot (x_{r_{ij}}^d - x_{r_i}^d)] \\
 x_{r_{ij}}^d &
\end{array} \right\} \quad (33) \]

In this moment, the cost values of the current individual and the test vector are computed. These are results of evaluation by the cost function \( f_{cost} \). The cost function represents a set of conditions which are basis of optimization.

\[ x_{r+1}^i = \left\{ \begin{array}{ll}
x_{r_{ij}}^{test} & \text{if } f_{cost}(x_{test}) \leq f_{cost}(x_{r}) \\
x_{r}^i & \text{else}
\end{array} \right\} \quad (34) \]

The one of the current individual and the test vector with better evaluation passes to the next generation.

As mentioned above, all individuals of the current generation are subsequently selected thus the evaluation of these individuals proceeds and the next generation is originated. The optimization process is similarly repeated in the next generations. The optimization ends when the number of generation is reached.

The Differential evolution has several parameters which are necessary to set and which were mentioned in (32-34). These parameters are:

- \( NP \) as a number of population / generation,
- \( F \) as a mutation constant,
- \( CR \) as a cross-over value,
- \( G \) which represents number of generation.

While designing transformer by DE there were tried different settings of mentioned parameters. The mostly used ones are \( NP = 1000, G = 10000, F = 0.8, CR = 0.4 \).

DE occurs in several variants. The differences of these variants consist of origination of the noise vector. This research utilizes DE/rand1/bin variation.
which considers origination of the noise vector as an equation (32).

2.2.1 The Cost function and its influence to the optimization

While designing transformer it is necessary to consider several conditions. The optimized parameters of transformer which are results of DE have to meet these conditions too. It can be ensured by the cost function which evaluates individuals which encodes parameters of transformer and equations (see section 3). The conditions which parameters of transformer have to accomplish have been defined as follows:

- \( f_{\text{min}} \) – min. transferred freq. in [Hz] – desired value,
- \( f_{\text{min-max}} \) – min. transferred freq. in [Hz] – the highest (worst) value that can be tolerated,
- \( m \) – optimal relative resistance of the windings,
- \( m_{\text{max}} \) – max. relative resistance of the windings that can be tolerated,
- \( r \) – primary to secondary WRR,
- \( r_{\text{min}} \) – min. acceptable primary to secondary WRR,
- \( r_{\text{max}} \) – max. acceptable primary to secondary WRR,
- \( B \) – optimal max. core induction in [T] to be reached at the frequency \( f_{\text{p-min}} \),
- \( B_{\text{max}} \) – max. tolerable core induction in [T] to be reached at the frequency \( f_{\text{p-min}} \),
- \( \text{att} \) – optimal attenuation of the transformer in [dB],
- \( \text{att}_{\text{max}} \) – max. tolerable attenuation of the transformer in [dB],
- \( t_{0} \) – core window height according to the type of the core (see [4]),
- \( t_{w} \) – total winding thickness.

The above mentioned parameters must comply with the set of constrains resulting from the design requirements.

3 Optimization processing and results

The optimization task was defined according to the schematics presented in [5] and the requirements defined in [4]. As can be seen in [4], the results achieved by a simple analytic algorithm led to a transformer with quite low resonant frequency. The aim of this task was to reach the optimized solution driven by the evolutionary algorithm under the same requirements and considerations. The basic requirements are as follows: \( R_{G} = 500 \, \Omega, R_{L} = 2 \times 200 \, k\Omega, C_{L} = 50 \, pF, N = 1 : 2 \times 10, f_{\text{min}} = 15 \, Hz, f_{\text{p-min}} = 25 \, Hz, f_{\text{p-max}} = 25 \, Hz, U_{\text{max}} = 2 \, V, S_{\text{min}} = 7.10^{-4} \, \text{mm}^{2}, \text{att}_{\text{min}} = 3 \, \text{dB}, \text{att}_{\text{max}} = 4 \, \text{dB}, m_{\text{min}} = 5, m_{\text{max}} = 20, f_{\text{p-std}} = 100 \, \mu m, f_{\text{std}} = 1, f_{\text{min}} = 0.25, f_{\text{max}} = 4, B_{\text{max}} = 0.3 \, T \). The wire parameters are defined in Table 2 in [4].

As well as in [4] the EI38/20 core has been chosen which results in the further parameters specified in Table 1.

### Table 1 – EI30/10 core parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Core mass cross-section</td>
<td>( S = 2.4 , \text{cm}^{2} )</td>
</tr>
<tr>
<td>Total height of the coil section</td>
<td>( t_{0} = 6.5 , \text{mm} )</td>
</tr>
<tr>
<td>Coil former material thickness</td>
<td>( t_{f} = 0.5 , \text{mm} )</td>
</tr>
<tr>
<td>Average length of a single turn</td>
<td>( o = 84 , \text{mm} )</td>
</tr>
<tr>
<td>Average length of a magnetic line of force inside the core mass</td>
<td>( l = 71.5 , \text{mm} )</td>
</tr>
<tr>
<td>Core window width</td>
<td>( h_{w} = 19 , \text{mm} )</td>
</tr>
<tr>
<td>Estimated air gap caused by the inaccuracy of the transformer sheets cutting</td>
<td>( l_{w} = 0.1 , \text{mm} )</td>
</tr>
<tr>
<td>Core mass relative permeability</td>
<td>( \mu_{r} = 1.000 )</td>
</tr>
</tbody>
</table>

A set of equations has been converted into the Wolfram Mathematica and DE was applied on this set. The set includes the equations (3) to (26) excluding (15), provided that at 11 the (35) is applied:

\[
\text{att} = -A \tag{35}
\]

According to the specifications described above, DE has provided several results. The most suitable one is enlisted in Table 2.

### Table 2 – One of possible results

<table>
<thead>
<tr>
<th>Param.</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_{\text{min}} )</td>
<td>Minimal transferred frequency</td>
<td>19.68 Hz</td>
</tr>
<tr>
<td>( S_{1} )</td>
<td>Primary winding wire cross-section</td>
<td>0.0153 mm²</td>
</tr>
<tr>
<td>( d_{w1} )</td>
<td>Prim. winding wire inner diameter</td>
<td>0.14 mm</td>
</tr>
<tr>
<td>( t_{p} )</td>
<td>Prim. to sec. isolation layer thickness</td>
<td>0.1 mm</td>
</tr>
<tr>
<td>( S_{2} )</td>
<td>Sec. winding cross-section</td>
<td>0.0007 mm²</td>
</tr>
<tr>
<td>( d_{w2} )</td>
<td>Sec. winding inner diameter</td>
<td>0.03 mm</td>
</tr>
<tr>
<td>( L_{P} )</td>
<td>Magnetizing inductance</td>
<td>1.83 H</td>
</tr>
<tr>
<td>( n_{1} )</td>
<td>Primary turns number</td>
<td>1,075</td>
</tr>
<tr>
<td>( n_{2} )</td>
<td>Secondary turns number</td>
<td>2 x 12,855</td>
</tr>
<tr>
<td>( R_{P} )</td>
<td>Primary winding resistance</td>
<td>105 Ω</td>
</tr>
<tr>
<td>( R_{S} )</td>
<td>Secondary winding resistance</td>
<td>2 x 13,729 Ω</td>
</tr>
<tr>
<td>( r )</td>
<td>Primary to secondary WRR</td>
<td>1.09</td>
</tr>
<tr>
<td>( B )</td>
<td>Core induction at ( U_{i} ) and ( f_{\text{p-min}} )</td>
<td>111.76 mT</td>
</tr>
<tr>
<td>( \text{Att} )</td>
<td>Attenuation</td>
<td>3.0 dB</td>
</tr>
<tr>
<td>( L_{L} )</td>
<td>Leakage inductance</td>
<td>7.8 mH</td>
</tr>
<tr>
<td>( C_{r} )</td>
<td>Primary winding capacity</td>
<td>367.71 pF</td>
</tr>
<tr>
<td>( C_{rS} )</td>
<td>Prim. to sec. winding capacity</td>
<td>444.59 pF</td>
</tr>
<tr>
<td>( C_{S} )</td>
<td>Secondary winding capacity</td>
<td>2 x 48.06 pF</td>
</tr>
<tr>
<td>( C_{SC} )</td>
<td>Sec. winding to transformer core capacity</td>
<td>26.24 pF</td>
</tr>
<tr>
<td>( f_{t} )</td>
<td>Estimated resonant frequency of the loaded transformer</td>
<td>30.73 kHz</td>
</tr>
<tr>
<td>( t_{w} )</td>
<td>Total winding thickness</td>
<td>6.28 mm</td>
</tr>
<tr>
<td>( n_{v1} )</td>
<td>Number of primary winding layers</td>
<td>11</td>
</tr>
<tr>
<td>( n_{v2} )</td>
<td>Number of sec. winding layers</td>
<td>2 x 36</td>
</tr>
<tr>
<td>( m )</td>
<td>Winding resistance to total resistances ratio</td>
<td>19.6 %</td>
</tr>
<tr>
<td>( R_{L'} )</td>
<td>Load resistance recalculated to the primary side of the transformer</td>
<td>706 Ω</td>
</tr>
<tr>
<td>( C_{L'} )</td>
<td>Load capacity recalculated to the primary side of the transformer</td>
<td>7,080 pF</td>
</tr>
<tr>
<td>( C_{tot'} )</td>
<td>Total parasitic capacity of the transformer recalculated to the primary side of the transformer</td>
<td>13,737 pF</td>
</tr>
</tbody>
</table>
The verification of the results is quite easy. The mechanical issues can be proven by calculating the thickness of the whole winding manually and checking whether the winding will fit into the core window. The electrical parameters may be proven by employing the equivalent transformer circuit (see [5] using the parameters found by DE. Because all the parameters must be recalculated in the primary side, C_{SC}' and C_{S}' parameters must be used instead of C_{SC} and C_{S}. Moreover, the RL, CL and RS must also be recalculated to the primary side according to the following equations:

\[ RL' = \frac{RL}{\left(1 + \frac{m}{100} \cdot N\right)^2} \]  
\[ RS' = \frac{RS}{\left(1 + \frac{m}{100} \cdot N\right)^2} \]  
\[ CL' = \left(1 + \frac{m}{100} \cdot N\right)^2 \cdot CL \]  

The frequency response and input impedance dependences on the frequency for the solution from Table 2 are depicted in Fig. 1 and 2.

Fig. 1 – Frequency response of the transformer

Fig. 2 – Input impedance response

4 Results discussion

As obvious from the results verification, the transformer should work properly within the boundaries required by the designer and should exhibit greater performance, namely the better resonant frequency, compared to the one constructed only upon the basic estimations described in [4].

The simulation of the results obtained by DE agreed with the requirements except of evaluation the f_{min} parameter that was usually higher than expected. This is caused by the recursions among the applied equations. This problem could be solved by changing the order of the sequence of the equations and/or by repeating the evaluation iteratively. This is a topic for the further research.

5 Conclusion

In this paper the description of how DE can be applied in the audio transformers design is provided. By means of DE a small step-up transformer with relatively high frequency range at high ohmic load on parallel secondary winding has been proposed. Because this subject seems to be quite perspective, the authors decided to continue in this research in order to create more general solution for different transformer types and configurations.

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