Computationally Efficient Simulation-Driven Design Optimization of Microwave Structures

Slawomir Koziel¹, Stanislav Ogurtsov¹, and Leifur Leifsson¹

¹ Engineering Optimization & Modeling Center, School of Science and Engineering, Reykjavik University, 101 Reykjavik Iceland, koziel@ru.is, stanislav@ru.is, leifurth@ru.is

Abstract
A simple yet computationally efficient methodology for design optimization of microwave structures is introduced. Our technique is based on sequential optimization of coarse-discretization electromagnetic (EM) models. The optimal design of the current model is used as an initial design for the finer-discretization one. The final design is then refined using a polynomial-based approximation model of the coarse-discretization EM data. The unavoidable misalignment between the polynomial and the fine model is corrected using space mapping. Our technique is straightforward to implement, it does not require a circuit-equivalent coarse model or any modification of the structure being optimized. It is also computationally efficient because the optimization burden is shifted to the coarse-discretization models. The proposed approach is demonstrated through the design of various microwave devices including microstrip bandpass filter, planar ultrawideband antenna and a coplanar-waveguide-to-microstrip transition.

Keywords: Computer-aided design (CAD), simulation-driven design, multi-fidelity optimization, microwave engineering, surrogate modeling.

1. Introduction
Due to increasing complexity of microwave structures, their analytical models can only be used—in many cases—to yield initial designs that need to be further tuned to meet given performance specifications and growing demands for accuracy. On the other hand, for some emerging classes of circuits such as ultrawideband (UWB) antennas [1] or substrate integrated circuits [2] there are no systematic design procedures available that would result in designs satisfying prescribed specifications. Therefore, EM-simulation-based design optimization becomes increasingly important. However, a serious bottleneck of simulation-driven optimization is its high computational cost. As a result, straightforward approaches employing electromagnetic solver directly in an optimization loop are impractical. Decomposition, i.e., breaking down an EM model into smaller parts and combine them in a circuit simulator to reduce the CPU-intensity of the design process is one possible approach to alleviate this problem [3], [4]. Still, this is only a partial solution as the EM-embedded co-simulation model is subjected to direct EM-driven optimization, and the “near-field” interaction or coupling between these smaller parts is then neglected thereby compromising accuracy of the final response and, in some situations, e.g., simulation of antenna radiators, making the solution unreliable.
Efficient simulation-driven design can be realized using a surrogate-based optimization (SBO) principle [5], [6], where the optimization burden is shifted to a surrogate model, computationally cheap representation of the structure being optimized (referred to as the fine model). The successful SBO approaches used in microwave area include space mapping (SM) [7]-[12], various forms of tuning [13]-[15] and tuning SM [16], [17], as well as response correction techniques [18]-[20]. Unfortunately, their implementation is not always straightforward: substantial modification of the optimized structure may be required (tuning), or additional mapping and more or less complicated interaction between auxiliary models is necessary (SM). Also, space mapping performance heavily depends on the proper selection of the surrogate model and its parameters [21]. On the other hand, tuning might not be directly applicable for radiating structures (antennas), whereas space mapping normally requires fast coarse model (e.g., physically-based circuit models). Such models might not be available for many important structures including broadband antennas, substrate-integrated and hybrid circuits.
Here, a simple yet computationally efficient design optimization methodology is introduced based on sequential optimization of coarse-discretization EM models. The optimal design of the current model is used as an initial design for the finer-discretization one. The final design is then refined using a polynomial-based approximation model of the coarse-discretization EM data. The unavoidable misalignment between the polynomial and the fine model is corrected using space mapping. The proposed methodology is very simple to implement, as unlike space mapping or other surrogate-based approaches it does not require a circuit-equivalent coarse model or any modification of the structure being optimized. It is also computationally efficient because the optimization burden is shifted to the coarse-discretization models. The proposed approach is demonstrated through the design of various microwave components including microstrip bandpass filter, planar ultrawideband (UWB) antenna, and a coplanar-waveguide-to-microstrip transition.
2. Multi-Level Design Optimization

In this section, we formulate the optimization problem (Section 2.1), describe the building blocks of the proposed optimization procedure (Sections 2.2 and 2.3), formulate the procedure (Section 2.4), and discuss some practical issues (Sections 2.5 and 2.6).

2.1. Design Problem Formulation

The goal is to solve the following optimization problem

\[ x^*_f = \arg \min U \left( R_f(x) \right) \]  

(1)

where \( R_{f}(x) \in \mathbb{R}^n \) denotes the response vector of a high-fidelity (fine) model of the device of interest; \( U \) is a given objective function (e.g., typically minimax), whereas \( x \in \mathbb{R}^2 \) is a vector of design variables. It is assumed that the computational cost of evaluating the fine model is high so that solving Eq. (1) directly is impractical.

2.2. Coarse-Discretization Models

The design optimization methodology described in this paper is based on a family of coarse-discretization models \{\( R_{c,j} \), \( j = 1, \ldots, K \} \) evaluated by the same EM solver as the one used for the fine model. Discretization of the model \( R_{c,j+1} \) is finer than that of the model \( R_{c,j} \), which results in better accuracy but also longer evaluation time. In practice, the number of coarse-discretization models is two or three.

2.3. Approximation-Based Design Refinement

Having the optimized design \( x^{(K)} \) of the last (and finest) coarse-discretization model \( R_{c,K} \), we evaluate it at all perturbed designs around \( x^{(K)} \), i.e., at \( x^{(K)} = [x^{(K)}_1 \ldots x^{(K)}_j + \text{sign}(k) \Delta_k d_k \ldots x^{(K)}_n]^T \), \( k = -m, -m+1, \ldots, n-1, n \). We use the notation \( R^{(k)} = R_{c,x}(x^{(K)}) \). This data can be used to refine the final design without directly optimizing \( R_f \). Instead, we set up an approximation model involving \( R^{(k)} \) and optimize it in the neighborhood of \( x^{(K)} \) defined as \([x^{(K)} - \Delta d, x^{(K)} + \Delta d] \), where \( \Delta = [\Delta_1, \Delta_2, \ldots, \Delta_n]^T \). The size of the neighborhood can be selected based on sensitivity analysis of \( R_{c,1} \) (the cheapest of the coarse-discretization models); usually \( \Delta \) equals 2 to 5 percent of \( x^{(K)} \).

Here, approximation is performed using a reduced quadratic model \( q(x) = [q_1, q_2, \ldots, q_m]^T \), defined as

\[ q_j(x) = q_j([x_k \ldots x_j]^T) = \lambda_{j,0} + \lambda_{j,1}x_k + \ldots + \lambda_{j,m}x_k^m \]  

(2)

Coefficients \( \lambda_{j,r}, j = 1, \ldots, m, r = 0, 1, \ldots, 2n \), can be uniquely obtained by solving the linear regression problems

\[
\begin{bmatrix}
1 & x^{(K)}_1 & \ldots & x^{(K)}_m & (x^{(K)}_1)^2 & \ldots & (x^{(K)}_m)^2 \\
\vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\
1 & x^{(K)}_1 & \ldots & x^{(K)}_m & (x^{(K)}_1)^2 & \ldots & (x^{(K)}_m)^2 \\
\end{bmatrix}
\begin{bmatrix}
\lambda_{j,0} \\
\vdots \\
\lambda_{j,2n}
\end{bmatrix}
= 
\begin{bmatrix}
R^{(0)}_{j,0} \\
\vdots \\
R^{(0)}_{j,n}
\end{bmatrix}
\]

(3)

where \( x^{(K)}_k \) is a jth component of the vector \( x^{(K)} \), and \( R^{(k)}_{j} \) is a jth component of the vector \( R^{(k)} \), i.e.,

In order to account for unavoidable misalignment between \( R_{c,k} \) and \( R_{c,j} \) instead of optimizing the quadratic model \( q \), it is recommended to optimize a corrected model \( q(x) + [(R(x^{(K)}) - R_{c,k}(x^{(K)}))] \) that ensures a zero-order consistency [22] between \( R_{c,k} \) and \( R_{c,j} \). The refined design can then be found as

\[ x' = \arg \min_{x^{(K)} - \Delta d \leq x \leq x^{(K)} + \Delta d} U(q(x) + [R(x) - R_{c,k}(x^{(K)})])] \]

(4)

where \( U(q(x) + [R(x) - R_{c,k}(x^{(K)})])] \) is the original model \( R_f \) is only evaluated at the final stage (step 4) of the optimization process. Operation of the algorithm in illustrated in Fig. 1. Coarse-discretization models can be optimized using any available algorithm. Here, depending on the problem, we use simple pattern search algorithm [23] or Matlab’s minimax routine fminimax [24].
Typically, the major difference between the responses of $R_f$ and coarse-discretization models $R_{f,j}$ is that they are shifted in frequency. This difference can be easily absorbed by frequency-shifting the design specifications while optimizing a model $R_{f,j}$. More specifically, suppose that the design specifications are described as \{\omega_k, \Delta \omega_k, s_k\}, $k = 1, ..., n_s$ (e.g., specifications $|S_2| \leq -3$ dB for $3$ GHz $\leq \omega \leq 4$ GHz, $|S_2| \leq -20$ dB for $1$ GHz $\leq \omega \leq 2$ GHz and $|S_2| \leq -20$ dB for $5$ GHz $\leq \omega \leq 7$ GHz would be described as \{3, 4; –3\}, \{1, 2; –20\}, and \{5, 7; –20\}). If the average frequency shift between responses of $R_{f,j}$ and $R_{f,j-1}$ is $\Delta \omega$, this difference can be absorbed by modifying the design specifications to \{\omega_k - \Delta \omega, \omega_k, s_k\}, $k = 1, ..., n_s$.

2.6. Selection of the Coarse-Discretization Models
As mentioned in Section 2.2, the number $K$ of coarse-discretization models is typically two or three. The first coarse-discretization model $R_{f,1}$ should be set up so that its evaluation time is at least 30 to 100 times shorter than the evaluation time of the fine model. The reason is that the initial design may be quite poor so that the expected number of evaluations of $R_{f,1}$ is usually large. By keeping $R_{f,1}$ fast, one can control the computational overhead related to its optimization. Accuracy of $R_{f,1}$ is not critical because its optimal design is only supposed to give a rough estimate of the fine model optimum. The second (and, possibly third) coarse-discretization model should be more accurate but still at least about 10 times faster than the fine model. This can be achieved by proper manipulation of the EM solver mesh density.

3. Design Examples
In this section we demonstrate the operation and performance of the proposed design methodology. We consider various types of microwave devices including a microstrip filter, an ultra-wideband band antenna, and a coplanar-waveguide-to-microstrip transition. In each case, the optimized design is obtained at a low computational cost corresponding to a few high-fidelity full-wave EM simulations of the structure in question.

3.1. High-Temperature Superconducting (HTS) Filter [25]
Filters are essential electronic components that are used to select specific frequency components from the signal. Microwaves filters are typically passive devices with electric properties depending on their geometry, dimensions, and, in cases of substrate integrated [2] and planar technologies [8], [14], dielectric constant of the substrate. Consider the high-temperature superconducting (HTS) filter shown in Fig. 2(a) [25]. The design parameters are $x = [L_1, L_2, S_1, S_2, S_3]^T$. The width of all the sections is $W = 8$ mil. A substrate of lanthanum aluminate is used with $\varepsilon_r = 23.425$ $H = 20$ mil. The filter is simulated in Sonnet em [26] using a grid of 0.5 mil $\times$ 0.5 mil (the $R_i$ model). The design specifications are $|S_2| \leq 0.05$ for $\omega \leq 3.966$ GHz, $|S_2| \geq 0.95$ for 4.008 GHz $\leq \omega \leq 4.058$ GHz, and $|S_2| \leq 0.05$ for $\omega \geq 4.100$ GHz. The initial design is $x^{(0)} = [196 196 190 20 92]$ mil. We are using two coarse-discretization models: $R_{f,1}$ (grid of 2 mil $\times$ 4 mil) and $R_{f,2}$ (grid of 1 mil $\times$ 2 mil). The evaluation times for $R_{f,1}$, $R_{f,2}$ and $R_i$ are about 2 min, 6 min and 51 min, respectively. Figure 2(b) shows the responses of $R_{f,1}$ at $x^{(0)}$ and at $x^{(1)} = [188 190 188 20 76 84]$ mil, its optimal design found using a pattern search, as well as the response of $R_{f,2}$ at $x^{(0)}$. Because of noticeable frequency shift between $R_{f,1}(x^{(0)})$ and $R_{f,2}(x^{(0)})$ (7 MHz on average) the design specifications were adjusted as described in Section 2.5 while optimizing $R_{f,1}$. Figure 3(a) shows the responses of $R_{f,2}$ at $x^{(0)}$ and at its optimized design $x^{(2)} = [188 189 188 20 76 86]$ mil, as well as the response of $R_i$ at $x^{(2)}$. Here, the average frequency shift between $R_{f,2}(x^{(0)})$ and $R_i(x^{(0)})$ is about 5 MHz and the design specifications are modified accordingly. Figure 3(b) shows the responses of $R_i$ at $x^{(2)}$ (specification error –0.01) and the refined design $x^* = [188 189 188 20.5 78 88]$ mil (specification error –0.02). Total optimization cost (Table 1) corresponds to only 10 evaluations of the fine-discretization model.
Table 1: Optimization cost of the HTS filter

<table>
<thead>
<tr>
<th>Algorithm Component</th>
<th>Number of Model Evaluations</th>
<th>Evaluation Time Absolute [min]</th>
<th>Relative to $R_f$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optimization of the coarse-discretization model $R_{c.1}$</td>
<td>104</td>
<td>195</td>
<td>3.8</td>
</tr>
<tr>
<td>Optimization of the coarse-discretization model $R_{c.2}$</td>
<td>26</td>
<td>152</td>
<td>3.0</td>
</tr>
<tr>
<td>Evaluation of the original (fine-discretization) model $R_f$</td>
<td>3</td>
<td>153</td>
<td>3.0</td>
</tr>
<tr>
<td>Total optimization time</td>
<td>N/A</td>
<td>500</td>
<td>9.8</td>
</tr>
</tbody>
</table>

Figure 2: HTS filter: (a) geometry [25], (b) responses of the coarse-discretization model $R_{c.1}$ at the initial design $x^{(0)}$ (dashed line) and at its optimized design $x^{(1)}$ (solid line), as well as the response of $R_{c.2}$ at $x^{(0)}$ (dotted line); design specifications are shifted by 7 MHz towards higher frequencies to absorb the frequency shift between $R_{c.1}(x^{(0)})$ and $R_{c.2}(x^{(0)})$.

Figure 3: HTS filter: (a) responses of the coarse-discretization model $R_{c.2}$ at $x^{(1)}$ (dashed line) and at its optimized design $x^{(1)}$ (solid line), as well as the response of $R_f$ at $x^{(1)}$ (dotted line); design specifications are shifted by 5 MHz toward higher frequencies to absorb the frequency shift between $R_{c.2}(x^{(1)})$ and $R_f(x^{(1)})$; (b) responses of the fine-discretization model $R_f$ at $x^{(2)}$ (dashed line) and at the refined final design $x^*$ (solid line); here the original design specifications are shown.

3.2. Ultra-Wideband (UWB) Planar Dipole Antenna

An antenna is an important part of any wireless communication channel. Differently from many state of the art antennas designed to work at narrow frequency bands [27], the UWB antenna [1] is required to support transmission and receiving communication signals over the ultra-wide frequency range, e.g. in North America the allocated spectrum for civil UWB communications is from 3.1 to 10.6 GHz [28]. Since the UWB antennas are to be an integral part of compact portable and even miniature devices [29], e.g., wires USB sticks, design of compact low cost antennas capable to support UWB communication in terms of matching, radiation characteristics, and introduced phase distortion over the UWB range is a challenging task.

Consider a planar dipole antenna (Fig. 4) consisting of the main radiator element and two extra “parasitic” strips [30].
The design variables are $x = [l_0, w_0, a_0, l_p, w_p, s_0]^T$. Other variables are: $a_1 = 0.5$ mm, $w_1 = 0.5$ mm. Rogers RT5880 laminate is used for the substrate dielectric, substrate height is $h = 1.58$ mm. The high-fidelity antenna model $R_f$ (10,250,412 mesh cells at the initial design) is simulated using the CST MWS transient solver [30]. The design objective is to obtain $|S_{11}| \leq -10$ dB for 3.1 GHz to 10.6 GHz. The initial design is $x^0 = [20 10 1 10 8 2]^T$ mm. We are using two coarse-discretization models: $R_{c,1}$ (108,732 mesh cells at $x^0$) and $R_{c,2}$ (427,548 mesh cells at $x^0$). The evaluation times for $R_{c,1}$, $R_{c,2}$ and $R_f$ are 43 s, 100 s and 44 min, respectively. Figure 5(a) shows the responses of $R_{c,1}$ at $x^0$ and at its optimal design $x^{(1)} = [18.66 12.98 0.526 13.717 8.00 1.094]^T$ mm. Figure 5(b) shows the responses of $R_{c,2}$ at $x^{(1)}$ and at its optimized design $x^{(2)} = [18.45 13.0 0.419 13.48 6.88 1.096]^T$ mm. Figure 6 shows the responses of $R_f$ at $x^{(2)}$ (specification error $-3.0$ dB) and the refined design $x^* = [18.64 12.96 0.362 13.40 6.38 1.158]^T$ mm (specification error $-4.1$ dB) obtained in two iterations of the refinement step (4). The optimization cost (Table 2) is low and corresponds to only 7 evaluations of the original, fine-discretization model $R_f$.

Figure 4. Dipole antenna geometry [27]: top and side views. The dash-dot lines show the magnetic (YOZ) and the electric (XOY) symmetry walls. The 50 ohm source impedance is not shown at the figure.

Figure 5: Dipole antenna: (a) responses of the coarse-discretization model $R_{c,1}$ at the initial design $x^0$ (dashed line) and at the optimized design $x^{(1)}$ (solid line), (b) responses of the coarse-discretization model $R_{c,2}$ at $x^{(1)}$ (dashed line) and at its optimized design $x^{(2)}$ (solid line).

Figure 6: Dipole antenna: responses of the original, high-fidelity model $R_f$ at $x^{(2)}$ (dashed line) and at the refined final design $x^*$ (solid line).

<table>
<thead>
<tr>
<th>Algorithm Component</th>
<th>Number of Model Evaluations</th>
<th>Evaluation Time Absolute [min]</th>
<th>Relative to $R_f$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optimization of the coarse-discretization model $R_{c,1}$</td>
<td>127</td>
<td>91</td>
<td>2.1</td>
</tr>
<tr>
<td>Optimization of the coarse-discretization model $R_{c,2}$</td>
<td>44</td>
<td>73</td>
<td>1.6</td>
</tr>
<tr>
<td>Setup of model $q$</td>
<td>13 ($R_{c,2}$)</td>
<td>21</td>
<td>0.5</td>
</tr>
<tr>
<td>Evaluation of the original (fine-discretization) model $R_f$</td>
<td>3</td>
<td>132</td>
<td>3.0</td>
</tr>
<tr>
<td>Total optimization time</td>
<td>N/A</td>
<td>317</td>
<td>7.2</td>
</tr>
</tbody>
</table>
3.3. Coplanar-Waveguide-To-Microstrip Transition

Consider the coplanar-waveguide-to-microstrip transition [32]. Transitions from the microstrip to the coplanar waveguide (CPW) find applications in many integrated microwave circuits where the use of both microstrip-based and CPW-based sub-circuits is beneficial [33]. Design specifications imposed on the reflection level and bandwidth are normally problem dependent. In many modern applications, e.g., UWB communication, high-speed signalling, overall performance of a circuit can be strongly affected and deteriorated by transitions which are improperly designed or fail to meet specifications.

Here, the signal traces of the input TLs are directly connected and TLs’ grounds are connected with through-hole vias. We consider three versions of this transition, with one (version I), two (version II) and three (version III) pairs of vias. Figure 7 shows version II of the transition. The lengths of the input TLs are 10 mm each. \( W_g \) is 6.4 mm.

The design objective is to obtain \( |S_{11}| \leq -25 \text{ dB} \) and \( |S_{22}| \leq -25 \text{ dB} \) from DC to 20 GHz. The design variables are \( x = [R, x, L_1, L_2, L_3]^T \) for version I and \( x = [R, x, s, L_1, L_2, L_3]^T \) for versions II and III. For each version we use two coarse-discretization models. Initial designs and simulation times for the models are: version I: \( x^{(0)} = [0.3 1.2 0.0 0.5 0.5]^T \), \( R_c.1 - 90 \text{ s}, R_c.2 - 200 \text{ s}, R_f - 16 \text{ min} \); version II: \( x^{(0)} = [0.3 1.2 1.0 0.0 0.5 0.5]^T \), \( R_c.1 - 80 \text{ s}, R_c.2 - 140 \text{ s}, R_f - 14 \text{ min} \); version III: \( x^{(0)} = [0.25 1.2 0.6 0.0 0.5 0.5]^T \), \( R_c.1 - 80 \text{ s}, R_c.2 - 190 \text{ s}, R_f - 33 \text{ min} \). It can be seen from Fig. 8 and 9 that the initial designs violate the specification level after 7 GHz.

The final designs obtained using our optimization procedure are the following: \( x^* = [0.541 1.470 0.015 -0.236 0.461]^T \) (version I), \( x^* = [0.20 1.153 1.343 0.180 0.013 0.687]^T \) (version II), \( x^* = [0.134 1.073 0.706 0.019 -0.040 0.622]^T \) (version III). The optimized design for version I satisfies \( |S_{11}|, |S_{22}| < -26 \text{ dB} \) from DC to 20 GHz; for versions II and III we have \( |S_{11}|, |S_{22}| < -30 \text{ dB} \) from DC to 20 GHz. The computational cost of the design corresponds to several full-wave simulations (Table 3). Figures 10 and 11 show the design process for version II of the transition. Figure 9 shows the reflection response of the fine model at the initial and final designs for versions I and III.

![Figure 7: Directly connected signal traces transition: (a) 3D view (substrate is not shown); (b) layout views.](image)

![Figure 8: Directly connected signal traces transition (version II): (a) responses of the coarse-discretization model \( R_{c.1} \) at the initial design (dotted line), optimized model \( R_{c.1} \) (dashed line), and the coarse-discretization model \( R_{c.2} \) at the optimized design of \( R_{c.1} \) (solid line). |\( S_{22} | \) distinguished from |\( S_{11} | \) using circles; (b) fine model responses at the initial design (dotted line), at the optimized design of \( R_{c.1} \) (dashed line), and at the final (refined) design (solid line). |\( S_{22} | \) distinguished from |\( S_{11} | \) using circles.](image)
Figure 12: Directly connected signal traces transition: fine model responses at the initial design (dashed line), and at the final (refined) design (solid line): (a) version I, and (b) version III. $|S_{22}|$ distinguished from $|S_{11}|$ using circles.

Table 3: Design cost of coplanar-waveguide-to-microstrip transitions

<table>
<thead>
<tr>
<th>Configuration</th>
<th>Design Procedure Component</th>
<th>Number of Model Evaluations</th>
<th>Evaluation Time</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>Absolute [min]</td>
</tr>
<tr>
<td>Version I</td>
<td>Optimization of $R_{c1}$</td>
<td>75</td>
<td>113</td>
</tr>
<tr>
<td></td>
<td>Setup of model $q$</td>
<td>11 ($R_{c2}$)</td>
<td>37</td>
</tr>
<tr>
<td></td>
<td>Evaluation of $R_f$</td>
<td>2</td>
<td>32</td>
</tr>
<tr>
<td></td>
<td>Total design time</td>
<td>N/A</td>
<td>182</td>
</tr>
<tr>
<td>Version II</td>
<td>Optimization of $R_{c1}$</td>
<td>63</td>
<td>84</td>
</tr>
<tr>
<td></td>
<td>Optimization of $R_{c2}$</td>
<td>23</td>
<td>53</td>
</tr>
<tr>
<td></td>
<td>Setup of model $q$</td>
<td>13 ($R_{c2}$)</td>
<td>30</td>
</tr>
<tr>
<td></td>
<td>Evaluation of $R_f$</td>
<td>2</td>
<td>28</td>
</tr>
<tr>
<td></td>
<td>Total design time</td>
<td>N/A</td>
<td>195</td>
</tr>
<tr>
<td>Version III</td>
<td>Optimization of $R_{c1}$</td>
<td>78</td>
<td>104</td>
</tr>
<tr>
<td></td>
<td>Optimization of $R_{c2}$</td>
<td>25</td>
<td>79</td>
</tr>
<tr>
<td></td>
<td>Setup of model $q$</td>
<td>13 ($R_{c2}$)</td>
<td>41</td>
</tr>
<tr>
<td></td>
<td>Evaluation of $R_f$</td>
<td>2</td>
<td>66</td>
</tr>
<tr>
<td></td>
<td>Total design time</td>
<td>N/A</td>
<td>290</td>
</tr>
</tbody>
</table>

4. Conclusion
Multi-fidelity algorithm for efficient design optimization of microwave structures is proposed. Our technique exploits coarse-discretization EM models that are sequentially optimized, and a polynomial-approximation-based final design refinement. It is very easy to implement, it does not require any auxiliary (e.g., equivalent-circuit) model or any modifications to the original structure. It is computationally efficient and reliable. As demonstrated through several test cases, the optimized design can be obtained at the low computational cost corresponding to a few full-wave electromagnetic simulations of the microwave structures of interest.

Acknowledgements
The authors thank Sonnet Software, Inc., Syracuse, NY, for em™ and CST AG, Darmstadt, Germany, for making CST Microwave Studio available. This work was supported by the Reykjavik University Development Fund under Grant T09009.

References


