Automated Design of Tunable Impedance Matching Networks for Reconfigurable Wireless Applications

Arthur Nieuwoudt†, Jamil Kawa‡, and Yehia Massoud†

†Dept. of Electrical and Computer Engineering
Rice University, Houston, Texas
{abnieu,massoud}@rice.edu

‡Advanced Technology Group
Synopsys, Inc., Mountain View, California
jamil@synopsys.com

ABSTRACT

In this paper, we develop a generalized automated design methodology for tunable impedance matching networks in reconfigurable wireless systems. The method simultaneously determines the fixed and tunable/switchable circuit element values in an arbitrary-order canonical filter for a general set of performance constraints over a discrete or continuous set of operating frequencies and source/load impedances. To solve the filter design problem, we combine deterministic nonlinear constrained optimization using Sequential Quadratic Programming with a systematic constraint relaxation approach to facilitate convergence. Using the proposed methodology, we successfully generate three different reconfigurable impedance matching networks with performance requirements that would be difficult to realize using manual design techniques.

Categories and Subject Descriptors

B.7.2 [Integrated Circuits]: Design Aids

General Terms

Design, Algorithms

Keywords

Tunable filter, impedance matching network, analog synthesis

1. INTRODUCTION

Wireless applications such as cellular telephones (GSM/CDMA), global positioning systems (GPS), Bluetooth, and wireless local area networks (WLAN) have become pervasive over the last decade. In the near future, ultrawideband (UWB) systems for both consumer electronics and vehicular applications will facilitate both low power and high bandwidth communication [1,2] while advances in RF CMOS technology will enable low cost millimeter wave circuits for high frequency wireless systems [3]. The increasing demand for these wireless applications has driven the need for a single communication system to support multi-band operation. Reconfigurable wireless solutions that reuse components for receiving and transmitting signals in the different frequency bands provide reduced power consumption and hardware complexity [4].

Permission to make digital or hard copies of all or part of this work for personal or classroom use is granted without fee provided that copies are not made or distributed for profit or commercial advantage and that copies bear this notice and the full citation on the first page. To copy otherwise, to republish, to post on servers or to redistribute to lists, requires prior specific permission and/or a fee.

DAC 2008, June 8–13, 2008, Anaheim, California, USA
Copyright 2008 ACM 978-1-60558-115-6/08/0006…$5.00

A major challenge associated with the realization of reconfigurable wireless systems is the design of tunable impedance matching networks (IMN) for circuits such as low noise amplifiers, power amplifiers, mixers, pre-processing filters, and antennas operating at multiple frequencies with variable load and source impedances [5–10]. While significant research efforts have demonstrated the potential performance of tunable IMNs through their physical realization using switches and varactors implemented with standard semiconductor, barium-strontium-titanate (BST), and RF MEMS based technologies [10–17], the design of tunable IMNs has primarily been a manual process that combines well-established design methods for standard non-tunable filters with the specific knowledge of the designer. Given the increasing complexity of reconfigurable wireless systems, systematic automated design techniques must be developed for tunable IMNs to enable greater design flexibility and performance, reduced cost, and shorter time-to-market.

In this paper, we develop a generalized automated design methodology for tunable impedance matching networks in reconfigurable wireless systems. The method simultaneously determines the fixed and tunable/switchable circuit element values in an arbitrary-order canonical filter for a general set of performance constraints over a discrete or continuous set of operating frequencies and source/load impedances. To solve the filter design problem, we combine deterministic nonlinear constrained optimization using Sequential Quadratic Programming (SQP) with a systematic constraint relaxation method to facilitate convergence. To demonstrate the efficacy of the proposed automated design method, we generate three tunable IMNs with a wide range of performance requirements. The results indicate that reconfigurable IMNs designed using the proposed methodology achieve a level of performance that would be difficult to realize using manual design techniques.

2. IMPEDANCE MATCHING NETWORKS

Impedance matching networks have important implications for RF receiver performance metrics such as noise, power consumption, and gain. The input and output impedance must be sufficiently matched to the source and load impedances. The insertion loss, which is the attenuation of a signal passing through the IMN, should be minimized to decrease the noise susceptibility of the RF signal and to reduce the number of amplifier stages required. The IMN must also be designed to satisfy the power handling constraints imposed by its circuit elements, which are especially important for tunable/switchable RF MEMS components [16].

Reconfigurable filters for tunable/switchable IMNs are significantly more challenging to design than their fixed valued counterparts since they must provide passband impedance matching and stopband rejection for different sets of frequencies and for potentially changing source/load impedances while only modifying a
small number of variable component values. Switchable IMNs utilize passive components with several discrete values to provide impedance matching at several sets of frequencies while tunable IMNs utilize passive components with continuously adjustable values to provide impedance matching over a continuous frequency range. In this paper, we refer to both tunable and switchable IMNs as tunable IMNs since the proposed design method treats both types of reconfigurable IMNs in a similar manner. For both lumped and distributed tunable IMNs, switches and varactors implemented using standard semiconductor, BST, or RF MEMS technologies can be utilized for the variable circuit elements [10–17].

In general, the microwave filters implementing tunable IMNs in wireless applications are typically realized using either lumped circuit elements or microstrips. Both lumped and microstrip filters are typically designed using canonical lumped filter representations such as the Butterworth/Chaplygin bandpass filter topology displayed in Figure 1 [18, 19]. Once the fixed and tunable circuit elements in the canonical filter are determined, the filter can be physically synthesized by employing lumped passive components for each circuit element [20] or by utilizing well-known techniques to transform the lumped circuit elements into their equivalent transmission line representations [18, 19]. Therefore, to develop a generalized technology-independent automated design method, we focus on the design of the fixed and tunable component values in the canonical filter topology displayed in Figure 1 in this paper.

3. MODELING OF TUNABLE IMPEDANCE MATCHING NETWORKS

We model the IMN topology of order \( n \) displayed in Figure 1 using its two port ABCD parameters. For the filter’s series circuit elements, the ABCD parameter formulation is

\[
\begin{bmatrix}
V_i \\
I_i
\end{bmatrix} = A_i \begin{bmatrix}
V_{i+1} \\
I_{i+1}
\end{bmatrix} = \begin{bmatrix}
1 & Z_i \\
0 & 1
\end{bmatrix} \begin{bmatrix}
V_{i+1} \\
I_{i+1}
\end{bmatrix}
\]

(1)

where \( Z_i \) is the impedance of the \( i \)-th set of series circuit elements, and \( V_i, I_i, V_{i+1}, I_{i+1} \) are the 2-port voltages and currents. Similarly, for the filter’s shunt circuit elements, the formulation is

\[
\begin{bmatrix}
V_i \\
I_i
\end{bmatrix} = A_t \begin{bmatrix}
V_{i+1} \\
I_{i+1}
\end{bmatrix} = \begin{bmatrix}
1 & 0 \\
Y_i & 1
\end{bmatrix} \begin{bmatrix}
V_{i+1} \\
I_{i+1}
\end{bmatrix}
\]

(2)

where \( Y_i \) is the admittance of the \( i \)-th set of shunt circuit elements. The entire filter’s ABCD parameters \( (A_{tot}) \) can be determined by multiplying the ABCD parameters for each stage of the filter \( A_i \)

\[
\begin{bmatrix}
V_i \\
I_i
\end{bmatrix} = A_{tot} \begin{bmatrix}
V_{i+1} \\
I_{i+1}
\end{bmatrix} = \prod_{i=1}^{n} A_i \begin{bmatrix}
V_{i+1} \\
I_{i+1}
\end{bmatrix}
\]

(3)

where \( V_i, I_i, V_{i+1}, I_{i+1} \) are the 2-port voltages and currents for the entire filter. The S-parameters \( (S_{11}, S_{21}, S_{22}) \) of the filter are determined based on the ABCD parameters \( (A_{tot}) \) calculated in (3) using standard 2-port parameter transformations such as the methods presented in [19] and [21], which can capture the effect of arbitrary complex load and source impedances.

The parasitic resistances associated with the \( i \)-th set of capacitors and inductors in the filter model are calculated using \( R_{C_i} = Q_{C_i}/(\omega C_i) \) and \( R_{L_i} = (\omega L_i)/Q_{L_i} \), where \( Q_{C_i} \) and \( Q_{L_i} \) are the quality factors of the \( i \)-th capacitor and inductor, respectively. Note that the quality factor encapsulates all of the resistive, capacitive, and inductive parasitics associated with the lumped passive components or microstrips at a given operating frequency. The worst-case apparent power at any node in the IMN can also be calculated using an ABCD parameter formulation similar to (3) based on the root mean squared input voltage and current.

4. AUTOMATED DESIGN OF TUNABLE IMPEDANCE MATCHING NETWORKS

4.1 Sensitivity Analysis

Understanding the sensitivity of IMN performance to changes in the filter circuit elements is crucial for tunable element selection. To probe the sensitivity of the input impedance of the filter \( (Z_{in}) \) to changes in the circuit element values, consider a simplified case where each branch of a 3-pole filter is replaced with a resistor \( (R_1, R_2, R_3, \text{ and } R_L) \). In this case, \( Z_{in} \) is

\[
Z_{in} = R_1 + \frac{R_2(R_3 + R_L)}{R_2 + R_3 + R_L}.
\]

(4)

The sensitivity of \( Z_{in} \) to \( R_1, R_2 \) and \( R_3 \) is

\[
\frac{\partial Z_{in}}{\partial R_1} = 1, \quad \frac{\partial Z_{in}}{\partial R_2} = \frac{R_2 + R_3}{R_2 + R_3 + R_L} - \frac{R_2(R_3 + R_L)}{(R_2 + R_3 + R_L)^2}, \quad \frac{\partial Z_{in}}{\partial R_3} = \frac{R_2}{R_2 + R_3 + R_L} - \frac{R_2(R_3 + R_L)}{(R_2 + R_3 + R_L)^2}.
\]

(5) (6) (7)

\( \frac{\partial Z_{in}}{\partial R_1} \) is always less than \( 1, R_2 \) and \( R_3 \) are always less than \( 1, \) and \( \frac{\partial Z_{in}}{\partial R_1} \) is greater than \( \frac{\partial Z_{in}}{\partial R_2} \) since the magnitudes of first and second terms of \( \frac{\partial Z_{in}}{\partial R_2} \) and \( \frac{\partial Z_{in}}{\partial R_3} \) are always less than \( 1, R_3 + R_L \) must be less than \( R_2 \) for \( \frac{\partial Z_{in}}{\partial R_3} \) to be greater than \( \frac{\partial Z_{in}}{\partial R_2} \). Consequently, \( \frac{\partial Z_{in}}{\partial R_2} \) will be greater than \( \frac{\partial Z_{in}}{\partial R_3} \) in the majority of the design space. In general, similar results will also apply to the general filter implemented with capacitors and inductors depending on the individual circuit element values. This confirms the intuitive notion that in general the elements closer to a given filter port will typically have the largest impact on the input impedance of that port. Therefore, circuit elements that are the same distance from the input and output ports will have the most balanced impact on \( S_{11}, S_{22}, \) and \( S_{21} \). The convexity of the objective and constraint functions associated with the tunable IMN plays an important role for determining if computationally efficient gradient-based methods can be employed for design automation. The \( S_{11} \) values associated with a
filter with an input impedance of \( Z_{in} = R_{in} + jX_{in} \) and a source impedance of \( Z_s = R_s + jX_s \) in dB are proportional to

\[
S_{11} \propto \log \left( \frac{Z_{in} - Z_s}{Z_{in} + Z_s} \right) \propto \log \left( \frac{(R_{in} - R_s)^2 + (X_{in} - X_s)^2}{(R_{in} + R_s)^2 + (X_{in} + X_s)^2} \right).
\]

In the regions where \( R_s \geq 0 \) and either \( X_s \geq 0 \) or \( X_s \leq 0 \), \( S_{11} \) only has one local minimum with respect to \( R_{in} \) and \( X_{in} \). This can also be shown for \( S_{22} \) and \( S_{21} \). Therefore, if \( R_{in} \) and \( X_{in} \) behave monotonically with changes to the circuit element values, then minimizing \( S_{11} \) or \( S_{22} \) or maximizing \( S_{21} \) will only yield one solution. For a typical fixed-valued narrow-band filter, the circuit elements associated with each series or shunt branch of the filter resonate near the narrow-band frequency, which causes changes in the circuit elements to behave in this manner.

To demonstrate this property, consider a typical three-pole narrow-band filter with \( Z_s = 25 \Omega \) and \( Z_t = 50 \Omega \) operating at 2.4 GHz. Figure 2 displays \( S_{11} \) versus a multiplication factor simultaneously applied to all of the circuit elements in the filter, which is normalized to the circuit element values of a three-pole tunable filter (described later in the paragraph). \( S_{11} \) only has one local minima. Now consider a tunable filter with two narrow-band frequencies, 2.4 GHz and 5.0 GHz, designed using the proposed design automation method. In both frequency bands, \( S_{11} \) has two convex regions with their own local minima as displayed in Figure 2. In this case, gradient-based optimization methods may fail to converge to the global minimum. However, if the fixed filter design is used as the start point for the optimization process, the tunable filter optimized using gradient-based techniques will converge to the optimal solution. We exploit this property in order to utilize gradient-based optimization techniques for tunable IMN design.

### 4.2 Frequency Mapping to Tunable Elements

Each set of operating frequencies and source/load impedances must be carefully mapped to a given set of tunable circuit elements. To facilitate this mapping, we propose a symmetric combinatorial mapping scheme where each of the \( M \) frequency ranges for the reconfigurable IMN is mapped to each possible combination of pairwise symmetric tunable circuit elements to balance their impact on \( S_{11}, S_{22}, \) and \( S_{21} \). Figure 3 displays a tree structure representing the frequency state to tunable element value mapping when two symmetric pairs of tunable elements are considered. The upper (\( \overline{T} \)) and lower (\( \underline{T} \)) bounds for each frequency range are denoted by \( \overline{T} = [T_1, T_2, \ldots, T_J, \ldots, T_M] \) and \( \underline{T} = [B_1, B_2, \ldots, B_j, \ldots, B_M] \) where \( T_j \) and \( B_j \) are the upper and lower bounds for the \( j \)th frequency range. \( E_{11} \) and \( E_{22} \) represent one of the \( J \) possible states (\( J = M \)), respectively, associated with two symmetric pairs of tunable circuit elements. For example, in a fifth order filter where the middle three capacitors are tunable circuit elements, \( E_{11} \) corresponds to the possible values of \( C_5 \), and \( E_{22} \) corresponds to pairs of possible values for the symmetric circuit elements \( C_2 \) and \( C_4 \). Each path from the root node to a leaf node in Figure 3 represents a different possible combination of tunable element values with the associated frequency range denoted below the leaf node. Since the variable states of \( E_{11} \), span a larger frequency range than the possible states of \( E_{22} \), the circuit element corresponding to \( E_{11} \) should have greater sensitivity to \( S_{11} \), \( S_{22} \), and \( S_{21} \). Based on the sensitivity analysis in Section 4.1, the circuit elements near the center of the filter (branch \( \approx n/2 \)) most effectively change both the input and output impedance. Note that we apply a similar mapping scheme for tunable IMNs with variable load/source impedances.

### 4.3 Design Optimization Problem Formulation

For the canonical filter displayed in Figure 1, the design variables are the individual fixed and tunable/switchable capacitors and inductors in each series and shunt branch of the \( n \)th order filter,

\[
\mathbf{\bar{x}} = \left[ C_1, C_2, \ldots, C_n, L_1, L_2, \ldots, L_n \right].
\]  

(8)

If \( C_i \) or \( L_i \) is a switchable or tunable circuit element, the design variable vector contains several component values that span the element value ranges associated with the switchable or tunable states.

We formulate the general tunable IMN design optimization problem for \( M \) frequency ranges as

\[
\begin{align*}
\text{Minimize} & \quad \| S_{211T} \|_2 \\
\text{Subject to} & \quad S_{111T} \leq S_{111C}, \quad S_{110T} \geq S_{110C} \\
& \quad S_{211T} \leq S_{211C}, \quad S_{220T} \geq S_{220C} \\
& \quad S_{211T} \geq S_{211C}, \quad S_{PT} \leq S_{PC} \\
& \quad \bar{x}_{min} \leq \mathbf{\bar{x}} \leq \bar{x}_{max}
\end{align*}
\]

(9)

where the insertion loss objective function vector is

\[
S_{211T} = \left[ S_{211}(T_1, f_1, Z_{S1}, Z_{L1}), \ldots, S_{211}(T_M, f_M, Z_{SM}, Z_{LM}) \right].
\]

\[
S_{211}(T_i, f_i, Z_{Si}, Z_{Li}) \text{ measures the insertion loss in dB for a set of frequencies } (f_i) \text{ spanning the range } [B_i, T_i] \text{ where } T_i \text{ is a vector of the fixed and tunable circuit elements mapped to the } f_i \text{ based on the frequency mapping scheme described in Section 4.2.}
\]

\( Z_{Si} \) and \( Z_{Li} \) are the source and load impedances when the tunable filter is in a state corresponding to the frequency range \([B_i, T_i]\).

The constraints on \( S_{111T} \) in each frequency band are

\[
S_{111T} \leq S_{111C} \implies \begin{cases}
S_{11} \left( T_1, f_1, Z_{S1}, Z_{L1} \right) \leq S_{11}^{max} \\
S_{11} \left( T_2, f_2, Z_{S2}, Z_{L2} \right) \leq S_{11}^{max} \\
\vdots \\
S_{11} \left( T_M, f_M, Z_{SM}, Z_{LM} \right) \leq S_{11}^{max} 
\end{cases}
\]

(10)
where $\tilde{S}_{11T}^T(t_i, f_i, Z_{St}, Z_{Li})$ measures the return loss in dB for a set of frequencies spanning the range $[B_i, T_i]$, and $S_{11\text{max}}$ is a vector of the maximum allowed values of $S_{11}$ for the frequencies in the $i$th frequency band. The in-band constraints on $S_{22}$ ($S_{22\text{IT}} \leq S_{22\text{IC}}$) and on $S_{21}$ ($S_{21\text{IT}} \geq S_{21\text{IC}}$) are constructed in the same manner as (10). To ensure that $S_{11}$ and $S_{22}$ are large enough outside of the frequency band for each tunable state to provide sufficient stopband rejection, we apply the constraints $S_{11\text{OT}} \geq S_{11\text{OC}}$ and $S_{11\text{OT}} \geq S_{11\text{OC}}$, which are defined in a similar manner to (10) except that the constraints are evaluated at a set of frequencies ($\tilde{f}_{oi}$) spanning the spectrum outside of the passband,

$$\tilde{f}_{oi} = [f_{min}, \ldots, B_i(1 - R_{Bi}), T_i(1 + R_{Ti}), \ldots, f_{max}],$$

where $R_{Bi}$ and $R_{Ti}$ determine the size of the buffer region associated with the stopband rejection constraint for the $i$th tunable frequency band, and $f_{min}$ and $f_{max}$ are the minimum and maximum frequencies that are significant for stopband rejection. The constraint $S_{FT} \leq S_{PC}$ limits the power at each node of the filter that to ensure that the power handling capabilities of the circuit components are not exceeded in any of the possible tunable states. The circuit element values are constrained by $x_{min}$ and $x_{max}$ to ensure that their values are suitable for on-chip or in-package integration.

### 4.4 Automated Design Methodology

To solve the optimization problem described in the previous section, we utilize Sequential Quadratic Programming (SQP), a state-of-the-art gradient-based nonlinear programming technique [22]. In practice, SQP typically does not converge to a feasible solution when attempting to solve (9) directly with an arbitrary start point due to the non-convex design space characteristics described in Section 4.1. However, based on the analysis performed in Section 4.1 and in Figure 2, the solution to a less complex related IMN optimization problem can provide a suitable start point for the complete IMN optimization problem. To formulate this less complex design problem, we relax the constraints associated with the original optimization problem in (9). Constraint relaxation for tunable IMN optimization can be achieved using three possible methods: (1) remove tunable frequency bands/elements from consideration, (2) relax the design requirements associated with the filter, (3) relax the quality factors and the circuit element value constraints.

Leveraging the aforementioned optimization and constraint relaxation techniques, we design the tunable IMN by solving a series of successively more constrained and complex optimization problems. Figure 4 displays the tunable IMN automated design methodology. To determine a start point for the optimization process, we first design a fixed-valued filter for one of the tunable frequency and source/load impedance combinations using traditional filter design techniques [19]. We then employ an iterative process where we repeatedly solve the following relaxed optimization problem:

Minimize $\| S_{22\text{IT}}^\text{red} \|_2$

Subject to $S_{22\text{IT}}^\text{red} \leq S_{22\text{IC}}$, $S_{21\text{IT}}^\text{red} \leq S_{21\text{IC}}$, $x_{\text{min}}^\text{red} \leq x^\text{red} \leq x_{\text{max}}^\text{red}$\hspace{1cm}(12)

where $S_{21\text{IT}}^\text{red}$, $S_{22\text{IT}}^\text{red}$, $S_{21\text{IC}}$, $S_{22\text{IC}}$, and $S_{22\text{IC}}$ are the performance metrics defined in Section 4.3 but with vectors that only contain a subset of the possible tunable frequency and source/load impedance combinations. We iteratively add additional tunable frequency and source/load impedance combinations based on the constraint application order depicted in Figure 3 and solve the optimization problem defined in (12) using the previous relaxed solution as the start point. Note that at this point in the process, we assume that the frequency bands have a narrow bandwidth at a frequency of $M_j = \sqrt{T_j/B_j}$ for the $j$th frequency band and that the components in the filter have large quality factors ($Q \approx 1000$).

Once all of the tunable frequency and source/load impedance combinations have been included in the design problem, we then add relaxed versions of the stopband rejection constraints $S_{11\text{OT}} \geq S_{11\text{OC}}$ and $S_{11\text{OT}} \geq S_{11\text{OC}}$ to (12) and solve the optimization problem for a feasible solution. We relax the stopband rejection constraints by increasing both $R_{Bi}$ and $R_{Ti}$. We then employ an iterative process where we reduce $R_{Bi}$ and $R_{Ti}$ and perform the design optimization using the results from the previous optimization problem as a start point. This process continues until $R_{Bi}$ and $R_{Ti}$ reach the values required by the design specification for stopband rejection. We then utilize a similar process to iteratively add the passband width constraints by changing the lower and upper bounds of each tunable frequency range from $M_j$ to $B_j$ and from $M_j$ to $T_j$, respectively, and then solving the tunable IMN optimization problem. Once the passband width constraints have been enforced, we perform a similar iterative optimization procedure to reduce the quality factors of the elements in the filter from their relaxed values of 1000 to their values assumed in the design specification. We then add the power handling and insertion loss constraints ($S_{FT} \leq S_{PC}$ and $S_{22\text{IT}} \geq S_{22\text{IC}}$) and solve the complete design optimization problem listed in (9), which yields the final fixed and tunable component values for the tunable IMN.

In terms of computational complexity, the proposed tunable IMN automated design methodology requires the solution to

$$N_{\text{opt}} = \sum_{i=1}^{N_1} (N_i) + N_{\text{del}} + N_{\text{pbo}} + N_\text{f} + 2 \hspace{1cm}(13)$$

Figure 4: Automated design methodology for tunable IMNs.
Table 1: Reconfigurable Impedance Matching Network Design Examples

| Design Example | Frequency Bands (GHz) | $Z_s$ (Ω) | $Z_L$ (Ω) | $S_{11_{\text{min}}}^{\text{opt}}$ & $S_{22_{\text{min}}}^{\text{opt}}$ | $S_{11_{\text{max}}}^{\text{opt}}$ & $S_{22_{\text{max}}}^{\text{opt}}$ | $R_{\text{RL}}$ & $R_{\text{RL}}$ | $L_{\text{min}}$ (nH) | $L_{\text{max}}$ (nH) | $C_{\text{min}}$ (pF) | $C_{\text{max}}$ (pF) | Quality Factor | $S_{11_{PC}}$ (mW) |
|----------------|----------------------|-----------|-----------|-------------------------------|-------------------------------|-----------------|-----------------|----------------|----------------|----------------|----------------|----------------|----------------|
| 1 (5th order)  | WCDMA (2.11-2.17 GHz) 802.11bg (2.405-2.484 GHz) 802.11a (5.15-5.35 GHz) 802.11a (5.725-5.825 GHz) | 20 to 50 | 50 | -10 dB | -3 dB | 0.085 | 0.05 | 5.0 | 5.0 | 0.05 | 5.0 | 100 | 400 |
| 2 (5th order)  | Continuous Narrow-Band: 20-55 GHz | 20 | 50 | -10 dB | -5 dB | 0.1 | 0.01 | 0.5 | 0.01 | 0.5 | 0.01 | 65 | 300 |
| 3 (9th order)  | Full UWB MB-OFDM: 14 Bands (528 MHz Width) Centered from 3.4 to 10.3 GHz | 78 | 50 | -10 dB | -5 dB | 0.063 | 0.05 | 2.0 | 0.05 | 2.0 | 0.05 | 20 (L) and 80 (C) | 300 |

Figure 5: S-parameters for design example 1 with 4 frequency bands and a source impedance of (a) 20 Ω, (b) 35 Ω, and (c) 50 Ω.

Figure 6: S-parameters for design example 2 with a tunable frequency ranging from 20 to 55 GHz.

numerical optimization problems. In (13), $N_{\text{te}}$ refers to the number of tunable element levels depicted in Figure 3, $N_i$ is the number of tunable element states in level $i$, and $N_{\text{nbr}}, N_{\text{phw}},$ and $N_{\text{qf}}$ are the number of iterations used to tighten the stopband rejection constraints, the passband width constraints, and the quality factors, respectively, to their required levels. We find that setting $N_{\text{nbr}}, N_{\text{phw}},$ and $N_{\text{qf}}$ to 4 provides a good balance between convergence and computational performance. For realistic IMN design problems, the $N_{\text{opt}}$ required is typically small. In design examples 1, 2, and 3 described in Section 5, $N_{\text{opt}}$ is 23, 21, and 28, respectively. Each optimization problem typically requires less than one thousand model evaluations to converge to a feasible solution. For the three design examples described in Section 5, the average CPU time was 12.9 minutes using our Matlab implementation of the proposed method on a Windows machine with a dual-core 2.4 GHz AMD Opteron processor and 2 GB of RAM. Consequently, the proposed automated design methodology provides a computationally feasible means for generating tunable IMNs that would otherwise be intractable to realize using manual design techniques.

5. RESULTS

To demonstrate the effectiveness of the proposed automated design methodology, we designed three IMNs with the design specifications and constraints listed in Table 1. In the first design example, we generate a 5-pole reconfigurable IMN with 4 operating frequency bands corresponding to WCDMA (2.11-2.17 GHz), 802.11bg (2.405-2.484 GHz), and 802.11a (5.15-5.35 GHz and 5.725-5.825 GHz) [5]. The IMN is designed to match $Z_s$ values that range from 20 Ω to 50 Ω, which corresponds to a typical range of variable sources impedances that may be possible due to changes in antenna position [10,16]. In the design optimization process, we utilize 5 discrete values of $Z_s$ to capture the entire range of anticipated $Z_s$ values. Therefore, $C_3$ has 4 possible values corresponding to the 4 frequency ranges, and $C_2/C_4$ each have 5 possible values corresponding to the 5 discretized $Z_s$ values.

Figure 5 displays $S_{11}, S_{22}$, and $S_{21}$ for the reconfigurable IMN generated in design example 1 when the filter is tuned to match source impedances of 20 Ω, 35 Ω, and 50 Ω. In each of the four operating frequency bands, the IMN exceeds the -10 dB constraint on $S_{11}$ and $S_{22}$ with an average in-band insertion loss of -3 dB for source impedances ranging from 20 Ω to 50 Ω. The tunable IMN also provides enough selectivity to effectively resolve each of the two closely spaced bands near 2.4 and 5.5 GHz. The canon-
In this paper, we created a generalized automated design methodology for tunable impedance matching networks in reconfigurable wireless systems. The method simultaneously determines the fixed and tunable/switchable circuit element values in an arbitrary-order canonical filter for a general set of performance constraints over a discrete or continuous set of frequencies and source/load impedances. The results indicate that reconfigurable IMNs designed using the proposed methodology achieve a level of performance that would be difficult to realize using manual design techniques. The proposed method provides a generalized technology-independent framework for the automated design of tunable IMNs, which will be an invaluable tool for RF circuit designers developing reconfigurable RF systems for multi-standard wireless applications.

7. REFERENCES


Figure 7: (a) Return loss (S11 and S22) and (b) insertion loss (S21) in design example 3 for the 14 switchable states covering the 3.1-10.6 GHz UWB frequency range.

In design example 2, we generate a 5-pole narrow-band tunable IMN with an adjustable operating frequency ranging from 20 to 55 GHz. In this example, the IMN has a single tunable element C1. As depicted in Figure 6, the generated IMN is able to provide continuous narrow-band impedance matching across the entire 35 GHz wide frequency range. The canonical filter’s inductor values are [L1, · · · , L5] = [0.37, 0.49, 0.05, 0.44, 2.99] nH, and the filter’s fixed capacitor values are [C1, C5] = [1.99, 1.50] pF. The continuously tunable elements C2 and C4 have capacitance values ranging from [0.05, 0.42] pF and [1.48, 1.35] pF, respectively, to achieve impedance matching when Zs varies from 20 Ω to 50 Ω. C3 is a switchable capacitor with four discrete values of [5.00, 3.88, 0.65, 0.41] pF that correspond to the four frequency bands of the filter with increasing frequency. To the best of our knowledge, this is the first proposed 4-band reconfigurable IMN for combined CDMA and wireless LAN applications that successfully incorporates variable source impedance matching.

In design example 2, we generate a 5-pole narrow-band tunable IMN with an adjustable operating frequency ranging from 20 to 55 GHz. In this example, the IMN has a single tunable element C1. As depicted in Figure 6, the generated IMN is able to provide continuous narrow-band impedance matching across the entire 35 GHz wide frequency range. The canonical filter’s inductor values are [L1, · · · , L5] = [0.318, 0.297, 0.149, 0.033] nH, and the filter’s fixed capacitor values are [C1, C2, C4, C5] = [0.500, 0.034, 0.010, 0.500] pF. C3 is a tunable capacitor with values ranging from [0.152, 0.013] pF. Compared to the two-band switchable 20 to 55 GHz filter with 6 switchable circuit elements presented in [15], the generated tunable IMN offers greater frequency selection capabilities and requires only 1 tunable circuit element.

In design example 3, we design a 9-pole switchable IMN covering the 14.528 MHz wide sub-bands proposed for multi-band (MB) OFDM UWB applications [1]. The filter has only one variable circuit element (C5). As displayed in Figure 7, the -10 dB S11 and S22 requirements are met over the entire 528 MHz bandwidth for each of the 14 states. The canonical filter’s inductor values are [L1, · · · , L9] = [1.82, 1.23, 0.16, 1.90, 0.28, 2.00, 0.59, 1.09, 0.88] nH, and the fixed capacitor values are [C1, · · · , C9, C6, · · · , C9] = [1.02, 0.05, 2.00, 0.05, 0.05, 2.00, 0.05, 1.68] pF. C5 is a variable capacitor that ranges from [2.00, 0.13] pF. Based on these results, the proposed automated design methodology provides an efficient means to generate reconfigurable IMNs with frequency responses that would be intractable to realize using manual design techniques.