Inductively-loaded RF MEMS Reconfigurable Filters

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ABSTRACT: This article presents an inductively loaded radio frequency (RF) microelectromechanical systems (MEMS) reconfigurable filter with spurious suppression implemented using packaged metal-contact switches. Both simulation and measurement results show a two-state, two-pole 5% filter with a tuning range of 17% from 1.06 GHz to 1.23 GHz, an insertion loss of 1.56–2.28 dB and return loss better than 13 dB over the tuning range. The spurious passband response in both states is suppressed below –20 dB. The unloaded Q of the filter changes from 127 to 75 as the filter is tuned from 1.06 GHz to 1.23 GHz. The design and full-wave simulation of a two-bit RF MEMS tunable filter with inductively loaded resonators and monolithic metal-contact MEMS switches is also presented to prove the capability of applying the inductive-loading technique to multibit reconfigurable filters. The simulation results for a two-bit reconfigurable filter show 2.5 times improvement in the tuning range compared with the two-state reconfigurable filter due to lower parasitics associated with monolithic metal-contact MEMS switches in the filter structure.

Keywords: microelectromechanical systems; RF MEMS; reconfigurable filter; inductive loading; microstrip

I. INTRODUCTION

Highly linear radio frequency (RF) microelectromechanical systems (MEMS) switches have been demonstrated in miniaturized tunable microwave filters for reconfigurable front-ends and wideband tracking receivers [1–4]. Combiner filters have been a popular choice for tunable filters due to their wide tunability and broad stopband [5], but show a large variation in fractional bandwidth as the filter is tuned [6]. Methods have been developed to simultaneously control the center frequency and bandwidth in a combline filter as it is tuned by placing variable couplings reducers between the filter resonators [7]. The resonances in the coupling reducers produce unwanted resonant peaks close to the desired passband. Fixed preselection filters can be used to suppress these unwanted resonances but the use of additional components increases the overall cost and complexity of the system.

An alternative method to maintain fixed relative bandwidth over the tuning range is to develop tunable filters with capacitively loaded resonators and shunt inductive inverters [8]. In this way, the values of the inductive inverters scale with the filter center frequency and help maintain a fixed relative bandwidth as the filter is tuned. This filter topology is well suited to coplanar waveguide implementation but becomes complex for microstrip designs in which shunt inductive inverters are difficult to implement. As constant...
relative bandwidth is maintained when loading and coupling mechanisms are different from each other, the dual of the topology with inductively loaded resonators and series capacitive inverters can also be used for microstrip filters. Inductive loading can be easily achieved by using a step in the line-width of a microstrip resonator. Also, a series capacitive inverter can be easily implemented with a microstrip gap.

However, filters with capacitive loading and inductive coupling or inductive loading and capacitive coupling have a spurious passband due to resonance between the loading and coupling elements which results in poor high frequency rejection [9]. It is possible to improve the upper rejection skirt by pushing the spurious response to higher frequencies using distributed loading [8] or by introducing a transmission zero at the spurious response using tapped-line input [10].

In this article, a two-state, two-pole microstrip RF MEMS reconfigurable filter with inductive-loading controlled by surface-mount metal-contact packaged RF MEMS switches is fabricated and measured. Quarter wavelength stubs at the filter input and output are used to improve the upper rejection skirt and suppress the spurious passband. The design and simulation of a two-bit RF MEMS tunable filter is also presented based on the inductive loading technique.

II. CIRCUIT ANALYSIS AND DESIGN FOR A TWO-STATE FILTER

A. Resonator Design

The proposed half-wavelength open-ended microstrip resonator and its equivalent circuit model are shown in Figure 1. The microstrip line step discontinuity provides an equivalent lumped inductance [11] which is used to load the resonator. Maximum tuning is achieved when the high-impedance transmission line is placed in the middle of the resonator (at the point of maximum current density). Simultaneously actuated series RF MEMS switches are used to control the resonator loading. In the up-state position, the effective length of the resonator is larger and hence tunes to a lower frequency. Assuming the switches provide an open circuit, the resonance condition for the loaded resonator is derived by replacing the symmetry plane with an electric wall (short circuit) and solving \( Y_{\text{in, eq}} = 0 \). The resonance condition is given by,

\[
\tan \theta_1 - Y_2 \cot \theta_2 = 0
\]

where \( \theta_1 \) and \( \theta_2 \) are the electrical lengths of \( l_1 \) and \( l_2/2 \), respectively, at the filter center frequency.

B. Filter Design

A two-pole 5% Chebyshev filter with 0.036-dB ripple is designed at \( f_{\text{min}} = 1.06 \) GHz based on shunt resonators and J-inverters. The required values of external \( Q \) \( (Q_{\text{ext}}) \) and inter-resonator coupling coefficient \( (k_{12}) \) can be found from [12],

\[
Q_{\text{ext}} = \frac{g_0 \cdot g_1}{\text{FBW}} = \frac{g_2 \cdot g_3}{\text{FBW}}
\]

\[
k_{12} = \frac{\text{FBW}}{\sqrt{g_2 \cdot g_3}}
\]

where FBW is the filter fractional bandwidth and \( g_0, \ldots, g_3 \) are the low-pass prototype values. Equation (2) results in \( Q_{\text{ext}} = 12.64 \) and \( k_{12} = 0.086 \) at \( f_{\text{min}} \).

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Figure 1. (a) Proposed resonator structure and (b) equivalent circuit of the resonator with ideal switches in the up-state position (\( Y_2 < Y_1 \)).
Figure 2(a) shows the proposed filter layout with one arm of the resonator bent at a right angle to achieve filter miniaturization. Each ideal MEMS switch is represented by “S” and all switches are actuated simultaneously. \( C_{io} \) represents input/output coupling capacitance and \( k_{12} \) represents the inter-resonator mixed coupling coefficient. Tapped resonator coupling at the input and output of the filter introduces a transmission zero which is used to suppress the spurious passband response. The tap-point on the resonator is chosen so that the distance from the open-end of the resonator to the tap-point acts as a quarter-wavelength open-ended stub at the spurious passband frequency.

The equivalent half-circuit model of the filter in the up-state condition is shown in Figure 2(b). The open-ended stub with characteristic admittance \( Y_1 \) is designed to suppress the second harmonic response \( (\phi_1 = 90^\circ \text{ at } f_0) \). Since the stub input admittance is not zero at the resonator center frequency \( (T_{in,eq} \neq 0 \text{ at } f_0) \), the new condition for resonance can be found by placing a magnetic wall (open circuit) at reference plane \( B \), neglecting coupling capacitances and finding \( \theta_{1,M} \) and \( \theta_{2,M} \) for which \( Y_p = 0 \), at the resonant frequency. \( \theta_{1,M} \) and \( \theta_{2,M} \) are extracted from ADS [13] simulations such that the new resonance condition is met. In the down-state condition, the electrical length of the overall unloaded resonator must be half wavelength long.

### Table I. Tunable Filter Model Element Values

| \( Y_1 = Y_1' = \frac{1}{30}S \), \( Y_2 = \frac{1}{100}S \), \( C_{io} = 6 \text{ pF} \), \( k_{12} = 0.086 \) | \( \theta_{1,M} \) | \( \theta_{2,M} \) | \( \theta_{d} \) |
|---|---|---|
| 20.4° | 31.7° | 10.8° |

Figure 2. (a) Proposed filter layout and (b) equivalent half-circuit model of the filter in the up-state condition.

Figure 3. Radant MEMS switch enhanced model [16].

The circuit element values for the model with ideal switches in Figure 2(b) are provided in Table I. The admittance values are chosen as \( Y_1 = Y_1' = \frac{1}{30}S \) and \( Y_2 = \frac{1}{100}S \), and electrical lengths related to the resonator are at \( f_{\text{min}} \). The filter layout is simulated using Sonnet [14] and the electrical lengths of Table I are mapped into their corresponding physical lengths after full-wave simulation. The input/output capacitance value and inter-resonator gap length are also found by full-wave simulation using the methodology described in [12].

The ideal RF MEMS switch is replaced by a packaged metal-contact switch from Radant MEMS Inc. [15]. The enhanced switch model for the metal-contact switch reported in [16] is valid but the bondwire inductance and landing pad capacitance are recalculated for RT/Duroid substrate between 0.5 and 3 GHz. The changes in microstrip line width around the MEMS switch and the transition from microstrip line to bondwire results in an additional external inductance (4 nH) which is in series with the switch (Fig. 3). Full-wave simulations for the reconfigurable filter using the enhanced switch model are discussed in Section III.

### III. Fabrication and Measurement

The microstrip filter is fabricated on RT/Duroid 6010LM substrate with \( \varepsilon_r = 10.2 \pm 0.25 \), \( \tan \delta = 0.0023 \), \( t = 2.54 \text{ mm} \) and with 35-\( \mu \text{m} \) thick copper. The filter is excited using 50 \( \Omega \) input/output microstrip lines. The input/output J-inverter capacitances \( (C_{io} = 6 \text{ pF}) \) are surface mount capacitors and the inter-resonator coupling coefficient \( (k_{12}) \) is implemented with a gap of 0.8 mm. The switch actuation voltage (90 V) is applied to the switch using copper bias lines. Each resonator has a 56 k\( \Omega \) resistor connected to ground to prevent any direct current (DC) floating nodes which could result in electrostatic discharge across the switch. Each switch has a separate DC coupling R-C filter \( (R = 10 \text{ k}\Omega \), \( C = 10 \text{ pF} \), \( f_c = 1.6 \text{ MHz} \) for added isolation between DC and RF signals [16].
Figure 4 shows the fabricated filter and Figure 5 shows the simulated and measured results. The enhanced switch model is included in full-wave simulations for up-state and down-state conditions. The filter shows center frequencies of 1.06 GHz to 1.23 GHz with fractional bandwidths of 5.1% and 5.3% when switches are in the up- and down-state positions, respectively. The insertion loss is 1.56–2.28 dB with return loss better than 13 dB over the tuning range. The measured unloaded $Q$ of the filter changes from 127 to 75 as the filter is tuned from 1.06 GHz to 1.23 GHz. Unloaded $Q$ is lower at 1.23 GHz because additional losses are introduced in the filter due to the down-state resistance (0.7 $\Omega$) of the MEMS switch.

The 4 nH external inductance in series with the switch limits the tuning range. Reducing the external inductance to 1 nH by tapering switch drain/source landing pads and using parallel bond wires will increase the passband tuning range from 17% to 25.5%. The presence of a spike at 1.74 GHz for both filter states is due to an unwanted coupling path from the input to the output of the filter through the bias circuitry. The rejection performance can be improved by providing a separate bias pad to each switch and also by adding extra RC filters to suppress RF leakage through the biasing circuitry.

Nonlinear characterization of a tunable filter using metal-contact switches from Radant MEMS Inc. has been described in [16] and the filter third-order intermodulation level has been found to be below the noise level of the spectrum analyzer. The measured IIP3 is shown to be greater than 65 dBm and proves that filters using RF MEMS switches have excellent linearity compared to Schottky varactor-diode tunable filters (IIP3 of 15–25 dBm) [17] and pin-diode switches filters (IIP3 of 25–35 dBm) [18].

IV. TWO-BIT INDUCTIVELY-TUNED FILTER

A. Resonator Design

The proposed half-wavelength open-ended microstrip resonator for two-bit inductive tuning is shown in Figure 6. Each of the sections “A” and “B” in Figure 6 are equidistantly placed at a distance of $L = 5.2$ mm from the open end of the resonator to provide a loading inductance determined by the length of each section ($L_A$, $L_B$) and the ratio of the width of each section ($W_A$, $W_B$) to the line-width of the resonator ($W_R$). In Figure 7(a), full-wave simulation is used to study the dependency of resonator center frequency to the dimensions of Section “A” ($L_A$, $W_A$), ignoring the presence of Section “B” where “$S_0$” is closed and “$S_1$” is open. Similarly, Figure 7(b) shows the dependency of the resonance frequency to the dimensions of Section “B” ($L_B$, $W_B$) ignoring the presence of Section “A” where “$S_1$” is closed and “$S_0$” is open. When both switches “$S_0$” and “$S_1$” are open,
the overall inductance provided by both Sections “A” and “B” results in the lowest resonance frequency of the four possible resonant frequencies of the two-bit resonator.

The minimum width of the line is chosen to be 0.1 mm for full-wave simulations, due to tolerance limitations of the fabrication process. The maximum line-width of the inductive section is determined by the line-width of the resonator ($W_R$). As $W_A$ or $W_B$ approaches $W_R$, the loading inductance of the section tends to be zero. The upper bound on line-width of each section is chosen to be 4 mm. Reasonable lengths of inductive sections are chosen ($L_A, L_B = 2–6$ mm) to ensure that each section provides a loading inductance which is sufficient to tune the resonator to lower frequencies.

When both switches are closed, the resonance frequency is determined by the overall electrical length of the unloaded resonator (half-wavelength) and is assumed to be 1.65 GHz. If two intermediate frequencies are considered to be 1.26 ($S_0: 01$), and 1.4 GHz ($S_1: 00$), respectively, Figures 7(a) and 7(b) show that many solutions for ($W_A, L_A$) and ($W_B, L_B$) exist for these resonant frequencies. A choice must be made based on a trade-off between maximum allowable mismatch at the filter input/output and filter miniaturization. Inductive sections with smaller widths can achieve the same resonance frequency with a smaller length and hence result in greater miniaturization. But the drastic change in the width of the transmission line from $W_R$ to $W_A$ or $W_B$ also results in greater mismatch at the input/output of the filter. This trade-off is not significant for the two-state tunable filter described in Section III but becomes important for the design of multibit tunable filters due to the presence of multiple discontinuities in the resonator structure. For 1.4 GHz resonance frequency, the dimensions of the resonator are chosen to be ($W_B, L_B$) = (2.2 mm, 5 mm) when “$S_0$” is open (“$S_1$” is closed). Similarly, ($W_A, L_A$) = (0.3 mm, 3 mm) gives a resonance frequency of 1.26 GHz when “$S_1$” is open (“$S_0$” is closed). When all switches are open, the resonator tunes to 1.15 GHz. By choosing different widths ($W_A < W_R$) for each section, acceptable return loss is guaranteed for all filter states. This will be demonstrated in Simulation Results Section.

B. Filter Design

Figure 8 shows the layout of a two pole, 5% tunable filter with 0.036 dB ripple with two bits of tuning control. The circuit is designed on RT/Duroid 6010LM substrate with $\varepsilon_r = 10.2 \pm 0.25$, $\tan \delta = 0.0023$, $t = 2.54$ mm at $f_{\text{min}} = 1.16$ GHz. The interresonator coupling coefficient ($k_{12} = 0.086$) is realized with a microstrip gap of 0.8 mm. Optimized values of input/output J-inverters are chosen to achieve best input matching over all four states ($C_{io}=2$ pF). The dimensions of the inductive sections be ($W_B, L_B$) = (2.2 mm, 5 mm) when “$S_0$” is open (“$S_1$” is closed). Similarly, ($W_A, L_A$) = (0.3 mm, 3 mm) gives a resonance frequency of 1.26 GHz when “$S_1$” is open (“$S_0$” is closed). When all switches are open, the resonator tunes to 1.15 GHz. By choosing different widths ($W_A < W_R$) for each section, acceptable return loss is guaranteed for all filter states. This will be demonstrated in Simulation Results Section.

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“A” and “B” (Fig. 6) are also optimized by electromagnetic simulation to account for parasitics in the filter structure. Better return loss is observed when the section with greater line-width (Section “B” in Fig. 6) is placed closer to the feed lines.

Monolithic metal-contact series switch banks are used to tune the filter. Table II shows the RF MEMS switch parameters for the metal-contact switch developed by Northeastern University and Analog Devices [1, 19]. A similar switch structure has been modified by Nishijima et al. for high-power RF application [20]. The packaged version of the switch in [19] has been tested by Radant MEMS to 100 billion cycles at 100 mW RF power [21] and has been used in the design of a two-state reconfigurable filter described in section 3.

Each RF MEMS switch bank represented by “S0” and “S1” in Figure 8 is substituted with three simultaneously actuated metal-contact switches connected in parallel with each other to decrease the loss associated with the down-state resistance ($R_s = 1 \, \Omega$) of each MEMS switch. Table III shows the trade-off between the filter insertion loss and the number of switches in each bank. Adding more than three switches to the bank does not improve the filter insertion loss significantly. Using a number of switches in parallel will result in greater up-state capacitance across the terminals of the switch bank. This up-state capacitance has almost no effect on the loading inductance provided by the step discontinuity and hence will have negligible effect on the tuning range of the filter at low GHz frequencies.

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**TABLE II. Parameters for the Northeastern University/Analog Devices RF MEMS Switch [1]**

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<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
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</thead>
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<td>Cantilever length (µm)</td>
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<td>Actuation voltage (V)</td>
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<tr>
<td>Cantilever width (µm)</td>
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<td>Switch resistance, $R_s$ (Ω)</td>
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<td>Up-state capacitance (fF)</td>
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</tr>
<tr>
<td>Cantilever thickness (µm)</td>
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<td>Cantilever inductance (pH)</td>
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<tr>
<td>Spring constant (N/m)</td>
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<td>Isolation at 4 GHz (dB)</td>
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</tr>
<tr>
<td>Actuation area (µm²)</td>
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<td>Loss (0.1–20 GHz) (dB)</td>
<td>-0.15</td>
</tr>
</tbody>
</table>

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Figure 8. Filter layout for a two-bit tunable filter using monolithic series metal-contact RF MEMS switch banks. All dimensions are in mm and given as: $L_1 = L_5 = L_7 = 1.5$, $L_2 = L_9 = 4.2$, $L_3 = L_4 = 5.3$, $L_6 = 3.4$, $L_8 = 7.5$, $W_1 = W_7 = 5.5$, $W_2 = W_3 = 2.3$, $W_4 = 2.7$, $W_5 = 0.3$, $W_6 = 2.1$, $W_8 = 0.25$, $W_9 = 0.8$, $C_{io} = 2$ pF.
Stepped-impedance stubs are placed at the input and output of the resonator to suppress the spurious passband and improve the upper rejection skirt. Independent control of two transmission zeros is achieved in this design by controlling the ratio of admittances of microstrip lines in each stepped-impedance stub. By placing the transmission zeros at two different frequencies, spurious suppression is achieved over a broader band of frequencies. The transmission zeros are placed around twice the midband frequency of the tuning range of the filter so that the spurious passband remains suppressed for all tuning states. The ratio of admittances in each stub is adjusted so that transmission zeros are placed at 2.84 GHz ($W_3/W_1 = 0.42$, $L_2 = 4.2$ mm) and 2.89 GHz ($W_4/W_1 = 0.5$, $L_9 = 4.2$ mm). However, if the transmission zeros are placed wider apart, the upper stopband performance is degraded due to the presence of a sidelobe between the two transmission zero frequencies.

C. Simulation Results

Figure 9 presents the obtained full-wave simulation results of the filter shown in Figure 8. The center frequency of the filter changes from 1.16 GHz to 1.67 GHz (44% tuning range) in four discrete states while maintaining a relatively constant bandwidth (Table IV). The simulated return loss is better than ~12 dB over the entire tuning range. The predicted insertion loss of the filter varies from 1.2 dB to 1.9 dB when all the switches are in the up-state and down-state positions, respectively. The down-state resistance of the switch dominates the loss mechanism in the “11” state.

The drastic change in the line-width of the resonator when “$S_1$” is in the up-state position results in higher radiation loss due to fringing fields at the discontinuity. When “$S_0$” is in the up-state position, lesser fringing fields at the discontinuity will result in lower filter insertion loss. The use of stepped-impedance stubs at the ends of the filter results in second harmonic suppression better than 30 dB for all filter states except the “11” state. To maintain spurious suppression for all states including “11” state, the quarter wavelength stubs should be tuned as the filter center frequency is changed.

V. CONCLUSION

This article demonstrates a two-pole reconfigurable microstrip filter using packaged RF MEMS switches. A novel inductive-tuning technique is used to tune the center frequency from 1.06 to 1.23 GHz (17% tuning range) with almost constant fractional bandwidth. Both simulated and measured results are in good agreement. The inductive-tuning technique is extended to design a two-pole, two-bit tunable filter

| TABLE III. Variation of Insertion Loss with Number of MEMS Switches in Each Bank, When All Switches in the Filter are in the Down-State Position |
|-----------------|---|---|---|---|---|
| Number of switches in parallel | 1 | 2 | 3 | 4 | 5 |
| Insertion loss (dB) | 2.7 | 2.2 | 1.9 | 1.8 | 1.74 |

| TABLE IV. Tunable Filter Characteristics (“0”: Up-state; “1”: Down-state) |
|-----------------|---|---|---|
| $S_1$ | Frequency (GHz) | I.L. (dB) | BW (MHz) | BW (%) |
| 00 | 1.16 | 1.20 | 58 | 5 |
| 01 | 1.27 | 1.60 | 70 | 5.5 |
| 10 | 1.42 | 1.43 | 75 | 5.3 |
| 11 | 1.67 | 1.90 | 95 | 5.6 |

Figure 9. Simulation results of the two-bit tunable filter. (a) Insertion loss. (b) Return loss.
using monolithic RF MEMS metal-contact switches. The two-bit filter is tunable from 1.16 to 1.67 GHz (44% tuning range) when maintaining almost constant fractional bandwidth. The inductive-loading technique can be used to design tunable filters with greater than four states by cascading inductive-sections in the resonator and using switches to control the overall loading inductance. Stepped-impedance stubs at the input and output of the filter have shown to suppress the spurious passband and improve the upper rejection skirt.

The tuning range of the filter fabricated with packaged RF MEMS switches is limited by the parasitic inductances associated with the package and bondwire. Also, packaged single-pole single-throw RF MEMS switches impose limitations on the realization of multibit tunable filters due to space constraints (each packaged switch has a size of 1.42 mm × 1.37 mm [15] compared with monolithic RF MEMS switches which have dimensions in the order of several hundred microns or less [1]). Tunable filters using monolithically fabricated MEMS switches show greater tuning range due to the reduced parasitic elements associated with each MEMS switch. Filters implemented with monolithically fabricated MEMS switches do not have the space constraints associated with a packaged switch and hence enable multibit tunability by employing multiple inductive sections. The minimum length of each inductive section in the resonator is limited only by the size of the MEMS switch. On-wafer hermetic packaging of MEMS switch banks ensures reliable operation of the switch bank without introducing significant parasitics that potentially limit the tuning range of the filter.

The use of capacitive RF MEMS shunt switches as tuning elements is more suited for high frequency (>10 GHz) implementations of the inductively loaded reconfigurable filters described in this article where the insertion loss and isolation of series metal-contact switches become unacceptable.

This study has demonstrated a class of multibit reconfigurable filters with low loss and constant fractional bandwidth over the tuning range, using a novel method to control resonator loading in microstrip filters.

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BIOGRAPHIES

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**Kamran Entesari** received the B.S. degree in electrical engineering from the Sharif University of Technology, Tehran, Iran, in 1995, the M.S. degree in electrical engineering from the Tehran Polytechnic University, Tehran, Iran, in 1999 and the Ph.D. degree from The University of Michigan at Ann Arbor, in 2005. In 2006, he joined the Department of Electrical and Computer Engineering, Texas A&M University, College Station, where he is currently an Assistant Professor. His research interests include the design of RF/microwave/millimeter-wave integrated circuits and systems, RF microelectromechanical systems (MEMS), related front-end analog electronic circuits, and medical electronics. Dr. Entesari was the recipient of the 2009 Semiconductor Research Corporation (SRC) Design Contest Second Project Award for his work on dual-band millimeter-wave receivers on silicon.