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High-Selectivity E-Band Image-Reject Filters in SiGe Technology

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Abstract — To alleviate practical limitations in the design of mm-wave on-chip image-reject filters, systematic design methods are presented. Two low-order filters with high-selectivity and low-loss characteristics are compared. The transmission zeroes are created by means of a quarter-wave transmission line (filter 1) and a series LC resonator (filter 2). Implemented on SiGe, they consume only 0.125 and 0.064 mm² chip area including pads. The measured transmission losses across 81-86 GHz E-band frequency range are 3.5-5 dB (filter 1) and 3-4.5 dB (filter 2) where rejection at the image frequency is greater than 30 dB.

Index Terms — E-band, image-reject filters, MMIC, mm-wave, selectivity, SiGe, transmission lines, transmission zero.

I. INTRODUCTION

While the spectra of low-GHz wireless systems have been overly crowded, the 71-76 GHz and 81-86 GHz E-band system offers a fresh opportunity to establish fiber-like high-speed wireless links with full duplex throughput of at least 1 Gbps. Unlike the 60-GHz band, the E-band is not penalized by the transmission loss due to atmospheric particles, i.e., oxygen and therefore it suits well for long-distance backhaul point-to-point communications. In addition, the availability of generous bandwidth within E-band enables deployment of simple modulation schemes such as OOK, BPSK, and FSK.

In the E-band heterodyne receiver employing a frequency quadrupler and a divide-by-two frequency divider, the LO and IF frequencies are respectively \( f_{\text{RF}}/9 \) and \( f_{\text{RF}}/9 \) and as a result the image frequency will be located at 73-76 GHz for \( f_{\text{RF}} = 81-86 \) GHz (upper E-band). Such heterodyne system typically requires high image noise rejection of at least 30 dB so as the overall system noise figure (NF) can be kept sufficiently low.

A simple notch filter typically incorporated within the LNA-Mixer circuits can only provide a modest image rejection of 10-15 dB. When compared to bulky waveguide-type filters which come with low-loss and high-selectivity characteristics, on-chip planar filters [1]-[5] offers more compact solution but at the expense of higher insertion loss. For example, a third-order bandpass filter with a center frequency of 9.45 GHz reported in [5] was fabricated in 130 nm CMOS technology and it exhibited an insertion loss of 15.6 dB. Higher-order filters provide high selectivity but require increased number of components leading to large chip area and more importantly high loss. In this paper, we present the design, realization, and comparison of two on-chip image-reject filters (IRFs) which exhibit low loss within 81-86 GHz frequency range and simultaneously facilitate high rejection across 63-66.9 GHz frequency range.

II. FILTER’S SYNTHESIS

A. Third-Order HPF with RC Parallel Loading

To start with, a third-order high-pass filter (HPF) is chosen due to its simplicity, i.e., it only requires few components. As illustrated in Fig. 1(a), the filter is comprised of a series capacitance \( C_{\text{X}}/2 \) shunted by an inductance \( L_{\text{X}} \) at both sides. The source and load impedances are represented by a parallel RC network where \( R \) is typically 50 \( \Omega \) and \( C \) represents the parasitic capacitance of input/output pads \( C_{\text{pad}} \) whose effects are detrimental at mm-wave frequencies and therefore have to be accounted for in the design. For given \( R, C \), and \( R_{\text{L}} \) values, the optimum values of \( L_{\text{X}} \) and \( C_{\text{X}} \) can be computed using (1) and (2) where \( G = 1/R \) and \( G_{\text{L}} = 1/R_{\text{L}} \). Fig. 2 illustrates the effect that parameter \( R_{\text{L}} \) has on the transmission and return losses of the HPF. Lower \( R_{\text{L}} \) values result in better rejection but at the cost of narrower band matching.

\[
L_{\text{X}} = \left[ \omega \left( \omega C + \sqrt{G(G_{\text{L}} - G)} \right) \right]^{-1} \quad (1)
\]

\[
C_{\text{X}} = G_{\text{L}} / \left( \omega \sqrt{G(G_{\text{L}} - G)} \right) \quad (2)
\]
behave like an inductance. Hence, the characteristic impedance of the stub, \( Z_0 \), is achieved at \( f_1 \) (mid-band frequency 81-86 GHz). The electrical length of the stub is \( \frac{\lambda}{4} \) stub in Fig. 1(b) is replaced by a series LC resonator, \( Y \), will be larger than 90° (3), and as a result, the stub will behave like an inductance. Hence, the characteristic impedance of the stub, \( Z_0 \), can be calculated using (4).

\[
\theta_2 = \left( \frac{f_2}{f_1} \right) 90° \quad (3)
\]
\[
Z_{in} = \frac{Z_0}{j \tan \theta_2} = \frac{Z_0}{j \tan(180 - \beta)} = \frac{j Z_0}{\tan \beta} \leftrightarrow j \omega L_X
\]
\[
\therefore Z_0 = \omega L_X \tan \beta \quad (4)
\]

C. High-Selectivity Filter with Series LC Resonator

Although in terms of component count, the circuit in Fig. 1(b) is as compact as that in Fig. 1(a), in the implementation the quarter-wave transmission line (TL) employed in Fig. 1(b) will consume a considerable amount of chip area, even at millimeter-wave frequencies. In order to miniaturize the filter in Fig. 1(b) while keeping its high selectivity characteristic, the \( \lambda/4 \) stub in Fig. 1(b) is replaced by a series LC resonator, Fig. 1(c). Here, the insertion zero at the image frequency is created by resonating \( L_Y \) and \( C_Y \) at \( f_1 \) (5). At \( f_2 \), this series LC network will present an inductive net reactance. Once \( L_X \) in Fig. 1(a) has been computed, the value of \( L_Y \) can be determined using (6).

\[
\omega^2 L_Y C_Y = 1 \quad (5)
\]

\[
Z_{LC} = j \left( \alpha^2 L_Y - \frac{1}{\alpha^2 C_Y} \right) = j \frac{\alpha^2 \omega^2 L_Y C_Y - 1}{\alpha \omega_1 C_Y} = j \frac{\alpha^2 - 1}{\alpha^2 \omega_2 L_Y} \leftrightarrow j \omega_2 L_X
\]

III. PRACTICAL DESIGN LIMITATIONS

The image-reject filters were designed and realized using Infineon B7HF200 SiGe technology. In this process, four copper metal layers (M1–M4) are available. The topmost metal, M4, is 2.8-µm thick. The smallest available pad measures 68x68 µm² and it has an equivalent parasitic capacitance \( C_{pad} \) of around 25 fF. Metal-insulator-metal (MIM) capacitors are also provided within this process.

The initial physical length of the \( \lambda/4 \) stub is estimated using \( c/(4 \times f \times \alpha_{eff}) \) where \( c \) is the speed of light and \( \alpha_{eff} \) is the effective relative permittivity of the substrate. It is then optimized in SONNET in such a manner that the lowest impedance is achieved at \( f_1 \). The initial width of the \( \lambda/4 \) stub is first calculated using (3)-(4). It is then carefully tuned until the equivalent inductance value of the stub at \( f_2 \) is equal to \( L_X \).

While the design procedure for the quarter-wave stub employed in Fig. 1(b) is quite straightforward, it is not the case for the series LC resonator employed in Fig. 1(c) since the values of \( L_X \) and \( C_Y \) are strongly dependent on the frequency. As a consequence, equation (6) is only valid under assumptions that the inductance values of \( L_Y \) at \( f_1 \) and \( f_2 \) are identical and the capacitance values of \( C_Y \) at \( f_1 \) and \( f_2 \) are also identical. However, in practice, their values at \( f_1 \) and \( f_2 \) would be largely different \((L_{Y2} \neq L_{Y1} \text{ and } C_{Y2} \neq C_{Y1})\). To overcome this problem, the following design method is proposed:

**Step 1:** For pre-determined \( R \), \( C \), and \( R_1 \) values, calculate \( L_X \) and \( C_X \) using (1) and (2).

**Step 2:** Compute \( \alpha \) and \( C_{Y1} \) (i.e. \( L_Y \) at \( f_1 \)) using (6) and then calculate \( C_{Y2} \) (i.e. \( C_Y \) at \( f_2 \)) using (5). This is to ensure that the \( L_X-C_Y \) resonator will provide a short circuit at \( f_1 \).

**Step 3:** At \( f_2 \), \( L_Y = L_{Y2} \) and \( C_Y = C_{Y2} \). The equivalent inductive reactance of the \( L_{Y2}-C_{Y2} \) resonator i.e. \( L_{LC} \) will be larger than \( L_X \); see Fig. 3. In order to mitigate the effect of this excess inductance \( L_X \), a shunt capacitance \( C_3 \) is added. This is accomplished by resonating \( L_3 \) and \( C_3 \) at \( f_2 \). As a consequence, the initial value of \( C \) in **Step 1** must be set larger than \( C_{pad} \) i.e. \( C = C_3 + C_{pad} \). If \( L_3 \) is not properly compensated, the filter’s insertion loss will be compromised.

IV. SIMULATION VS. MEASUREMENT RESULTS

Simulated transmission losses of the filters in Figs. 1(a) and (c) are compared in Fig. 4. When compared to the filter in Fig. 1(a), the series-LC resonator employed in Fig. 1(c) proves effective for improving the rejection at the image frequency.
Simulated using EM models, the circuit in Fig. 1(c) results in rejection levels higher than 30 dB from 53.7 GHz to 66 GHz with a notch of -54 dB occurred at 61.3 GHz (graph c). Transmission loss within 81-86 GHz frequency band varies between 1.5 and 2.1 dB, i.e., better than that using λ/4 TLs. Return loss is higher than 10 dB from 76 GHz to 99.1 GHz.

Two chip prototypes have been fabricated and their microphotographs are shown in Fig. 5. The IRF with a series LC resonator occupies only 268 × 238 µm² chip area whereas the IRF with λ/4 TLs image-signal trap measures 468 × 266 µm². Small-signal measurements were undertaken using an Agilent 110 GHz general-purpose network analyzer (PNA) and GSG Cascade probes with 100 um pitch.

The measured and simulated S-parameter results of the IRFs with λ/4 TLs and a series LC resonator are compared in Figs. 6 and 7, respectively. They are in reasonably good agreement. Since the filters are symmetrical, only S_{21} and S_{22} are shown. From Fig. 6, it can be observed that the λ/4-TL filter exhibits higher than 30 dB rejection from 49.5 GHz to 66.5 GHz with a notch of -46 dB occurred at 58.3 GHz and transmission loss ranging from 3.5 dB to 5 dB across the 81-86 GHz frequency band. Return loss was better than 10 dB within a frequency range 80-102 GHz. On the other hand, from Fig. 7, the more compact IRF that employs a series LC network offers rejection greater than 30 dB from 54 GHz to 67 GHz with a notch of -46 dB at 60.5 GHz. The transmission loss varied between 3 dB and 4.5 dB across the frequency band of interest and return loss higher than 10 dB was obtained over a wide frequency range 80-110 GHz. Higher insertion losses observed in the measurements compared to the simulations could be due to the presence of parasitic capacitive coupling to the low-resistivity substrate stronger than what had been predicted. The performances of the two filters reported in this paper are summarized and compared with previously published mm-wave on-chip filters in Table I.

### Performance Summary and Comparison

<table>
<thead>
<tr>
<th>Ref</th>
<th>Process</th>
<th>Freq (GHz)</th>
<th>IL (dB)</th>
<th>Attenu. Slope (dB/10 GHz)</th>
<th>Area (mm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[1]</td>
<td>CMOS</td>
<td>44.5-75.5</td>
<td>&gt;3.9</td>
<td>~4/-4</td>
<td>0.85</td>
</tr>
<tr>
<td>[2]</td>
<td>CMOS</td>
<td>77</td>
<td>9.3</td>
<td>~11/-9</td>
<td>0.11</td>
</tr>
<tr>
<td>[3]</td>
<td>SiGe</td>
<td>71.3-83.3</td>
<td>6.4-9.4</td>
<td>~20.5/-12</td>
<td>0.107</td>
</tr>
<tr>
<td>[4]</td>
<td>SiGe</td>
<td>50.5-66</td>
<td>4-6</td>
<td>~16/-6</td>
<td>0.185</td>
</tr>
<tr>
<td>Fig. 5(a)</td>
<td>SiGe</td>
<td>81-86</td>
<td>3.5-5</td>
<td>~18.5</td>
<td>0.125</td>
</tr>
<tr>
<td>Fig. 5(b)</td>
<td>SiGe</td>
<td>81-86</td>
<td>3-4.5</td>
<td>~20.5</td>
<td>0.064</td>
</tr>
</tbody>
</table>

**References**


