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The differential output current of the Gilbert mixer is depicted as highly linear and only performs the frequency translation. Make the current fully switch. The current commutation mechanism with only about 0.1-V LO voltage swing requirement to offset voltage for the two differential pairs of $M_1-M_2$ and $M_3-M_4$.

The drain saturation voltage is less than the gate-overdrive voltage if the short channel effect takes place. However, if the gate-overdrive voltage is still small, the MOS transistors are still in the long channel regime because the electric field is not large enough to saturate the drift velocity of the electrons. Because the PMOS transistor has a lower mobility than the NMOS transistor for the same size, the PMOS transistor is almost in the long channel regime even if the gate-overdrive voltage is large. The gate length of the PMOS transistors $M_1-M_4$ in the input stage of the first chip is 0.5 μm, whereas the gate length of the PMOS transistors $M_1-M_4$ of the other chip is 1 μm. For the lower mobility of the PMOS transistors, the widths of the PMOS transistors are designed much wider to achieve similar transconductance gain when compared with the NMOS upconverter. As a result, the IF bandwidth of the upconverter with a folded PMOS TCA is narrower than that with an NMOS TCA.

### 2.2. Gilbert Mixer Core with an LC Current Combiner

Instead of using the MOS transistors in the Gilbert switching quad with a large LO switching voltage requirement of $\frac{\sqrt{2}V_m}{2}$ (gate-overdrive voltage), the bipolar transistors $Q_1-Q_4$ are utilized in this work with only about 0.1-V LO voltage swing requirement to make the current fully switch. The current commutation mechanism is highly linear and only performs the frequency translation. The differential output current of the Gilbert mixer is depicted as $I_+$ and $I_-$ with the LC current combiner as shown in Figure 2. At the resonant frequency of $\omega_0 = \frac{1}{\sqrt{2LC}}$, the input current $I_+$ reverses and thus the total output current doubles when compared to the single-ended output current of each path [7, 8].

![Figure 1](image1.png) Schematic of the Gilbert upconverters with NMOS/PMOS-type bias-offset TCA's and a single-band LC current combiner

![Figure 2](image2.png) Block diagram of the LC current combiner and its equivalent circuit at the resonant frequency

![Figure 3](image3.png) Schematic of the 2.4/5.7 GHz dual-band Gilbert upconverter with a dual-band LC current combiner

#### 2.3. Shunt–Shunt Feedback Transimpedance Amplifier

The output shunt–shunt feedback TIA is used to translate the combined current output $I_{\text{tot}}$ to the voltage signal. Moreover, the output impedance is reduced by the factor of $(1 + A\beta)$ for the shunt-type feedback and thus the output matching is easily

![Figure 4](image4.png) (a) dual-band prototype of the LC current combiner and (b) its current combining performance
achieved. The 20-GHz bandwidth of the feedback TIA is designed with strong feedback to reduce the nonlinearity effect [9]. Moreover, the schematic of the 2.4/5.7 GHz dual-band upconverter using dual-band LC current combiner is shown in Figure 3. Figure 4 shows a dual-band resonator consisting of two parallel resonators \( L_p (= L_{p1} = L_{p2}) \) and \( C_p \) and a series resonator \( (L_s \) and \( C_s) \). The differential input currents \( (I_+ \) and \( I_-) \) are generated by the Gilbert mixer. By definition, the element \( D \) of the ABCD matrix (1) represents the current gain from port-1 to port-2 when port-1 is short-circuited.

\[
D = \left. \frac{I_2}{I_1} \right|_{V_{1s}=0} = 1 + \frac{L_s}{L_p} + \frac{C_p}{C_s} - \omega^2 (C_s L_s) - \frac{1}{\omega^2 C_s L_p}
\]

The current reversal occurs when \( D = -1 \) and \( I_2 = -I_+ = -I_1 \). Thus, the total combined current \( (I_{tot}) \) is doubled, that is, ideally 6 dB gain improvement, when compared with each output node of the Gilbert mixer. This mechanism is achieved at two frequencies by solving \( D = -1 \).

\[
D = \frac{I_2}{I_1} \bigg|_{V_{1s}=0} = 1 + \frac{L_s}{L_p} + \frac{C_p}{C_s} - \omega^2 (C_s L_s) - \frac{1}{\omega^2 C_s L_p}
\]

Figure 5  (a) Photograph of the SiGe BiCMOS high linearity Gilbert upconverter using an NMOS TCA and an LC current combiner (b) using a PMOS TCA and the same current combiner (c) dual-band Gilbert upconverter with dual-band LC current combiner.

Figure 6  (a) Conversion gain with respect to RF frequency and (b) power performance of the high linearity Gilbert upconverters using NMOS/PMOS bias-offset TCAs and the LC current combiner.
are shown in Figures 5(a) and 5(b), whereas photograph of the dual-band upconverter is shown in Figure 5(c).

3. MEASUREMENT RESULTS AND CONCLUSION

The SiGe BiCMOS high linearity upconverters facilitate on-wafer RF measurements. The peak conversion gain of the upconverters with NMOS/PMOS input TCAs occurs at both RF = 4.4 GHz, IF = 20 MHz, and the LO power is −5 dBm as shown in Figure 6(a). The power performance of the upconverters with NMOS/PMOS TCAs is shown in Figure 6(b). The OP1dB and OIP3 are −11/−11 dBm and 5.5/9.5 dBm for the upconverters with NMOS and folded PMOS TCAs, respectively. Therefore, the difference between the OIP3 and OP1dB of the upconverters with NMOS/PMOS TCAs are 16.5 and 20.5 dB, whereas a conventional Gilbert upconverter has only 10-dB difference between the OIP3 and OP1dB [10, 11]. The IF bandwidth of the upconverters with NMOS and PMOS TCAs are 500 MHz and 100 MHz, respectively. The output RF return loss is better than 16 dB for both circuits over 20 GHz. The LO-to-RF isolation of each upconverter is better than 28/26 dB when LO frequency ranging from 4–5.2 GHz. The power consumption of each circuit is 40/46 mW, respectively.

Figure 7(a) shows the conversion gain with respect to RF frequency when IF = 100 MHz. The measured conversion gain at desired frequencies of 2.4/5.7 GHz is −3/−3.5 dB with the LO input power of −4 dBm. However, the peak conversion gain is 0/−3 dB at 2.7/5.9 GHz due to the process variation. As shown in Figure 7(b), the dual-band upconverter has the OP1dB of −11.5/−11 dBm, and the IIP3 of 0.5/1 dBm when IF = 100 MHz, RF = 2.4 GHz, and 5.7 GHz, respectively. The power consumption is 49.5 mW at a 3.3-V supply.

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A MINIATURE X-BAND-EMBEDDED MULTILAYER BAND-PASS FILTER USING LTCC TECHNOLOGY

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ABSTRACT: This article proposes a miniature embedded multilayer interdigital bandpass filter (BPF) based on low temperature co-fired ceramic (LTCC) technology, which is used to suppress fundamental and harmonic components of local-oscillator signal in a LTCC transceiver module. A practical design methodology corresponding to LTCC multilayer configuration is described. A five-order Chebyshev interdigital BPF is developed and verified by full-wave simulation. The proposed filter is fabricated using multilayer LTCC technology and measured using vector network analyzer. Good agreement between simulated and measured response is observed. Covering area of the fabricated BPF is only 8.1 × 7.4 mm² (including via fences).

Key words: BPF; LTCC; LO; transceiver; multilayer configuration

1. INTRODUCTION

Increasing demands of making communication systems lighter, more compact, and with better functionality, set many challenges for packaging and structure configuration. LTCC technology gives a new design conception and method to solve these problems [1, 2]. By embedding passive components into multilayer substrate and mounting active elements on the surface layer, LTCC technology offers opportunity to realize very compact systems, such as LTCC transmitters, receivers, or complicated transceiver modules [3, 4].

LTCC bandpass filter (BPF) is a key component in these modules [5]. When the frequency is not high, LTCC filters are usually implemented by LC-element type. With increasing of frequency, the effects of parasitics deteriorate performance of such filters. So, coupling resonator or cavity bandpass filter realized in LTCC multilayer substrate was widely used in the LTCC microwave and millimeter-wave modules [6].

In this article, a feed-in RF signal (4.75 GHz) was octupled up to millimeter-wave frequency (38 GHz), as local-oscillator (LO) signal in the LTCC transceiver module, as shown in Figure 1. After first doubling, second-order harmonic is needed, but the leaking fundamental (4.75 GHz) and 3rd order harmonic (14.25 GHz) components should be suppressed at least 40 dB by BPF. At the same time, BPF must be incorporated with the whole LTCC module. Therefore, an embedded LTCC BPF was designed for this application.

2. ANALYSIS

The classical design theories of filter have been consummated since long time ago [7–9]. For symmetric coupling structure, even- and odd-mode analyze is widely used. However, this method is not applicable for asymmetric coupled line buried in different layers. In this section, a practical design method is presented, which is based on EM simulation extraction of external quality factor \( Q_e \) and coupling coefficient \( K_{ij} \). First of all, the element values for Chebyshev lowpass prototype can be calculated as follows [10]:

\[
g_{0} = 1.0, g_{1} = 2 \sin \left( \frac{\pi}{2n} \right)
\]

\[
g_{i} = g_{i-1} \cdot \frac{(2i-1)\pi}{2n} \sin \left( \frac{(2i-3)\pi}{2n} \right)
\]

\[
g_{n+1} = (n_{odd}) \text{ or } \cot \left( \frac{\beta}{4} \right) (n_{even})
\]

Where \( \beta = \ln \left[ \cot \left( \frac{L_{0}}{17.37} \right) \right], \gamma = \sinh \left( \frac{\beta}{2n} \right) \)

then, the design parameters of BPF, that is, the coupling coefficient \( K_{ij} \) and external quality factor \( Q_{e} \), can be determined by the formulas:

\[
Q_{e} = Q_{w} = \frac{g_{0}g_{1}}{FBW}
\]

\[
K_{i+1} = \frac{FBW}{\sqrt{g_{i}g_{i+1}}} \text{ for } i = 1 \text{ to } (n - 1)
\]

Where \( g_{i} \)'s are the element values of Chebyshev lowpass prototype filter, \( FBW \) is the fractional bandwidth, and \( n \) is the order of the filter.

Relationship between these parameters and filter’s physical dimensions can be made by

\[
Q_{e} = \frac{g_{0}}{\Delta \omega_{200}^{2}} \left( \frac{f_{2} - f_{1}}{f_{2}^{2} - f_{1}^{2}} \right)
\]

\[
K = \frac{g_{0}}{f_{2}^{2} - f_{1}^{2}}
\]

\( \omega_{o} \) is center frequency of BPF, \( \Delta \omega_{200} \) can be determined from the frequency at which the phase of \( S_{11} \) shifts ± 90° with respect to