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Design of Lange-Couplers and Single-Sideband Mixers Using Micromachining Techniques

Chen-Yu Chi and Gabriel M. Rebeiz

***Abstract*—This paper reports on the design and performance of micromachined Lange-couplers and single-sideband mixers (SSB) on thin dielectric membranes at *Ku*-band. The micromachined Lange-coupler results in a 3.6±0.8 dB coupling bandwidth from 6.5 to 20 GHz. The Lange-coupler and an interdigital filter are used in a 17-GHz SSB. The SSB mixer requires 1–2 mW of local oscillator (LO) power without dc bias and achieves a 30 dB upper-sideband (USB) image rejection for an IF frequency of 1 GHz and above. The micromachined membrane technology can be easily scaled to millimeter-wave monolithic microwave integrated circuits (MMIC’s) to meet the low-cost requirements in automotive or portable communication systems.**

***Index Terms*—Micromachining, single-sideband mixers.**

I. Introduction

Silicon micromachined technology has been used to build low-loss lumped elements, filters, and Wilkinson power dividers [1]–[3]. Recent research on micromachined stripline resonators and filters has demonstrated a conductor-loss limited performance at *K*- and *Ka*-band [4]. The micromachined components, suspended on thin dielectric membranes, do not suffer from dielectric loss and dispersion and can be used in the design of high-quality low-loss circuits at millimeter-wave frequencies. The purpose of this paper is to demonstrate the capability of micromachining technology to the design of integrated circuit modules. A single-sideband (SSB) mixer (also called an image-rejection mixer) is built and tested at *Ku*-band frequencies with applications in satellite communication systems and future millimeter-wave radiometers and automotive systems.

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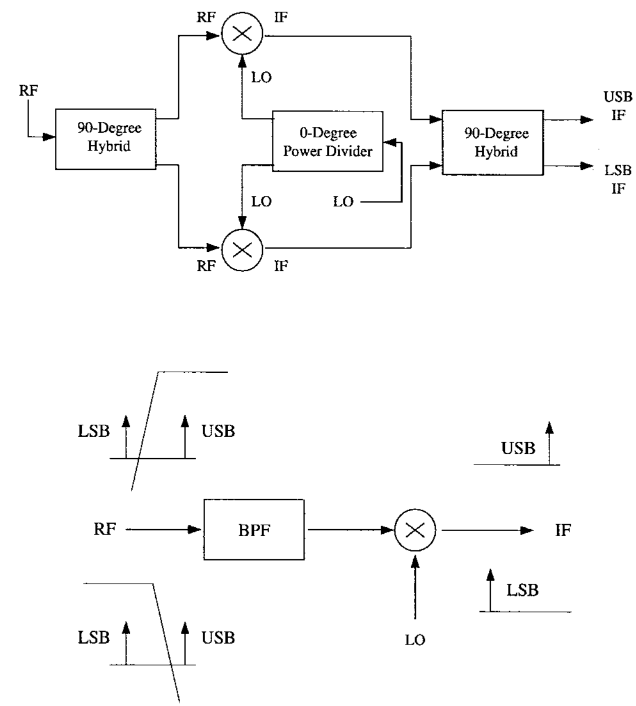

Fig. 1. Two different topologies in the design of SSB’s (image rejection mixers).

There are two different topologies in the design of single-sideband mixers (Fig. 1). The first topology is to use hybrid circuits to provide the required amplitude balance and phase shift to achieve the image rejection mixing. The advantage of this method is that a good isolation between the local ocsillator (LO), RF, and IF ports can be achieved and a low IF (0.01–2 GHz) can be extracted. However, the disadvantage of this design is that the performance of the SSB mixer strongly depends on the amplitude and phase balance of the hybrid circuits used and since this is difficult to achieve at millimeter-wave frequencies, this method is therefore seldom used in this frequency range. The second SSB mixer topology is to place a filter in front of a balanced mixer. The filter is designed to pass the RF frequency and reject the image frequency. This design employs only one balanced mixer and does not use an RF hybrid, LO power divider and IF-hybrid, resulting in a much smaller unit. A SSB mixer employing micromachining technology based on the second topology is presented here. The SSB mixer can be easily scaled to *V*-band (60 GHz) and *W*-band (77 GHz, 94 GHz) frequencies.

II. Design of the Micromachined Lange-Coupler

The analysis of a Lange-coupler is very similar to the analysis of interdigital filters and starts from the even- and odd-mode analysis of a pair of coupled-lines as described by [5]. In [5]’s analysis, the mutual coupling from nonadjacent lines is assumed to be small and is ignored to simplify the analysis. For membrane supported coupling structures, the electric field is spread out between all the fingers due to the low dielectric constant of the membrane ($\epsilon_r \simeq 1.0$), and therefore, a strong mutual coupling between the nonadjacent fingers is expected. This implies that the design equations obtained by [5] are not accurate for the membrane case. However, with the help of the microwave scale-modeling technique, the problem associated with the mutual coupling can be solved experimentally. A scaled microwave model with a scaling factor of 25.6 was built on a 0.91-cm-thick Plexiglas and a 152-μm-thick polyethylene sheet to simulate the behavior of a 15-GHz 3±0.7-dB six-fingers membrane Lange-coupler. The finger length in the scaled model is 12-cm long and is equal to a quarter-wavelength at 625 MHz. The finger width and gap are first calculated using the design equations developed by [5], and are adjusted experimentally until the correct coupling is achieved. Coaxial cables are used to feed the microwave model, and folded transition areas near to the coaxial feeding points are used to compensate for the dimensional change of the feeding structures. The final finger width and gap for the scaled model are 4.5 and 0.9 mm, respectively. The measured results of the scaled model show an amplitude balance of 3.4±0.6 dB over the 300–850-MHz frequency range with a phase balance of $9^\circ \pm 3^\circ$ from 100 MHz to 1.1 GHz.

Next, a 15-GHz (585 MHz × 25.6) Lange-coupler was fabricated on a 355- μm-thick high resistivity silicon wafer. Details of the fabrication process are presented in [1]. The finger length, width, and spacing of the 15-GHz design were translated directly from the scaled model, and are 4688 μm, 178 μm, and 36 μm, respectively. On-chip thin-film Titanium (Ti) resistors with resistance values of 55±15 Ω are used to terminate the micromachined Lange-coupler for measurement purposes. The fabricated micromachined membrane Lange-coupler is shown in Fig. 2. The measurement system is calibrated using the Short-Open-Load-Thru (SOLT) on the wafer calibration routine. The Lange-coupler is measured from 5 to 20 GHz and the measured results are shown in Fig. 3. The input return loss is 16 dB at 12 GHz which is due to the nonperfect thin-film resistors. The phase difference between the coupled port and direct port is $90 \pm 3^\circ$ from 5 to 20 GHz. The length of the input/output feed line is 3 mm (see Fig. 2) and contributes about 0.3-dB loss from each section [4]. The measured 4.2±0.8-dB coupling, therefore, includes 0.6-dB loss from the grounded coplanar waveguide (GCPW) feeding structures. The total ohmic and radiation loss of the coupler can be calculated to be around 0.5 dB. The coupling bandwidth covers the 6.5–20 GHz and is centered at 13.25 GHz. Compared to the scaled model, the center frequency shifts 1.7 GHz lower. One of the factors that cause this frequency shift are the folded transition areas. In the scaled model, the sidewalls of the Plexiglas sheet are covered by copper tape. In the micromachined circuit there is no metal coated at the etched sidewalls. This results in a higher effective dielectric constant in the micromachined coupler which, in turn, results in a longer electric length and lower center frequency. However, due to the wide-band characteristic of the Lange-coupler, this shift can be tolerated in most circuit applications. Future numerical modeling of the micromachined structure using FEM or FDTD techniques should solve this problem.

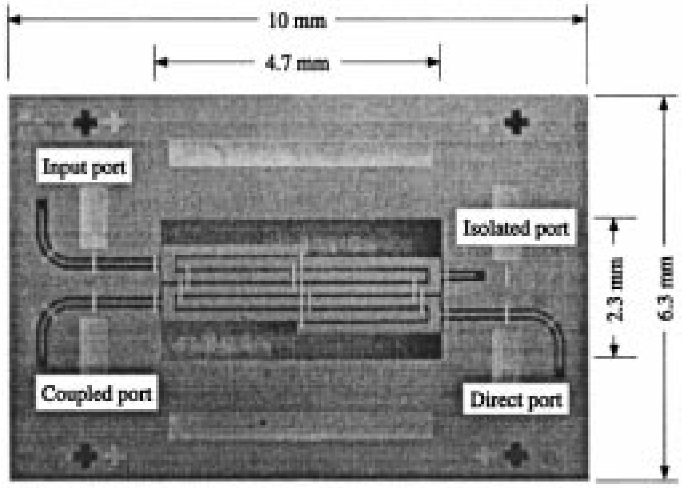

Fig. 2. Fabricated micromachined membrane Lange-coupler. Dimensions of the chip are 10 mm × 6.3 mm.

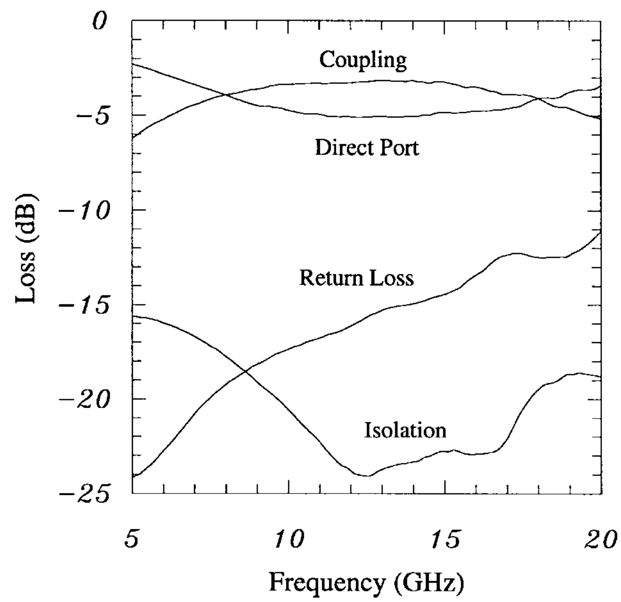

Fig. 3. Measured response of the micromachined six-fingers Lange-coupler.

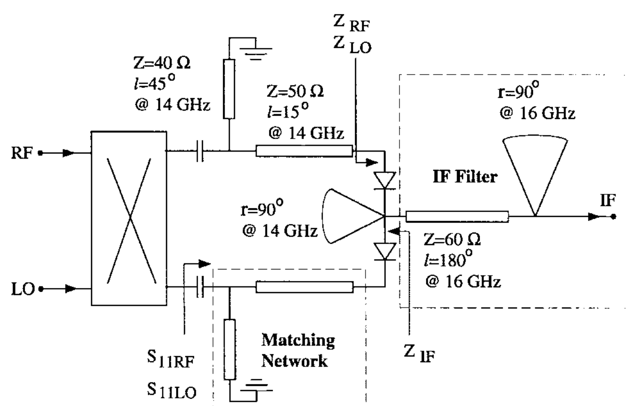

Fig. 4. Equivalent circuit of the balanced mixer.

III. Design of the SSB Mixer

A micromachined bandpass filter is placed in front of the balanced mixer to reject the image signal and achieve a SSB mixer. The filter, presented in [1], is a six-stage interdigital filter with a 10.3–16.8 GHz 3-dB passband response and a 0.7-dB port-to-port insertion loss at the center frequency of 13.5 GHz. To avoid the internal impedance mismatch, usually a circulator or an amplifier is added between the filter and mixer to isolate the two circuits and improve the impedance matching. However, in order to simplify the design, neither a circulator nor an amplifier is used in the work. A single-

TABLE I
SUMMARY OF THE MEASURED AND CALCULATED LOSS PERFORMANCE OF THE MICROMACHINED LANGE-COUPLER FROM 6.5 TO 20 GHz

Measured Coupling Ratio (dB)	GCPW Feed-line Loss (dB)	S_{11} Loss (dB)	Coupling Loss (dB) (Scaled model)	Extracted Ohmic Loss (dB)
4.2±0.8	0.6	0.15	3±0.8	0.5

balanced mixer is designed using the Lange-coupler outlined above. A Metelics MSS-30 254-B20 low barrier Tee-type silicon Schottky beam lead diode pair is used. The dc parameters of the diode are $R_s = 3\ \Omega$, $C_{j0} = 0.22$ pF, $C_p = 0.25$ pF, $I_s = 64$ nA, and $\phi = 0.41$ eV resulting in a figure-of-merit cutoff frequency of 240 GHz. Due to the physical layout of the diode chip, a series configuration is chosen in the single-balanced mixer design. The mixer and the associated matching networks are optimized using EESof Libra (Fig. 4 and Table II). The IF filter is achieved by an open quarter-wavelength radial stub design centered at 14 GHz at the diodes/IF connection. Another radial stub and a 180° transmission line both centered at 16 GHz are used to increase the bandwidth of the synthesized RF/LO short at the Tee-junction of the diodes. The radial-stub IF filter achieves an RF/LO rejection better than 30 dB at *Ku*-band. Since the membrane Lange-coupler is fed by GCPW structures, it is more convenient to use the GCPW structure for the rest of the circuit to avoid any other transition design. The physical dimensions of the GCPW transmission lines are calculated using the EESof Linecalc and the design of the GCPW radial stubs follows Simons' design method [6].

IV. MICROWAVE MEASUREMENTS

The SSB mixer is fabricated on a 355 μm-thick high-resistivity silicon wafer (Fig. 5). For dc testing purposes, the two shorted stubs which are designed for the dc return path are open in this fabricated chip and later are shorted to ground with silver epoxy. The total dimensions of the chip are 1.2 cm × 1.8 cm and can be reduced by 66% (or more) in future monolithic designs. The membrane filter is first intentionally broken to test the performance of the balanced mixer. In the measurement, the IF frequency is fixed at 100 MHz and 1 GHz, the RF frequency is swept from 12–18 GHz, the LO frequency is set at $f_{\rm RF} + f_{\rm IF}$, and the LO power is adjusted until the best conversion performance is achieved. The measured conversion loss versus RF frequency for the two IF frequencies is 8.2±1 dB from 12–18 GHz. The best port-to-port conversion loss is measured at 16.4 GHz with a 100 MHz IF, and is equal to 7.2 dB (Fig. 6). It is seen that only 1–2 mW (0–3 dBm) of LO power is needed for good conversion losses showing the low-loss aspect of the Lange-coupler. The port-to-port conversion loss includes a 0.5 dB loss from the Lange-coupler (see Table I) and around 1.2-dB loss from the GCPW transmission lines. (The distance from the RF feeding point to the Lange-coupler is 6.4 mm and the distance from the Lange-coupler to the diodes is 4.9 mm. The GCPW line has an insertion loss around 0.1 dB/mm at *Ku*-band resulting in a total transmission loss of 1.2 dB [4].) The isolation between the RF/LO to IF is better than −26 dB and the LO to RF isolation is better than −12 dB from 12–17 GHz. Due to the Lange-coupler used in the design, any LO mismatch at the diodes will result in an LO-reflected power at the RF port. The value of −12 dB is in agreement with the $S_{11\rm LO}$ values listed in Table II when line losses are included.

For the single-sideband mixer, a shielding cavity is first placed on top of the interdigital filter to result in a stripline structure. Any RF signal between 17 and 37 GHz will be rejected by the bandpass filter. The SSB mixer is first measured with a fixed IF frequency, the RF signal is swept from 12 GHz to 17 GHz, and the LO frequency

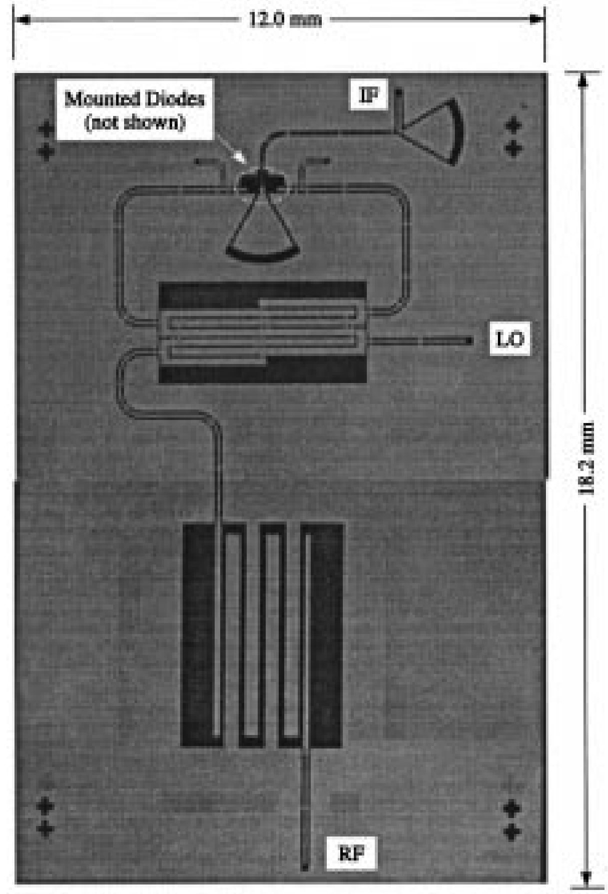

Fig. 5. The fabricated micromachined 18 mm× 12 mm SSB mixer. An extra shielding cavity (not shown in this picture) is required to provide the stripline environment for the interdigital filter.

TABLE II
CALCULATED DIODE IMPEDANCES AND CONVERSION LOSS. DEFINITION OF $Z_{\rm RF}, Z_{\rm LO}, Z_{\rm IF}, S_{11\rm RF}$, AND $S_{11\rm LO}$ IS SHOWN IN FIG. 4. IF FREQUENCY IS SET AT 1 GHz; LO FREQUENCY IS SET AT $f_{\rm RF} + f_{\rm IF}$ WITH A 5-mW PUMPING POWER. $Z_{\rm IF}$ IS THE EFFECTIVE PARALLEL IMPEDANCE FROM THE TWO DIODES, $Z_{\rm RF}$ AND $Z_{\rm LO}$ ARE THE INPUT IMPEDANCE OF A SINGLE DIODE

RF Freq. (GHz)	$Z_{\rm RF}$ (Ω)	$Z_{\rm LO}$ (Ω)	$Z_{\rm IF}$ (Ω)	$S_{11\rm RF}$ (dB)	$S_{11\rm LO}$ (dB)	Diode Conversion Loss (dB)
16	21-j30	17-j28	38-j4	-10	-7	5.0

is set at $f_{\rm RF} + f_{\rm IF}$. The conversion loss is around 1.5-dB higher than the single-balanced mixer. The 1.5-dB extra loss includes the 0.7-dB bandpass filter loss and about 0.8-dB mismatch loss between the bandpass filter and the single-balanced mixer. The SSB mixer is then tested as an image-rejection mixer. The LO frequency is fixed at 17 GHz and the RF frequency is swept from 15 to 19 GHz to result in a maximum of 2 GHz LSB (RF: 15–17 GHz) and USB (RF: 17–19 GHz) IF bandwidth. The measured conversion loss for the LSB and USB signals is shown in Fig. 7 and a 30-dB image rejection is achieved at 18 GHz and above, which translates to an IF frequency of 1 GHz and above. The port-to-port conversion loss (includes a total of 1.2-dB GCPW line-loss) of the LSB remains lower than 10 dB for the entire 2-GHz IF range.

V. CONCLUSION

This paper has demonstrated the capability of employing the membrane micromachined techniques in the design of an integrated-

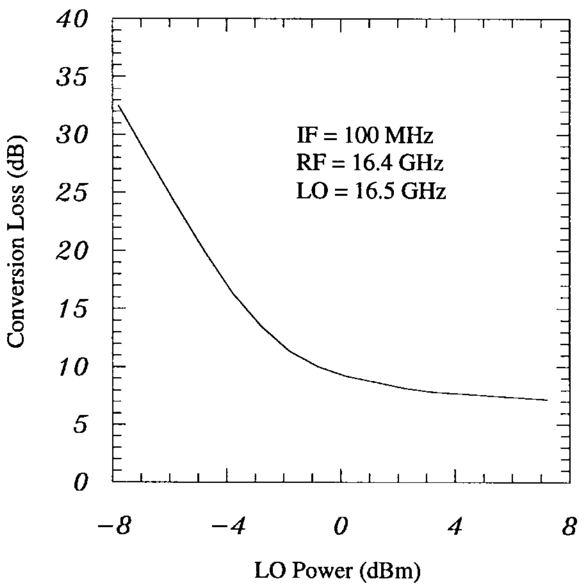

Fig. 6. Measured conversion loss versus LO pump power for the balanced mixer at 16.4 GHz.

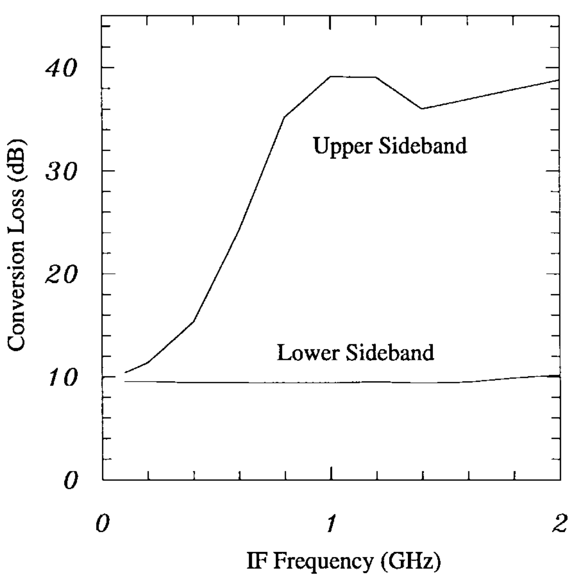

Fig. 7. Measured conversion loss versus the IF frequency for the single-sideband balanced mixer. The LO frequency is fixed at 17 GHz during the measurement.

circuit module. To our knowledge, the micromachined SSB mixer presents state-of-the-art conversion loss and image rejection. While this chip is physically larger than competitive GaA's FET-based MMIC's, it requires no dc power and only 1–2 mW of LO power. The membrane technology is fully compatible with the via-hole process in silicon, SiGe, GaA's and InP and can result in high performance IC's suitable for high volume productions for *Ka*, *V*, and *W*-band frequencies.

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A Method for Investigating a Class of Inhomogeneous Stripline Circulators

Giuseppe Macchiarella, Gian Guido Gentili, and Alberto Lobina

***Abstract*—Broad-band stripline circulators are studied by means of a mixed numerical technique which employs both Boundary Integral and Segmentation methods; this technique allows the analysis of planar circuits where the substrate is constituted by several regions with arbitrary shapes and different electrical properties.**

It is known that tracking circulators require matching structures because they present a low-gyrator impedance (real and almost constant in an octave frequency band). The matching structures (generally tapers or multisection's transmission line transformers) must be realized on a reciprocal substrate. The overall device (circular disk on ferrite substrate and matching structure on dielectric substrate) constitutes a planar circuit with an inhomogeneous medium.

The method of study presented here allows the determination of the overall impedance matrix of the planar circuit constituted by the nonreciprocal disk with sections of striplines connected to each port; in this way, the discontinuities between reciprocal and nonreciprocal medium are included in characterization of the overall device. Moreover the accuracy of the representation is increased. In fact, the coupling ports of the overall device may be located at a suitable distance from the disk boundary where higher-order modes excited by the discontinuities have been sufficiently attenuated and only the TEM mode is present on the striplines (which is the only one considered in the design of the matching structures).

***Index Terms*—Boundary elements, planar circulators, stripline devices.**

I. Introduction

Intrinsic wide-band stripline circulators were proposed in [1] and since that time this kind of device (often referred to as a tracking circulator) has been deeply investigated [2]–[5]. It is now well known

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