A Mm-Wave Gm-Assisted Transformer-Based Matching Network 2x2 Phased-Array Receiver for 5G Communication and Radar Systems

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Abstract—This paper describes a 50-58GHz 2x2 phased-array receiver (RX) for 5G communication and radar systems. The RX utilizes a Gm-assisted matching network (MN) to reduce the noise figure (NF) of a conventional passive mixer-first RX by placing an extra gain of ~5dB prior to the down-conversion mixers with minimal additional components. This prototype phased-array RX integrates a proposed Gm-assisted MN, mixer-first RX with translational feedback, Cartesian phase shifters as a baseband beamformer, poly-phase filters, and LO generation in TSMC 28nm CMOS, occupying a total area of 0.53mm². This RX achieves a NF of 7dB, gain of 26dB, input P1dB of -20dBm with an 8-GHz 3-dB BW and a S11 lower than -10dB over 22-GHz BW.

Keywords—5G mobile communication, CMOS receivers, millimeter-wave, phased-arrays, radar systems

I. INTRODUCTION

The pervasive nature of future 5th Generation (5G) wireless systems has motivated research to explore low-cost, ultra-broad bandwidth (BW), and low power wireless systems which exploit millimeter-wave (mm-Wave) frequencies. Although CMOS mm-Wave phased-array solutions hold potential to address concerns of cost and silicon size, a key challenge associated with mm-Wave receiver (RX) front-end modules (FEM) relates to the power inefficiency of producing gain at high carrier frequencies. Assuming the RX FEM gain is scaled to exploit the ADC’s full range, several points along the RX chain can potentially provide signal amplification, see Fig. 1.

An LNA is traditionally used for signal amplification prior to the down-conversion mixers to improve the RX noise performance [1], [2] (Fig. 1). However, in the mm-Wave bands (above ~30GHz), the gain at the carrier frequency, prior to the down-conversion mixers, becomes extremely power-inefficient due to the inherent gain roll-off. As shown in Fig. 1, the $G_{\text{max}}$ of a 10-um-wide NMOS transistor in 28nm technology with a current density of 125µA/µm, is only 10.5dB at 60GHz. By comparison, the same transistor with an identical bias current integrates a proposed Gm-assisted MN, mixer-first RX with translational feedback, Cartesian phase shifters as a baseband beamformer, poly-phase filters, and LO generation in TSMC 28nm CMOS, occupying a total area of 0.53mm². This RX achieves a NF of 7dB, gain of 26dB, input P1dB of -20dBm with an 8-GHz 3-dB BW and a S11 lower than -10dB over 22-GHz BW.

II. GM-ASSISTED MATCHING NETWORK

Matching networks (MN) can serve a multi-purpose role which includes converting a single-ended signal to differential and impedance matching. Thus, MNs are necessary, independent of the RX architectural choice. Fig. 2 (a) shows a single-element block diagram of a 2x2 phased-array RX with the proposed Gm-assisted MN. Similar to [3], this RX includes a translational feedback loop to significantly reduce the power consumption of the LO driver by using smaller switches, while maintaining a low mixer input impedance. However, in contrast...
to [3], this RX utilizes the on-chip MN not only to preserve the advantages of wideband input match of a passive mixer-first RX, as was done in [3], but also functions as an amplifier when combined with a feedforward \( G_m \) stage to deliver a modest amount of gain prior to the noisy down-conversion mixers.

The \( G_m \)-assisted MN uses an auxiliary \( G_m \) common-source amplifier inserted in parallel with a transformer-based MN to provide active gain through a feedforward path from the antenna port to the mixer input, as shown in Fig. 2 (a). With the feedforward amplifier, both the RX gain and noise performance can be improved. The MN is implemented using an inverting transformer. Both the transformer and common-source \( G_m \) have the same inverting polarity between the input and output, which is in contrast to traditional drain-to-gate transformer-based feedback amplifiers introduced in [5], see Fig. 2 (b). Moreover, there are significant differences between the two topologies which include: 1) the transformer used in the \( G_m \)-assisted MN has the opposite polarity as compared to [5], 2) the \( G_m \)-assisted MN has a weak amount of positive feedback with a loop gain of \(<10dB\), allowing a modest gain when the outputs drive a relatively low-input-impedance mixer and maintains a high-linearity. In contrast, [5] uses negative shunt-shunt feedback and can only achieve gain when loaded with a high-input-impedance mixer. A key advantage of this feedforward amplifier topology relates to the use of an already existing MN to resonate out the extra parasitic capacitance introduced by the feedforward amplifier without requiring an extra passive component.

In short, the transformer-based MN serves a dual purpose of providing both matching to the antenna impedance, and as a load to the feedforward \( G_m \) amplifier realized as a gain stage, while requiring negligible additional silicon area. Fig. 2 (c) compares the simulated gain from the antenna to the mixer input with the \( G_m \)-assisted MN and a conventional 4\(^{th}\)-order low-\( k \) transformer described in [2]. The \( G_m \)-assisted MN shows a ~5dB increase in gain while consuming an additional current of 7-mA for the \( G_m \) amplifier. Note that more gain can be added by further increasing the size and current of the \( G_m \) amplifier. However, this will also add more the capacitive load to both sides of the transformer and make it difficult to satisfy both high gain and wide BW.

III. CIRCUIT IMPLEMENTATION

The block diagram of the 2x2 phased-array RX is shown in Fig. 3. This work employs Cartesian phase shifters in the BB to perform current-mode signal combining. The mixers are driven by 4-phase IQ LOs which are generated locally using poly-phase filters (PPF). A single-ended LO is converted to a differential signal with a matching transformer, then routed to each of the 2x2 arrays using an H-tree LO distribution.

Fig. 2. (a) The proposed 2x2 RX with \( G_m \)-assisted matching network. (b) The \( G_m \)-assisted matching network. (c) Gain comparison between \( G_m \)-assisted matching network and conventional low-\( k \) transformers.

![Block-level diagram of the 50-58GHz 2x2 phased-array RX.](image)

Fig. 3. Block-level diagram of the 50-58GHz 2x2 phased-array RX.

![Detailed schematic of one receiver element.](image)

Fig. 4. Detailed schematic of one receiver element.

IV. EXPERIMENTAL RESULTS

This work was fabricated in TSMC 28nm HPC+ CMOS process with an area of 0.53mm\(^2\). The die photo is shown in Fig. 5. The 4 inputs of the RX elements are on the North and South of the die. The external LO comes in on the left, passing through the LO matching transformer before being distributed to the 2x2 phased array. The differential quadrature BB outputs exit on the chip's right side. This chip was measured using on-chip probing of all external I/O signals and DC bias. The active die area of one RX element is 0.049mm\(^2\) where the \( G_m \)-assisted stage
occupies an insignificant area of 0.0004mm². One RX element consumes 28.5mA from a 0.95-V supply.

![Die photo of mm-Wave 2x2 phased-array RX.](image)

Fig. 5. Die photo of mm-Wave 2x2 phased-array RX.

The single-element RX performance was measured using a standalone structure on the same chip, results are shown in Fig. 6. The RX achieves a 26-dB peak gain and a 7-dB NF with a 3-dB BW from 50-58GHz and $S_{11} < -10$dB from 56-78GHz. The input $P_{1dB}$ is -20dBm. The RX core consumes a total power of 15mW (excluding phase shifters and buffers).

![Measured results of one RX element.](image)

Fig. 6. Measured results of one RX element: (a) RX conversion gain, (b) cascaded NF, (c) input return loss ($S_{11}$), and (d) input $P_{1dB}$.

The setup for the 2x2 phased-array measurement is shown in Fig. 7 (a). An arbitrary waveform generator (AWG) card with 4-ch outputs was used to generate phase differences between the 4 RX inputs to emulate different delays associated with incident angles of an incoming signal. The 4 BB signals were then up-converted and passed through an external PA and a 4-way power divider. Two dual probes (GSGSG) were used to provide the 4 signals for each of the 4 RXs (Fig. 7). The RX LO was generated through an active X4 Frequency Multiplier. All signal generators were synchronized by sharing a common 10-MHz reference clock to ensure the frequency alignment.

To measure the beam pattern, the phase shifters of 4 RXs elements were first set to a steering angle with a fixed static phase error among all RX elements being compensated. The 4 input signals were provided by manipulating the phase difference among the 4-ch outputs of the AWG card with different incident angles from -90° to +90°. Fig. 7 (b) and (c) show two beam patterns with steering angles of 0° and 45°, respectively.

V. CONCLUSION

This paper presents the design and measurement of a 50-58GHz 2x2 phased-array RX in 28nm process which demonstrates the feasibility for compact, low-power, wide BW, and high-performance mm-Wave radios. Table 1 summarizes the performance of the RX and compare with prior-art RXs.

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<td>RX BW (GHz)</td>
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<td>Single NF (dB)</td>
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<td>Power (mW)</td>
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<td>0.42</td>
<td>0.6</td>
<td>0.39</td>
<td>0.085</td>
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Table 1. Comparison to Prior-art MM-Wave RX Publications

- Estimated from die photo, without pads.
- $f_t$ or $f_m$ BW.
- The power of buffer driving off-chip is not included.
- Calculated from reported IP3.
- Single element.
- Frequency range based on $S_{11}$ response.
- Including T/R SW loss.

REFERENCES