A compact H-band Power Amplifier with High Output Power

Ahmed S. H. Ahmed*#1, Utku Soylu#, Munkyoo Seo*, Miguel Urteaga§, and Mark J. W. Rodwell#

*Marki Microwave Inc., 215 Vineyard Court, Morgan Hill, CA 95037
#Department of Electrical and Computer Engineering, University of California, Santa Barbara, USA
§Teledyne Scientific Company, Thousand Oaks, CA 91360, USA

*a_s_ahmed@ucsb.edu

Abstract—We report a compact H-band power amplifier with high output power in 250nm InP HBT technology. Stacking and parallel power combining together provide the desired output power. Common-base stages with base capacitive degeneration act as stacked power cells. Four power cells are combined by a compact low-loss 4:1 transmission line network. At 270GHz, the four-stage amplifier has 16.8dBm saturated output power with 4% power-added efficiency (PAE). Over 266-285GHz, the amplifier's saturated output power is 14-16.7dBm with an associated 2.2-4%PAE. The 3-dB small-signal bandwidth extends from 233GHz to 281GHz with a peak gain of 20.5dB at 264GHz. Power amplifier's saturated output power is 14-16.7dBm with an associated 2.2-4%PAE. The 3-dB small-signal bandwidth extends from 233GHz to 281GHz with a peak gain of 20.5dB at 264GHz. Power amplifier's saturated output power is 14-16.7dBm with an associated 2.2-4%PAE. The 3-dB small-signal bandwidth extends from 233GHz to 281GHz with a peak gain of 20.5dB at 264GHz.

Keywords—H-band, millimeter wave, high-efficiency, stacking, parallel combining, compact combiner, 250-nm InP.

I. INTRODUCTION

Millimeter-wave communications permit high data rates due to the available wide spectrum. Power amplifiers are key components in the transmitter since they dominate the efficiency and limit the communication range. CMOS shows only -3.9dBm at 260GHz [1]. Advanced technologies with high power gain cut-off frequency (fmax) are necessary to generate moderate power at H-band frequencies [2]-[11].

Here we report a compact and high output power H-band power amplifier. The four-stage amplifier uses stacking and parallel power combining techniques. The key design features are 1) a compact and low loss 4:1 power combiner, 2) common base stages with base capacitive degeneration, and 3) driver scaling. The compact amplifier shows a broadband operation with a 48GHz small-signal bandwidth (BW) and 14-16.7dBm saturated output power from 266GHz to 281GHz.

II. AMPLIFIER DESIGN

A. Amplifier Block Diagram

The amplifier is fabricated in Teledyne 250nm InP HBT technology [12]. The amplifier (Fig. 1b) delivers the desired output power by series and parallel power combining. The common-base (CB) stages have significant RF voltage swings on the base nodes, hence design is similar to series-connected cells. Four power cells (Stage4 96-µm HBT periphery) are parallel-combined by a low-loss compact power combiner.

The required gain is achieved by cascading four stages. Conservative driver scaling, (Stage1 24-µm), (Stage2 48-µm), (Stage3 96-µm) HBT periphery, ensures proper amplifier saturation.

B. Unit Cell for Driver and Power Cells

We evaluated a common emitter (CE) (Fig. 2a), a grounded CB (Fig. 2b), and a capacitive base degeneration (Fig. 2c), i.e. with significant capacitive reactance presented to the base. The maximum base capacitance value is limited by the self-resonance frequency (SRF) due to the parasitic inductance associated with the capacitor while the minimum base capacitance is limited by the minimum acceptable gain. Lossless schematic large-signal simulations are performed for the same transistor size of (4·0.25µm·6µm) with the load impedance tuned for max PAE, and under the same bias conditions (1.8mA/µm and VCE=2.2V). Though all topologies have the same peak PAE (Fig. 2d), CE shows 13.1dBm of OP1dB with 13.6% PAE and 4.9dB associated gain while grounded CB shows 9.6dBm of OP1dB with 8.4% PAE and 10.8dB associated gain. The CB with 208F base capacitance shows OP1dB=13.7dBm with 16%PAE and 5.6dB associated gain. These results show that CB with finite base capacitance shows superior output power and PAE at 1dB gain compression which is consistent with the study in [13].

Fig. 3 shows the schematic diagram, used in drivers and power cells, of the capacitively degenerated CB stages with the matching and biasing elements. The base capacitance is implemented using MIM capacitors and symmetrically connected with a minimum parasitic inductance to the four-finger transistor (4·0.25µm·6µm). The simulated impedance presented to the base is ~j2.8Ω at 270GHz.
Fig. 2. Schematic diagram of: (a) CE; (b) CB with grounded base; (c) CB with 208fF base capacitor; (d) $P_{\text{out}}$, gain, and PAE for CE, grounded CB, and CB with a base capacitor at 270GHz.

It is difficult to accurately simulate the base inductance in multifinger mm-wave transistors. If the operating frequency approaches the capacitor SRF, the stage gain becomes sensitive to small modeling errors in the base inductance; further, strong gain peaking or even instability can arise. In the capacitively degenerated CB stage, the base capacitance is kept small so that the parasitic inductance is negligible and a simulated SRF of ~400GHz, well above the operating frequency.

The cell can be viewed as a stacked power cell [16], [17] since there is a significant AC voltage swing on the base capacitance. With proper load tuning, the total output voltage swing exceeds that at the transistor collector-base junction. As in a stacked stage, the delivered output power is greater than that of a grounded-base CB stage having the same load impedance, with the additional output power being contributed by the driver stage.

The transistor parasitic capacitance is tuned by shunt inductive transmission lines terminated by bypass capacitors. Interconnects are microstrip lines with the signals routed on the top metal layer (Met4) and the ground plane on Met1. This provides lower losses than inverted lines (signals on Met1, Met4 ground plane), but the necessary ground plane openings at transistor locations reduce EM modeling accuracy. After shunt inductive tuning, the required load impedance for each power cell is ~35Ω. The transistor input is matched to ~50Ω.

Given uncertainties in modeling stage gains at 270GHz, the drivers are conservatively scaled to ensure that the output stages can be driven into compression. The gain cells (Fig. 1b) are all identical, but while each cell in stages 1 and 2 drives two cells, each cell in stage 3 drives only one cell. To be compact, the splitters between stages 2 and 3 are implemented by series transmission line sections and distributed capacitors, while that between stages 1 and 2, with larger available space, uses $\lambda/4$ sections.

C. Proposed 4:1 Power Combiner

The desired output power is achieved by combining four power cells (schematic in Fig. 4a and chip micrograph in Fig. 4b). Each cell requires 35Ω load impedance. 50Ω transmission line sections with negligible electrical lengths combine two cells. Therefore, the required load impedance for the pair is 17.5Ω. The load impedance is 50Ω and by symmetry, it could be split into two parallel 100Ω resistors. A quarter-wave transmission line section with 29.5Ω characteristic impedance transforms the 100Ω load impedance (for half circuit) into the required 17.5Ω. The combiner (Fig. 4) has ~0.6dB simulated loss. Because signals pass through a single $\lambda/4$ section of 29.5Ω characteristic impedance, the combiner is compact and low loss, while the low-Z lines are wide hence can carry large DC and RF currents. The drawback of this combiner is smaller bandwidth than a 4:1 Wilkinson, because the impedance transformation occurs at a single step. Further details of similar transmission line combiners could be found in [13]-[15].

Fig. 3. Schematic diagram of the unit cell

Separate supply lines ($V_{\text{CC}}$ and $V_{\text{BB}}$) for the stage 4 and the remaining stages. With 140GHz designs of similar topology [13], radio-frequency oscillations have been observed if the supply is not adequately externally bypassed. Radio-frequency oscillations have been observed if the supply is not adequately externally bypassed. Addressing this, the present IC has many on-wafer supply bypass capacitors, all with series resistors to damp LC resonances. With this IC, no evidence of supply-induced oscillation has been observed. All passive elements, multi-finger transistors, and matching networks are simulated by Ansys HFSS, 3-D EM simulation, and compared to ADS momentum. The presented simulations use HFSS.

Fig. 4. proposed 4:1 power combiner: (a) schematic and (b) die photo.
The measurement is performed on the 3-mil thinned die without heatsink (Fig. 1a). S-parameters are measured using a Keysight network analyzer with 220-325GHz Oleson frequency extender modules and 325GHz GGB wafer probes. There are several calibration standards [18]. Given that the required precision in circuit measurements is relaxed compared to device modeling, a short-open-load-thru (SOLT) calibration standard is arbitrarily used. SOLT standard calibration on an external substrate moves the reference plane to the probe tips. Despite extensive EM simulations during design, a ~20GHz frequency shift is observed between the amplifier’s measured and simulated S-parameters (Fig. 5) biased at $(V_{CCPA}=2.2V, V_{BBPA}=2.3V, I_{CCPA}=172mA, I_{BBPA}=9.4mA, V_{CCdriver}=2.5V, V_{BBdriver}=2.1V, I_{CCdriver}=275.5mA,$ and $I_{BBdriver}=14.8mA$). The peak measured small-signal gain $(|S_{21}|)$ is 20.5dB at 264GHz and it is in good agreement with simulation. The measured 3dB-BW is 48GHz. The input reflection coefficient $(|S_{11}|)$ is better than -10dB from 261GHz to 291GHz. Though the output network is tuned for maximum saturated power, the output reflection coefficient $(|S_{22}|)$ is better than -5dB from 266GHz up to 320GHz.

In large-signal measurements, a 110-170GHz VDI extender drives a 270-290GHz AMC doubler from Virginia diodes. The doubler is followed by an H-band coupler. The coupled port goes to an H-band harmonic mixer and spectrum analyzer. The through port goes to the Erickson power meter in the calibration phase (Fig. 6a) and a 325GHz-GGB probe followed by the amplifier in the measurement phase (Fig. 6b).

![Fig. 5. Measured (solid) and simulated (dashed) S-parameters.](image)

**III. MEASUREMENT RESULTS**

The spectrum analyzer reading represents the output power from the coupler’s through port by applying the appropriate correction factor at each frequency. The correction factors are the power difference, in dB, between the spectrum analyzer readings and the PM4 in the calibration phase (Fig. 6a). The power is swept by adjusting the drive power to the 110-170GHz VDI extender.

The amplifier is biased by DC probes with bypass capacitors at $(V_{CCPA}=2.2V, V_{BBPA}=2.5V, I_{CCPA}=173.3mA, I_{BBPA}=10.8mA, V_{CCdriver}=2.4V, V_{BBdriver}=2.2V, I_{CCdriver}=274.8mA,$ and $I_{BBdriver}=16.2mA$). The last stage’s output voltage is coupled to the RF output, so the on-wafer voltage can be monitored through the probe’s bias T. At 270GHz frequency (Fig. 7), the amplifier has 16.8dBm, after calibrating the probe losses, saturated output power with 4% PAE and 10.2dB associated gain. We expect that the power could further increase with proper heatsinking. The amplifier demonstrates wideband operation. Fig. 8 shows the output power at different frequencies. Over 266GHz-285GHz, the output power is 14-16.8dBm with 2.2-4% PAE and 9.6-10.9dB associated gain (Fig. 9).

**IV. CONCLUSION**

The paper presented a record PAE and output power at the H-band frequency range. Over 266-285GHz, the amplifier demonstrates 14-16.8dBm saturated output with 2.2-4%PAE. The amplifier used stacking and parallel power combining techniques to achieve the desired power. The amplifier could be integrated into mm-wave transmitters or high-power signal sources. Table 1 shows the state-of-the-art high-efficiency H-band amplifiers. The present work demonstrates record power and $P_{sat}/mm^2$ with reasonable PAE at 270GHz.

![Fig. 7. Measured output power, PAE, and gain versus input power at 270GHz.](image)

![Fig. 8. Measured output power versus input power at various frequencies.](image)

Table 1. State-of-the-art H-band amplifiers.

<table>
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<tr>
<th>Ref</th>
<th>Freq, GHz</th>
<th>240</th>
<th>185-255</th>
<th>265</th>
<th>325</th>
<th>275-320</th>
<th>338</th>
<th>300-305</th>
<th>300</th>
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<th>301</th>
<th>280-328</th>
<th>266-285</th>
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<tr>
<td>Power, dBm</td>
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<td>20-23.9</td>
<td>17.2</td>
<td>11.3</td>
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<td>10</td>
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<td>7.8-10</td>
<td>13.5</td>
<td>9.6-13.7</td>
<td>14-16.8</td>
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<tr>
<td>Gain at P_{sat} (dB)</td>
<td>15</td>
<td>12.2-17</td>
<td>11.7</td>
<td>9.4</td>
<td>13.5-15</td>
<td>3.3</td>
<td>7.5-7.8</td>
<td>11</td>
<td>10-12</td>
<td>11.8</td>
<td>11.5-13.8</td>
<td>9.6-10.9</td>
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<tr>
<td>PAE at P_{sat}%</td>
<td>5</td>
<td>4.1</td>
<td>0.95</td>
<td>1.1</td>
<td>2.3^{d}</td>
<td>1.8</td>
<td>1.1</td>
<td>2.97</td>
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<td>2.2-4</td>
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</tr>
<tr>
<td>BW_{sat}, GHz</td>
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<td>53^{c}</td>
<td>9</td>
<td>~100^{c}</td>
<td>10</td>
<td>40</td>
<td>57</td>
<td>21</td>
<td>15^{c}</td>
<td>48^{c}</td>
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This work

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REFERENCES


