Coupled Embedding Networks for 7-dB Gain-per-Stage at 130–140 GHz in a 20-dBm Gallium Nitride Power Amplifier

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(Regular Paper) 
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This work was supported in part by the Semiconductor Research Corporation JUMP Program (ComSenTer), in part by DARPA DSSP (Tom Kazior), and DARPA MGM Program (Tim Hancock). 

ABSTRACT We propose a coupled embedding network (CEN) as a portable, robust design technique to simultaneously introduce gain boosting and loadline matching in millimeter-wave power (mm-wave) amplifiers (PAs). The approach uses directional couplers as an abstraction for the embedding network and produces straightforward design equations to achieve optimal input-stability margins for any specified gain. We demonstrate the design process and implementation using a 40-nm Gallium Nitride (GaN) process technology through simulated and measured performance of the PA with and without the embedding network. The output power is greater than 20 dBm and remains above 18 dBm over an 8 GHz bandwidth. Peak power added efficiency (PAE) is 6.5%. The 3-stage PA achieves 21 dB of gain at 137 GHz offering 7.1-dB gain per device. The measured gain is 8 dB higher than the common-emitter design with the same number of stages. To our knowledge, this is the highest gain per stage for a GaN PA in D-band. 

INDEX TERMS Embedding network, millimeter-wave, D-band, power amplifier, Gallium Nitride. 

I. INTRODUCTION Radar and communications in millimeter-wave bands above 100 GHz demand higher PA output power to compensate path loss and maintain practical range. A key challenge is how to realize the maximum output power in frequency bands that approach \( f_{\text{max}} \) with enough efficient gain. Furthermore, GaN HEMT modeling introduces uncertainty above bands (67 GHz) where direct model/hardware correlation is performed. This paper proposes a robust PA design technique based on passive embedding networks that are simultaneously optimized for gain and output power. 

GaN PAs above 100 GHz frequently feature low gain per stage [1], [2], [3]. Improving the gain per stage increases the areal power density and PAE by reducing power dissipated in oversized driver stages. Furthermore, allowing for embedding networks enlarges the design space to reduce passive component loss in matching networks. Approaches have been explored in recent CMOS work using cross-coupled neutralization for high PAE CMOS gain cells [4] and employing gain-plane techniques [5], [6], [7]. This earlier work requires matching networks to be designed after identifying the gain-boosting network as illustrated in Fig. 1. Internal loadline matching is not generally achieved under matched conditions, necessitating costly nonlinear power simulations across the embedding design space. 

In this work, a directional coupler is used to define the embedding network around a pre-matched transistor. The
The proposed coupled embedding network (CEN) is proposed as shown in Fig. 2 and uses the directional coupler to embed the amplifier. The CEN achieves gain boosting and also overcomes three shortcomings of earlier work. First, the CEN technique intrinsically accounts for the loadline condition. Second, the CEN technique decouples matching-network loss from the embedding network design resulting in simplified analysis. Third, the CEN technique can be broadly applied to any device technology where the wiring environment allows low-loss couplers. In Section II, the coupled embedding network design methodology is presented. In Section III the methodology is applied to design a 140 GHz PA in a 40-nm GaN HEMT process with an $f_{\text{max}}$ of 400 GHz. In Section IV, measurements are presented for the 2-stage core and 3-stage CEN PA to achieve an output power of 20 dBm with more than 7-dB gain per stage and compares this measurement with the state of the art.

**II. COUPLED EMBEDDING NETWORK METHODOLOGY**

In this section, the analysis of a coupled embedding network for arbitrary matching conditions is presented. A power cell should be terminated by a loadline impedance, $Z_L$, that delivers optimal power performance in some metric such as peak PAE or saturated output power. The input is typically gain matched. The corresponding target S-parameters for an input-matched power cell are

$$S_t = \begin{bmatrix} 0 & S_{12} \\ S_{21} & S_{22} \end{bmatrix}. \quad (1)$$

These S-parameters are achieved with matching networks to a 50-$\Omega$ port impedance or can be forced by choosing the output port-impedance convention to be the loadline impedance and the input port-impedance convention to be conjugate matched when port 2 is loadline terminated. The loadline condition implies no incident power wave on port 2, i.e., $\Gamma_L = 0$.

The design procedure is captured in Fig. 2 (top), where the power cell is first embedded in a unilateralization embedding network. Next, a second embedding network is applied for gain boosting. The following discussion explains how a directional coupler such as in Fig. 2 (middle) operates as the unilateralization network, the gain-boosting network, and a composite combination of the two networks. The two-step design approach introduces simpler design equations when compared with the equivalent one-step procedure that directly produces a composite result.

**A. COUPLER EMBEDDING PROPERTIES**

Fig. 2 (bottom) shows the signal flow graph for embedding the amplifier in a directional coupler with S-parameters, $S_c$. The coupling ratio is determined with a complex coupling coefficient, $\alpha$.

$$S_c = \begin{bmatrix} 0 & -\alpha^* & \sqrt{1 - |\alpha|^2} & 0 \\ -\alpha^* & 0 & 0 & \sqrt{1 - |\alpha|^2} \\ \sqrt{1 - |\alpha|^2} & 0 & 0 & \alpha \\ 0 & \alpha & 0 & 0 \end{bmatrix} \quad (2)$$

The complex coupling factor relates $S_{33} = S_{34}$ between the input and output of the power cell. For clarity the $S_{31} = S_{13}$ and $S_{42} = S_{24}$ terms are real and positive and the unitary property forces their magnitude to be $\sqrt{1 - |\alpha|^2}$. Choosing $S_{31} = S_{42}$ to be positive and real forces $S_{21} = S_{12} = -\alpha^*$ by the unitary property. The isolated-port pairs enforce $S_{41} = S_{14} = 0$ and $S_{32} = S_{23} = 0$.

Varying the phase of $S_{31}$ and $S_{42}$ only adjusts the phase relationships between the input, output, and internal ports. The amplitudes remain the same and depend only on $\alpha$. This is observed by repeating the following analysis, allowing phase shifts and applying the unitary property of lossless networks. Conceptually, the phase of $S_{31} = S_{13}$ and $S_{42} = S_{24}$ represent the effect of adding delay transmission lines on ports 1 and 2, and, therefore, do not require additional analysis.

From the flow graph in Fig. 2 (bottom), there is no power wave incident on port 2 of the embedded amplifier (a power wave exiting port 4 of the embedding network) when port 2 of the embedding network is terminated with its reference impedance. In (1), the absence of the incident power wave defines the loadline condition, $\Gamma_L = 0$ and results in optimal power performance.
B. UNILATERALIZATION

Using Mason’s gain formula to solve the signal flow graph [8], the S-parameters of the embedded PA, $S_e$, are

$$S_e = \begin{bmatrix} 0 & \frac{S_{12} - a_s^*}{1-a_s S_{12}} \\ \frac{S_{21} - a_s^*}{1-a_s S_{21}} & 0 \end{bmatrix}. \quad (3)$$

The unilateralization embedding network is described with a primary directional coupler $S_c^u$ with a coupling coefficient, $\alpha_u$. From (3), the unilateralization condition occurs when $\alpha_u = S_{12}^u$, which is valid if $S_{12} < 1$. The resulting unilateralized S-parameters are

$$S_u = \begin{bmatrix} 0 & \frac{S_{21} - S_{12}}{1-S_{12} S_{21}} \\ \frac{S_{21} - S_{12}}{1-S_{12} S_{21}} & 0 \end{bmatrix}. \quad (4)$$

As shown above, the unilateralization directional coupler forces $S_{u21} = 0$.

C. GAIN BOOSTING FOR MAXIMUM STABILITY CIRCLE RADIUS

Next, a second directional coupler, $S_c^b$, is applied to produce gain boosting with a separate coupling coefficient, $\alpha_b$. $\alpha_b$ is chosen to have the phase of $S_{u21}^b$ so that gain-plane position of the measure of reciprocity ($|S_{b12}/S_{b21}|$) is on the real axis. The magnitude of $\alpha_b$ is chosen to achieve the desired gain.

The S-parameters of the gain-boosted amplifier are

$$S_b = \begin{bmatrix} 0 & -a_b^* \\ \frac{S_{21} - a_b^*}{1-a_b S_{21}} & \frac{1}{1-a_b^* S_{21}} \end{bmatrix}. \quad (5)$$

For $|S_{u21}| \gg 1$, CEN PA gain is approximately

$$|S_{b21}| \approx \left| \frac{S_{u21}}{1 - \alpha_b S_{u21}} \right|. \quad (6)$$

While this expression indicates the existence of a solution with arbitrary gain, the choice of $\alpha_b$ will be comprised by stability considerations. The magnitude of gain $|S_{b21}|$ is

$$|S_{b21}|^2 = \frac{|S_{u21}|^2 + |\alpha_b|^2 - 2 |S_{u21}| |\alpha_b| \cos (\phi)}{1 + |S_{u21}|^2 |\alpha_b|^2 - 2 |S_{u21}| |\alpha_b| \cos (\phi)} \quad (7)$$

where $\phi = \phi_u + \phi_{b1}$ is the sum of the phase of the coupling coefficient and the phase of $S_{u21}$. By selecting the phase of $\alpha_b$, gain maxima can be found. The derivative with respect to the phase of $\phi_u$ is

$$\frac{d |S_{b21}|^2}{d \phi_u} = \frac{(1 - |S_{u21}|^2) \left(1 - |\alpha_b|^2\right) 2 |S_{u21}| |\alpha_b| \sin (\phi)}{|1 - S_{u21} \alpha_b|^4}. \quad (8)$$

The denominator is always greater than zero as long as $\alpha_b \neq 1/S_{u21}$ and $|S_{u21}| > 1$. The multiplier of $\sin(\phi)$ in the numerator will be negative. Consequently, maxima of (7) with respect to the phase of $\alpha_b$ are achieved when $\phi_u + \phi_{b1}$ sum to full rotations of 0° and minima are achieved when they sum to integer rotations of 180°. It is therefore shown that choosing $\alpha_b$ to have the phase of $S_{u21}$ results in the maximum gain for a given $|\alpha_b|$.

The extrema of $|S_{b21}|$ suggest three domains of interest for $\alpha_b$ as illustrated in Fig. 3 under the condition that $S_{u21} = 2$. In the top plot, the maxima of (7) are plotted as a function of $\alpha_b$. In the bottom plot, the minima of (7) are plotted.

In region 1, $|\alpha_b|$ ranges from 0 to 1 as the minimum value of $|S_{b21}|$ ranges from $|S_{u21}|$, e.g. 6 dB, to 0. Because minimum $|S_{b21}|$ decreases monotonically with increasing $|\alpha_b|$, the inverse represents the minimum value of $|\alpha_b|$ that can produce a given $|S_{b21}|$.

In region 2, $|\alpha_b|$ ranges from 0 to $1/|S_{u21}|$ and maximum $|S_{b21}|$ ranges from $|S_{u21}|$ to $\infty$. Because maximum $|S_{b21}|$ increases monotonically, the inverse represents the minimum $|\alpha_b|$ that can produce a given $|S_{b21}|$. In region 3, $|\alpha_b|$ ranges from $1/|S_{u21}|$ to 1 and maximum $|S_{b21}|$ ranges from $\infty$ to 1. Because maximum $|S_{b21}|$ decreases monotonically, the inverse represents the maximum $|\alpha_b|$ that can produce a given $|S_{b21}|$.

Choosing the phase of $\alpha_b$ to be the conjugate of the phase of $S_{u21}$ is optimal because it maximizes the radius of the input stability circles for any specified gain. Because the amplifier is input-matched, the input-stability circles are centered at the origin with a radius of

$$r_s = \frac{1}{|S_{b21}^*|} \approx \frac{1}{|\alpha_b S_{u21}|} - 1. \quad (9)$$

For fixed gain $|S_{b21}|$, the stability circle radius is maximized for the minimum $|\alpha_b|$ that can produce the desired gain. As shown above, the minimum $|\alpha_b|$ required occurs when $\alpha_b$ has the phase of $S_{u21}^*$ for the case of gain-boosting (region 2) and the phase of $-S_{u21}^*$ for the case of lossless degeneration. Because the stability circles are centered, CEN amplifiers remain stable into the designed termination impedances even at high gains. Choosing high gain with stability circle radii less than one is acceptable as long as the impedance presented to that port remains inside the stability circle.

D. COMPOSITE EMBEDDING NETWORKS

While the previous analysis describes the unilateralization and gain-boosting as two separate embedding networks, directional couplers form a group under the action of embedding. Consequently, the embedding of a directional coupler in a
FIGURE 4. Plot of coupled embedded 2-stage amplifier $S_{21}$ (dB) contours over complex coupling factor $\alpha_t$ with the real $S_{b12}/S_{b21}$ contour and conditionally-stable region. A cross-section along the optimal $\alpha_t$ values of the real $S_{b12}/S_{b21}$ contour would produce a plot similar to Fig. 3 and identical to the low $|S_{11}|$ limit of Fig 6. Note that $\alpha_t$ represents the entire proposed design space.

directional coupler is itself a directional coupler and can be consolidated into a single coupler with

$$\alpha_t = \frac{\alpha_u + \alpha_b}{1 + \alpha_u^* \alpha_b}. \quad (10)$$

The choice of coupling coefficient, $\alpha_t$ is applied to the HEMT device technology in the next section.

III. 140 GHZ AMPLIFIER DESIGN

The CEN PA is implemented in a 40-nm GaN HEMT technology ($f_{\text{max}} = 400$ GHz and $P_{\text{sat}} = 1.8$ W/mm) and illustrated in Fig. 9. The PA consists of a buffer stage and core PA that is embedded in a 12-dB directional coupler ($|\alpha| = 0.063$).

PDK HEMT models are based on extracted devices in 2x25, 4x37.5, and 6x50um width. The larger 6x50um HEMT would produce the highest power but was not selected because it has lower gain and less modeling accuracy at 140 GHz. A 4x37.5um HEMT was selected for the core power cell and a 2x25um device was chosen for the buffer. The 4x37.5um HEMT have previously been reported to reach 20.3 dBm of output power at 28 GHz [9]. The buffer ensures relatively constant impedance to the coupler input and improves the stability margin. At 140 GHz, the buffer can only present output reflection coefficients with magnitude less than 0.25.

A. GAIN-STABILITY TRADEOFFS AND LOADPULL BEHAVIOR

The gain-stability tradeoff is illustrated in Fig. 4. The 2-stage amplifier gain in the absence of embedding (8.3 dB) is located at the origin. The contour representing real $S_{b12}/S_{b21}$ is nearly straight because $S_{112}$ is very small. As the value of the $\alpha$ increases along the real $S_{b12}/S_{b21}$ contour, gain increases to infinity and then decreases back to zero at the unit circle. Fig. 4 also identifies Singhakowinta’s maximum gain of 14 dB at the intersection of the stability region and the real $S_{b12}/S_{b21}$ contour [5].

The analysis in Section II expects identically zero transistor $S_{11}$, coupler return loss, and coupler isolation. To check the consequences of non-ideal parameters in a practical circuit we show the effects of nonzero $S_{11}$ for the 8.3 dB amplifier in Fig. 6. The change to $S_{11}$ was achieved by adding lossless reciprocal input matching network to the transistor input with $S_{21}$ chosen to have zero phase shift. Each vertical cross-section uses the same coupled embedding. Fig 6 shows that nonzero $S_{11}$ results in reduced gain and degraded stability margin.

The simulated loadpull is plotted in Fig. 5 for the 4x37.5 um HEMT, core PA, and core PA embedded in an ideal 6.6 dB coupler. The large signal gains are 3.6 and 5.7 dB, respectively, for the core PA and CEN PA. The input stability
circle radius is 0.75. Fig. 5 demonstrates the coupler preserves the optimal loadline match to the core PA since the power contours remain centered at the origin under embedding. The output power of the CEN PA is lower under lossless embedding because a fraction of the core PA output power at port 4 couples back to its input at port 3. For a 6.6 dB coupler, theory would predict a $10 \log_{10} |1 - 10^{-0.66}| = -1.07$ dB degradation in output power. Interestingly, the embedded PA (7.25 dB gain per stage) retains good tolerance to load impedance variation.

While a single 4x37.5 um device has the potential to generate 5.6 dB of gain per stage, the gain drops to 3.6 dB as the PA compresses at peak efficiency. Consequently, a PA capable of 17 dB of compressed gain might require as many as 5 stages while the CEN PA of Fig. 5 requires only 3.

**B. COUPLER DESIGN**

While the theoretical result suggests a 6.6-dB coupler, a 12-dB coupling ratio was selected as a conservative choice in case the HEMTs yielded higher than modeled gain, a possibility indicated by our previous measurements of the process. The 12-dB coupler is a quarter-wave coupled line with a phase shift line added to set the phase of $\alpha$ as described in Section II.

The design equations in Section II require that the embedding network have ideal directional-coupler characteristics. Consequently, achieving good isolation and return loss is critical for design success. Poor isolation would feed power from the core-amp output to the circuit input, possibly resulting in negative input impedances and sacrificing unconditional stability.

Coupled-line microstrip directional couplers have poor isolation when the even and odd modes have different phase-velocities. The high dielectric contrast between SiC and air exacerbates the problem. Designers often employ lumped capacitors at the ends of the coupled line sections in order to compensate the phase-velocity difference [10].

Inspired by these approaches, this work employs two different line spacings in order synthesize the effect of an end-capacitor and maximize the isolation. The two line widths were optimized over AXIEM EM simulations using the AWR simplex algorithm. The objective function attempted to minimize return loss and isolation in-band, while setting coupling equal to 12 dB. The final coupler layout is shown in Fig. 7. The optimized S-parameters are shown in Fig. 8. The insertion loss of the coupler is 0.5 dB at 140 GHz. The 3dB-coupling bandwidth is 113 GHz.

The delay lines used to set the feedback-phase employ aggressively mitred corners. The mitres were found to preserve directional coupler properties better than fixed radius bends.

Amplifier bandwidth is limited by the phase response of coupler’s added feed lines. The feed lines of the coupler are a full-wavelength longer than necessary to achieve the desired phase of $\alpha$. The additional phase allows the coupler to have the length necessary to connect with both sides of the core-amplifier. The coupler with feed-lines has $S_{13}$ phase roll-off of 4.9 degrees per GHz while the core coupler as shown in Fig. 7
has phase roll-off of 1.6 degrees per GHz. A future design might achieve wider bandwidth and lower loss by folding the amplifier to bring the input closer to the output and remove a wavelength of feed line.

IV. MEASUREMENT RESULTS

The die microphotograph is plotted in Fig. 10. The chip occupies 1.2 mm by 2.1 mm. Significant area is dedicated to on-chip dc bypass networks that target stabilization of the PA below the band. The buffer PA occupies 430um X 520um and the core PA occupies 740 um X 470um.

The measurements use a Keysight N5247 A PNA-X with VDI WR6.5 network extender calibrated with a PM5B power meter. The probe losses are de-embedded using on-chip TRL calibration standards in conjunction with shim/short waveguide calibration.

A. S-PARAMETERS

To demonstrate the comparative performance of the CEN PA relative to the core PA, the S-parameters of the core PA were measured and indicate that the 2-stage PA reaches 11 dB of gain (or 5.5 dB per stage) over a 3-dB bandwidth of 10 GHz, reaching significantly higher gain than was expected. Core-PA S-Parameters are shown in Fig. 11. The measured two-stage amplifier differs from the embedded two-stage amplifier in that the embedded variant includes a 90° turn in the input and output matching networks. Furthermore, the measured two-stage was on a different wafer from a different processing run, meaning some differences in performance should be expected.

The S-parameters for the CEN PA are plotted in Fig. 12. To highlight the importance of development of device models for operation at D-band, we plot the S-parameters with respect to the PDK device model (top) and S-parameters with respect to small-signal device measurements at D-band (bottom). Notably, the accuracy of the gain and return loss between the
B. STABILITY VERIFICATION

The $\mu$ factors of stability are plotted in Fig. 13 to determine the minimum distance from the Smith chart origin to the unstable region of the load and source stability circles. The minimum measured $\mu = 0.55 = -5.2$ dB means a return loss greater than 5.2 dB on either port would guarantee stability. The PA is highly stable under all bias-conditions due to the multistage bias networks for lower frequency stability.

The validity of the Rollett condition [11], assessed here via $\mu$, requires that the open loop transfer functions have no poles in the right half-plane (RHP) [12]. Prior work on gain-boosting [4], [5], [6], [7] tacitly assumes that the Rollett criterion is sufficient to demonstrate the stability of a 2-port, whose transfer functions have no poles in the RHP, embedded in a passive reciprocal 4-port. We find the validity of that approach unclear and therefore must assess the open loop stability of the system.

The transfer functions of the core amp embedded in a the designed coupler can be calculated via Mason’s gain formula. Both the transfer function numerators and the network determinant in the denominator only have poles found in one of the original systems. By design the 2-stage core amp has no poles in the RHP. By passivity the coupler also has no poles in the RHP. All new poles in the connected system must come from RHP zeros of the determinant $\Delta$ in the denominator. Because the determinant can have no RHP poles, all clockwise encirclements of the origin by the determinant’s Nyquist plot reflect RHP zeros and therefore poles of the embedded system. Take $S_c$ and $S_t$ to be the S-parameters of the coupler and core-amp respectively using the numbering scheme from Fig. 2. The network determinant is

$$\Delta = 1 - \Delta_1 + \Delta_2,$$

where

$$\Delta_1 = S_{c33}S_{t11} + S_{c44}S_{t22} + S_{c34}S_{t21}$$

$$+ S_{c43}S_{t12} + S_{c43}S_{t22}S_{c34}S_{t11}$$

$$\Delta_2 = S_{c33}S_{t11}S_{c44}S_{t22} + S_{c43}S_{t12}S_{c34}S_{t21}$$

C. POWER

The large signal compression characteristics are plotted in Fig. 15. At the peak gain (137 GHz), the CEN gain boosting effects are strong but compression is observed at relatively low output power. The saturated output power is 18.5 dBm. At 133 GHz, the gain is reduced from the peak value but the saturated power exceeds 20 dBm. In the 133 GHz measurement the available source power limited our ability to identify $P_{sat}$ and peak PAE. In both cases, the peak efficiency shows similar behavior. The peak PAE at 137 GHz is 4.7% while the peak PAE for 133 GHz is 6.5%. At all measured frequencies below
TABLE 1. Comparison With Prior D-Band GaN PAs

| Ref. | Technology     | Freq. (GHz) | $|S_{21}|$ (dB) | $|S_{21}|$ (dB) per stage | OP1dB (dBm) | $P_{sat}$ (dBm) | Peak PAE (%) | Die Area (mm²) |
|------|----------------|-------------|----------------|--------------------------|-------------|----------------|--------------|----------------|
| [1]  | 100nm GaN HEMT | 120         | 34 4.3 (8.5/cascade) | 23                        | 26.4        | 16.5*         |              | 7.5            |
| [3]  | 150nm GaN HEMT | 140         | 12 4 11                | 14                        | 18.2        | 1.7           |              | —              |
| [2]  | 40nm GaN HEMT  | 180         | 4.5 4.5                | 11                        | 13.8        | 3.5           |              | —              |
| [1]  | 100nm GaN HEMT | 20           | 13 3.3                  | —                         | 15.8        | 2.4*          |              | 2.47           |

*Peak PAE not achieved due to limited generator power

133.6 GHz, input power was insufficient to achieve maximum PAE.

The fact that peak power performance is achieved at 133 GHz while peak gain performance is found at 137 GHz is a tuning error due to the design’s sensitivity to device modeling. Fig. 12 demonstrates that amplifier performance was significantly changed due to the difference between the devices in the PDK and the devices as measured. Furthermore the core-amp was measured to have peak gain at 133 GHz as shown in Fig. 11. It seems that model errors downshifted the peak power performance frequency from 140 GHz to 133 GHz, but only had a minor impact on the effective frequency of the passive feedback loop.

To demonstrate the trade-offs between the peak gain, output power, and PAE, these measurements are shown as a function of frequency in Fig. 16. The output power remains greater than 18 dBm from 131 to 139 GHz while the gain increases from 10 to 21 dB. Compared to the cascade of the buffer and measurement of the two-stage core PA, the CEN boosts the gain at 137 GHz by more than 8 dB or 4 dB per stage.

A comparison with prior work is shown in Table 1. All OP1dB figures for prior work were obtained from compression curve plots. Power density and peak PAE are comparable to the state of the art. Gain per transistor stage of 7 dB is the highest reported for GaN PAs in D-band.

V. CONCLUSION

We present a new gain boosting technique based on coupled embedding networks that also achieves optimal loadline matching in a PA. We describe an analytical method for designing the embedding network while controlling amplifier stability. Measured results of the PA implemented in a 40-nm GaN HEMT process indicates that the embedded PA realizes 8 db more gain than a similar common-emitter approach while achieving an output power of 18 dBm across the frequency band. Future work will investigate the application to efficiency optimization. Other important future work for passive embedding networks includes refining the stability conditions and quantifying the gain-bandwidth trade offs.

ACKNOWLEDGMENT

The authors would like to thank Matt King and Florian Herraault at HRL for their engaged support.

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