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Performance Analysis of a Low-Noise, Highly Linear Distributed Amplifier in 500-nm InP/InGaAs DHBT Technology

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Abstract—This article is an extension of the previous report on an ultrawideband distributed amplifier (DA) in the InP double heterojunction bipolar transistor (DHBT) technology. With the choice of a tricode unit cell, the Miller capacitance is reduced and a larger bandwidth (BW) is obtained. Measured S-parameters show a 3-dB BW of near dc to >170 GHz with a gain of 12 dB. In addition, the circuit operates in low dc bias conditions, thus yielding a low measured noise figure (NF) of 8 dB over the frequency range of 5–50 GHz and 12 dB at 135 GHz. Within the 5–50-GHz frequency range, the NF is only 0.5 dB higher than the NF of a similar DA with cascode unit cell. When the circuit is measured for large-signal condition, it reaches a 1-dB output compression point, $P_{1\text{dB}}$, of 8.4 dBm at 150 GHz, a saturated output power of approximately 10 dBm, and an associated maximum power-added efficiency (PAE) of 6%. This is the best linearity as well as the highest saturated output power and PAE reported at this frequency for DAs. When measured at 20 GHz, the circuit shows a $P_{1\text{dB}}$ of 9 dBm and output referred third-order intercept point of 22 dBm, a saturated output power of 11 dBm, and an associated maximum PAE of 6%, thereby proving its uniform large-signal character within the full band. In short, this circuit brings in the unique combination of low NF, highly linear characteristics, and high PAE simultaneously together over large BW. Furthermore, the circuit consumes 180 mW of dc power only.

Index Terms—Distributed amplifier (DA), InP double heterojunction bipolar transistor (DHBT), monolithic microwave integrated circuit (MMIC), traveling-wave amplifier (TWA), ultrawideband low-noise amplifier.

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I. INTRODUCTION

RECENT trends in high data rate in optical communications, broadband spectroscopic, and broadband measurement systems are pushing the limits of the current state of the art in broadband amplifiers. Highly uniform ultrawideband operation, low deviation from linear phase, high output dynamic range, and high efficiency are the key requirement in these applications. For receiver applications within the aforementioned systems, low-noise characteristic is additionally required over the large bandwidth (BW). Usual low-noise topologies require low bias condition, which is contradictory to achieving high output dynamic range. Therefore, an ultrawideband, highly linear system with low-noise characteristics is a challenge to achieve both from circuit topology and choice of transistor technology.

A useful first approach is to choose a circuit topology that offers gain within higher than an octave BW. As such, distributed amplification has been a promising architecture in monolithic microwave integrated circuits (MMICs), which was first proposed in GaAs MESFET technology [1]. Several reports have demonstrated the implementation of this architecture in various technologies: InP HEMT [2], InP double heterojunction bipolar transistor (DHBT) [3], CMOS [4], and SiGe HBT [5]–[7]. With increased scaling possibilities, amplifiers up to 235-GHz BWs have been presented [2], [3], [8], [9], but with low output power levels. In order to achieve the state-of-the-art small- and large-signal properties, a 500-nm InP DHBT with f_{max} of 490 GHz can give high-dynamic-range possibilities. Using these devices as subcells in a distributed architecture, ultrawideband, large dynamic range, and low-noise amplifiers can be designed [10].

This article reports an ultrawideband distributed amplifier (DA) that addresses this demand. This article offers a large BW of >170 GHz, low deviation from linear phase of only $\pm 10^\circ$ up to 165 GHz, and a noise figure (NF) of 8 dB when measured between 5 and 50 GHz. In terms of large-signal characteristics, the circuit has a measured output power with 1-dB compression point $P_{1\text{dB}} = 8.4$ dBm at 150 GHz, a saturated output power $P_{\text{sat}} = 10$ dBm, and an associated maximum power-added efficiency (PAE) of 6% at 150 GHz. These are the highest $P_{1\text{dB}}$, P_{sat} , and PAE values

reported at these frequencies for amplifiers with BWs larger than 110 GHz. The high value of $P_{1\text{ dB}}$ and PAE is measured at the high end of the BW that indicates similar behavior for the whole frequency range. In terms of noise characteristics, it exhibits among the lowest NF among DAs with BW greater than 110 GHz.

II. TECHNOLOGY

The circuit presented in this article is based on the Ferdinand-Braun-Institut (FBH) transferred-substrate InP-DHBT technology. The cross-sectional schematic of the layer stack is shown in Fig. 1.

The large-signal model structures for the transferred-substrate InP/InGaAs DHBTs are shown in Fig. 2. It consists of the FBH HBT model [13] embedded in an external parasitic network associated with via transitions and device electrodes. The external parasitic network is extracted using a dedicated 3-D electromagnetic (EM) simulation approach, as described in [14]. The 3-D EM simulation approach allows accurate modeling of the distribution along the device access structure and skin effect due to field penetration into the conductor. In particular, frequency-dependent impedances

where f is the frequency, are included in series with each device terminal. Furthermore, thin-film microstrip (TFMS) lines are added to model the transition to the external interconnect environment. The parasitic network modeling approach has been verified experimentally on a small-signal model up to 220 GHz [14]. The device in a transferred-substrate technology lends itself well to a two-terminal configuration with the emitter grounded. For added flexibility, a device in a three-terminal configuration is also available. This device requires an opening of the ground plane underneath the HBT and, thus, a modified device layout. Therefore, the large-signal model structure is modified as shown in Fig. 2 to be able to capture the effect of the modified device layout.

The FBH HBT model parameters are determined from multibias S-parameters and static I - V characteristics measured over temperature. The S-parameter data have been corrected using on-wafer multilines TRL calibration setting the reference plane at the boundary of the device layout. Details in the FBH HBT model extraction procedure for transferred-substrate InP/InGaAs DHBTs can be found in [13].

Fig. 3. Measured (dashed lines) and modeled (lines with symbols) (a) forward output characteristics at temperature 17 °C, (b) unity current gain frequency versus collector current, and (c) base-collector capacitance versus collector current.

Fig. 3 shows the measured and modeled characteristics for a $0.5 \times 6 \mu\text{m}^2$ transferred-substrate InP/InGaAs DHBT in two-terminal common-emitter configuration. The measured and modeled forward I - V characteristics at 17 °C are shown in Fig. 3(a). The model fitting at other temperatures is similar. It is well known that HBTs in the InP/InGaAs material system are prone to current blocking at the heterointerface in the collector leading to a softening of the knee region, drop in current gain, and increase in the charge storage. The FBH HBT model has no dedicated parameters to model this effect, and hence, the fitting of the softening of the knee region is only approximated, as shown in Fig. 3(a). Fig. 3(b) and (c) shows the measured and modeled unity current gain frequency, f_t , and base-collector capacitance, C_{bc} , versus collector current with collector-emitter voltage as a parameter. In the FBH HBT model, a unified collector charge description is responsible for modeling both. Therefore, f_t and C_{bc} must be fit simultaneously during the parameter extraction process. In general, the model fitting shown in Fig. 3(b) and (c) is excellent, with the exception of the lowest collector-emitter voltages where current blocking is noticeable. The delicate bias dependence of f_t and C_{bc} is important for the accurate prediction of the maximum available gain/maximum stable gain (MAG/MSG) for the InP/InGaAs DHBT, which would not be possible using models dedicated to silicon-based bipolar transistors.

As a final verification of the developed transferred-substrate InP/InGaAs DHBT model, the large-signal performance measured at 94 GHz is considered. Fig. 4(a) and (b) shows the measured and modeled large-signal performance under class-A operation for devices in two- and three-terminal configuration (with an externally grounded emitter), respectively. In the large-signal model for the three-terminal configuration, only the external parasitic network is changed. The model prediction of the measured large-signal performances remains

Fig. 4. Measured and modeled large-signal performance at 94 GHz for (a) two-terminal device and (b) three-terminal device (with externally grounded emitter).

Fig. 5. (a) Measured (symbols) versus simulated (solid lines) resistance values versus force current. Increase in resistance indicates self-heating of NiCr TFRs. (b) Simulated S_{21} versus input power level for a 20-Ω and a 30-Ω NiCr resistance.

accurate well beyond the 1-dB compression point and confirms the proposed modeling strategy.

C. Frequency and Power Handling Characteristics of the Passives

Thermal stability due to high RF and dc currents in the passive elements of the circuit plays a pivotal role in the functionality of any THz MMIC. The current technology offers SiN-based MIMs and NiCr TFRs for the realization of passive networks. The MIM capacitors were designed to withstand high operating voltage way above the transistor rating, with dielectric breakdown fields reaching 1.5 MV/cm.

On the other hand, the thermal stability of the NiCr resistors is of utmost importance to ensure long-term operation. To investigate this, a dc current injection measurement technique was employed. In this test measurement, a force current was varied and the NiCr TFR resistors were measured using a four-point contact method on a process control monitor (PCM) test structure. The output resistance was plotted versus the force current (for two nominal resistor values of 40 and 60 Ω) shown in Fig. 5(a). The results show a slight increase in the resistance value with an increase in the force current. This is an indication of the self-heating phenomena taking place in the TFR. Nevertheless, the values of the resistors show a maximum increase of 4.5% over the complete current range.

The 2-D electrothermal simulations were performed to estimate the effect on the RF performance [15]. The temperature dependence of the NiCr strip was modeled according to [16]. The resulting simulated resistance agrees with the measured values. A temperature rise of approximately 35 °C at 1 mW of input power was predicted by the simulations.

Fig. 6. Simulation of MAG and MSG for CE, cascode, and tricode subcells.

Since the upper bound of frequency is beyond 150 GHz, S-parameters were simulated for the NiCr resistors versus input power to ensure the stability of the resistor from the design point of view. Considering the self-heating effects of the TFRs, small-signal parameter, S_{21} , was simulated using TCAD for the resistor values of 20 and 30 Ω in Fig. 5(b). These two values were picked to match the similar ones used in the design later on. From Fig. 5(b), it is evident that the resistors behave ideally up to 10 dBm of transmitted RF power. This is reasonable enough to be used on the RF path of the circuit since the maximum achievable RF output power is within this limit.

III. CIRCUIT ANALYSIS

A. Subcell Selection for Distributed Amplifier

Since distributed topology is a potential candidate for operating beyond 100% of the fractional BW, the next step in the design is to choose appropriate subcell that satisfies the challenges stated in Section I. A few topologies exist as reported in the literature: common emitter (CE), cascode, stacked tricode, and buffered cascode. Each of the different subcells varies in terms of wideband property. A comparison in MAG and MSG is shown in Fig. 6.

The CE has the least gain performance among the three competitors with the tricode subcell clearly showing its advantage in terms of both MAG and MSG. Although the gain in BW is marginally higher than for the cascode, it has higher gain. This is clearly advantageous in the design of an ultrawideband DA because it directly contributes to the maximum gain-BW product of the DA.

The tricode structure can provide the best performance among the three also in terms of RF output power handling. A simplified tricode structure and its small-signal equivalent circuit are shown in Fig. 7. Since a classic cascode or tricode cell has the base directly connected to ground, an ac signal imbalance, namely, the respective V_{ce} , I_b , and I_c , occurs between the T_{CE} transistor and the upper transistors, T_{CB1} and T_{CB2} . To enhance equal voltage swing across all the three transistors, a base-ground capacitor is added to each of T_{CB1} and T_{CB2} . As a result, after a small optimization, the total voltage swing of all the three transistors can be added to provide three times the output swing. Therefore, a tricode transistor outperforms its competitors in terms of both wideband property and high RF power handling capability.

Fig. 7. Small-signal model of a tricode structure optimized with inductive peaking.

Fig. 8. Comparison of NF for a CE, cascode, and tricode subcell.

The choice of using CE, cascode, and tricode subcell can also be analyzed in terms of their respective NF versus frequency characteristics. Fig. 8 shows the NF of CE, cascode, and tricode cell, at the same dc bias conditions, i.e., the same I_b , I_c , and V_c across each of the identical transistors. It is seen that NF increases when choosing tricode over a cascode or a cascode over a CE configuration. Thus, the disadvantage of using tricode is the highest NF among the three, particularly at higher frequencies of operation. The NF of cascode is higher than a CE transistor but lower than that of a tricode. Hence, it is beneficial to use a cascode over a tricode for low noise at high frequencies.

From the detailed analysis of the subcell, it is evident that the tricode offers the widest BW and highly linear performance at an expense of NF. With a choice of a 500-nm technology node, a tricode is, therefore, the optimum choice to design a DA with a BW of one-third of f_{max} of the transistor. To avoid its adverse effect on NF, the dc bias is kept low, however, just high enough to obtain a good linearity performance.

After the choice of the tricode cell, an appropriate broadband topology needs to be selected. Since the uniformly DA topology provides ultrawideband, low deviation from group delay, highly linear characteristics, and moderately low NF, this topology is applied for the amplifier design.

Fig. 9. Simplified schematic of the DA (detailed dc bias omitted).

Fig. 10. Chip photograph of the tricode DA (0.65 × 1.5 mm²).

B. Distributed Amplifier Design With Tricode Subcell

The simplified circuit schematic is given in Fig. 9. The circuit topology is based on a five-cell DA. Each identical cell consists of three $0.5 \times 6 \mu\text{m}^2$ transistors arranged in the tricode configuration.

At the initial design phase, the tricode cell was optimized for maximum BW, maximum uniform gain, highest achievable linearity, and output power over the whole frequency band. To achieve this, an emitter degeneration resistance of 6.5Ω and a capacitance of 200 fF were added to the bottom transistor T_{ce} , as shown in Fig. 9. Between T_{ce} and T_{cb1} , and T_{cb1} and T_{cb2} , there is an inductive transmission line, which improves the flatness of the gain characteristics at the higher end of the frequency range. Another inductive line L_{cb} improves the output matching at higher frequencies.

In a second step, the entire DA is designed using the optimized cell. Since the input and output impedances of the cell are different, lines $L_a = 210 \mu\text{m}$ and $L_b = 180 \mu\text{m}$ are chosen to be of different lengths and widths, with a targeted characteristic impedance of 50Ω and 62Ω , respectively. The termination resistances, $R_{term1} = 31.25 \Omega$ and $R_{term2} = 32 \Omega$ help to match the circuit to 50Ω at both ends. Finally, to improve the input match, two stubs are added at the input. Introduction of these stubs provides a good input match throughout the BW of operation. The fabricated circuit is shown in Fig. 10. It measures $0.65 \times 1.5 \text{ mm}^2$.

A standalone tricode cell has high impedance looking at the output because of the reduction of Miller capacitance. From simulations, the capacitance looking at the output is 7 fF . The input impedance is much lower compared to the output with an input capacitance value of 50 fF . This capacitance values are absorbed by the input and output lines, L_a and L_b ,

Fig. 11. Simulated (dotted line) and measured (solid lines) S-parameter of the circuit.

Fig. 12. Measured (dotted line) and simulated (solid line) stability factor (k) in logarithmic scale versus frequency. (b) Simulated (dotted line) and measured group delay and deviation from linear phase versus frequency.

which finally makes the input/output of DA matched to 50Ω within the BW of interest.

IV. MEASUREMENTS AND DISCUSSION

A. Small-Signal Measurements

The small-signal measurements for this circuit were performed in two steps, namely, on a $0.1\text{--}110\text{-GHz}$ and a $140\text{--}220\text{-GHz}$ setup. The probes for the 110- and $140\text{--}220\text{-GHz}$ measurements feature pitches of 100 and $50 \mu\text{m}$, respectively. An on-wafer multiline TRL calibration was applied for both setups. The carefully selected line lengths ensure the validity of the calibration over the whole frequency range. The reference impedance of the multiline TRL was transformed to 50Ω which is supported by a transmission line capacitance determination.

The results of the S-parameter measurements are presented in Fig. 11. A forward gain S_{21} of 12 dB was achieved with a 3-dB BW from almost dc to $>170 \text{ GHz}$, with measured data available up to 110 GHz and from 140 to 200 GHz , due to the different bands of the setup. The input of the circuit is well matched, with reflections below -10 dB and a slight increase between 70 and 110 GHz . The output reflection parameter is below -5 dB up to 105 GHz and in the $140\text{--}160\text{-GHz}$ frequency range. With further precise modeling of the transistors and transmission lines at these higher frequencies, S_{22} could be improved further. With reference to Fig. 9, the following dc values were $V_{b3} = 1.0 \text{ V}$, $I_{b3} = 1.2 \text{ mA}$, $V_{b2} = 2.8 \text{ V}$, $I_{b2} = 1.1 \text{ mA}$, $V_{b1} = 3.8 \text{ V}$, $I_{b1} = 1.4 \text{ mA}$, and $V_c = 5.9 \text{ V}$, and $I_c = 29 \text{ mA}$.

As seen from the small-signal data in Fig. 12(a), the circuit is stable throughout the measured frequency region, with a

Fig. 13. (a) NF measurement setup up to 50 GHz. The noise source is needed for calibration purposes only. (b) NF measurement setup for 110–150 GHz.

worst case k factor of 3 at 165 GHz, which indicates its overall stable operation. There is a low-frequency peaking in S_{21} in Fig. 11, but the circuit itself is stable ($k > 10$) at these peaking frequency. Therefore, this phenomenon is attributed to the insufficient decoupling of dc feed during the S-parameter measurements. This can be eliminated using appropriate blocking capacitors at the bias feeds.

A constant group delay characteristic is highly desirable for optical communication and measurement systems. Fig. 12(b) shows the measured versus simulated group delay and the deviation from linear phase for the measured frequency ranges. The simulated group delay is uniform and matches the circuit results within the measured frequency ranges and the deviation from the linear phase is within $\pm 10^\circ$ throughout the measured data and increases beyond 165 GHz, which also closely matches the simulation. This confirms the circuit's excellent group delay characteristics.

B. Noise Figure Measurements

The NF measurements are added as an unexplored feature of the circuit as presented in [11]. To be used as a component in a receiver or spectroscopic application, a low NF over the whole frequency band is highly desirable. In order to investigate this further, two similar five-cell DAs were compared with respect to their NF, one using a cascode and the other with a tricode subcell, both based on the same technology node (500-nm InP/InGaAs DHBT). This aimed to identify how much NF degrades when using a tricode subcell instead of a cascode one.

To estimate the NF, measurements up to 50 GHz were carried out. The setup shown in Fig. 13(a) includes a Keysight PNA-X vector network analyzer as the noise receiver and a Keysight 346CK01 50-GHz noise source for calibration purposes. Due to the built-in, dedicated noise receiver, the NF of the system was fairly low and an additional low-noise amplifier was not required.

The on-wafer NF measurements between 110 and 150 GHz were performed at the Fraunhofer Institute IAF, Freiburg, Germany, as shown in Fig. 13(b). The setup consists of a commercial mixer module, whose local oscillator input is driven by a multiplied synthesizer signal. The NF analyzer controls the synthesizer as well as the noise source and evaluates the downconverted signals [10].

The NF measurements of the tricode-based DA for the frequency range up to 50 and 110–150 GHz are plotted

Fig. 14. Simulated (dotted) and measured (solid line) NF for the two DAs, one with a five tricode unit cells and the other with five cascode units.

in Fig. 14. The measured NF is almost constant with frequency, with an average value of 8 dB up to 50 GHz and a variation of ± 1 dB within this frequency range. Between 110 and 150 GHz, the NF follows a similar trend as the simulation, going from 10 to 14.6 dB up to 150 GHz. This increase in NF is caused by its nearness to the upper limit of the BW (i.e. 175 GHz). Downscaling of the transistors and rightly adjusted linewidths of the amplifier will generate a higher BW and, consequently, a lower NF within this frequency range.

The NF is initially very high in the low-frequency range because of the absence of gain at very low frequencies. Lack of gain causes low SNR values which, in turn, increase the NF at low frequencies. The amplifier has a lower gain below 3 GHz mainly because of low on-chip input–output decoupling capacitance (1 pF). Higher values would increase the gain at lower frequencies, and the NF would be lower and closer to the value of 8 dB in the lower frequency end.

Using the same setup, NF and small-signal gain of a cascode-based DA have been measured, which is shown in Fig. 14. It is seen that the cascode-based DA offers a slightly lower NF with an average of 7.5 dB, i.e., 0.5 dB better compared to its tricode counterpart. This was also illustrated by simulation in Fig. 8, which shows that the NF of a cascode subcell is slightly lower than its tricode variant.

Even though there is a slight increase of NF using the tricode unit cell in a DA, the gain-BW product is much higher. This enables certain ultrawideband optoelectronic and spectroscopic applications, where the BW offered by a simple cascode unit cell of the same technology might not be enough and the compromise in NF is acceptable.

C. Large-Signal Measurements

The large-signal measurements were carried out in three steps, within 5–55 GHz with a PNA-X spectrum analyzer, from 80 to 110 GHz with an HP power meter, and within 145–165 GHz with an Erikson PM4 power meter. The measurements data were corrected by deembedding of the on-wafer probe losses in both cases.

Fig. 15 summarizes the output large-signal performance within the aforementioned frequency bands. At 150 GHz, the circuit exhibits an output power at 1-dB compression of $P_{1\text{dB}} = 8.4$ dBm and a saturated output power of $P_{\text{sat}} = 10$ dBm, at a maximum PAE of 6% at 150 GHz. Slightly higher values were obtained in the lower frequency

TABLE I
STATE-OF-THE-ART DAS WITH LARGER THAN 110-GHz BANDWIDTH

Ref.	BW* (GHz)	Gain (dB)	Technology	Circuit Topology	GBP (GHz)	P_{DC} (mW)	P_{1dB} (dBm) @ Freq (GHz)	MaxPAE (%)	Noise Figure (dB)
[19]	110	11	50 nm InGaAs mHEMT	Cascode	390	450	7 @ 75	1.02	2.5-5 @ <90 GHz
[9]	>110	13	500 nm InP DHBT	Cascode	>491	129	10 @ 5-110	8	6.5-8 @ 5-50 GHz
[18]	>110	19.7	50 nm mHEMT	4-stacked	>1062	541	10 @ 90	6	2.5-6.4 @ 0-98 GHz
[20]	120	7.3	40 nm GaN DHFET	Cascode	278	448	15.5 @ 20	6.5	5-7 @ 5-40 GHz
[7]	160	19	130 nm SiGe HBT	Tricore	1515	560	9 @ 135	1.4	NA
[5]	170	10	130 nm SiGe HBT	Tricore	537	108	7.5 @ 50	4.9	6 @ 50-67 GHz
This work	>170	12	500 nm InP DHBT	Tricore	697	180	8.4 @ 150 / 6.2 @ 165	6/4	8@5-50 GHz 10-14.6@110-150 GHz
[17]	180	18.7	130 nm SiGe HBT	Tricore ^a	1550	86	0 @ 100	1.2	6 @ 60 GHz
[8]	>200	13.5	250 nm InP DHBT	Cascode ^b	980	210	3.7 @ 100, 1.9 @ 195	NA	NA
[3]	235	16	250 nm InP DHBT	Cascode ^c	1483	117	NA	NA	9 @ 70-110 GHz 10-12 dB@110-150 GHz

*Table is arranged in the order of increasing BW, ^a3-cascaded tricore DA, ^cmitter follower cascode, ^b2-cascaded cascode DA

Fig. 15. Measured P_{1dB} , P_{sat} , and PAE_{max} within 5–165 GHz.

region, as expected from a DA according to the literature, with a maximum output power of 11 dBm and P_{1dB} of 9 dBm around 20–25 GHz, at a maximum PAE of 6%. These power measurements for the lower and higher bounds of the frequency BW indicate that the circuit delivers uniform large-signal behavior over the whole BW. The dotted lines in Fig. 15 are only a predictive line to indicate continuous large-signal behavior. The actual measured data are plotted with symbols.

At the higher end of the frequency band, P_{1dB} is relatively constant at 8.4 dBm up to frequencies of 150 GHz and then drops to around 6 dBm at 165 GHz, while the saturated output power lays 1 dB higher across this frequency range. Due to the decrease in linearity and saturated output power, the corresponding maximum PAE reduces from 6% to 4% when measured from 145 to 165 GHz. Power measurements at lower frequencies provided consistent data to that of the higher edge of the BW. P_{1dB} is approximately 9 dB between 20 and 35 GHz and within 80–110 GHz. There are certain frequencies, namely, 15, 45, and 90 GHz, where there is a dip in the output power performance. This effect could

Fig. 16. Measured third-order intercept point within 10–50 GHz.

originate from the measurement system, which requires further investigation.

To further confirm the linearity of the amplifier, two-tone measurements have been performed within the frequency range of 10–50 GHz with the help of a PNA-X. The two tones are generated directly from the PNA-X and the power of the fundamental and third-order intermodulation products are measured after calibrating out the path loss for each measurement frequency. For each fundamental tone, a second tone 1 MHz higher in frequency was chosen and fed into the amplifier. The power of the output of the fundamental tones and third-order intermodulation products is extrapolated, which intersects at points shown in Fig. 16. Theoretically, the 1-dB compression point, P_{1dB} , is expected to lay ~ 10 dB below the third-order intercept point, this graph confirms the earlier measurements on linearity in Fig. 15. The linear behavior of the circuit is, therefore, further established. Two-tone measurements in the W-band and beyond are currently limited by lack of measurement equipment.

The comparison to the state-of-the-art DAs is presented in Table I, arranged in the order of increasing BW.

Our circuit [11] can be compared with the state-of-the-art in three aspects, namely, low deviation from linear phase within very large BW, low NF, and highly linear characteristics even up to the higher end of the BW and correspondingly high efficiency.

Since one of the intended applications is in optical receivers, a low deviation from linear phase over the whole BW is a necessity. This article provides this feature up to 165 GHz with a variation of $\pm 10^\circ$. Moreover, in receiver applications, a low NF is another necessary merit. When compared with the state of the art, the circuit provides a relatively low NF among DAs with a BW of more than 17 GHz. Despite the fact that HBT-based amplifiers have the technological disadvantage of higher NF compared to its FET-based counterparts, the circuit presented here achieves a competitive low NF within its BW. In terms of large-signal characteristics, it has the highest linearity and PAE at frequencies beyond 140 GHz for such class of circuits. In the literature, most ultrawideband amplifiers either report a lower NF or higher linear performance, whereas our circuit performs the highest linearity with a reasonably good NF value. Only [3] and [8] show a larger BW than this article, however, at considerably lower RF output power and PAE levels. This article is the first report of an amplifier that combines all the abovementioned necessary merits in a single design. Additionally, all these merits of the circuit have been achieved at a larger technology node when compared to the others in the table. This indicates that the BW of the circuit could be further improved using a smaller InP transistor node. The circuit consumes a moderate dc power of 180 mW, which is lower than HEMT-based devices and comparable with other HBT technologies. The circuit, therefore, can fulfill the demand when a highly linear amplifier is required with an acceptable NF value.

V. CONCLUSION

This article presents the characterization results of an ultrawideband DA from both small- and large-signal perspectives. The circuit achieves a 3-dB BW of > 170 GHz. Measurements show a forward gain of 12 dB and a maximum deviation from the linear phase of $\pm 10^\circ$ up to 165 GHz. In terms of NF, the circuit exhibits a value of 8 dB within 5–50 GHz and 10–14.6 dB within 110–150-GHz frequency range. In terms of output power, it shows a $P_{1\text{dB}}$ of 8.4 dBm at 150 GHz and a corresponding maximum PAE of 6%. The measured $P_{1\text{dB}}$ output power decreases to $P_{1\text{dB}} = 6.2$ dBm when measured at 165 GHz. At the lower frequency end, the circuit maintains its highly linear and efficient properties. Considering both small- and large-signal merits, this circuit is suitable for application in measurement systems and optical modulator and demodulators with the data rates of 200 Gb/s and beyond.

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