

Residual AM and PM Noise of Selected X-Band Amplifiers

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Summary—In microwave systems design, amplitude modulation (AM) noise is often ignored relative to phase modulation (PM) noise. However, in high-performance applications requiring ultra-low PM noise, AM noise becomes a critical limiting factor due to AM-to-PM conversion mechanisms. Because AM noise specifications are rarely, if ever, provided for commercial microwave components, this study characterizes the residual AM and PM noise of several 10 GHz amplifiers. We present performance data for various Heterojunction Bipolar Transistor (HBT) amplifier designs and demonstrate the noise reduction achieved through power-summed configurations. For example, a discrete-die GaAs HBT amplifier (labeled GA-HBT-1) achieves residual PM noise below -160 dBrad²/Hz and AM noise of approximately -155 dB/Hz at a 1 kHz offset in a quad power-summed configuration. A notable finding is that the majority of amplifier designs surveyed in this study exhibit higher AM noise relative to their PM noise, with only a few exceptions. The impact of bias supply quality on residual noise is also quantified, demonstrating that the choice of voltage regulator is a first-order design parameter below 1 kHz offset.

Keywords—amplifier; amplitude noise; phase noise; X-band; HBT; cross-spectrum; power summing; frequency synthesis; voltage regulator

I. INTRODUCTION

Residual amplitude modulation (AM) and phase modulation (PM) noise are fundamental metrics that set the noise floor of radar waveform generators, high-speed communication transmitters, and precision frequency synthesis systems [1]–[10]. In the design community, residual PM noise has been extensively studied and characterized over several decades [11]–[20]. Residual AM noise, however, has received substantially less attention in the primary literature [21]–[27] and is almost universally absent from commercial component datasheets.

In any component exhibiting even modest amplitude-to-phase nonlinearity, AM noise ($S_{\alpha}(f)$) at the input is converted to PM noise ($S_{\varphi}(f)$) at the output through the AM-to-PM conversion. This mechanism is particularly consequential in frequency multiplication chains, where AM noise from an amplifier stage enters a subsequent multiplier or mixer and reappears as PM noise at the multiplied output.

The practical consequences for phase noise metrology were documented by Cibiel et al. [28], who showed that AM noise from amplifiers in a cross-spectrum measurement bridge can directly corrupt PM noise floors below -167 dBrad²/Hz. Rubiola and Boudot [25] extended this analysis to quantify the

cross-talk floor introduced by finite AM rejection in homodyne detector bridges. The present work builds on the prior study by Hati, Howe, and Nelson [24], which characterized both AM and PM noise in X-band oscillators and amplifiers circa 2006. The intervening years have seen the introduction of a new generation of Gallium Arsenide Heterojunction Bipolar Transistor (GaAs HBT) and Monolithic Microwave Integrated Circuit (MMIC) devices; whether they represent a substantive improvement in AM noise performance over legacy devices is an open and practically important question. With the wider availability of commercial phase noise analyzers, more manufacturers now provide PM noise information in their datasheets, but to the best of our knowledge AM noise specifications remain universally absent.

The present study was motivated directly by the stringent PM and AM noise requirements of a frequency synthesis development program at National Institute of Standards and Technology (NIST). Identifying X-band amplifier designs that meet or exceed the noise budget requires measured residual AM and PM data that does not exist in any manufacturer documentation. The scope is: (a) 10 GHz center frequency; (b) mostly GaAs HBT technology (discrete die, MMIC, and module-level packages); (c) single, dual, and quad power-summed array configurations; and (d) characterization across a range of input drive levels up to and including mild compression. The central questions are which amplifier designs and noise mechanisms yield the lowest combined residual AM and PM noise, and what improvements in AM and PM noise can be achieved with array-summing.

II. MEASUREMENT SYSTEM AND METHODOLOGY

A. System Architecture

Figure 1 shows the cross-spectrum homodyne measurement system, referred to hereafter as the NIST phase noise analyzer (NIST PNA), used throughout this study [29]–[31]. The cross-spectrum technique is employed because it allows the system noise floor to be suppressed below the residual noise of the device under test (DUT) by cross-correlating the outputs of two independent measurement channels.

Let $x(t)$ and $y(t)$ denote the outputs of channels 1 and 2, respectively, and let $X(f)$ and $Y(f)$ denote their Fourier

transforms. Each channel output can be decomposed as

$$x(t) = d(t) + n_x(t), \quad y(t) = d(t) + n_y(t) \quad (1)$$

where $d(t)$ represents the common DUT noise contribution and $n_x(t)$ and $n_y(t)$ are the uncorrelated noise contributions of the individual channels (mixer, IF amp, etc.). Under the assumption that n_x , n_y , and d are mutually uncorrelated, the cross-power spectral density S_{xy} is estimated as the ensemble average, over m acquisitions, of the product $X(f)Y^*(f)$, where $*$ denotes complex conjugation:

$$\langle S_{xy} \rangle_m = \frac{1}{T} \langle X(f)Y^*(f) \rangle_m \approx S_d(f) + \frac{S_{n_x}(f) + S_{n_y}(f)}{\sqrt{m}}. \quad (2)$$

Here, T is the measurement or truncation time normalizing the PSD to 1 Hz. This means that the cross-spectrum estimator converges to the true DUT noise power spectral density (PSD), free of the individual channel noise floors, as the number of averages $m \rightarrow \infty$. In practice the residual channel-noise bias scales as $1/\sqrt{m}$. For m cross-spectral averages, the noise floor improves by $5 \log_{10}(m)$ dB, enabling characterization of devices whose residual noise would otherwise be masked by the measurement noise floor. Unlike residual PM noise measurements, where reference oscillator PM noise cancels to a high degree, AM noise does not. Therefore, valid residual AM noise measurements require the source AM noise to be substantially below that of the DUT (see Appendix A). In the present configuration (Fig. 1), the cross-spectrum floor is confirmed to be 5 dB to 15 dB below the residual noise of the lowest-noise device under test. Furthermore, to minimize the anti-correlated thermal noise contribution of the power splitters [32], it is important not to attenuate the signal at the splitter input, and rather set the detector power level directly prior to the AM/PM detector as shown in Fig. 1.

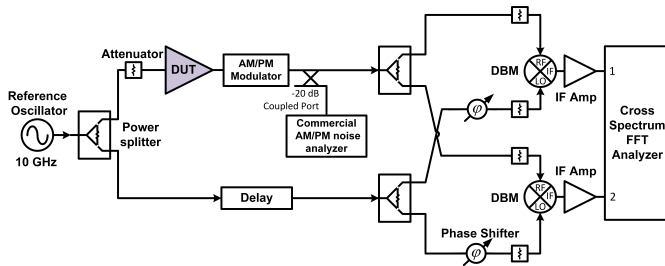


Fig. 1. Block diagram of the custom-built cross-spectrum measurement system for residual AM and PM noise. DUT: Device Under Test, FFT: Fast Fourier Transform, DBM: Double Balanced Mixer. The DBM acts as either a phase or amplitude detector.

B. Reference Oscillator

The system reference is an air-dielectric cavity resonator in a simple loop configuration operating at 10 GHz (Fig. 2). This oscillator exhibits an exceptionally low AM noise level, as illustrated in Fig. 3, $S_{\alpha}(1 \text{ kHz}) \approx -162 \text{ dB/Hz}$, which was set by an obsolete commercial eight-element HBT

array amplifier (IGP-HBT-3, Table I) operating in deep compression of about 7 dB. This performance ensures that the reference source AM noise in the flicker region is well below the residual AM noise of the amplifiers selected for this study.

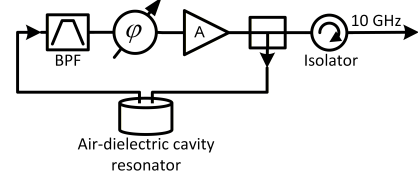


Fig. 2. Block diagram of the simple loop oscillator used as a reference frequency source. BPF - Bandpass filter, A - Low AM noise loop amplifier.

C. AM/PM Detector Configuration

Rather than employing conventional diode-based AM detectors [25], [33], [34], this system uses double-balanced mixers (DBMs) as homodyne detectors [35]. The primary reason for this configuration was the ease of switching between AM and PM measurements with the same set-up. The mixer sensitivity to AM or PM is selected by setting the relative phase between the local oscillator (LO) and radio frequency (RF) ports of each mixer: 0° phase yields AM detection, 90° phase yields PM detection. For PM detection both mixer ports are driven at or above saturation, whereas for AM detection the RF port power is backed off into its linear regime. The AM-to-PM and PM-to-AM leakage and its dependence on saturation are analyzed in Appendix A. An integrated AM/PM modulator provides the reference modulation tone required for absolute calibration and alignment of both channels. Cross-talk between the AM and PM sensitivity, arising from imperfect phase alignment, is suppressed by adjusting the phase shifter until the orthogonal alignment tone nulls in the detector channel.

D. Verification Instrument

A commercial cross-correlation phase noise analyzer (PNA) with AM noise measurement capability was connected as a monitor channel for all measurement setups. It enables quick and convenient alignment of the modulator for pure AM and PM tones. The commercial PNA is not used as the primary measurement instrument; the NIST PNA (Fig. 1) provides the reported AM and PM noise spectral densities. However, the commercial PNA was occasionally used to verify the accuracy of the AM and PM measurement. Although the AM noise floor of the commercial PNA was insufficient for characterizing some of the low-noise amplifiers, it allowed for the validation of analog measurements on a higher-noise device as shown in Fig. 3.

E. Calibration Procedure

Calibration of residual AM and PM noise spectral density is performed by injecting a known modulation from the integrated AM/PM modulator [36]–[41] at multiple offset frequencies and recording the detector output level. The modulation index is set to a known value, establishing the detector transfer

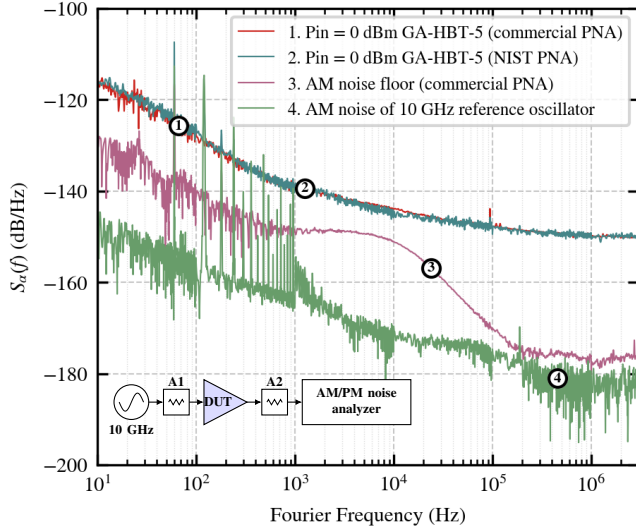


Fig. 3. Comparison of AM noise measurements for amplifier GA-HBT-5 using a commercial PNA and the NIST PNA. (i) Traces 1 and 2: Excellent agreement between the two systems for the DUT. (ii) Trace 3: AM noise of the 10 GHz reference measured directly by the commercial PNA (DUT and attenuator A2 bypassed), illustrating the instrument's internal AM noise floor at 10 GHz. (iii) Trace 4: Residual AM noise of the 10 GHz reference oscillator (DUT bypassed), measured via the Fig. 1 configuration.

coefficients k_φ in V/rad for PM and k_α in V for AM. This calibration is performed at each input power setting for all amplifiers to account for any operating-point dependence in the mixer conversion gain. Attenuators at the RF input of the mixers were adjusted for each amplifier to achieve roughly constant conversion factors across all PM measurements; this same procedure was repeated for AM measurements, albeit at a different mixer operating point.

Accurate k_α calibration necessitates placing the AM/PM modulator after the DUT in Fig. 1. Because AM noise measurements require the mixer's RF port to operate strictly within its linear regime, this specific placement prevents compression of the AM calibration tone which would otherwise occur if the modulator is placed in the saturated reference arm, yielding erroneous k_α values. Conversely, for PM noise and k_φ calibration, the modulator can be positioned either on the reference arm or post-DUT.

F. Array Configuration

This study evaluates the noise performance of both commercial and custom-built amplifier arrays. All custom power-summed architectures utilize Wilkinson power dividers and combiners to distribute and recombine signals across n parallel paths. For an array of n amplifiers driven by split copies of a common input signal, the summed output amplitude and phase noise spectra are given by [17], [42]

$$S_\alpha(f) = \frac{S_{\alpha-f}(f)}{n} + \frac{N_F k_B T}{P_{in}} \quad (3)$$

$$S_\varphi(f) = \frac{S_{\varphi-f}(f)}{n} + \frac{N_F k_B T}{P_{in}} \quad (4)$$

where $S_{\alpha-f}$ and $S_{\varphi-f}$ are the flicker amplitude and phase noise of a single amplifier, N_F is the noise figure, $k_B T$ is the thermal noise power in 1 Hz bandwidth (≈ -174 dBm/Hz at room temperature), and P_{in} is the input power to the array amplifier. The multiplicative flicker noise is reduced by the array factor n because the correlated carrier amplitudes add linearly while the uncorrelated flicker noise variables add in power. The additive thermal noise, though also uncorrelated across amplifiers and thus subject to the same $1/n$ power averaging, receives no net benefit because the input splitter reduces the signal power delivered to each amplifier by the same factor n , leaving the thermal noise unchanged relative to a single amplifier.

In an ideal case of an amplifier array, the gain and noise figure remain the same while the 1 dB compression point increases. In practice, several factors cause performance to deviate from this theoretical limit. First, the excess insertion loss of the Wilkinson power splitters attenuates the signal, thereby reducing gain, the thermal signal-to-noise ratio (SNR), and total power output. Second, imbalances in amplitude and phase between channels prevent perfect coherent recombination, further degrading gain and inducing AM $\leftrightarrow$ PM conversion. Additionally, unit-to-unit variations in amplifier parameters and noise correlation, often arising from shared power supplies or environmental coupling, preclude the ideal $10 \log_{10}(n)$ improvement. To mitigate correlated power supply noise, which can limit the statistical advantage of uncorrelated noise addition of the array, individual voltage regulators were used for each amplifier. As shown in Fig. 4, the custom-built array consists of four amplifiers, with each element utilizing an independent regulation stage to ensure maximum isolation and noise performance.

For all array amplifiers discussed, the input power (P_{in}) refers to the total power at the splitter input, rather than the power entering each individual element of the array.

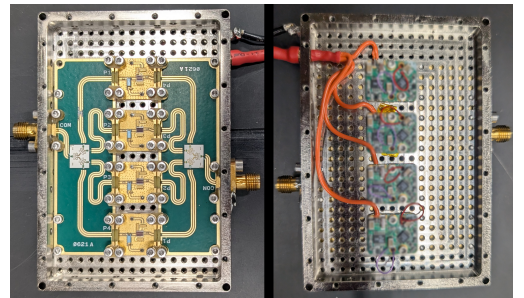


Fig. 4. Picture of the custom-built quad array of amplifier GA-MMIC-2. Four voltage regulators (right), directly mounted underneath the amplifiers, provide independent bias voltage to each element of the array (left).

III. DEVICES UNDER TEST

Table I lists all amplifiers characterized in this study. In accordance with NIST policy on the evaluation of commercial products, each device is identified only by a label that encodes its semiconductor technology and package or topology class; manufacturers and part numbers are not given, and the order

TABLE I
AMPLIFIERS CHARACTERIZED IN THIS STUDY

Device label	Package / topology	Technology (from datasheet)	BW (GHz)	Gain (dB)	P1dB (dBm)	OIP3 (dBm)	N_F (dB)	S_φ (1 kHz) (dBrad ² /Hz)	S_α (1 kHz) (dB/Hz)
GA-HBT-1 (S,Q)	Discrete die	GaAs HBT	2–30	12	21	33	5.0	−157	−149
GA-HBT-2 (S,Q)	Discrete die	GaAs HBT	6–12	15.6	20	32	5	−153	−147
GA-MMIC-1 (S,Q)	Packaged MMIC	GaAs MMIC	2–20	15	20	31	3.2	−150	−142
GA-MMIC-2 (S,Q)	Packaged MMIC	GaAs MMIC	DC–18	14.4	24	34	3.5	−150	−143
GA-MMIC-3 (S)	Packaged MMIC	GaAs MMIC	6–18	18	20	29	4.5	−152	−143
IGP-HBT-1 (D)	Array (2)	InGaP/GaAs HBT	DC–12	9.7	13	23	4.9	−150	−149
IGP-HBT-2 (S,Q)	Distributed die	InGaP/GaAs HBT	2–18	14	15	27	5	−152	−145
GA-HBT-3 (S)	Hermetic module	GaAs HBT	6–12	11	23	34	4.5	−156	−152
GA-HBT-4 (S)	Hermetic module	GaAs HBT	7–11	9	22	33	6	−158	−155
GA-pHEMT (S)	Distributed	GaAs pHEMT	DC–22	13.5	28	39	3.5	−152	−140
GA-HBT-5 (S)	Connectorized module	GaAs HBT	2–30	11	20	29	4.5	−152	−140
IGP-HBT-3 (O)	Array (8)	InGaP HBT	8–12	18	21	29	8	−159	−155

S = Single, D = Dual, Q = Quad, O = Octa. Devices are anonymized; labels denote package/technology class only.

Bandwidth (BW), Gain, Noise Figure (N_F), 1-dB compression point (P1dB), Output Third-Order Intercept Point (OIP3) at 10 GHz.

AM and PM noise are for the single amplifier configurations for a $P_{in} = 0$ dBm.

IGP-HBT-1 and IGP-HBT-3 are measured as array configurations for a $P_{in} = 0$ dBm.

AM and PM noise are measured at NIST, while all other parameters are sourced directly from the manufacturer’s datasheet.

of presentation does not imply a ranking. The devices were sourced from five manufacturers and span several Gallium Arsenide Heterojunction Bipolar Transistor (GaAs HBT) subtypes. GA-HBT-1 and GA-HBT-2 are discrete GaAs HBT die in chip carrier packages. GA-MMIC-1, GA-MMIC-2, and GA-MMIC-3 are packaged devices from a common product family and are listed by the manufacturer as GaAs MMICs. IGP-HBT-1 is an Indium gallium phosphide (InGaP)/GaAs HBT, supplied and measured as a two-element array. IGP-HBT-2 is a GaAs InGaP HBT MMIC distributed amplifier die. In contrast, GA-HBT-3 and GA-HBT-4 are GaAs HBT ultra-low phase noise amplifiers housed in hermetically sealed modules; both are now obsolete. GA-pHEMT is a GaAs pseudomorphic high-electron-mobility transistor (pHEMT) distributed amplifier and is included for comparison with the HBT devices despite its different technology; it represents a distinct noise mechanism baseline. GA-HBT-5 is a broadband GaAs HBT amplifier in a connectorized package. IGP-HBT-3, though now obsolete, consists of an array of eight parallel amplifiers utilizing InGaP HBT technology. The technical data in Table I are based on available manufacturer documentation. Experimental results indicate performance variations between units of the same model, particularly for GA-MMIC-1. The amplifiers integrated into the quad arrays were distinct units from those utilized in the single-amplifier characterization measurements.

IV. RESULTS

A. Single-Amplifier PM Noise Comparison

Figure 5 shows the residual PM noise of all characterized single amplifiers operating at input power, $P_{in} = 0$ dBm. The spread in performance at 1 kHz offset is almost 10 dB across the full set and exceeds 15 dB at 10 Hz offset. Amplifier GA-HBT-1, together with the obsolete modules GA-HBT-3 and

GA-HBT-4, measured residual PM noise below approximately −155 dBrad²/Hz at 1 kHz offset.

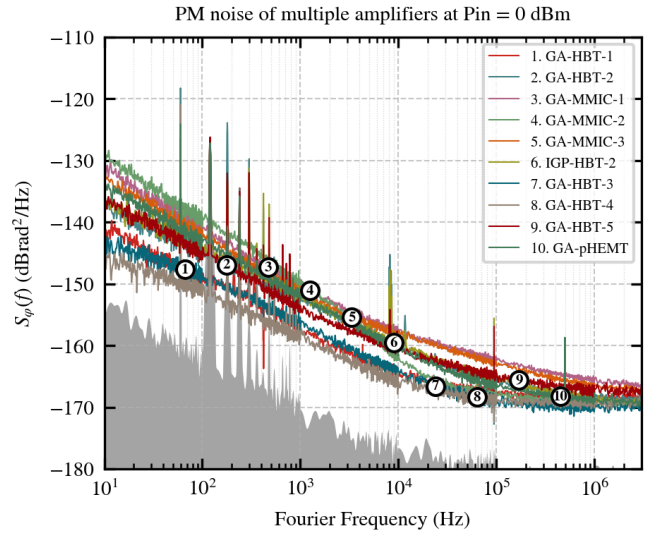


Fig. 5. Measured PM noise of multiple amplifier designs (single configuration) at 10 GHz, $P_{in} = 0$ dBm. Device details are in Table I; traces are numbered in table order, not by performance. Gray shaded area is the system PM noise floor for $P_{in} = 0$ dBm. The number of FFT averages (m) for different frequency spans are as follows: 10 Hz to 100 Hz = 500, 100 Hz to 1 kHz = 1000, 1 kHz to 10 kHz = 3000, 10 kHz to 100 kHz = 5000, and 100 kHz to 3 MHz = 10,000.

GA-pHEMT, which is a 1 W pHEMT rather than an HBT device, exhibits a residual PM noise floor that is comparable to the mid-power HBT devices despite its fundamentally different transport mechanism. The pHEMT flicker noise behavior is generally attributed to interface trap states at the channel heterointerface rather than to bulk recombination, and its mea-

sured PM noise is consistent with its high 1 dB compression point (P1dB) margin at the 0 dBm operating point.

B. Single-Amplifier AM Noise Comparison

Figure 6 shows the corresponding residual AM noise. The performance broadly follows the PM noise results, but two observations are quantitatively significant.

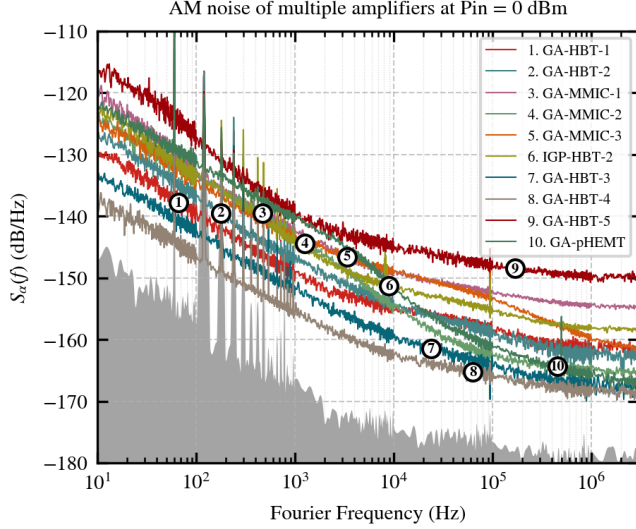


Fig. 6. Measured AM noise of multiple amplifier designs (single configuration) at 10 GHz, $P_{in} = 0$ dBm. Device details are in Table I; traces are numbered in table order, not by performance. Gray shaded area is the system AM noise floor for $P_{in} = 0$ dBm. The number of FFT averages (m) for different frequency spans are as follows: 10 Hz to 100 Hz = 500, 100 Hz to 1 kHz = 1000, 1 kHz to 10 kHz = 3000, 10 kHz to 100 kHz = 5000, and 100 kHz to 3 MHz = 10,000.

First, the spread across devices is larger in AM than in PM. Second, several devices show AM noise 5 dB to 10 dB above their own PM noise in the flicker region below 1 kHz. Also, in the thermal region there is an assumption, widely held in system-level noise modeling, that AM noise is equal to the PM noise of the same device due to additive noise. However, as Fig. 5 and Fig. 6 show, the AM noise is higher than the corresponding PM noise by approximately 1 dB (GA-HBT-4) to 17.5 dB (GA-HBT-5) at 1 MHz offset, also indicating that the white AM noise mechanism is not additive.

C. Measurement Uncertainty

Rather than a full “International Organization for Standardization - Guide to the Expression of Uncertainty in Measurement (ISO GUM)” budget [43], a streamlined approach is used to bound the dominant error terms of the AM and PM measurement system. The accuracy of the detector constants (k_φ and k_α) is verified using a NIST AM and PM white noise calibration standard [44] with a 2-sigma uncertainty of ± 0.7 dB. Combining this uncertainty with the statistical variance from the $m = 1000$ cross-spectrum FFT averages yields a total 2-sigma uncertainty of ± 1.0 dB at a 1 kHz offset in Fig. 5 and Fig. 6.

AM-to-PM and PM-to-AM conversion terms are excluded from this budget because the measurement configuration suppresses each by more than 30 dB. For PM measurements, a pure AM tone is injected while mixer PM sensitivity is optimized, rejecting DUT AM noise at the RF port, and operating the mixer in saturation further reduces source AM noise. For AM measurements, the complementary configuration is used: a pure PM tone is injected while mixer AM sensitivity is optimized, rejecting DUT PM-to-AM conversion, and the source’s common-mode PM noise cancels between the two ports.

Finally, a worst-case noise floor bias of approximately +0.4 dB occurs at a 1 kHz offset on the lowest noise traces in Fig. 5 and Fig. 6; this bias has been left uncorrected to maintain a conservative presentation of the raw data.

D. GA-HBT-1: Array PM Noise Performance

Figure 7 shows the residual PM noise of GA-HBT-1 (Discrete-Die HBT) in single and quad configurations at several input power levels. In the single configuration at $P_{in} = 0$ dBm, GA-HBT-1 achieves approximately -157 dB rad^2/Hz at 1 kHz offset. In the quad configuration, the flicker region improves by approximately 3 dB to 5 dB, reaching below -161 dB rad^2/Hz at 1 kHz offset. The varying improvement with offset frequency may be attributed to unit-to-unit performance variations, correlation from power supply, etc. As expected and discussed earlier, the thermal white phase noise, $S_\varphi(f) \approx N_F k_B T / P_{in}$ [45], does not improve with the quad configuration at $P_{in} = 0$ dBm relative to the single amplifier. However, for $P_{in} = +7$ dBm, where per-element input power is roughly 0 dBm, a 5 dB to 6 dB improvement is observed at all offset frequencies relative to single amplifier at 0 dBm.

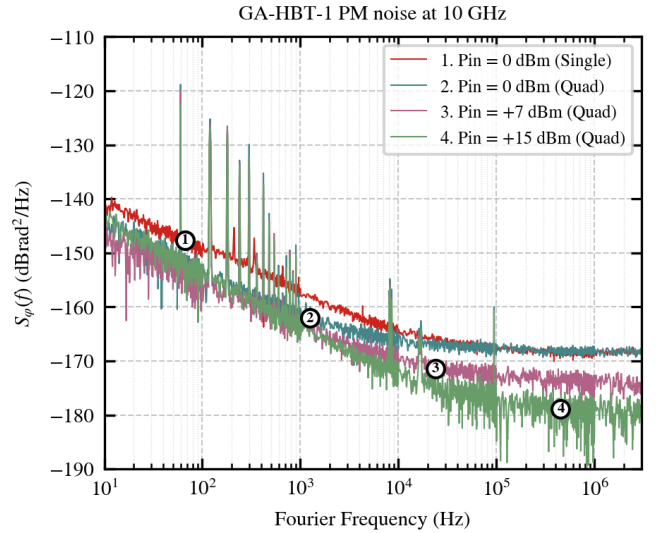


Fig. 7. Residual PM noise of amplifier GA-HBT-1 in single and quad configurations at 10 GHz.

E. GA-HBT-1: Array AM Noise Performance

Figure 8 reveals the most technically significant result in this study. Unlike the PM noise, which shows an improvement in flicker but no change in the white phase noise, the AM noise of GA-HBT-1 improves in both the flicker and white noise regions when converting operation from single to quad configuration for a fixed input power level. The observed improvement in the AM white noise is approximately 3 dB to 5 dB while the PM white noise remains unchanged (see Fig. 7 and Fig. 8, $P_{in} = 0$ dBm). This suggests that the white AM noise in these devices is dominated by internal excess noise rather than being set by the $50\ \Omega$ SNR thermal limit. Additionally, in Fig. 8, the three quad AM traces nearly overlap in the white region. This power independence provides stronger proof that the AM floor is not additive-thermal.

The amplitude noise of the single and quad GA-HBT-1 is significantly higher than its phase noise across both the flicker and thermal frequency regions. Notably, in the quad configuration at an input power of +15 dBm, the difference between phase noise and amplitude noise exceeds 10 dB in the white noise region.

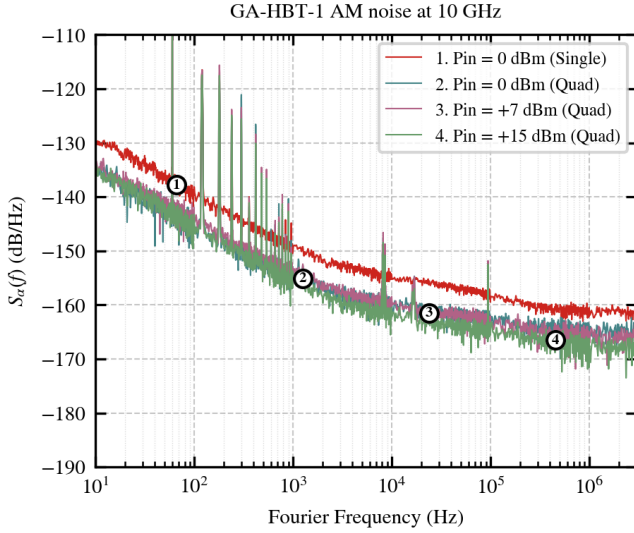


Fig. 8. Residual AM noise of amplifier GA-HBT-1 in single and quad configurations at 10 GHz. Note improvement in both flicker and white noise regions for the quad configuration.

F. GA-MMIC-1: Array AM and PM Noise Performance

Figures 9 and 10 show the PM and AM noise of GA-MMIC-1 in single and quad configurations. At 1 kHz offset, its flicker noise level is higher than that of GA-HBT-1 by approximately 7 dB. The quad configuration provides the expected flicker improvement. The AM noise of GA-MMIC-1 exhibits the AM>PM behavior in both the flicker and thermal regions.

G. Effect of Voltage Regulator (GA-MMIC-2, Single Amplifier)

Figure 11 shows perhaps the most practically actionable result in the study, illustrating the residual PM and AM

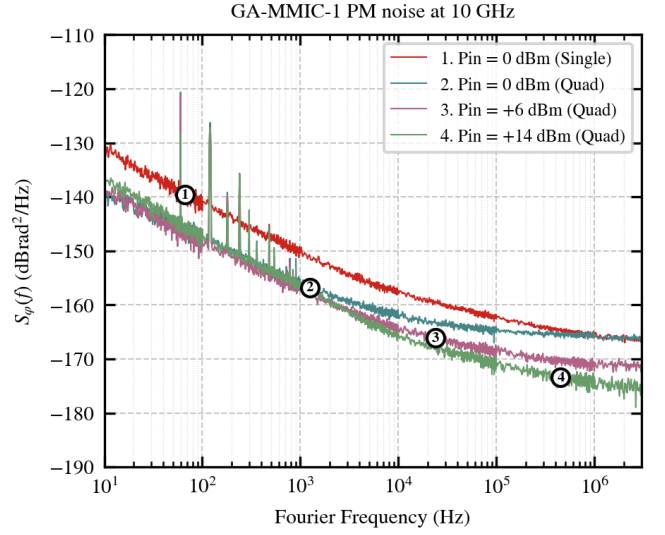


Fig. 9. Residual PM noise of GA-MMIC-1 in single and quad configurations.

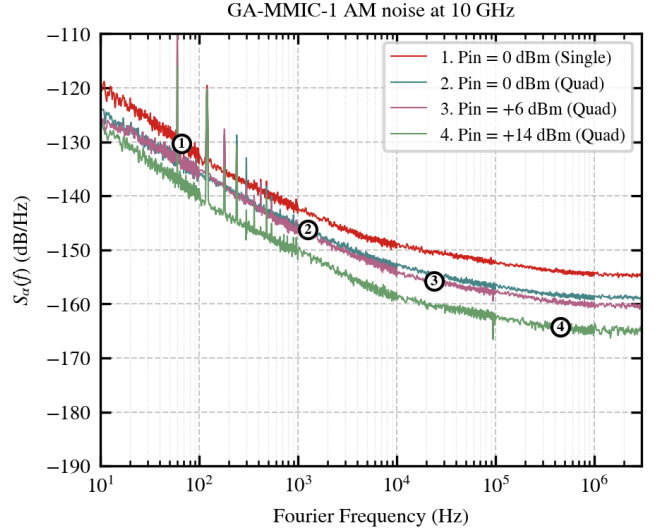


Fig. 10. Residual AM noise of GA-MMIC-1 in single and quad configurations.

noise of GA-MMIC-2 under two bias supply conditions. The baseline measurement uses a commercially packaged GA-MMIC-2 amplifier with the supplied commercial bias module, which provides the drain voltage (V_{DD}) and gate voltage (V_{GG}) through low-dropout (LDO) regulators. The improved measurement replaces this module with a NIST custom-designed low-noise supply built around ultra-low-noise LDO regulators. The dominant noise contribution of the commercial bias module is its V_{GG} regulator, whose voltage noise density at 1 kHz is ≈ 1700 times higher than that of the corresponding regulator in the custom supply. This excessive bias noise degrades AM and PM by 10 dB and 17 dB, respectively, relative to nominal residual noise levels, and the degradation profile mirrors the noise spectrum of the commercial V_{GG} reg-

ulator. Because the power supply-induced AM and PM noise levels are identical, the respective supply pushing conversion coefficients are likely equal, emphasizing the critical role of low-noise amplifier biasing.

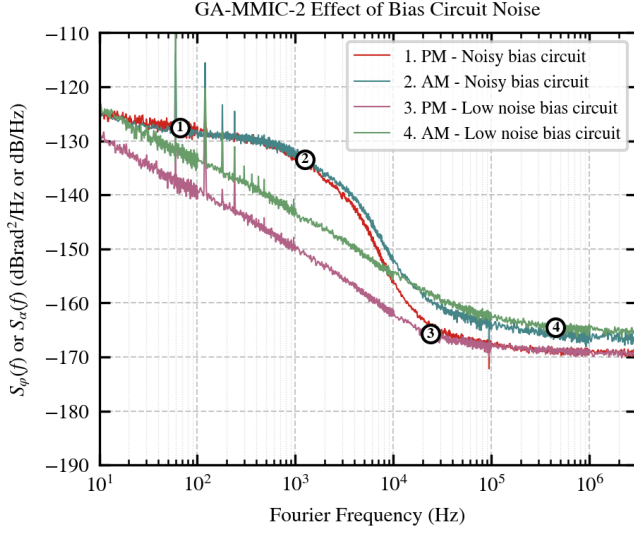


Fig. 11. Effect of voltage regulator on residual AM and PM noise of the GA-MMIC-2 (single configuration) at 10 GHz. Commercial noisy bias module vs. NIST custom low-noise bias supply.

H. Remaining Devices

Figures 12–20 show the residual AM and PM noise for the remaining devices tested in this study. Key observations follow.

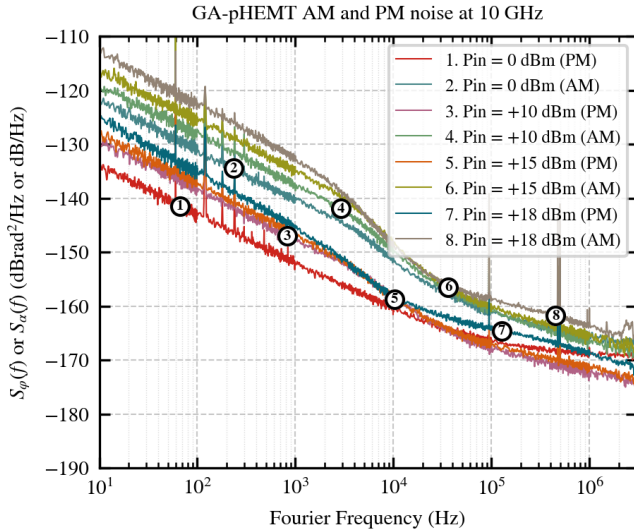


Fig. 12. Residual AM and PM noise of GA-pHEMT (GaAs pHEMT) in single configuration at various input power levels. The band-limited excess $1/f$ noise under 3 kHz offset may be a contribution of the manufacturer supplied bias circuit.

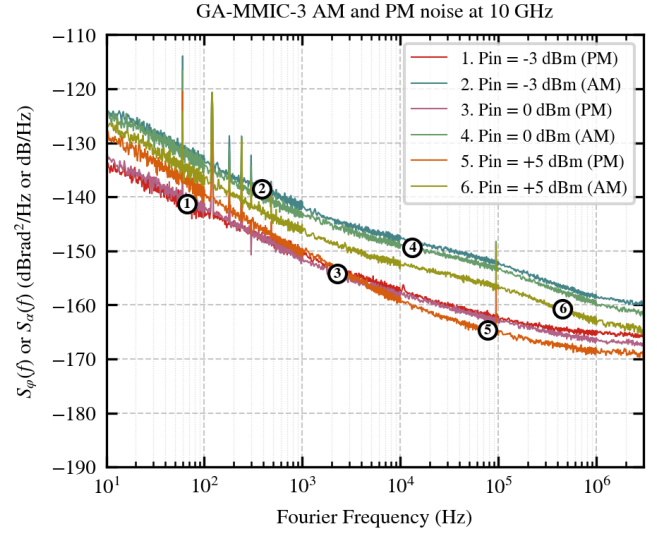


Fig. 13. Residual AM and PM noise of GA-MMIC-3 in single configuration. An increase in input power results in a degradation of the residual flicker PM, whereas the AM performance exhibits a corresponding improvement.

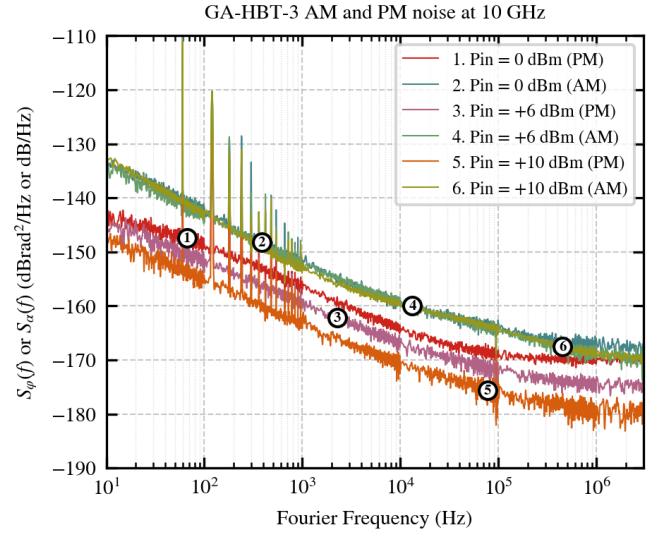


Fig. 14. Residual AM and PM noise of GA-HBT-3 in single configuration. At an input power of +10 dBm, AM noise is significantly higher than PM noise, exceeding it by more than 10 dB.

GA-pHEMT: This pHEMT amplifier is included as a process outlier. As a GaAs pHEMT distributed amplifier with $P_{1dB} = +28$ dBm, it operates at 0 dBm approximately 15 dB below its compression point, the largest headroom margin of any device in this study. Notably, for a pHEMT, the PM noise is still comparable to that of the mid-power HBT devices. For this device AM and PM noise degrades as it enters compression (Fig. 12). Also, excess $1/f$ noise near the carrier may be attributable to the commercial bias circuitry.

GA-MMIC-3: Consistent with the other devices of the same product family (GA-MMIC-1, GA-MMIC-2), GA-MMIC-3 shows AM noise above PM noise in the flicker as well as

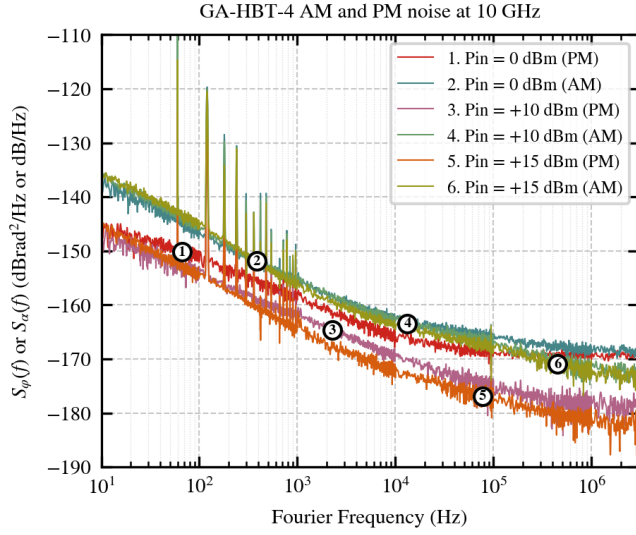


Fig. 15. Residual AM and PM noise of GA-HBT-4 in single configuration. The PM noise characteristics align closely with GA-HBT-3; however, GA-HBT-4 shows a marginal improvement in AM noise, measuring a few decibels lower across the offset range.

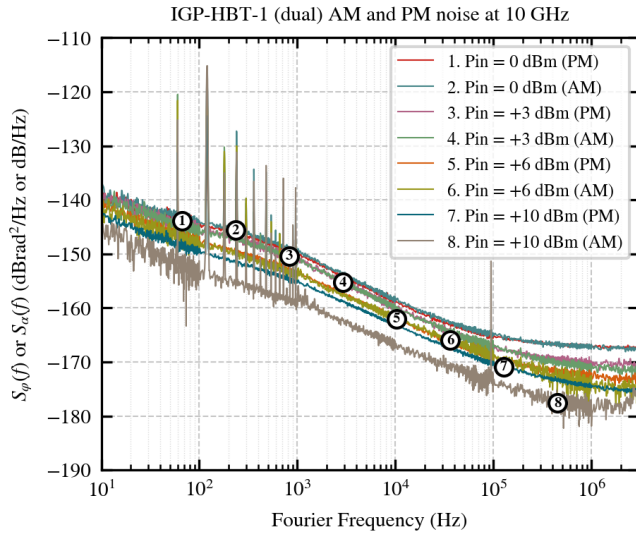


Fig. 16. Residual AM and PM noise of IGP-HBT-1 in dual configuration. For this amplifier, the AM and PM noise are similar for any given input power level less than +10 dBm. AM noise measurement at +10 dBm is biased by a comparable reference oscillator AM noise.

thermal regions (Fig. 13). Its gain of ≈ 18 dB is the highest of the three packaged MMICs tested. While entering saturation leads to an improvement in AM noise, it causes a degradation in PM noise. Within the flicker region, the AM and PM noise characteristics of a single GA-MMIC-3 are comparable to those of GA-MMIC-1 and GA-MMIC-2 (Figs. 5 and 6).

GA-HBT-3 and GA-HBT-4: These obsolete hermetic modules were characterized in single configuration (Figs. 14 and 15). GA-HBT-4, for instance, exhibits a phase noise of less than -163 dBrad²/Hz at 1 kHz offset, measured at P_{sat} at

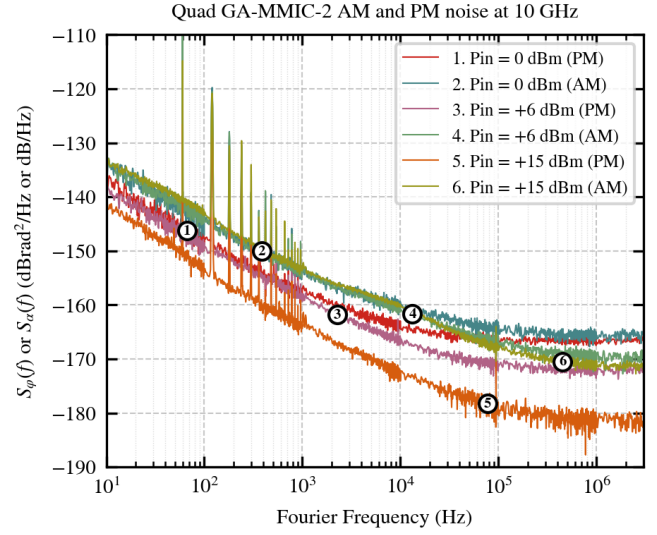


Fig. 17. Residual AM and PM noise of GA-MMIC-2 in quad configuration. For an input power level of +15 dBm, the AM and PM noise performance is comparable to that of quad GA-HBT-1.

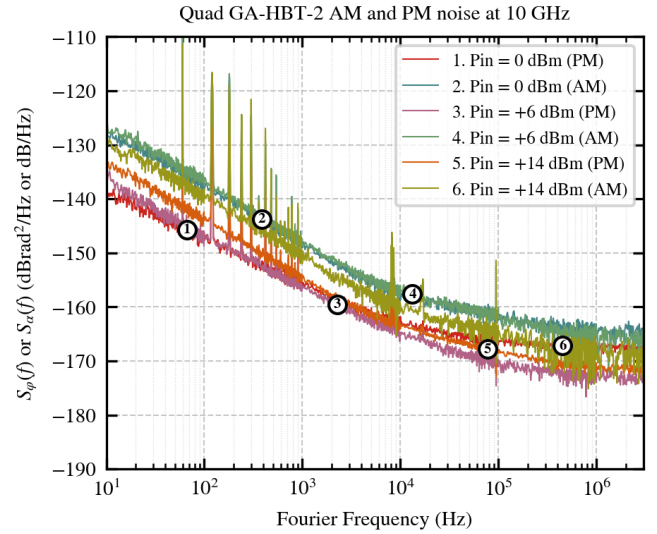


Fig. 18. Residual AM and PM noise of GA-HBT-2 in quad configuration. The noise is approximately 5 dB higher than that of the quad GA-HBT-1.

10 GHz (Fig. 15). For these amplifiers, PM improves with compression while AM noise stays relatively constant. Their noise is comparable to that of the GA-HBT-1 quad amplifier.

IGP-HBT-1 (Dual) array: The IGP-HBT-1 unit examined in this study is configured as a dual-amplifier array, utilizing a shared voltage regulator to bias both amplifiers in the array. In its linear operating region, this array demonstrates similar AM and PM noise levels, with both improving as input power increases. However, as the device enters deeper compression with higher input power, the AM noise performance begins to surpass PM noise, eventually approaching the reference oscillator's AM noise (Fig. 16).

GA-MMIC-2 (Quad): GA-MMIC-2 in quad configuration

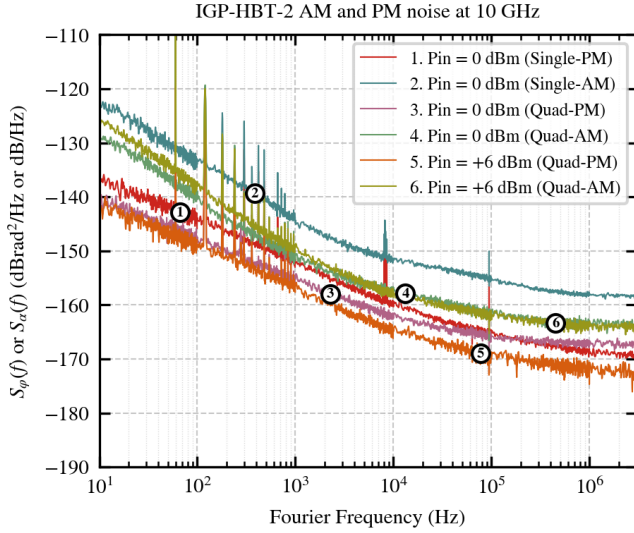


Fig. 19. Residual AM and PM noise of IGP-HBT-2 in single and quad configurations.

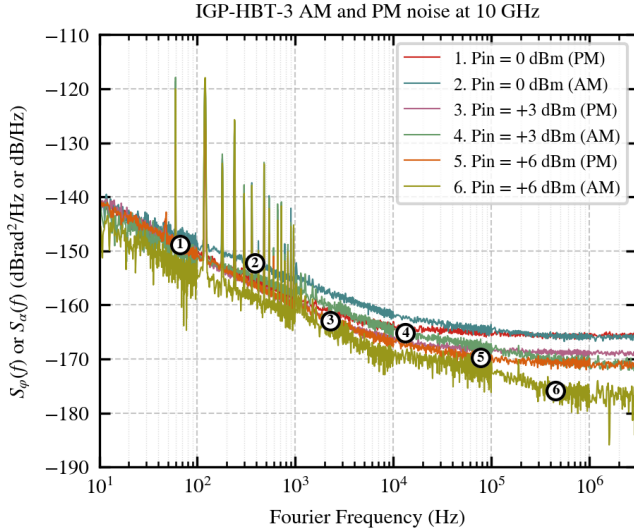


Fig. 20. Residual AM and PM noise of IGP-HBT-3 in array configurations. In linear operation the white AM and PM noise both match the thermal limit $N_F k_B T / P_{in}$; in deep compression the AM noise falls below this value. AM noise measurement at +6 dBm is biased by a comparable reference oscillator AM noise.

is presented in Fig. 17. It can produce almost 0.5 W of output power and its PM noise improves significantly in compression approaching $S_\phi(100 \text{ kHz}) = -180 \text{ dB/Hz}$. The flicker AM noise is independent of input drive level or compression.

GA-HBT-2 (Quad): Figure 18 shows the AM and PM noise of GA-HBT-2 in quad configuration. GA-HBT-2 is designed for the 6 GHz to 12 GHz band specifically, in contrast to the broadband (2 GHz to 30 GHz) GA-HBT-1. Despite its narrower operating range, its quad noise is approximately 5 dB higher than that of the quad GA-HBT-1 at a 1 kHz offset.

The expected 6 dB flicker improvement for AM and PM

noise is not fully met by the quad configuration unless the amplifiers are operating in saturation. However, saturation leads to degraded flicker PM but improved flicker AM noise (by a few decibels).

IGP-HBT-2 (Single and Quad): This InGaP HBT distributed amplifier in single and quad configuration is presented in Fig. 19. The IGP-HBT-2 quad configuration achieves $\approx 6 \text{ dB}$ improvement in AM noise over the single. Between 100 Hz and 1 kHz offset frequencies, this difference is approximately 7 dB, exceeding the ideal 6 dB improvement due to unit-to-unit performance variation. The close-to-the-carrier noise is comparable to the GA-MMIC-1 quad at comparable operating points. At a power level of 0 dBm, the PM thermal noise of the quad configuration is higher than that of the single amplifier, which suggests excess power splitter insertion loss.

IGP-HBT-3 (Octa) array: This discontinued amplifier consists of an array of eight parallel, dual-cascaded InGaP HBT amplifiers. It draws 27 W of DC power. The corresponding PM and AM noise characteristics are shown in Fig. 20. Notably, its flicker PM noise is comparable to GA-HBT-4 and the quad GA-HBT-1, while its AM noise aligns closely with that of the dual IGP-HBT-1. This similarity in AM noise suggests that a higher-order array of IGP-HBT-1 units could potentially achieve the AM noise level of IGP-HBT-3.

V. CONCLUSION

This study has characterized the residual AM and PM noise of twelve X-band, mostly HBT, amplifier designs at 10 GHz, identified by technology and package class, in single, dual, quad, and octa power-summed configurations. The principal findings are:

- 1) In a quad power-summed configuration, the discrete-die HBT amplifier GA-HBT-1 reaches below -160 dB/Hz residual PM noise and approximately -155 dB/Hz AM noise at 1 kHz offset, independent of input power, meeting the requirements of the target frequency synthesis application. The discontinued GA-HBT-4, GA-HBT-3, and IGP-HBT-3 modules measured residual PM noise between -156 dB/Hz and -159 dB/Hz at 1 kHz offset (Table I).
- 2) IGP-HBT-3, GA-HBT-4, GA-HBT-3, and IGP-HBT-1 measured AM noise between -149 dB/Hz and -155 dB/Hz at 1 kHz offset (Table I). Several devices show AM noise 5 dB to 10 dB above their own PM noise. This runs counter to the common system-budget assumption that AM and PM noise are equal in the thermal region, and it matters because AM noise enters the phase noise budget through AM-to-PM conversion at downstream nonlinear stages. However, under saturated conditions, IGP-HBT-3 and IGP-HBT-1 exhibit a reversal with AM noise falling below their PM noise.
- 3) The $10 \log_{10}(n)$ array improvement is approached but not fully reached: against the ideal 6 dB for a quad configuration. We typically observed 3 dB to 5 dB, limited primarily by excess splitter insertion loss, with unit-to-unit performance variation and impedance match

also contributing. Typically, the performance of a power summed amplifier improves upon the noise of the individual amplifiers in the flicker region. For amplifiers with excess AM noise, the improvement extends into the white noise region as well because this excess AM noise is not additive in nature.

- 4) Both AM and PM noise depend on input drive level, but the trends are device specific (Figs. 12, 14 and 16). In the GA-pHEMT both components increase with input power; GA-HBT-3 shows improving PM noise while its AM noise remains constant; and the dual IGP-HBT-1 improves in both AM and PM as input power increases.
- 5) The voltage regulator is a first-order noise parameter, not a secondary effect. Replacing the supplied commercial bias module with a NIST low-noise supply (Sec. IV-G) reduces both the PM and AM noise of GA-MMIC-2 by more than 10 dB at 1 kHz offset. The dominant contribution originates from the V_{GG} rail, where the supplied regulator's noise density exceeds that of the low-noise replacement by three orders of magnitude. This is critical because V_{GG} couples directly into the bias current. For any application requiring noise below -150 dBrad²/Hz at sub-kilohertz offsets, the choice of voltage regulator is as consequential as the choice of amplifier, and the lowest-noise regulator should be allocated to the V_{GG} rail first.
- 6) No commercial manufacturer currently specifies residual AM noise on a standard datasheet. This study reports measured X-band HBT amplifier AM and PM noise organized by device technology and topology, and is intended to serve as a reference for noise budgeting in precision frequency synthesis, radar, and timing applications.

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DISCLAIMER

In accordance with NIST policy, the commercial devices characterized in this work are identified only by semiconductor technology and package class. No manufacturer or part number is given, the results are not a ranking, and no endorsement of any commercial product by NIST is implied.

APPENDIX

ANALYTICAL MODEL OF RESIDUAL AM AND PM NOISE MEASUREMENT

This appendix demonstrates the AM $\leftrightarrow$ PM leakage paths in mixer-based residual noise measurements by use of a minimal arctangent model of mixer port saturation, chosen over more complete models [46], [47] for mathematical simplicity. The phase-to-voltage and amplitude conversion factors are

treated as calibration constants, and the results are expressed in terms of the port overdrive ratios and the phase error $\delta\theta$. A memoryless odd nonlinearity transmits carrier phase exactly and therefore contains no intrinsic AM-to-PM conversion. Physical mixers possess additional mechanisms, notably envelope-dependent junction capacitance and drive-dependent self-bias that convert AM to PM even at exact quadrature [47]. For simplicity, this analysis omits these additional non-linear effects. Consequently, the leakage terms derived here represent lower bounds only due to the quadrature phase error.

The reference oscillator output voltage after the power splitter (Fig. 1) is modeled as a carrier with small fractional amplitude fluctuations, α_s , and phase fluctuations, φ_s ,

$$v(t) = A[1 + \alpha_s(t)] \cos[2\pi f_0 t + \varphi_s(t)], \quad (\text{A1})$$

with $|\alpha_s| \ll 1$ and $|\varphi_s| \ll 1$ rad. Here, A is the nominal peak amplitude and f_0 is the nominal frequency of the oscillator. After passing through the DUT, the signal acquires the additional residual fluctuations $\alpha_d(t)$ and $\varphi_d(t)$,

$$v_1(t) = A_1[1 + \alpha_s(t) + \alpha_d(t)] \cos[2\pi f_0 t + \varphi_s(t) + \varphi_d(t)]. \quad (\text{A2})$$

The reference arm delivers a copy of the source with a phase offset θ set by the quadrature phase shifter,

$$v_2(t) = A_2[1 + \alpha_s(t)] \cos[2\pi f_0 t + \varphi_s(t) + \theta]. \quad (\text{A3})$$

Both arms carry the identical source fluctuations $\alpha_s(t)$ and $\varphi_s(t)$; we assume a delay-matched system throughout.

The saturating voltage response of a mixer port to an input voltage x is modeled as

$$F(x) = \frac{2V_f}{\pi} \arctan\left(\frac{\pi x}{2V_f}\right), \quad (\text{A4})$$

normalized to unit small-signal slope and saturating asymptotically to the hard limits $\pm V_f$, where V_f is the saturation voltage of the port. Each port is characterized by its overdrive ratio,

$$\xi = \frac{A}{V_f}, \quad \xi_{\text{RF}} = \frac{A_1}{V_{f,\text{RF}}}, \quad \xi_{\text{LO}} = \frac{A_2}{V_{f,\text{LO}}}, \quad (\text{A5})$$

with $\xi \ll 1$ corresponding to linear operation and $\xi \gg 1$ to hard limiting. Because F is odd and memoryless, the carrier zero crossings, and hence the instantaneous phase, are transmitted exactly at any drive level, whereas the envelope is not. The detector response is dependent on the fundamental Fourier amplitude $a_1(\tilde{A})$ of the port output, evaluated at the quasi-static envelope $\tilde{A} = A[1 + \alpha]$. Expanding a_1 to first order in α identifies the fractional AM transmission coefficient, η ,

$$a_1(A[1 + \alpha]) \approx a_1(A) \left[1 + \underbrace{\frac{A a_1'(A)}{a_1(A)}}_{\eta} \alpha \right]. \quad (\text{A6})$$

Evaluating a_1 for the model (A4) gives the closed form

$$\eta(\xi) = \frac{1}{\sqrt{1 + (\pi\xi/2)^2}}, \quad (\text{A7})$$

which approaches unity in linear operation, falls as $2/(\pi\xi)$ in hard limiting, and vanishes only for an ideal limiter. Saturation therefore compresses, but never eliminates, the AM transmitted to the detector.

The phase detector multiplies its two port signals and after low-pass filtering the product retains the difference-frequency term. With the transmitted port envelopes written as $m_1(t)$ and $m_2(t)$, the baseband output is

$$v_{\text{out}}(t) = k m_1(t) m_2(t) \cos[\varphi_d(t) - \theta], \quad (\text{A8})$$

where k is the operating point dependent conversion constant obtained by calibration (k_φ in V/rad for the PM configuration, k_α in volts for the AM configuration), $m_1 = 1 + \eta(\xi_{\text{RF}})(\alpha_s + \alpha_d)$ and $m_2 = 1 + \eta(\xi_{\text{LO}})\alpha_s$. The source PM noise φ_s enters both ports identically and cancels exactly in (A8), which is the fundamental advantage of the homodyne measurement topology. In practice the phase setting contains a small phase error, $\theta = \theta_0 + \delta\theta$ with $|\delta\theta| \ll 1$ rad.

In the PM configuration both ports are saturated, so setting $\theta_0 = \pi/2$ in (A8) and expanding to first order gives

$$v_{\text{out}}^{\text{PM}}(t) = k_\varphi [\varphi_d(t) - \eta(\xi_{\text{RF}})\alpha_d(t)\delta\theta - (\eta(\xi_{\text{RF}}) + \eta(\xi_{\text{LO}}))\alpha_s(t)\delta\theta]. \quad (\text{A9})$$

The DUT AM noise leaks only through the RF port, as indicated by the second term in (A9). The source AM noise leaks through both ports, and because both contributions are driven by the same fluctuation they add coherently in amplitude. For our simplified saturation model, both incomplete saturation ($\eta \neq 0$) and a phase error ($\delta\theta \neq 0$) are required simultaneously for AM noise to contaminate the PM channel at first order.

In the AM configuration the LO port is saturated and the RF port is linear ($\eta \rightarrow 1$), so $m_1 = 1 + \alpha_s + \alpha_d$. Setting $\theta_0 = 0$ and expanding similarly gives

$$v_{\text{out}}^{\text{AM}}(t) = k_\alpha [1 + (1 + \eta(\xi_{\text{LO}}))\alpha_s(t) + \alpha_d(t) + \varphi_d(t)\delta\theta]. \quad (\text{A10})$$

The PM-to-AM leakage term $\varphi_d\delta\theta$ carries no η factor, the in-phase operating point sits at the maximum of the detector characteristic rather than at a zero crossing. Therefore, a small phase error introduces PM-to-AM leakage regardless of the saturation state of the ports. The source AM noise enters with coefficient $1 + \eta(\xi_{\text{LO}})$, as indicated by the second term in (A10). Unlike PM measurements there is no differential cancellation mechanism for source AM noise.

Collecting the first-order contributions, and for simplicity assuming all noise terms are uncorrelated, the measured one-sided double sideband (DSB) noise PSDs are

$$S_\varphi^{\text{meas}}(f) = S_{\varphi_d}(f) + \delta\theta^2 \eta^2(\xi_{\text{RF}}) S_{\alpha_d}(f) + \delta\theta^2 [\eta(\xi_{\text{RF}}) + \eta(\xi_{\text{LO}})]^2 S_{\alpha_s}(f), \quad (\text{A11})$$

$$S_\alpha^{\text{meas}}(f) = S_{\alpha_d}(f) + (1 + \eta(\xi_{\text{LO}}))^2 S_{\alpha_s}(f) + \delta\theta^2 S_{\varphi_d}(f), \quad (\text{A12})$$

with $\eta(\xi)$ from (A7) and the port overdrive ratios from (A5). Even in the hard-limiting limit $\eta(\xi_{\text{LO}}) \rightarrow 0$, the source AM

term in (A12) retains unit coefficient, so valid residual AM characterization requires

$$S_{\alpha_s}(f) \ll S_{\alpha_d}(f) \quad (\text{A13})$$

across the offset range of interest, a constraint with no analogue in the PM measurement.

Equations (A11) and (A12) constitute the principal results of this appendix, leading to four primary conclusions. First, the source PM noise is absent from both expressions at all saturation levels. Second, AM noise contaminates the PM channel when both $\eta(\xi) \neq 0$ and $\delta\theta \neq 0$ simultaneously. Third, PM noise contaminates the AM channel whenever $\delta\theta \neq 0$, regardless of saturation depth. Finally, even under ideal conditions, the AM noise measurement retains the source term S_{α_s} through the linear RF port. This highlights a fundamental asymmetry between the two modes: source PM cancels differentially, whereas source AM does not. Consequently, a valid residual AM noise measurement using Fig. 1 strictly requires $S_{\alpha_s} \ll S_{\alpha_d}$. Alternative measurement schemes do exist that utilize differential architectures to effectively suppress reference-oscillator AM noise [34], [48].

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