PA150 - 150W+ HF Linear Power Amplifier - EVA02

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Engineering Reporting - Linear Amplifiers | ROWAVES®

Abstract
The primary goal of this document is to provide essential experimental and theoretical data (performance measurements) generated by the evaluation process of the power amplifier test board (EVA, setup by ROWAVES®) in discussion, along with similar results obtained by related work. The design and original reference files do not belong to ROWAVES® and the referenced designs have been clearly mentioned in the reference section of this paper. The current paper's work is based on Steve Drury G6ALU modifications of the AN762 design [1]. Communication Concepts, Inc. also offers a great construction hints paper related to this work, on their website [2], paper that can be downloaded freely, along with other papers.

I. BACKGROUND

The original design files related to this paper are referring to Motorola Application Note AN762, by Helge O. Granberg, RF Circuits Engineering Dep. at Motorola, back in 1993, later transferred to M/A-COM. AN762 design was using MRF4xx series bipolar high-power transistors, delivering different power levels to the load and different IMD performance (see reference for details). Substrate of the original design was also FR-4, architecture is class-AB, biased at 150mAdc/transistor [3]. The paper AN762 extensively discusses the DC and AC performance of various Motorola MRF4xx series transistors, used to build 3 (three) versions of the amplifier. All details can be found in the mentioned paper. We will focus our work in this document on:

- general design consideration details
- practical implementation of the G6ALU work in ROWAVES® version
- IMD performance measurements

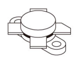
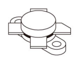
Impedance matching networks are employed in the circuitry, therefore power gain and linearity are usually sacrificed. Input correction networks can be designed with RC/RLC combinations to give better than 1 dB gain flatness across the band with low input VSWR [3].
Contrary to popular belief, there are several precautions and design hints to be taken into consideration regarding transistor amplifiers [2]:

- suppress circuit oscillation (incl. self oscillation), as it can lead to excessive power dissipation or cause the device to exceed its breakdown voltage limits, for active (e.g. transistors) components
- limit the power supply current for biasing the power transistors (if there is a major design failure on exceeded bias current, the malfunction can be avoided by limiting the current)
- implement protective circuitry, such as fast acting ALC
- ensure proper mechanical attachment of the device to a heatsink, using thermal paste (comparison details will be presented further in this paperwork).

II. HG2078 / HG1487 RF TRANSISTORS

Regarding the bipolar NPN transistors used in the design (HuaGao versions of 2SC2879 / SD1487) the recommended minimum collector quiescent current for class-AB operation is approx. 150 mA/each transistor. Higher values may be used, but this will reduce collector efficiency. Alternatively, the device may be operated in Class-A [...] with a quiescent current roughly equal to one-fourth of the maximum rated collector current.

Table 1 *- Gain vs. IMD3, multiple manufacturers*

Parameters	**HG2879 / 2SC2879** 	**HG1487 / SD1487** 
Gp (HuaGao DS)	15.2dB	15.2dB
Gp (Toshiba DS)	13dB	-
Gp (ST DS)	-	12dB
IMD3 max. (HuaGao DS)	-24dBc	-24dBc
IMD3 max. (Toshiba DS)	-24dBc	-
IMD3 max. (ST DS)	-	-30dBc

The original Toshiba 2SC2879 (even NOS) were completely unavailable to us while evaluating this design. The same is valid for the ST SD1487 version. So we have used the HuaGao versions of the famous transistors mentioned, this manufacturer being reliable and verified by ROWAVES while using over 340 pieces of 2SC2879s, for our PA100-D 100W+ amplifier, between 2019 - 2025 [4].
While this guideline is not formally specified for class-AB operation, it generally applies to most RF power transistors. Typical IMD distortion products are −31 to −33dBc below one of the two test tones [3] with a 13.6Vdc supply. Greater the figure of merit (ratio of emitter periphery and base area), the greater the power gain [2].

For a pair of HG1487, this design/evaluation paper demonstrates that 150…180W can be easily obtained with a gain of approx. 11…12dB typ. (2…30MHz). For a pair of HG2879, also 130…180W of power can be obtained with a typ. gain of approx. 12.5dB, with higher gain levels of up to 15dB at lower freq domain. (160m…80m).

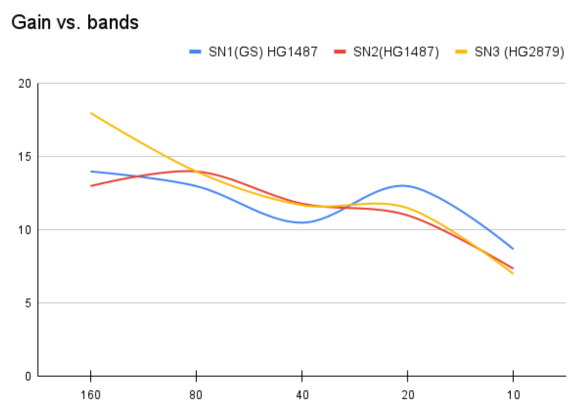


Figure 1 *- Gain vs. bands, for 3 x PA150+ amplifier units*

The design was evaluated based on 3 (three) samples, 2 (two) of which use HuaGao HG1487 devices, and one uses a pair of HG2879s. The purpose of variation was to compare and evaluate different types of transistors manufactured by the same vendor (HuaGao). Before starting to describe the architecture used in the original design of AN752 App Note and G6ALU approach [1], we would like to point out the performance obtained in terms of IMD3 at P1dB (approx. 160W) is within the legal limit of -30dBc from 160m to 40m and better than -25dBc from 20m to 10m. There is a slight variation of the IMD3 performance: 160 to 80m IMD3 for HG2879 is better than IMD3 of HG1487 with 2…3dB (same test conditions).

III. SCHEMATIC / ARCHITECTURE

For complete bias circuitry the G6ALU design uses a DIL-14 package of MC723CN voltage regulator (**see Annex 1**). The advantages of using such a circuit are line voltage regulation capability, low stand-by current, (1.0 mA) and wide range of voltage adjustability and stability [5]. This stability with thermal tracking is crucial in class-AB designs where even small shifts in bias voltage can impact linearity and IMD performance. With the component values shown, the bias voltage is adjustable from 0.5 to 0.9 Volts, which is sufficient from class-B to class-A operating conditions.
For the following calculation of collector current, bias current and other parameters, we will be starting with the following h_{FE} table summary:

Table 2 *- Typ.* h_{FE} *factor values, used for calculations*

h_{FE}	SD1487	2SC2879	MRF421
min.	20	15	15
typ.	45	35	30
max.	70	60	40

As a general rule, in class-B the bias voltage is equal to the transistor V_{BE} , and there is no collector idling current present (except small collector-emitter leakage, ICES), and the conduction angle is 180°. In class-A the bias is adjusted for a collector idling current of approximately one-half of the peak current in actual operating conditions, and the conduction angle is 360°. In class-AB (common for SSB modulations) the bias is set for a low collector quiescent current, and the conduction angle is [...] higher than 180° [5].
Some basic calculations for approximating the transistor's base bias current:

$$I_B = \frac{I_C}{h_{FE}} \text{ (eq.1)}$$

where:
I_c = desired collector quiescent current (bias current)

h_{FE}= DC current gain (typical for RF BJTs: 15–60 depending on the part)

Assuming an efficiency of abt. 50% and Pout of approx. 160W per amplifier the collector current is:

$$I_c = \frac{2 * P_{out}}{V_{CC}} = \frac{320}{13.8} = 23.2Adc \quad \text{(eq.2)}$$

$$I_B = \frac{I_C}{h_{FE}} = \frac{23.2}{45} = 0.51Adc \quad \text{(eq.3)}$$

R21 shares the dissipation with TR1, and its value must be such that the collector voltage never drops below approximately 2Vdc. R21 was approximated with 22R /5W, based on the calculations below:

$$R_{21} = \frac{V_{CC} - 2}{I_B} = \frac{13.8-2}{0.51} \sim 23.2\Omega \quad \text{(eq.4)}$$

This value will cover also high power transistors with h_{FE} values close to 40...50, at the higher end.

R11 determines the current limiting characteristics of the LM723, and 1 Ω will set the limiting point to approx. 0.65 Adc, ± 10%.

Temperature sensing "diode" Q3 is added for bias tracking with the RF power transistors. The base-emitter junction of a 2N5190 or similar device (e.g. BD135) can be used for this purpose. The temperature tracking within 15°C to 60°C is achieved [...] Extensive detailed design aspects can be found in reference [5]

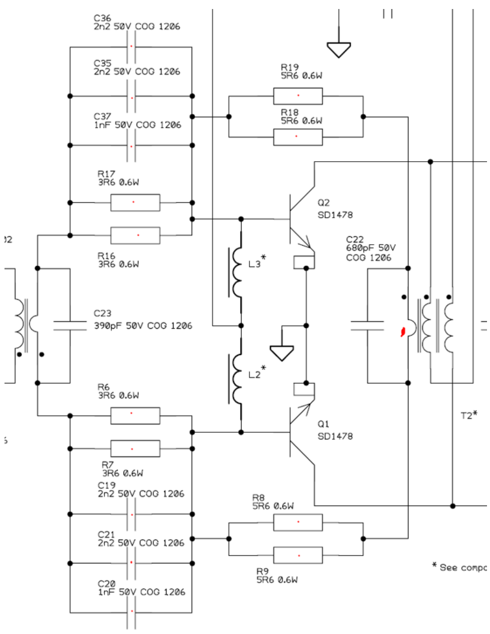


Figure 2 - *Close view of the input matching network*

Moving forward with the input frequency correction network implemented (R6||R7, R16||R17, C19||C20||C21, C35||C36||C37) together with the negative feedback network (single winding of T2, with R8||R9 and R18||R19) forms an attenuator with frequency selective characteristics (Fig. 2). All components involved in this network have been marked with a red dot in Fig.2. The design of this input attenuator compensation network is extensively discussed in reference [8].

Regarding the broadband transformers design, extensive notes and practical design and implementation is discussed of course in [5] and more on [8]. The input transformer T1 and the output transformer T3 are of the same basic type, with the low impedance winding consisting of two pieces of metal tubing, electrically shorted in one end and the opposite ends being the connections of this winding. Similar type of transformers that can be used:

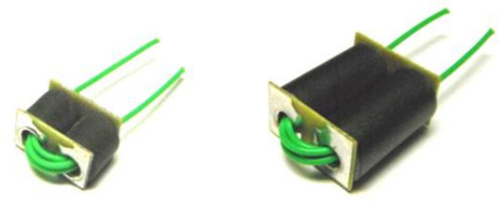


Figure 3 *- Close view of the input T1, 16:1 (left) and output (T3), 25:1 transformers © Communication Concepts, Inc. [2]*

The multiturn high impedance winding is threaded through the tubing so that the low and high impedance winding connections are at opposite ends of the transformer [...] The coupling coefficient between the primary and secondary windings is determined by the length-to-diameter ratio of the metal tubing [...], and the gauge and insulation thickness of the wire used for the high impedance winding. For high impedance ratios (36:1 and higher), miniature coaxial cable where only the braid is used, leaving the inner conductor disconnected, gives the best results.

The high coefficient of coupling is important only at the high-frequency end of the band, e.g. 20 to 30 MHz. [5]

Simplified design formulas are necessary to approximate the minimum necessary inductance of the T1 and T3 transformers. Both are loaded with ferrite materials to provide sufficient low-frequency response. The minimum required inductance in the one turn winding can be calculated as:

$$L = \frac{R}{2\pi F} \quad \text{(eq. 5)}$$

where:
L = required inductance in µH
R = base-to-base or collector-to-collector impedance
F - lowest application frequency, in MHz

For example, in a 160W application amplifier, the input transformer impedance ratio is 16:1, making the secondary impedance 3.13Ω, on a 50Ω load:

$$L = \frac{3.13}{6.28\,x\,1.6MHz} = 0.31\mu H \text{ (eq. 6)}$$

For the output transformer having a 25:1 impedance ratio to a 50 Ω interface:

$$L = \frac{2}{6.28\,x\,1.6MHz} = 0.2\mu H \text{ (eq. 7)}$$

It should be noted that in the lower power versions (e.g. 100...120W), where the input and output impedances are higher and the transformers have lower impedance ratios, the required minimum inductances are also higher. T2, the collector choke supplying the DC to each collector, also provides an artificial center tap for T3. This combination functions as a real center tapped transformer with even harmonic cancellation. T2 provides a convenient low impedance source for the negative feedback voltage, which is derived from a separate one turn winding. T3 alone does not have a true AC center tap, as there is virtually no magnetic coupling between its two halves [...]. Additional information on these transformers can be found in reference [9].

IV. IMPLEMENTED SCHEMATIC MODIFICATIONS

The original AN762 design does not use an input attenuator "pi-matching network". However, G6ALU design has implemented such an attenuator on the input stage. The claimed value is 3dB/8...10W/50Ω, according to Annex 1. We had to increase the power capabilities of the resistors currently used in the input attenuator. The schematic is showing a 8..10W/3dB attenuator network but the actual real power capability is at 40% (approx. 3..4W). As a result, for drive levels of 37...40dBm, the excessive overheating of the resistors have determined us to recalculate/redesign the input attenuator for 1dB/6...8W drive level. This was ensured by using 3W resistors. A 10µH inductor has been installed to supply the bias point (12 Bias 12V to Tx, see Annex 1) from the 12Vdc main supply line to simplify the biasing of the transistor stage.

V. EXTENSIVE PRACTICAL CONSIDERATIONS

The original PCB layout (Annex 3) was designed for direct coaxial cable connections at the board level, without any RF connector footprints, which made interconnections challenging during testing.

a) we have placed SMA-F connectors, straight (for RF In) and edge, long variant (for RF Out) for ease of interconnection and measurements
b) we changed the original design of the T1 (BN43-202) transformer from coaxial braid to copper tubing, with improved performance and easy implementation, in terms of transformer inductance mutual coupling
c) we have installed male-faston tabs for easy and fast high-current capability DC interconnection with power cables (+/- lines)
d) a simple DIL-14 AUGAT DIL adapter installation can prevent difficulty of IC1 replacement
e) C0G/NP0 1206/0805 SMD capacitors have been installed on the bottom side of the PCB for proper RF performance

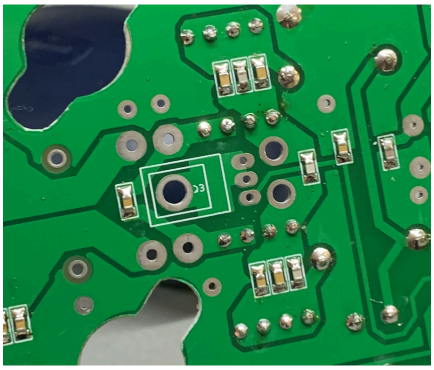


Figure 4 *- Close view of the bottom of the PCB, C0G/NP0 capacitors*

One of the recommended (see Chap. VI) heatsink should be marked, drilled, and tapped before construction begins on the PC board. Using the bare PC board as a pattern, lay the PC board on the heatsink in the desired position and mark the four mounting holes and the mounting hole for D1. Drill and tap the four mounting holes and one hole for D1, for M3 screws. Then bolt the bare PC board flush on the heatsink and place the RF transistors through the PCB so they rest on the heatsink surface. Mark the mounting holes for the

RF transistors. Remove the PC board from the heatsink and drill and tap the mounting holes for the RF transistors (M3 screws). The heatsink can also be fitted with connectors, cover etc. at this time if so desired [2]. The resistive losses in the collector DC voltage path should be minimized and the G6ALU variant has implemented modifications in terms of copper trace width for the power line traces of the BJT's collector paths.

VI. THERMAL DESIGN CONSIDERATIONS

The thermal design is based on worst-case operating conditions. An amplifier efficiency of 48% is assumed, resulting in a total power dissipation of 162.5 W. Considering that the amplifier operates in telegraphy mode where each word may contain up to six elements (five dashes with pause between them), the effective duty cycle is estimated at 65.3%. This leads to an average dissipated power of 107 W, or approximately 53.5 W per transistor.

$$P_{DC = 65.3\%} = \frac{0.653 * 162.5}{2} = \frac{106.11}{2} = 53.05\,W$$
(eq. 8)

The junction-to-ambient thermal resistance is calculated using the following formula:

$$R_{\theta JA} = \frac{T_J - T_A}{P} = \frac{150-25}{53.05} = 2.35°C/W \text{ (eq. 9)}$$

$R_{\theta JA}$= junction-to-ambient thermal resistance (°C/W)
T_J = maximum junction temperature (150°C, from datasheet)
T_A = ambient temperature (25°C)

The heat sink-to-ambient thermal resistance, the following formula is used:

$$R_{\theta SA} = \frac{R_{\theta JA} - (R_{\theta JC} + R_{R\theta CS})}{2} = \frac{2.35-(0.6+0.1)}{2} = 0.825°C/W$$
(eq. 10)

$R_{\theta SA}$= junction to ambient thermal resistance (°C/W)

$R_{\theta JC}$= device junction to case thermal resistance (0.6 °C/W from datasheet)
$R_{\theta CS}$ = case to heatsink thermal resistance (0.1°C/W)

The obtained value $(0.815°C/W)$ represents the maximum thermal resistance of the heat sink at which

this amplifier can operate without failure, with the heat sink temperature reaching 112°C.

$$T_{sink} = T_{amb} + P * R_{\theta SA} = 25 + 107 * 0.825 = 113.2°C$$
(eq. 11)

T_{sink} = heatsink temperature (°C)
T_{amb}= ambient temperature (°C)
P = total power (W)

Based on this, a maximum junction to ambient thermal resistance of 2.35°C/W was calculated to ensure that the junction temperature remains below the safe limit of 150°C. Subtracting the known junction-to-case and case-to-heatsink resistances yields a maximum allowable heatsink thermal resistance of 0.825°C/W.
To meet these thermal requirements and ensure reliable long term performance, heatsink(s) with equal or lower thermal resistance such as the ones listed below should be used:

1. Fischer SK-580-100-SA ($R_{\theta SA}$ =0.55°C/W)
2. Aavid Engineering 60140 ($R_{\theta SA}$ =0.55°C/W)
3. Fischer SK121-150SA ($R_{\theta SA}$ =0.62°C/W)

VII. PERFORMANCE AND MEASUREMENTS

Some important measurements have been performed to analyze the behaviour of the amplifier in terms of power output, IMD, harmonics, input matching / return loss, gain / gain flatness etc. For each parameter we will present full scale spectrum analyzer or VNA plots, or excel computed plots etc.

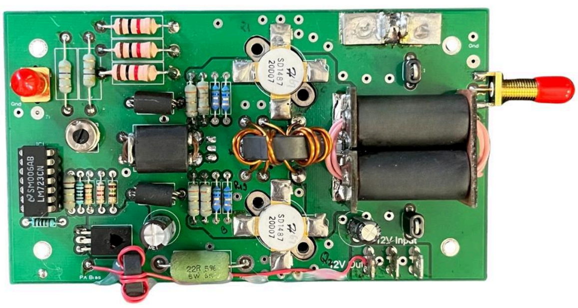


Figure 5 *- 2 x SD1487, ROWAVES variant - final assembly*

Summary of the performance results:

- P1dB ~ 180W CW
  Pout typ. 160W CW (IMD3 meas. performed at this power levels, 160-20m)
- IMD@ 10m - performed at 125…150W
- IMD3: > 30dBc (160m...40m) and > 25dBc, (20m-10m)
- gain, 12...13dB typ. (160...10m)
- power supply: 13.8Vdc/20Adc (typ.), max. 25Adc (peaks)
- input levels: 1…10W max.

Gain distribution over frequency has been presented in Fig.1 at the beginning of this paper. IMD3 behaviour is presented in Fig. 6 below. For this variant, total gain flatness is around approx. ± 5dB which is not a good number for such amplifier(s), and it can be significantly improved by both frequency and gain compensation methods. [3] [5]

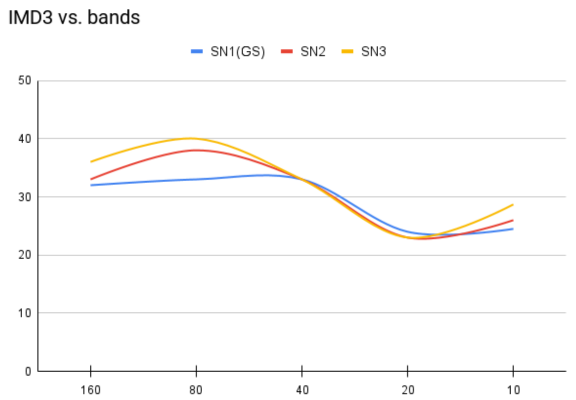


Figure 6 *- IMD3 vs. bands, for 3 x PA150+ amplifier units*

A simple power level plot in Fig. 6 below is reflecting the fact that up to 160…170W can be easily obtained with this kind of amplifier architecture.

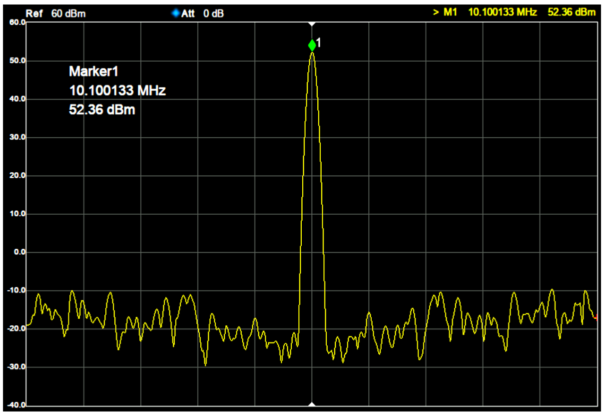


Figure 7 *- 170W power levels at 10MHz, 13.8Vdc/22Adc*

At 30℃ ambient temperature, for a test of 60s full CW at 160W, the external temperature of the heatsink reaches 50℃ and it stabilizes at 60℃ after 5 min. This condition is crucial for the amplifier to be operated at normal parameters.
We have used a portable digital thermal camera, the Ulefone Power Armor 19T with FLIR™ Lepton 3.5 Sensor, for thermal measurements during tests and to evaluate the full load thermal capabilities. The top ceramic temperature of the BJTs surface reaches 60..70℃, after 5 min. CW full power capability, for all 3 samples that were used for testing and evaluation.

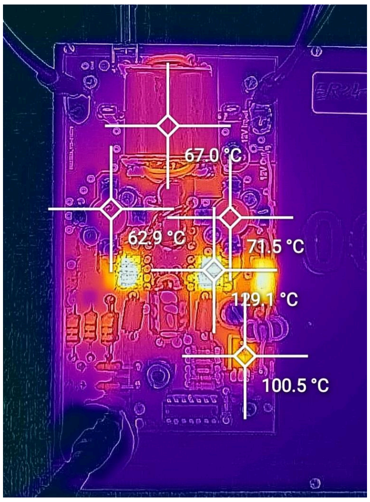


Figure 8 *- Thermal image of SN2 after 5 min. of CW testing at 14MHz*

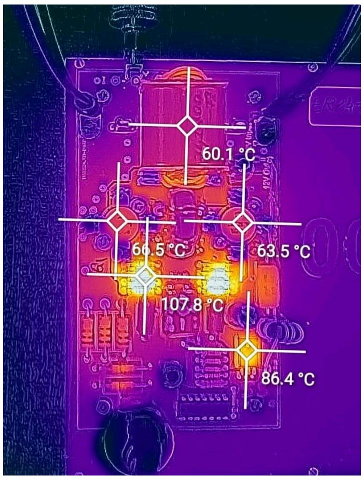


Figure 9 *- Thermal image of SN3 after 5 min. of CW testing at 14MHz*

The IMD measurements have been performed using our versatile IMD-TP1 testing platform that has been extensively used for PA1000 (HF+6m Linear Amplifier) testing and can generate reliable results in terms of repeatability and performance. Currently, the IMD-TP1 can perform IMD measurements for linear PAs up to 1500W, HF to VHF, ranging 1MHz to 220MHz.

Figure 10 *- IMD-TP1 ROWAVES testing platform used for IMD characterisation*

The test platform is able to perform 80/20/6m tests now based on current filtering capabilities, and with external LPF it can perform any frequency/band testing from 630m to 1.25m bands, being an advanced automatic test system (USB PC controlled + recording/plotting interface) [10].
Complete test setup for the PA150+ IMD3 measurements is illustrated in picture below:

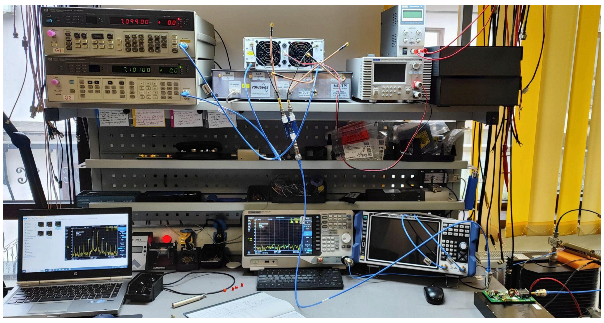


Figure 11 *- Complete IMD test setup, similar with the one used for PA1000+ testing [10]*

As for PA1000+, the test setup consists of:

1. 2 x HP865x series low phase noise signal generators [11]
2. RTLPF-520-xxx, 160/80/40/20/10m, 5th or 7th order LPF set e.g. [12]
3. ULTA-10R3, ultra linear Class-A power amplifier (2 x 10W modules inside), -65dBc 2nd harmonic / -57dBc 3rd harmonic, with IMD3 at -70dBc [13]
4. JUMA SC-2XD high power combiner, >33dB P-to-P isolation, max.0.35dB loss@ 100MHz [14]
5. PA150+ modules - available for sale also [15]
6. 30…50dB DDC / DC - Directional Coupler - e.g. Werlatone™ C5425 for high power RF sampling
7. spectrum analyzer, at least 100dB dynamic range e.g. Siglent SA3021X Plus - with various front high-power attenuators (e.g. 20…40dB) for preventing overloading of the SA front-ends
8. high-performance 50Ω dummy load (e.g. Spinner BN537761)

IMD performance on 160m is reaching -37dBc, 160W+ of CW power:

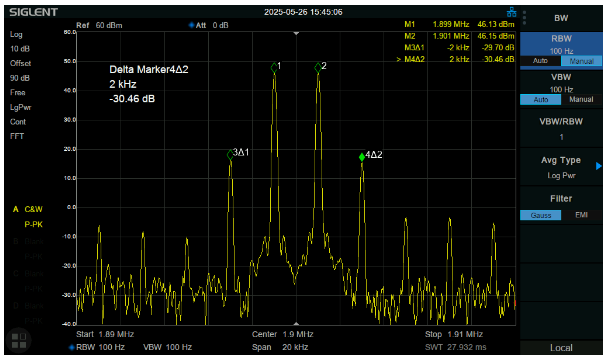


Figure 12 *- IMD3 level on 160m: -37dBc @ 160W Pout*

IMD performance on 40m is reaching -33dBc, 160W+ of CW power:

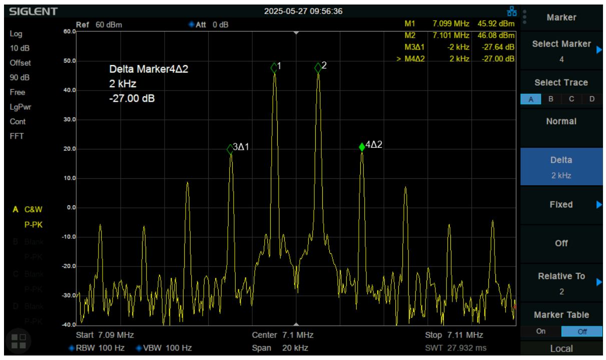


Figure 13 *- IMD3 level on 40m: -33dBc @ 160W Pout*

IMD performance on 10m is reaching -28dBc, 125W+ of CW power:

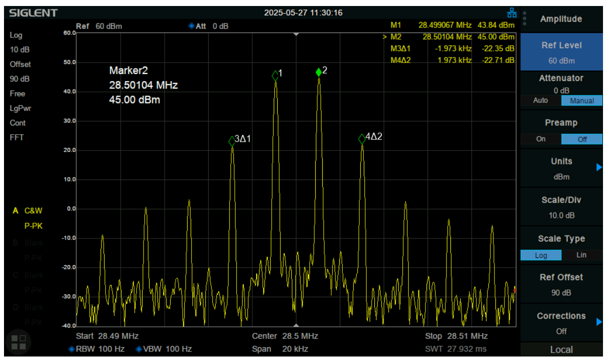


Figure 14 *- IMD3 level on 10m: -28dBc @ 125W Pout*

We have also tested different versions of ferrite materials (#43, #61, #77) for the input and output transformers (T1 & T3) of this amplifier. The comparison aimed to validate the influence of the ferrite materials on the IMD3 / gain performance. This will be presented in rev. 2 of this document, along with the calculations of the T1 & T3 transformers, impedance transformation ratios, etc.

VIII. FURTHER WORK / IMPROVEMENTS

Our main concern for the next revision of this paper is to improve the total gain flatness of the amplifier by using several well-known methods or acclaimed techniques:

- correctly determined input compensation networks
- improved transmission line transformers
- ferrite-core Bal-Un architectures on input and output
- negative voltage feedback approaches
- improved active bias stabilization (incl. temperature compensation techniques) etc.

The feedback loop calculation might be an important subject in the next revision of this document or in a similar one. Also, the collector phasing transformer is a "hot" topic taking into account the context of this paper, alongside with the feedback loop in the base-collector network.
Since we have missed the correct determination of the minimum inductance of the T1 and T3 transformers, in the current version of the paper, we are aiming to cover this important matter.
In terms of load mismatch, full power load mismatch of VSWR of 1.5:1, 2:1 and 3:1 are highly required for the context of this paper.

IX. CONCLUSIONS

This paper demonstrates that a classic design such as AN-762 – *Linear Amplifiers for Mobile Operation* – or its adaptations (e.g., G6ALU, see [1]) can still be successfully reproduced today using NOS, second-hand, salvaged, or decommissioned medium-power BJTs from series like MRF, SD, or 2SC. As evident, the PCB layout remains straightforward and easy to replicate. Our goal was to study, evaluate, and validate the functionality of this design as part of our steep learning curve in power amplifier R&D and business development. We would greatly appreciate any feedback, corrections, or suggestions, which can be sent to support@rowaves.com or shared via WhatsApp at +40 742 854 185. Thank you!

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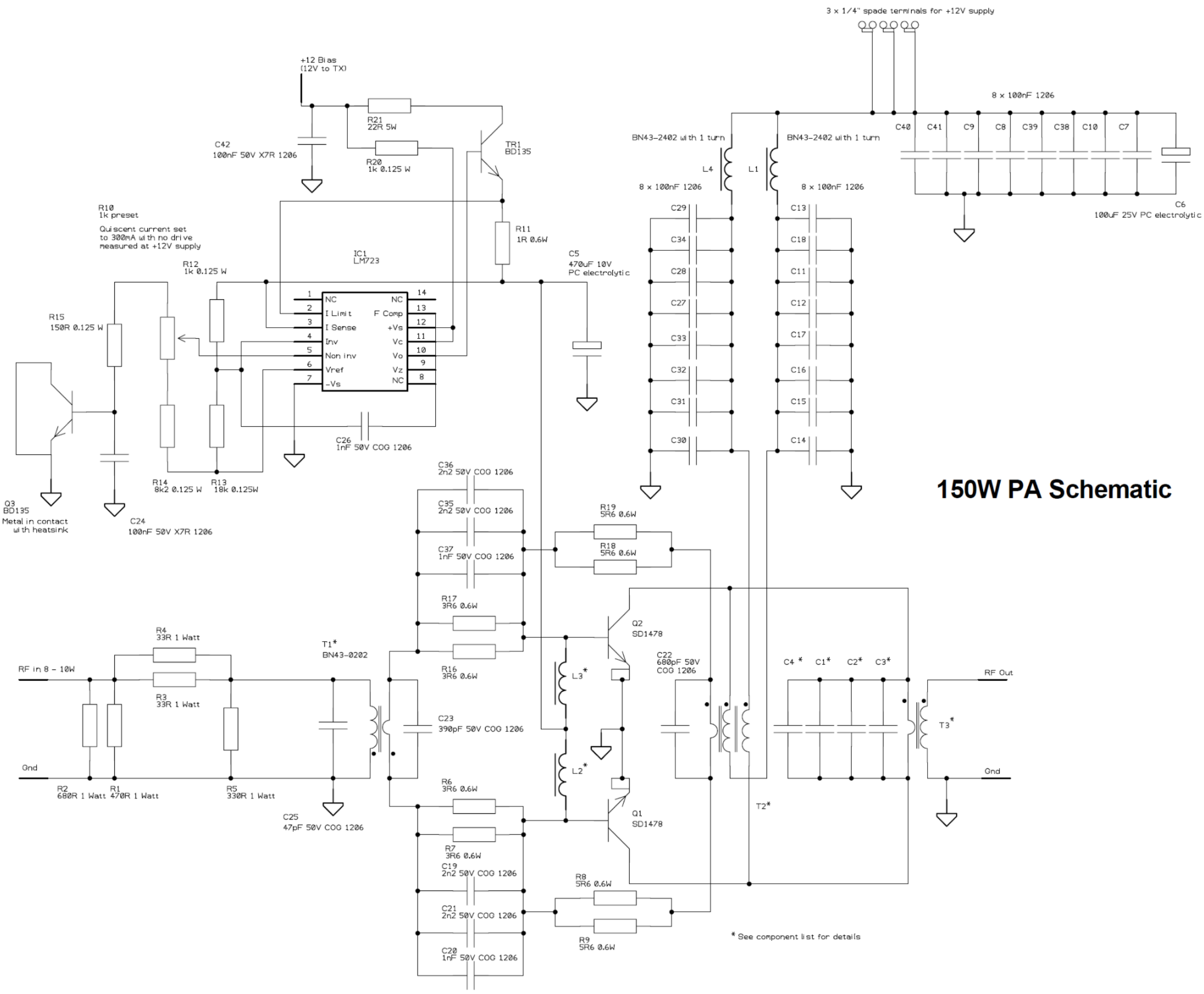
10) *Introducing the IMD-TP1: A Comprehensive Solution for IMD Characterization*, https://rowaves.com/rowaves-imd-tp1-testing-service/
11) list of used equipments: https://rowaves.com/
12) *RTLPF-520 | 5th Order 25W Low Pass Filter*, https://rowaves.com/rtlfp-520-5th-order-25w-test-lpf/
13) *ULTA-10 | 0.25-220MHz Ultra Linear Laboratory Amplifier*, https://rowaves.com/ulta-10/
14) *JUMA SC-2 | 3dB | 100MHz Power Divider / Combiner*, https://rowaves.com/juma-sc-2-100khz-100mhz-power-splitter-combiner/
15) *PA150 | 160-10m 150W+ Linear Power Amplifier Module*, https://rowaves.com/pa150/

Annex 2 - Original BOM for G6ALU

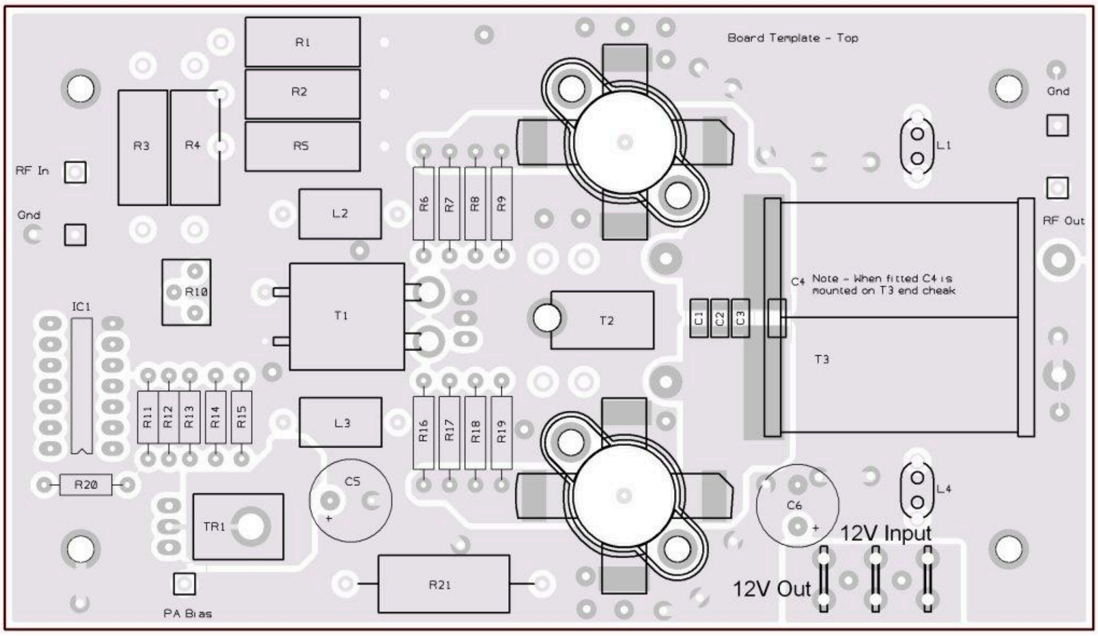
Qty	Component Name	Value	Part Name	Suggested Supplier	Part number	Item cost	Line cost
1	C5	470uF 10V PC electrolytic				0.06	0.06
1	C6	100uF 25V PC electrolytic				0.04	0.04
4	C19,21, 35, 36	2n2 50V COG 1206				0.12	0.48
3	C20, 26, 37	1nF 50V COG 1206				0.083	0.249
1	C22	680pF 50V COG 1206					0
1	C23	390pF 50V COG 1206					0
1	C25	47pF 50V COG 1206					0
4	C1, 2, 3, 4	470pF 500V COG 1206		Farnell	121-6453	0.16	0.64
26	C7-18, 24, 27-34, 38-42	100nF 50V X7R 1206					0
1	IC1	LM723				0.3	0.3
2	L1, 4	BN43-2402	One turn of heavy guage wire through holes (see picture), need not be insulated wire.	RS	467-3573	0.12	0.24
2	L2, 3	6 Hole ferrite choke	Ready wound choke	RS	260-6824	0.37	0.74
3	12V DC supply	0.25" spade					0
2	Q3, TR1	BD135				0.25	0.5
2	Q1, 2	SD1487					0
1	R1	470R 1 Watt				0.055	0.055
1	R2	680R 1 Watt				0.055	0.055
1	R5	330R 1 Watt				0.055	0.055
1	R10	1k preset				0.33	0.33
1	R11	1R 0.6W				0.04	0.04
2	R12, 20	1k 0.125 W				0.01	0.02
1	R13	18k 0.125W				0.01	0.01
1	R14	8k2 0.125 W					0
1	R15	150R 0.125 W				0.01	0.01
1	R21	22R 5 Watt				0.2	0.2
2	R3	33R 1 Watt				0.055	0.11
4	R6, 7, 16, 17	3R6 0.6W				0.04	0.16
4	R8, 9, 18, 19	5R6 0.6W				0.04	0.16
1	T1	BN43-0202	3 turns of PVC insulated hook up wire wound inside a 1 turn primary of coax cable braid	RS	467-3545	0.75	0.75
1	T2	T68-61 (if available T50-61 will probably be OK but not tested)	6 bifilar turns of 1.25mm (18SWG) enameled copper wire - feed-back winding is 1.25mm wire passed through the core once	RS	467-4352	0.37	0.37
2	T3	Type 43 402 size bead	5 turns wound inside primary of 1/4" outside diameter brass tube - see separate drawing / photo	RS	467-2750	0.62	1.24

Annex 1 - G6ALU schematic modification of the AN-762 application note © G6ALU [1]



Annex 3 - Top and bottom view of G6ALU design, based on modification of the AN-762 application note © G6ALU [1]

top view of the PCB



bottom view of the PCB: