Optimizing Shortwave Wideband RF Amplifier: Study of Transmission Line Transformer Construction Methods

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Abstract. *This paper presents a study of six physical transmission line transformers (TLTs) designed to provide wideband output matching for laterally diffused metal oxide semiconductor (LDMOS) transistors within a push-pull amplifier operating in the 1.8–30 MHz spectrum with an output power of 600 W. While the mathematical model of TLTs is well described in the literature, the impact of physical construction methods on impedance matching and real amplifier performance is more challenging to ascertain. This paper compares six different TLTs built on various ferrite cores and employing different implementations of transmission lines. Return loss below –14 dB was achieved from 1.65 to 37.4 MHz, with most of the tested transformers exhibiting return loss better than –10 dB up to 50 MHz. The study also presents the impact of transmission line implementation on impedance matching using both special-purpose low impedance coaxial cable and a combination of general-purpose coaxial cables connected in parallel. Comparison of three chosen transformers in a real RF amplifier shows that using parallel transmission lines can lead to a return loss comparable to that of a special-purpose coaxial cable, although at the cost of lower efficiency and output power. Second harmonic cancellation effect was also investigated for three transformers.*

Keywords

Amplifier, shortwave, Transmission Line Transformer (TLT), wideband, LDMOS

1. Introduction

In a conventional transformer, a signal in the primary winding generates a magnetic field, which induces voltage in the secondary winding, providing coupling and an impedance transformation ratio that is determined by the square of the primary to secondary turns ratio [1]. For a conventional transformer, the lower frequency cutoff is determined by decreasing permeability of the core, while the high frequency cutoff is determined by parasitic components such as series inductance and interwinding capacitance [2]. Therefore, such magnetically coupled transformers have quite limited bandwidth, especially when larger cores are used to couple final stages of high power RF amplifiers.

In 1944, G. Guanella published a novel concept of an impedance transformer consisting of a pair of coiled transmission lines [3]. The transmission line transformer (TLT), introduced by Guanella, provides transformation of input voltage and current ratio, and thus transformation of impedance by choking unwanted common mode current. The transmission line transformer utilizes the transverse electromagnetic (TEM) mode of a transmission line as a method of coupling instead of magnetic coupling [4], making it more similar to a special case of an auto-transformer. Connecting coils in series and parallel results in a circuit that can sum in-phase voltage and choke the common mode current between input and output by the high reactance of the windings.

Choking can be further improved at lower frequencies by loading the coils with ferrite material. Because of the equal lengths of the used transmission lines and the ideally frequency-independent behavior of the used transmission lines, the Guanella transformer can exhibit very wide bandwidth operation [5].

Figure 1 shows that the Guanella transformer sums voltages of equal delay at the output by employing two interconnected coils. This theoretically increases the upper cutoff frequency, but it also increases the footprint of the transformer.

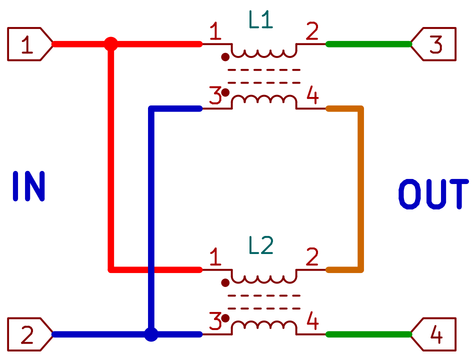

Fig. 1. Schematic of a 1:4 Guanella TLT.

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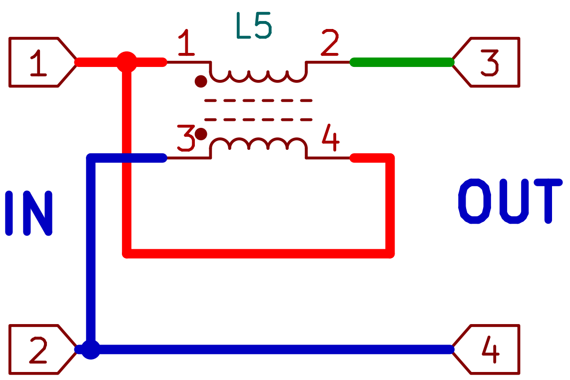


Fig. 2. Schematic of a 1:4 Ruthroff TLT.

Another wideband TLT solution using a simpler topology was proposed in 1959 by C. L. Ruthroff [6]. The schematic of a Ruthroff transformer with a 1:4 impedance transformation ratio is presented in Fig. 2. It can be observed that the Ruthroff transformer sums a delayed voltage to a direct voltage which introduces inherent upper frequency cut off [7]. For both TLTs, windings can be implemented using twisted wire pairs with the correct twisting pitch [8] or using coaxial cable [9], [10]. Above VHF TLTs can be implemented using planar technology. However, for HF and lower part of VHF, lumped elements are used [11].

In the case of a 1:4 Ruthroff transformer, the input signal is split into two transmission lines of equal length and characteristic impedance of 25 Ω, assuming an input impedance of 50 Ω. Waves are then recombined at the 12.5 Ω output port [9]. This transformer can be implemented using a coaxial cable in such a case, the core of the coax is used as the primary winding of L1 and the shield is used as the secondary winding. Current on the secondary winding must be minimal thus, coiling and magnetic loading of the outer shield must provide impedance much higher than 25 Ω.

Due to the canceling out of flux in the core, an efficiency of more than 95% (or 0.02 to 0.04 dB of insertion loss) is possible, losses in TLTs also increase with permeability of the used ferromagnetic core [5]. Therefore, a limiting factor is not the size of the core but its permeability in order to provide a sufficient choking at lower frequencies while providing low losses and the conductor's power handling capability.

A disadvantage of TLTs is also lower efficiency at low frequencies, which require more inductance from the winding [4], making them less practical at low frequencies. TLTs are also unilateral devices, a TLT can transform impedance in only one way. Therefore, a 1:4 transformer works as a 12.5 to 50 Ω transformer (or vice versa) but not as a 50 to 200 Ω transformer. Moreover, transmission line transformer efficiency is higher for lower impedances [5]. However, this is not a problem in the case of high power RF amplifiers since the output impedance of modern LDMOS transistors tends to be low resulting in higher currents and lower voltages at the drains of the transistors [12].

1.1 Evaluation of Current Solutions

In recent years, thanks to the development of LDMOS technology, compact high-power amplifiers have become possible without necessity for combiners and many amplifier cells. The design proposed in [13] for a 200 W amplifier operates in the 9 kHz to 400 MHz spectrum and relies on ferrite cores and coaxial cable based transmission line transformers. This amplifier spans 13 octaves with a gain higher than 53 dB, however, its efficiency is around 35%. Another work demonstrated the wide capability of TLT using ferrite cores and coaxial cables, achieving a bandwidth of 30 MHz to 1000 MHz with a power output of 4 W [14]. A solution operating in the VHF and lower part of the UHF band is presented in [15], where a 110 W LDMOS push-pull amplifier with discrete TLT input and output matching working in the 150 MHz to 400 MHz range achieved a power added efficiency of up to 42%. Although the authors did not achieve better efficiency compared to related amplifiers, the worst case third harmonic distortion of –26 dBc was a good result when compared to single ended amplifiers.

If wide bandwidth is not a priority and the amplifier is to be designed for frequency band operation, such as the 13.56 MHz or 27 MHz ISM (Industrial Scientific Medial) band, the matching network can be composed of discrete L and C elements. Such a solution is presented in [16], where a 1300 W LDMOS amplifier with an efficiency of 74% and a gain of 24 dB was constructed. Similar approach was used to built a 1 kW class E amplifier using SiC-MOSFETs [17].

LDMOS technology was also evaluated in terms of intermodulation distortion. In a 97 W amplifier cell, a third-order product for a dual-tone test was noted at 21.46 dBc [18]. In the same work, the return loss for the output of the amplifier was better than –14 dB, which is a very good result. This value will be used as a benchmark for the 1–30 MHz spectrum in this work.

The above examples of LDMOS amplifiers can be split into two main groups. The first group prioritizes bandwidth at the cost of more harmonics and potentially lower efficiency, requiring wide matching networks such as TLT. The second group uses more narrowly tuned networks, allowing for better rejection of harmonics. A comparison of related works and designs in terms of parameters such as bandwidth, peak power output, maximal efficiency, gain, harmonic suppression and used transistors is presented in Tab. 5.

1.2 Design Goals

The MRF-300 LDMOS transistors used in this project are rugged devices that, according to the manufacturer, can withstand a VSWR of 65:1 at all phase angles for a pulsed 100 µs 20% duty cycle signal with no device degradation [19]. It will be assumed that the minimum return loss provided by the output matching is 10 dB. This value is expected to be used mainly for the higher end of the HF or even VHF spectrum.

Since efficiency plays an important role in modern applications, the optimal operating bandwidth of the output matching network will be specified for return loss of –15 dB, which will present less than 3.3% mismatch loss and provide a VSWR smaller than 1.44 for the final transistors at the center of the design frequency. The datasheet of the MRF-300 claims a power gain of 28.1 dB at 13.56 MHz. NXP provides the typical performance characteristics based on the evaluation boards that use discrete LC components for narrow input and output matching networks. While designing the amplifier, the goal was to achieve an output power of 400 W with a transceiver that has an output power of 10 W, requiring a minimum gain of 16 dB. The amplifier should also achieve a best case efficiency of 70%.

1.3 Output Matching in Tested Amplifier

The TLTs introduced so far represent transformers of the second order since their input to output voltage transformation ratio is 1:2 and the impedance ratio is the squared ratio of effective turns, or 1:4. Although not every ratio can be obtained with transmission line transformers, ratios of 1:4 and 1:9 are the most popular in HF amplifiers, and they can be implemented by TLTs. A transformer with an impedance ratio of 1:9 represents a third order transformer, which can be built as a combination of lower order TLTs [20]. The schematic of a 1:9 TLT is presented in Fig. 3.

It can be noticed that this TLT is a combination of Ruthroff transformers. This circuit will be the main point of interest in further measurements, modification of this circuit in a way that port 2 of L3 is connected to port 4 of L4 (see Fig. 1) will also be tested in one TLT since it is an easy way to create a 1:4 TLT.

The designed TLTs were intended for use in a linear RF push-pull amplifier operating in the 1.8–30 MHz spectrum. The amplifier uses two MRF300 LDMOS transistors, which have an output impedance module of approximately 5.2 Ω at 13.56 MHz, which is typical for most modern 50 V transistors in this power range. The closest integer transformation ratio to 50 Ω that can be realized with a TLT is 1:9. The amplifier presented in Fig. 4 utilizes a balanced output configuration of transistors, while the output from the PCB is an unbalanced coaxial cable.

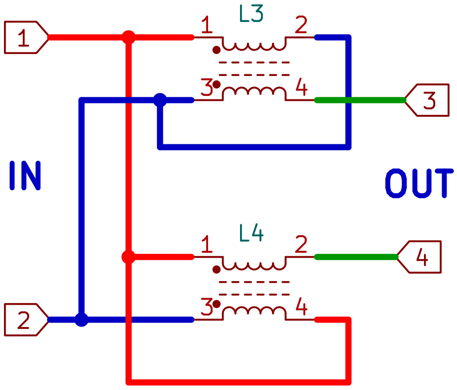


Fig. 3. Schematic of 1:9 TLT.

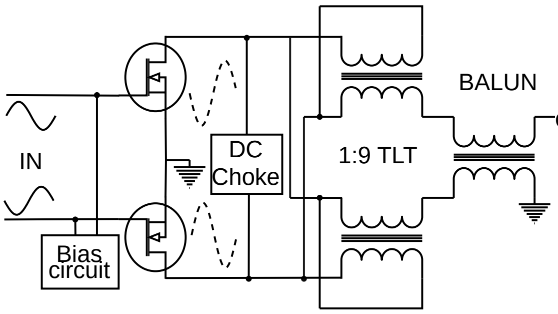


Fig. 4. Simplified schematic of amplifier with TLT output.

For wideband operation, a transmission line 1:1 transformer can be used to change the output from balanced to unbalanced (BALUN) [21]. Such a circuit attenuates common-mode current that would occur on the shield of a coaxial cable if it were connected to a balanced output.

Besides proper connections, each TLT needs to utilize a transmission line with characteristic impedance that is a geometric average of the input and output impedance in order to present proper match [4]:

$$Z_{\mathrm{tr}} = \sqrt{Z_{\mathrm{in}} Z_{\mathrm{out}}}. \tag{1}$$

For example, the 1:4 transmission line transformer presents a 12.5 Ω to 50 Ω transformation and should use a transmission line with a characteristic impedance of 25 Ω. For this case, it is still relatively easy to obtain a 25 Ω coaxial cable. However, for a 1:9 transformer, the characteristic impedance of the necessary transmission line is approximately 17 Ω. Although there are coaxial cables with such low impedance, they are much more difficult to obtain, and their prices are significantly higher. Therefore, in this paper there is also investigated the method of improving such a special-purpose cable with a parallel connection of more common cables. Fulfilling (1) is especially important as the length of the transmission line used in the TLT is larger than small fraction of wavelength, this effect will be demonstrated in measurement of 1:4 TLT that was rewired to act as 1:9 TLT.

1.4 LDMOS Bias Circuit

MOSFETs are voltage controlled devices in which drain current increases with the gate voltage. The amplifier is designed for class AB operation, in which the gate voltage is slightly above the threshold, allowing for idle drain current. The idle current of transistors (collector or drain) depends on many factors. MOSFET's drain current is quite sensitive to temperature changes, so the bias circuit needs to account for those changes so that the gate voltage is inversely proportional to the temperature, with a ratio of approximately $1\,\frac{\mathrm{mV}}{^{\circ}\mathrm{C}}$ to $2\,\frac{\mathrm{mV}}{^{\circ}\mathrm{C}}$ [22], [23]. In order to ensure stable and reliable operation of the amplifier, the relationship between transistor temperature and gate voltage should be characterized. A proper bias circuit that can maintain a stable idle drain current of a LDMOS transistor can be characterized by low output noise and low output impedance.

Bias circuit used in this project was described in [1]. This circuit was chosen because it uses very low noise linear regulator the LM723 which offers lower noise when compared to the 7805 regulator often employed in this role.

In bias circuit from Fig. 5, a thermistor is used to sense the temperature, and together with a network of potentiometers, a reference voltage for the linear regulator is adjusted. Potentiometer RV4 sets the bias voltage, which should be adjusted for proper quiescent drain current, and RV5 sets the slope of the gate voltage to temperature. Input voltage to the regulator was 12 V. Keeping the manufacturing spread of semiconductor devices in mind, the elevated bias voltage is further split into two arms, each having an additional potentiometer so that the gate voltage is adjusted separately for each MOSFET, schematic is presented in Fig. 6. Resistors and potentiometers in those arms were chosen to be of lower value in order to minimize thermal noise while still maintaining a relatively low output impedance. Multiple decoupling capacitors were used to provide low impedance across the wide frequency range. Output resistors R45–R47 are included to ensure stability at the lower frequency of operation [22].

In order to test the circuit across a wide temperature range and tune it for a desired slope, an assembled board without transistors was mounted onto a heat sink. Both gates were connected to ground using 1 MΩ resistors and parallel 100 pF capacitors. A DS18B20 thermometer was mounted on a brass lug screwed to the copper heat spreader plate. A simple data acquisition system was built using an Arduino Uno, which logged the temperature and converted the value from ADC to gate voltage. A hot air station was used to heat up the copper plate with the thermometer attached to it.

From the measured data, it can be seen that the relationship between gate voltage and temperature is nonlinear for lower operating temperatures and then becomes mostly linear. This is an acceptable result because the working amplifier very quickly heats up above 30°C. Furthermore, Figure 7 shows that for lower settings of RV5, the slope becomes steeper, giving a larger change in gate voltage for every °C compared to the high value of RV5. The exact ratio of gate voltage change to temperature change should be chosen based on the LDMOS manufacturer's datasheet and design resources, similar to quiescent drain current, which will also depend on the desired output power. For a future experimenter wanting to implement this circuit, a Tab. 1 was prepared that gives exact ratios for different settings. These values were taken as the difference between two extreme points of the linear part of traces. These values represent the relationship between the gate voltage and temperature change for different settings of RV5 in the tested bias circuit.

For the amplifier designed during this project, a total quiescent drain current of 400 mA was chosen. The ratio of gate voltage to temperature change was chosen as 1.25 $\frac{\mathrm{mV}}{^{\circ}\mathrm{C}}$.

RV5 [kΩ]	$\boldsymbol{\Delta V_{GS}}$ [mV]	$\boldsymbol{\Delta T}$ [°C]	[mV/°C]
10	54	80.18	0.673
9	78	81.19	0.961
8.5	112	88.75	1.262
8	122	89.69	1.360
7.5	161	87.06	1.849
7	249	94.57	2.633
6.5	225	79.87	2.817
5	294	83.38	3.526
2.5	567	63.19	8.973

Tab. 1. Example settings of RV5 for given ratio of gate voltage change to 1°C.

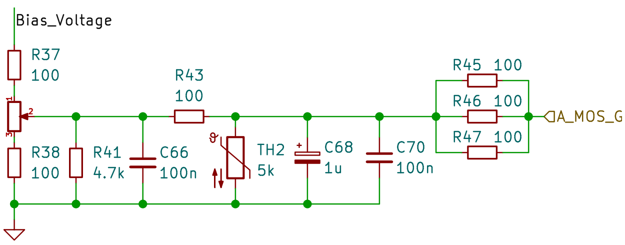


Fig. 6. Individual circuit of bias supply for one of the LDMOS transistors, the TH2 was not populated.

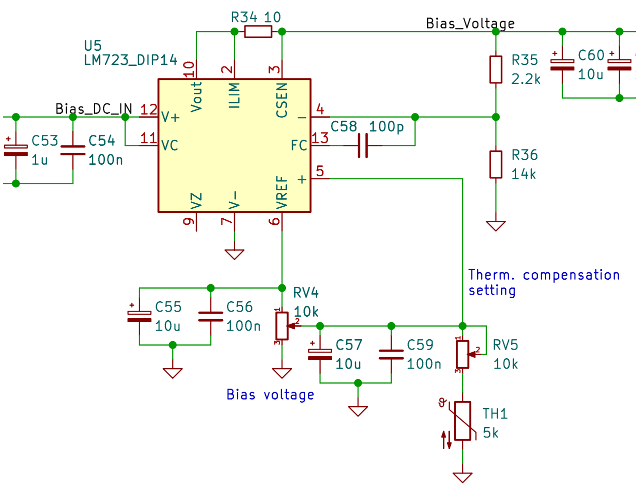


Fig. 5. Circuit of the bias regulator with thermal compensation.

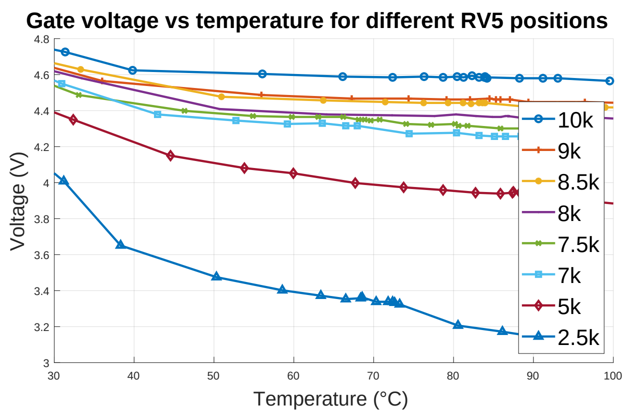


Fig. 7. Bias circuit test.

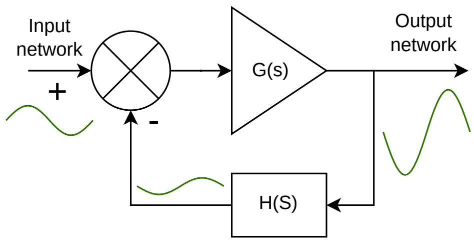


Fig. 8. Block diagram of negative feedback loop.

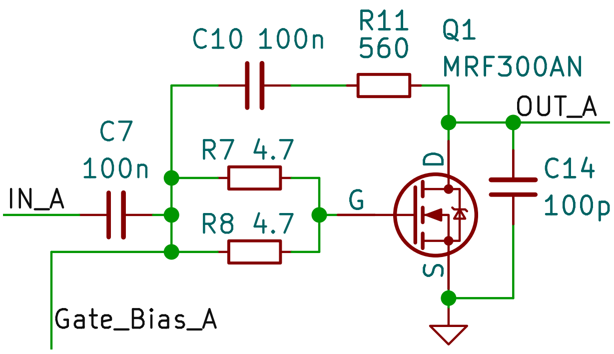


Fig. 9. Feedback network and connection of each MRF300.

1.5 Feedback Network

Several methods are implemented in the amplifier to improve its stability. One of the most important is the use of a negative feedback circuit. It samples part of the output signal from the drain of the MRF300s and feeds it back to the gates. Because the signal at the drain is 180° out of phase with respect to the gate, the feedback signal decreases the input signal, causing degradation of gain due to superposition.

The reduction of gain also reduces efficiency however, negative feedback can improve the stability of the amplifier and flatten the gain over the spectrum of operation. For the feedback network, a solution tested in [24] was used. The network consists of a series DC blocking capacitor (C10 in Fig 9) and a 560 Ω resistor R11 from drain to gate. The resistor is realized as a combination of two parallel 1.2 kΩ non-wire wound resistors to reduce parasitic inductance and increase power handling capability.

2. TLT Measurement Setup

2.1 Balun Measurement Setup

Setup used for testing each TLT included a balun that was put between TLT output and VNA. To make sure that the balun provides common mode attenuation two different baluns were compared. To test the 50 Ω balun, a spectrum analyzer with a tracking generator was used in a configuration presented in Fig. 10. The role of this setup was to measure the common mode attenuation created on the coax shield.

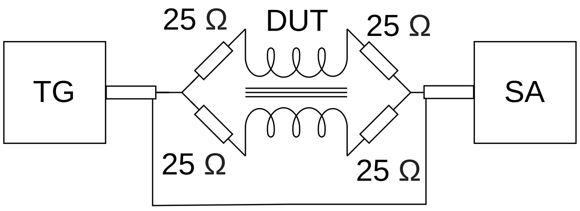


Fig. 10. BALUN measurement test setup.

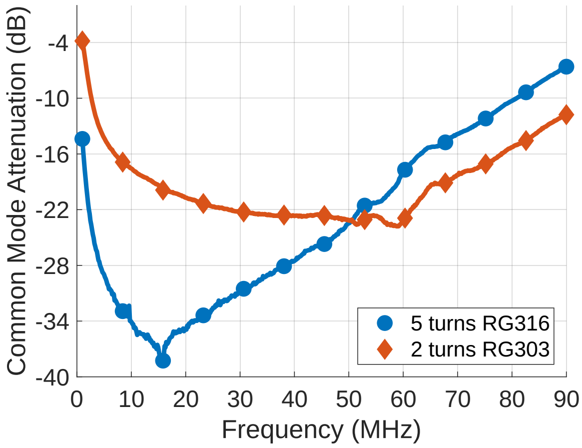


Fig. 11. Comparison of two different baluns in terms of common mode choking.

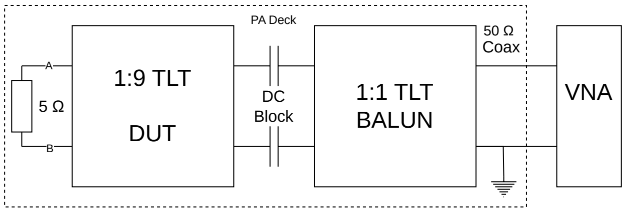


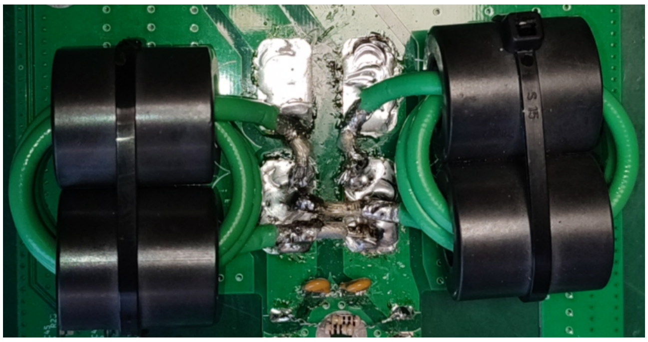
Fig. 12. Measurement setup of return loss measurement.

Two 43-material ferrite cores (Fair-Rite 2643102002) were used in both tested baluns. The first balun had two turns wound with RG303/U, while the second balun utilized much thinner RG316 with 5 turns.

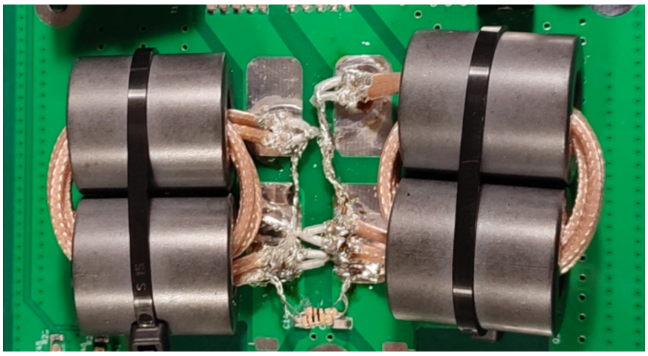
From Fig. 11 it can be seen that the balun wound with more turns provided better choking of common mode current at lower frequencies. However, most likely due to larger parasitics, its performance above 50 MHz is worse when compared to a balun wound with fewer turns of thicker cable. Ultimately, the balun with two turns of RG303 was used in further tests because it presented a more flat characteristic. This balun also used a shorter length of coax with smaller insertion loss.

2.2 Return Loss Measurement Setup

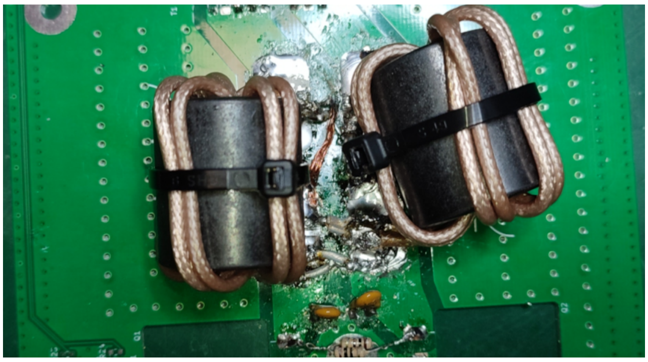
In order to measure the impedance matching provided by the tested TLTs, a setup presented in Fig. 12 was used. The dashed line box represents the PCB of the amplifier. Instead of transistors, two parallel 10 Ω resistors were used as a load. The device under test (DUT) was the TLT.



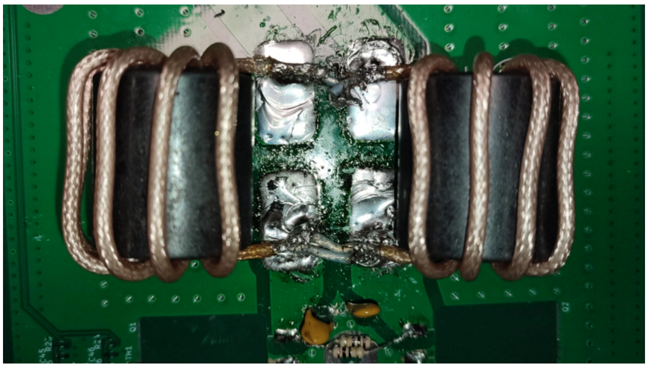
(a) 1:9 TLT, 17 Ω coax, 2×61 core



(b) 1:9 TLT, 50||25 Ω coax, 2×61 core



(c) 1:9 TLT, 50||25 Ω coax, 1×43 core



(d) 1:4 TLT, 25 Ω coax, 1×43 core

Fig. 13. Photos of tested TLTs.

TLT No.	Ratio	Z_0 [Ω]	Coax	Turns	Core (material / mfr. nr)
1a, 1b	1:9	17	HF141-17-FEP	3, 2	61 /Fair-Rite 2661102002
2	1:9	17*	RG316-25-Flex || RG316-50	3	61 / Fair-Rite 2661102002
3	1:9	17*	RG316-25-Flex || RG316-50	4	43 / Fair-Rite 2643102002
4a, 4b	1:4, 1:9	25	RG316-25-Flex	5	43 / Fair-Rite 2643102002

'*' denotes the theoretical impedance of parallel coaxes. Comma denotes variations.

Tab. 2. Built comparison of tested TLTs.

3. Tested TLTs

In this measurement six TLTs were tested in the place of DUT while the rest of the setup remained the same. Used VNA was the Lite VNA 64 connected to a computer with NanoVNA Saver software on it.

Table 2 summarizes the built transformers that were then tested. Four distinctive transformers were constructed, with modifications made to the number of coax turns and impedance ratio for two of them. Two different round ferrite cores were utilized for the tested TLTs. The first core, 2661102002, was made of type 61 material, specified for higher frequencies. At the frequency of 100 MHz, the manufacturer claims a typical impedance of 216 Ω. The second core, 2643102002, used material 43, designed for lower frequencies. Its impedance at 100 MHz is specified as 204 Ω. These specific cores were chosen not only because the materials used are commonly applied in the construction of high frequency transformers, especially in the case of 43 material, but also because they are of the same size and can accommodate a wide range of different combinations of windings. Both cores have an outer diameter of 25.9 mm and an inner diameter of 12.8 mm, and a length of 28.6 mm. A smaller core could likely have a positive effect on bandwidth by reducing the length of the used transmission line. However, winding the 17 Ω coaxial cable chosen for this project on a smaller core would be difficult. Alternatively, a thinner and possibly more flexible 17 Ω coaxial cable, such as the TC-18, might be considered.

The 17 Ω coaxial cable used is a special purpose cable manufactured by QAxial. It employs non-magnetic conductors and is designed to operate up to 24 GHz, capable of handling 1200 W average power at 100 MHz, thus it is a suitable option for this application. Disadvantage of this cable is its higher price and the difficulty in winding due to the dense shield used which results in relatively high stiffness.

TLTs using 17 Ω characteristic impedance transmission lines were wound with three turns for transformers 1a and 2, as indicated in Tab. 2. Additional cores were employed to increase the inductance, aiming to enhance performance at lower frequencies, particularly since these transformers utilized type 61 material with lower permeability. For transformers 1 and 2 in Tab. 2, two cores were used in parallel for each side of the transformer, resulting in a total of four ferrite cores per transformer. TLTs utilizing 43 type material (transformers 3 and 4 in Tab. 2) were wound on a single core per side of the transformer, resulting in only two cores in total per transformer, additional turns were employed in order to compensate for smaller number of pass through winding through the cores when compared to transformers 1 and 2.

In addition to the closest 1:9 transformation ratio, a 1:4 TLT was also tested (transformer 4a in Tab. 2). According to the theory presented in Sec. 1, its input should see a 12.5 Ω load. Based on (1), a transmission line with 25 Ω coax should be utilized. This 1:4 transformer was tested to evaluate the performance using more affordable and available coax, such as RG316-25 (the 25 Ω version of typical RG-316). The 1:9 transformer 3 utilizes a parallel connection of 50 and 25 Ω cables, and employs only two 43 cores. This version of the 1:9 transformer represents the second cheapest option after the 1:4 transformer.

4. Impact of TLT on Return Loss

Six variations of physical transmission line transformers were tested using the setup presented in Fig. 12. VNA was calibrated for a load of 50 Ω. A series of S11 measurements was plotted on a Smith chart in Fig. 14 for TLTs selected to be tested in the real amplifier. Rectangular graph in Fig. 15 presents return loss for all six tested transformers. The legend denotes the transformation ratio, type of transmission line used (single or parallel), number of turns around the core, and type of ferrite core material.

In Fig. 15, it is shown that the 1:4 transformer exhibits the worst match, a conclusion also supported by the Smith chart in Fig. 14. The trace corresponding to the 1:4 transformer deviates the furthest from an ideal match across the entire HF spectrum, displaying a worse than 2:1 VSWR. Tthis indicates low efficiency in impedance matching and relatively high reflection to the power transistors.

Using a single 25 Ω coaxial cable and employing the same winding as in the 1:4 transformer but connecting it in a 1:9 configuration results in much better behavior at lower frequencies. Return loss of –14 dB, which we will further assume as a minimum goal for the entire spectrum (1.49 VSWR, 3.98% mismatch loss), is achieved between 1.6 to 20.8 MHz. This transformer violates (1) for characteristic impedance. The consequence of this can be seen with rapidly decreasing return loss at higher frequencies, with the worst observed return loss above 30 MHz for all tested configurations. We hypothesize that this configuration works very well below 8 MHz because the length of the used coaxial cable is very short at this wavelength and does not exhibit significant transmission line effects. Thus, the characteristic impedance does not have as much of an effect. The length of used coax for this 1:4 transformer was 46 cm, which at the minimum return loss of –22 dB at 3.26 MHz equates to 0.005λ.

All other 1:9 transformers performed well in this test, confirming the theoretical background. The transformer wound on more 61 cores with 3 turns and 17 Ω special-purpose cable actually did not perform as well as the improvised transformer with a parallel combination of 50 and 25 Ω general purpose cables, at least according to return loss measurements.

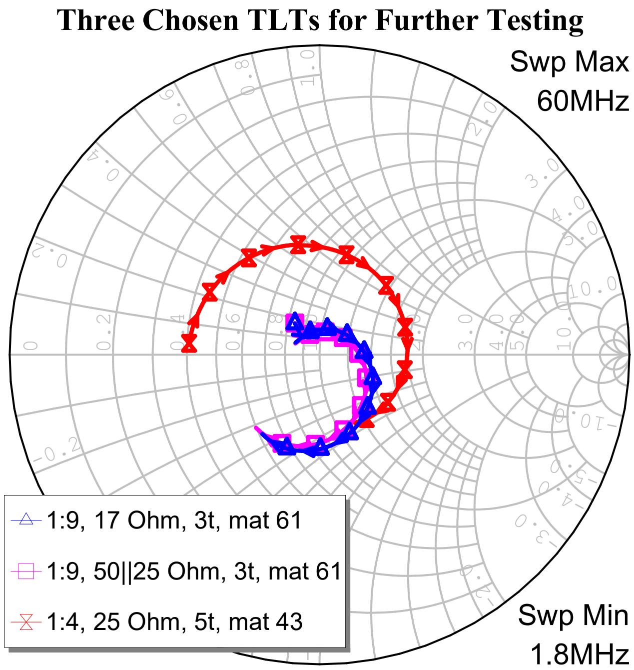


Fig. 14. S11 of three TLTs chosen for a further tests in amplifier.

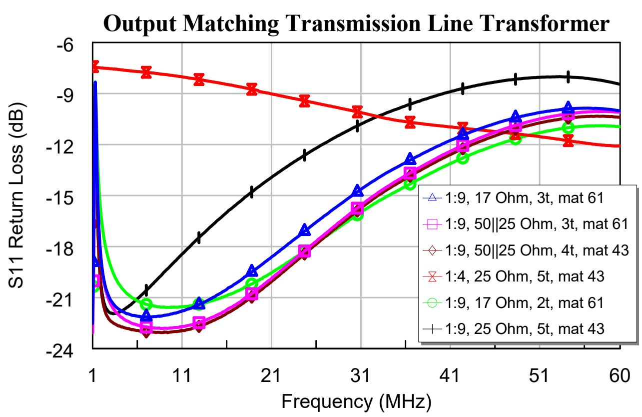


Fig. 15. Return loss for all six physical TLTs from 1 to 60 MHz in form of rectangular plot.

Return loss can be positively impacted by a higher insertion loss, so it should not be taken as the only metric. The test has also confirmed that the transformer made with a single 43-type core but more turns performs better at lower frequencies, but the difference is not significant. Ultimately, both transformers with a parallel connection of transmission lines resulted in a bandwidth of –14 dB return loss (RL) from 1.6 MHz to 36 MHz.

The last variation is the original 1:9 TLT with 17 Ω coaxial cable but with 2 turns instead of 3. As expected, this has moved the minimum return loss up in frequency from about 6.8 MHz to 10.3 MHz. This shift of the S11 trace has allowed for better RL at higher frequencies. This is most likely due to a smaller ratio of coax length to λ and also smaller leakage. Such a configuration offers –14 dB RL from 1.65 MHz to 37.4 MHz and results in a more flattened RL throughout the entire HF spectrum. Furthermore, this last variation also offers the lowest RL at 52 MHz of –11.2 dB, which is still an acceptable match resulting in a VSWR of 1.76 potentially allowing for operation at lower portion of VHF band.

Insertion loss or the S21 parameter of the transformers was not tested because such a measurement would require a second transformer. Connecting transformers back to back would bring the impedance back to 50 Ω, allowing for an S21 DUT measurement setup [2]. However, this would introduce uncertainties since a professionally characterized 1:9 wideband transformer was not available in the laboratory. In the future, such a test with two transformers and de-embedding could be employed to measure the insertion loss of an unknown transformer.

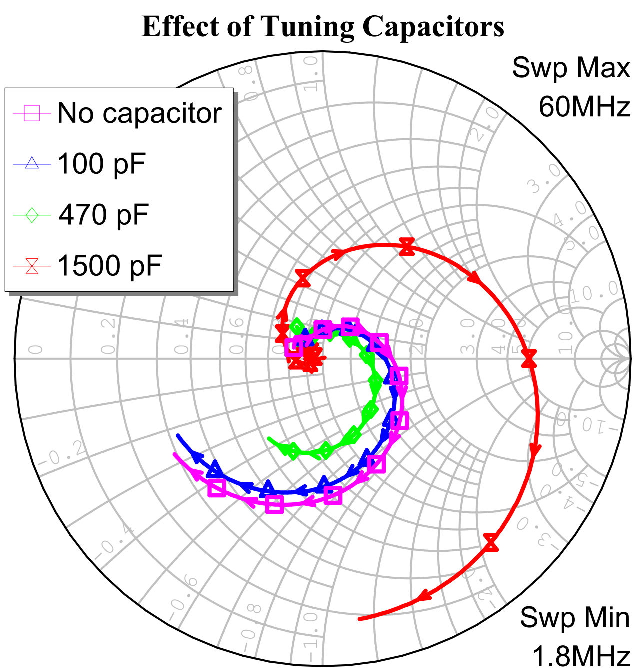


Fig. 16. Effect of tuning capacitor placed in parallel to the input of each side of 1:9 TLT made with special purpose coaxial cable.

Figures 14 and 15 present return loss for a transformers after a tuning capacitors were added. Every transformer has both inductive and capacitive components. In order to tune the output transmission line transformer for maximum bandwidth, two capacitors were used, each connecting one balanced input of the transformer to the ground. A VNA was used to choose the optimal value. Effect of the tuning capacitor on a 1:9 transformer with 3 turns of 17 Ω coax is shown in Fig. 16.

More values were tested, but the four presented in Fig. 16 show an effect of canceling out the capacitive part of the reflection coefficient as the frequency increases. A value of 470 pF was chosen because it significantly improves the match across a wide range of frequencies. This value was also used in all the other tests. While using this method of improving impedance matching in a high-power RF amplifier, it should be noted that the capacitors used have to withstand both the voltage and high current present at the low impedance output of the push-pull amplifier.

5. Impact of TLT on Real Amplifier

The amplifier was tested with three transformers: A 1:9 special purpose 17 Ω TLT on four 61 material cores (No. 1a in Tab. 2), the 1:9 TLT made using a two 43 material cores and parallel coaxial cables (No. 3 in Tab 2), and lastly, the 1:4 transformer on two 43 material cores with 25 Ω coax (No. 4a in Tab. 2). Amplifier used for this test was the Forte 600, a linear RF amplifier developed by one of the authors of this paper. The MRF300 LDMOS transistors used in the amplifier can provide up to 600 W of peak power for continuous wave signal. According to NXP these transistors are specified for a frequency range from 1.8 to 250 MHz and can withstand a VSWR of 65:1 at all phase angles. MRF300 has a maximal drain voltage of 50 V although in this project a 48 V power supply was used. RF input to the drains of the output transistors is matched using conventional 1:4 transformer on a binocular ferrite core. Amplifier board also included its own 3 dB attenuator at the input.

5.1 Measurement Setup

In order to measure the impact of TLT construction on the real RF amplifier, a test setup presented in Fig. 17 was used. The transceiver (TRX) used was a Xiegu G90, which is an amateur radio software defined HF transceiver. Because the Xiegu G90 allows for control of output power from 1 to 20 W in 1 W incremental steps, a 6 dB attenuator was used when the input power to the power amplifier was lower than two Watts. This allowed for a more precise characterization of the linear region of the amplifier. The output from the power amplifier was connected through an inline power meter (Maas RX-600) to a dummy load, which also included a 50 dB sampler. The coupled port from the sampler presented a signal level that was safe to connect to a spectrum analyzer (Rigol DSA-815).

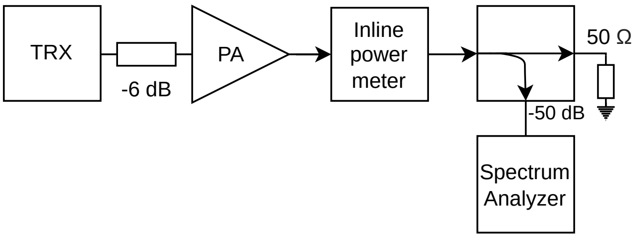


Fig. 17. Measurement setup for the power amplifier output.

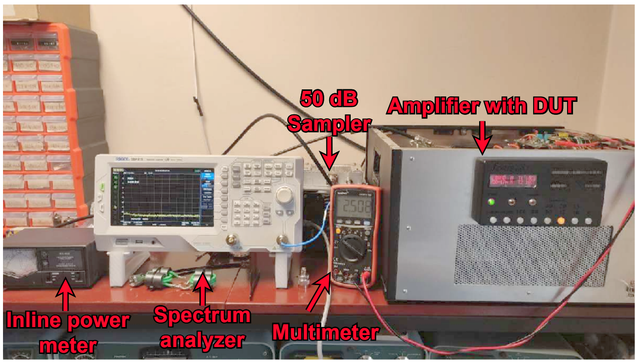


Fig. 18. Photo of the main part of experimental setup.

The inline power meter used has three ranges: 30 W and 300 W with an accuracy of ±5%, and 3 kW with an accuracy of ±7.5%. Average power was measured in all tests, switching from the 300 W to the 3 kW range when the power exceeded the scale, ensuring better accuracy for lower power levels. The uncertainty specified by the manufacturer for the spectrum analyzer is ±0.4 dB. However, in terms of absolute magnitude, there is some uncertainty stacking, as our results were normalized to the S31 measurement of the 50 dB sampler, requiring the spectrum analyzer to perform a series of measurements. This should not affect the most important results obtained with the spectrum analyzer, which are the spurious free dynamic range measurements, as they rely on relative measurements.

5.2 Impact of TLT on Output Power

In a first test the input power of a 14.15 MHz CW signal from the transceiver was increased from 24 dBm (0.25 W) to 38.45 dBm (7 W). Output power was measured using the inline power meter. Frequency of 14.15 MHz was chosen because it was within the amateur radio band and corresponded to the middle of the design frequency.

In Fig. 19, it can be observed that for every tested TLT, the amplifier has a linear region after which it saturates. The input power for which the output drops by 1 dB compared to the extrapolated linear region is known as the 1 dB compression point. Above this point, the amplifier exhibits more nonlinear effects, resulting in an increase of spurious emissions. Compression points were determined and summarized in Tab. 3 for each TLT at 14.15 MHz.

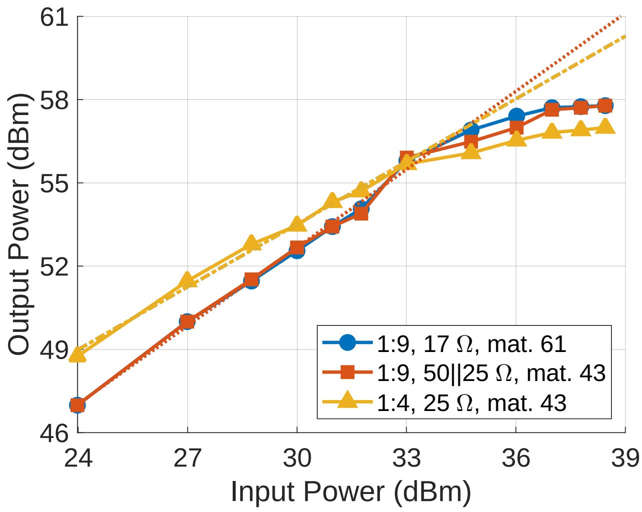


Fig. 19. Increasing output power from the amplifier for three transformers.

Test condition: 14.15 MHz CW		
TLT	**P1dB** [dBm]	P_{OUT} [dBm]
1:9, 17 Ω, mat. 61	36	57.2
1:9, 50‖25 Ω, mat. 61	35.4	56.8
1:4, 25 Ω, mat. 61	34.7	56

Tab. 3. P_{1dB} for three tested TLTs at 14.15 MHz.

TLT: 1:9 17 Ω, mat. 61		
Frequency [MHz]	P_{1dB} [dBm]	P_{OUT} [dBm]
3.6	36.3	56.1
10.1	35.6	57
14.15	36	57.2
28.5	37.5	56.2

Tab. 4. P_{1dB} for 1:9 TLT at different bands.

Highest compression point and output power was achieved for the TLT using a special purpose low impedance coaxial cable. The improvised TLT with parallel connection of more common coaxial cables exhibited very similar output power with slightly worse linearity. The 1:4 transformer presented much lower output power because the presented load to the drains was too high.

Next the output power was measured for maximal input power of 7 W for different amateur radio bands.

While Table 4 shows that the compression point is not constant and changes with the frequency it is usually close to 36 dBm or 4 W of input power. This variation might depend on the amount of the negative feedback used in the amplifier, larger feedback can be used to flatten the gain by the cost of lower output power. Figure 20 shows that the highest output power and most flat gain in the HF spectrum was achieved with the 1:9 TLT built with special purpose 17 Ω coaxial cable. Improvised TLT achieved slightly lower output power but its performance below 6 MHz and above 20 MHz has decreased much faster leading to less flat response.

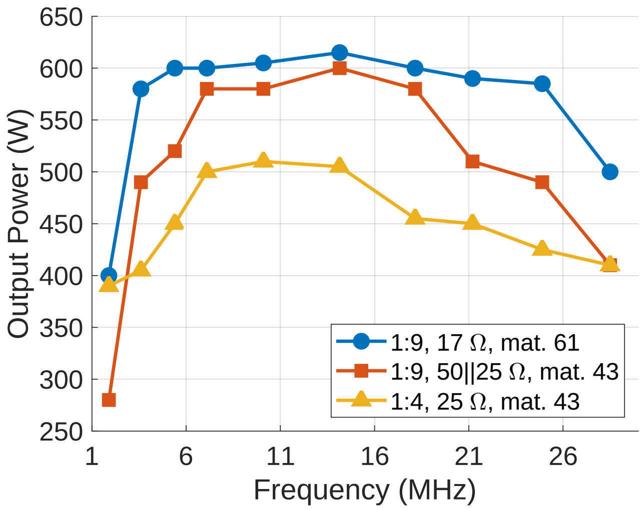


Fig. 20. Maximum CW output power from the amplifier for three transformers across frequency spectrum.

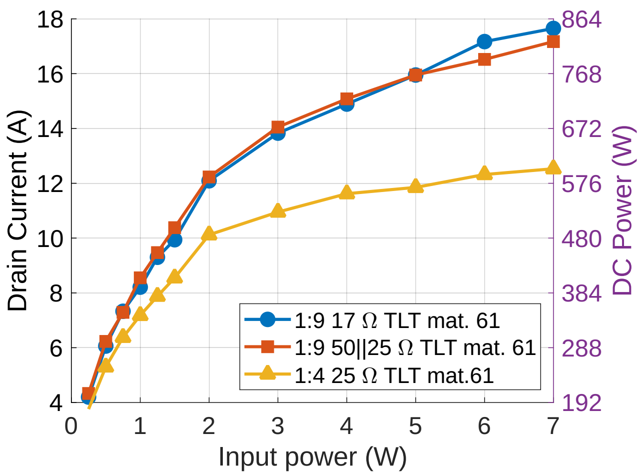


Fig. 21. Current drawn at 14.15 MHz for three different transformers.

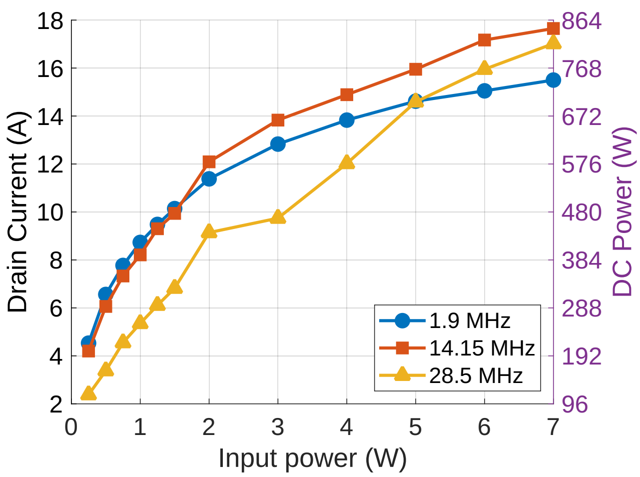


Fig. 22. Current drawn at three different frequencies for the TLT made with 17 Ω coaxial cable and type 61 material core.

5.3 Power Consumption and Supply

The power amplifier board consumes up to 18 A DC at 48 V, which amounts to 864 W. To efficiently deliver this power from the mains, a switching mode power supply is used. The unit chosen for this project is a server power supply produced by Delta, model AWF-2DC-2100W. It provides one 12 V rail at 20 A and a 48 V rail at 41 A for 200–240 V AC input or 20 A for 100–127 V AC input. The 48 V rail can thus supply a sufficient amount of drain current with a large margin, assuming a high-line AC supply. This PSU is designed for server use, so some modifications were necessary. First, 14 AWG wires were soldered to the output board of the PSU. The original fans were disconnected from the PCB and connected to the custom controller of the amplifier. Finally, the AC plug was modified, and pins responsible for startup were shorted so that the power supply turns on the 48 V rail whenever AC power is present. The 12 V rail is used for powering all the logic and other analog circuits that use further local regulation (such as the gate bias circuit).

Current was measured using an ACS712 Hall-effect based ratiometric sensor that was part of the custom controller board. Voltage at the output of the sensor was measured using a digital multimeter (Anneng AN870). An RF choke for each transistor was wound separately using 5 turns of 12 AWG copper cable in silicone insulation on a 43 ferrite core, the same type that was used in some of the TLTs. Drain voltage was measured on the DC side of the chokes. For a 24 dBm of input power, the drain voltage was measured as 48.2 V, while under heavier load (38 dBm RF input) the measured voltage sagged to 47.9 V.

Drain current and DC power drawn by the drains of the two LDMOS transistors were measured first at a fixed frequency of 14.15 MHz for three transformers, and then for three different frequencies for the same transformer.

In Fig. 21, it can be observed that the amplifier consumed almost identical amount of power for both 1:9 transformers. Only the 1:4 transformer consumed significantly lower current, which was expected based on its lower output power. In terms of the various frequencies tested in Fig. 22, it can be noticed that for the lowest tested frequency, the maximum drain current reached less than 16 A for a maximum safe input power. This was also expected since the TLT at this frequency likely provides insufficient choking, so the amplifier does not produce as high output power as it does in the middle of the spectrum.

5.4 Impact of TLT on Efficiency

There are several ways to express the efficiency of an RF amplifier. Common method is to use the collector efficiency for BJT or drain efficiency for MOSFET, defined as the ratio of RF output power to the DC input power of the transistor. However, the collector or drain efficiency (usually denoted as η_C or η_D) does not account for the input power necessary to drive the amplifier.

Another measure known as power added efficiency (PAE) is also used, it is defined as the difference between output and input RF power divided by consumed DC power. PAE is naturally smaller than drain efficiency, but it will be utilized in this work because it provides a more applicable measure. Power added efficiency was calculated using (2):

$$\mathrm{PAE} = \left(\frac{P_{\mathrm{OUT}} - P_{\mathrm{IN}}}{P_{\mathrm{DC}}}\right) 100\%. \quad (2)$$

Used amplifier was equipped with a Hall effect based ratiometric current sensor (ACS712). Output from the sensor was measured using digital multimeter (Anneng AN870) and converted to current. Drain voltage without load on the power supply was measured as 48.2 V while under maximum power output this value has dropped to 47.9 V. Average value of 48.05 V was used in the DC power calculation. Total drain current noted was 17.92 A which resulted in total drain power of 861 W.

Power added efficiency was measured for amplifier with each transformer at 14.15 MHz and for peak power output across amateur radio bands. Tests were conducted for a continuous wave signals. Amplifier was biased for the AB class operation with quiescent drain current of 200 mA per transistor.

TLT with 1:9 ratio made with 17 Ω coax achieved the most flat characteristic in Fig. 23, maintaining more than 65% PAE from 10 MHz to 25 MHz. Observations indicate that the amplifier with this transformer exhibits a very similar PAE to a 1:9 TLT built using two parallel coaxial cables. Furthermore, it can be noted that the transformer with a 17 Ω cable performs in a more predictable manner and, according to measurements, achieves approximately 7% better PAE for medium input power and only slightly worse PAE for high input power at 14.15 MHz (see Fig. 24) when compared to the improvised 1:9 TLT.

Both 1:9 transformers exhibited rather poor performance at 1.9 MHz. This confirms the information from the TLT handbook [5], where the problem of high reactive inductance required for lower frequencies was presented as a limiting factor in the practical application of this type of transformer.

The transformer with a 1:4 impedance ratio managed to achieve the highest efficiency out of the three tested transformers. The peak power added efficiency of w 87% was observed at 7.1 MHz. However, while this result is very good, the linearity of such an amplifier can be brought into question. Additionally, it can be noticed that this configuration achieves excellent efficiency of 72% at 1.9 MHz, which is most likely due to the larger reactive inductance of 5 turns on 43 material type cores when compared to only 3 turns in the case of other 1:9 transformers that used 61 type material. Ferrite mix 43 material also offers higher permeability which leads to higher choking at low frequencies. Amplifier with this transformer also consumes much less current which combined with higher efficiency should lead to a lower heat dissipation if reflected power is neglected.

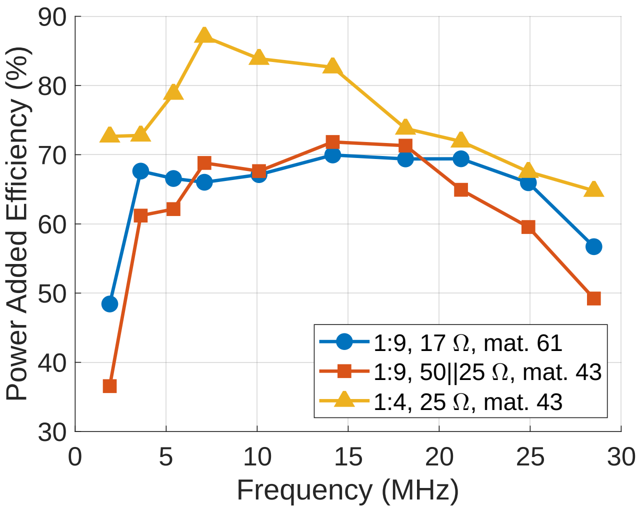


Fig. 23. Power added efficiency for three transformers across spectrum.

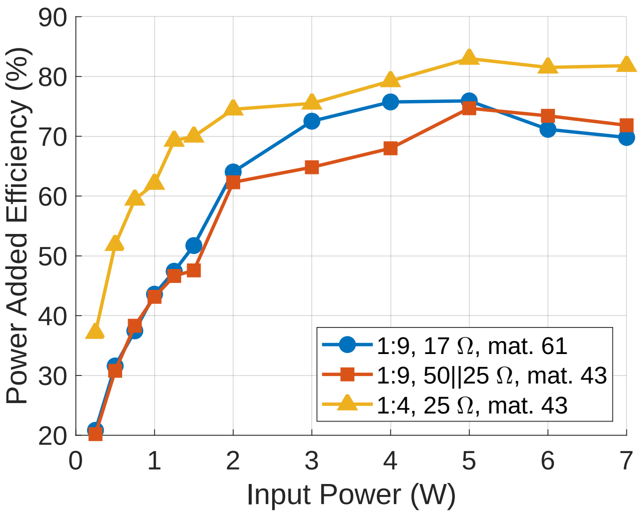


Fig. 24. Power added efficiency for three transformers at 14.15 MHz.

5.5 Impact of TLT on Second Harmonic

Topology of the push-pull amplifier should allow for the cancellation of the second harmonic because an ideal balanced transformer will present a short circuit at two times the fundamental frequency [25]. This idea was demonstrated in wide bandwidth VHF LDMOS push-pull power amplifier that utilized transmission line balun [26] and in wideband UHF GaN amplifier with planar TLT [27]. Comparison of three transformers should reveal which one is the closest to being ideal by providing the largest Spurious Free Dynamic Range (SFDR) between the fundamental signal (first harmonic) and the spurious second harmonic. Here, SFDR is defined as the free dynamic range between the fundamental and a specific harmonic. When a specific order of harmonic is considered, the other harmonics are ignored, even if they would normally decrease the SFDR by standard definition. Harmonics of the amplifier with a 1:9 TLT made with 17 Ω coaxial cable were measured for increasing levels of a 14.15 MHz CW input signal.

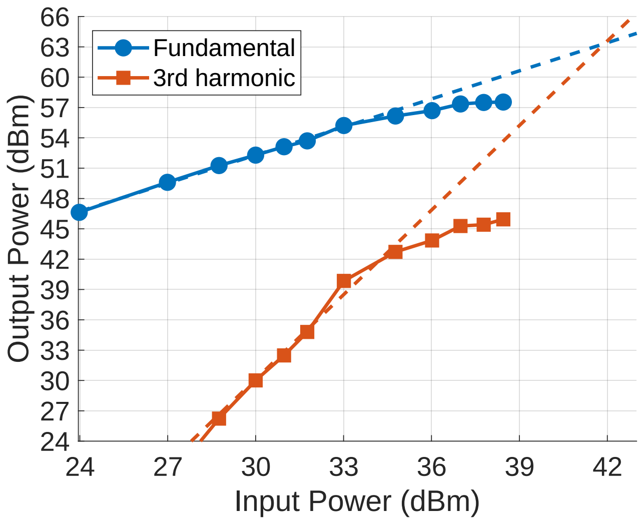


Fig. 25. Third order intercept based on harmonic.

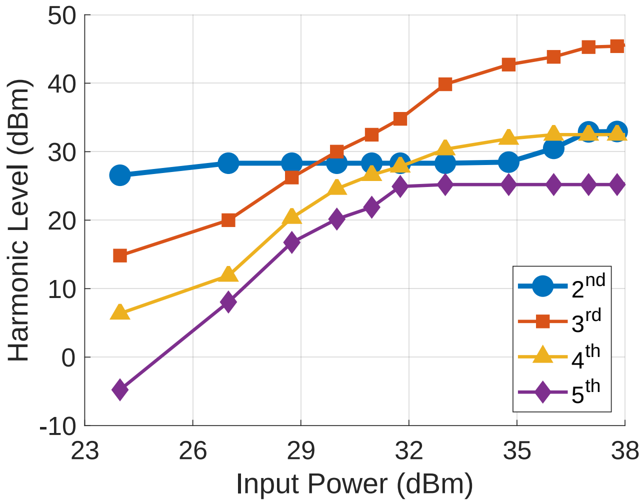


Fig. 26. Level of harmonics for increasing input power.

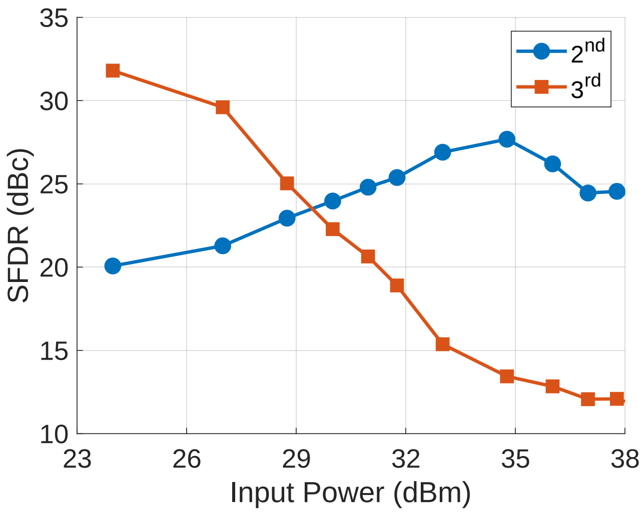


Fig. 27. Free dynamic range between fundamental and harmonic.

One of the desired characteristics for this amplifier is high linearity. Third order intercept is a theoretical point at which the third order distortion product is equal to the power of fundamental signal. Since achieving such condition is usually not possible because of saturation and related P_{1dB}, the linear region of 14.15 MHz fundamental and third harmonic was extrapolated, and intersection point was noted. In this case we can read from Fig. 25 that the IP3 = 42 dBm.

In the Fig. 26 it can be seen that the level of the second harmonic stays mostly constant, and for input levels above 28.75 dBm, it is actually lower than the third harmonic, which hints at the canceling effect of a balanced configuration. Second important observation from Fig. 26 is that the level of the second harmonic started to increase only after 34.77 dBm of input power, right before the previously measured P_{1dB} of 36 dBm. On the other hand, the odd third harmonic has consistently increased in power, similar to the first harmonic (fundamental), reaching almost 46 dBm or 40 W.

Because different transformers presented different power output levels, the power of harmonics was plotted as the level below the carrier in dBc. This can be seen as a spurious free range (SFDR) between fundamental and n-th order harmonic.

In Fig. 27, it can be seen that the SFDR increases between the fundamental and the second harmonic up to 34.77 dBm of input power, after which the SFDR drops. This shows that after this point, the fundamental grows slower than the second harmonic. This point, even lower than the P_{1dB} compression, is worth considering when linear operation is critical. On the other hand, the third harmonic results in an SFDR of only 12 dBc, but it is much easier to filter out with low pass filter since third harmonic is further away from fundamental signal.

Second harmonic SFDR in Fig. 28 varies between different transformers. Even though the 1:9 TLT made with 17 Ω cable and parallel connection of 50 and 25 Ω cables had similar characteristics before, here a significant difference in performance can be seen. Surprisingly, the improvised 1:9 transformer with parallel coaxial cables has exhibited the best suppression in the range from 29 dBm to 34.77 dBm of input power with a peak SFDR of 33 Bc. Although the characteristic is not flat, this is a very good result. The transformer made with custom 17 Ω cable, on the other hand, has performed relatively poorly in comparison to the other transformers except above 34.77 dBm input, where it was better than the 1:4 transformer by 1.47 dB, but it was still almost identical to the improvised transformer. This was not expected since the improvised transformer had most likely larger leakage capacitance and its windings were less organized because of difficulties during winding of two parallel coaxial cables at the same core. Perhaps two parallel transmission lines have minimized the losses leading to a better cancellation effect, or perhaps this is an exception occurring at 14.15 MHz.

Ref.	Operating frequency [MHz]	Output power [W]	Efficiency [%]	Gain [dB]	S22 [dB]	2nd harmonic [dBc]	3rd harmonic [dB]	LDMOS
This work	1.6–30	600	69.96 (PAE)	≥18*	≤–15	24.07	10.65	MRF300
[18]	22–24	975	66.5 (PAE)	36	≤–15	24.1	33.98	MRFX1K80H
[16]	27.12	1380	74	24.72	–4.18	30	35	MRF1K50H
[28]	2–30	1300	81.4	29.1	-	17.21	9.92	BLF188XR
[24]	1.8–51	600	70	21	-	-	-	MRF300

*at the input of tested amplifier 3 dB attenuator was present

Tab. 5. Comparison of results from this work to related works.

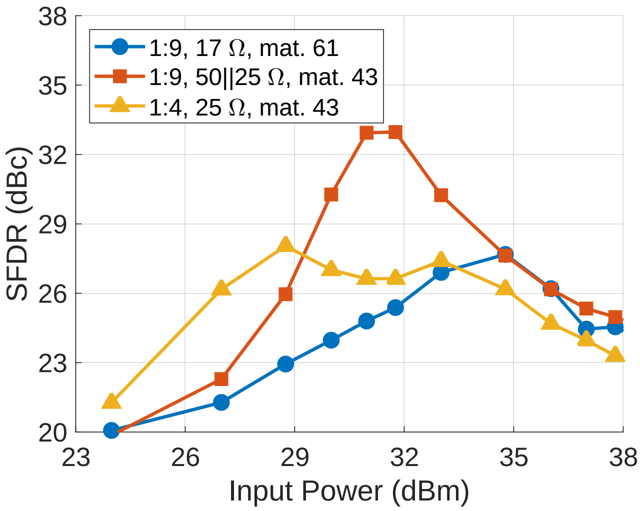


Fig. 28. Second harmonic SFDR for three TLTs at 14.15 MHz.

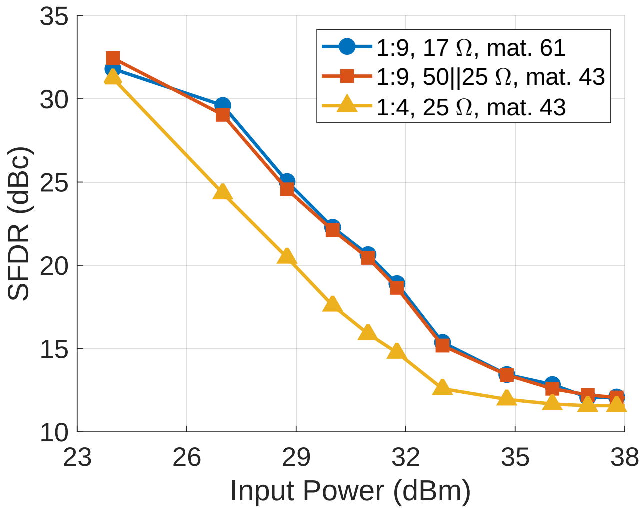


Fig. 29. Third harmonic SFDR for three TLTs at 14.15 MHz.

The cheapest transformer, the 1:4 TLT, has also provided very good results, with a wider range of attenuation higher than 26 dB. Although this transformer performed the worst for inputs above 34 dBm, it has outperformed every other construction between 24 and 28 dBm. This once again shows that this transformer, despite poor RL, has some advantages for balanced configuration.

For the third harmonic, the difference in SFDR between the two 1:9 transformers was negligible. On the other hand, the 1:4 transformer presented a SFDR better by about 5 dB until the input power reached 33 dBm, after which it converged to the result of the 1:9 transformers. From the plot in Fig. 29 showing the SFDR between the fundamental and the third harmonic for increasing power, it can be noticed that the SFDR for large input power is much smaller than in the case of the second harmonic, demonstrating the proper behavior of the push-pull amplifier.

5.6 Discussion of Results

Table 5 compares the most important measurements of the amplifier presented in this work when fitted with a 1:9 TLT made with 3 turns of 17 Ω coaxial cable on 61 type ferrite cores, to other RF amplifiers presented in recent years. A direct comparison of this amplifier to other constructions is difficult since consumer grade amplifier manufacturers don't always publish in-depth parameters such as return loss. A major motivation for this work was to measure various TLTs that can be utilized in a shortwave power amplifier. Designs referenced from other works also use different metrics and operate at different output power levels. We have gathered some data that is close to the amplifier presented in this work to visualize the advantages of our solution, demonstrating that the presented effort contributes valuable data for the design of RF amplifiers in HF spectrum.

Our solution has exceeded the minimal acceptable gain. In fact, the value in Tab. 5 is measured with a 3 dB input attenuator since the input source could not be controlled in steps smaller than 1 W. Therefore, the gain in the linear region is indeed the same as in [24]. We have also met the expected return loss for the entire HF spectrum. A return loss of less than –15 dB was achieved between 1.568 MHz and 30.02 MHz for the TLT made with a special type of coaxial cable which was the desired goal. Our measurements have also demonstrated that with a decreased number of turns, the same transformer can operate with the same minimal return loss from 1.8 MHz to 34 MHz.

It was impossible for us to evaluate the performance of the amplifier above 30 MHz due to a lack of an RF source with power high enough to drive the amplifier. We expect that the power would drop significantly. However, based on return loss measurements in the 1 MHz to 60 MHz spectrum in Fig. 15 and the work done by Razvan Fatu, it can be assumed that the amplifier would work well up to 50 MHz, especially when fitted with the TLT made with two turns of 17 Ω coaxial cable on 61 type cores which achieved –11.3 dB return loss at 51 MHz. In fact, the amplifier presented here is very similar to [24], though in Razvan Fatu's design, the TLT used TC-18 cable and no return loss measurements were provided. In our measurements, we have tried to address this gap by measuring both improvised TLTs made with a parallel connection of general-purpose coaxial cables and TLTs made with special-purpose cable, both in terms of return loss, maximal power and linearity.

While the measurement of return loss has shown that all properly built 1:9 transformers provided very good return loss in the 2–30 MHz spectrum, with values as low as –23.06 dB noted at 8.12 MHz, it is worth mentioning that operation with return loss below –10 dB is often considered acceptable. Such a return loss would most likely be safe for the amplifier, especially if it was operated below the 1 dB compression point, as no degradation or failure of transistors was observed after the tests with the 1:4 transformer were conducted.

Indeed, the 1:4 transformer allowed reliable operation with very good efficiency of up to 87% however, it did not allow for maximal power to be transferred to the load, and therefore we cannot ultimately recommend such a solution where low output impedance is present. The linearity of such a configuration can also be brought into question. As expected, the measurement of the 1:9 TLT made with improper impedance of the transmission line has shown a rapidly degrading return loss as the ratio of used transmission line length to wavelength started to increase.

At a frequency of 14.15 MHz, our amplifier also produced lower harmonics compared to [28], which was measured at the 13.56 MHz ISM band. This is most likely due to the fact that NXP used 200 mA of quiescent current, and we chose 400 mA. This difference in bias of AB class is another factor showing that comparison of results is more complicated. The amplifiers that had the best level of harmonics in Tab. 5 were the narrow-band amplifiers that were tuned to a particular part of HF spectrum. In other solutions TLTs were used, and so the output network was able to sustain the higher order harmonics. The narrow-band 975 W amplifier presented in [18] was tested for harmonics with an output power of 438.5 W and achieved almost the same suppression of the second harmonic as our wide bandwidth amplifier with 600 W output. This result was unexpected, but it shows a good ability for second harmonic cancellation. What is disadvantageous is the still high level of the third harmonic. However, that order is always easier to attenuate using a low pass filter.

6. Conclusions

Our measurements demonstrated a wide bandwidth behavior of a real transmission line transformers. Measurements of return loss for six constructed transformers in Sec. 4 have shown that for most broadband operation the characteristic impedance of used transmission line has to be a geometric average of input and output impedance, confirming the theory presented in Sec. 1. Special case of transformer using a wrong impedance of transmission line shows also the importance of length of transmission line with respect to operating wavelength. For a very small length of used transmission line when compared to wavelength a good return loss was observed, but the bandwidth was very small when compared to a properly built transformer. This might also hint at the fact that smaller transformer in general could exhibit even larger fractional bandwidth.

While both the improvised 1:9 TLT made with two parallel transmission lines and the well-made 1:9 TLT with special purpose 17 Ω coaxial cable have shown similar characteristics, both in terms of return loss and circuit performance, the 1:9 TLT with special purpose coaxial cable performed better. It presented higher output power across all the bands, and at the bottom and the end of the spectrum, it achieved better power added efficiency. This improved PAE at the ends of the spectrum was the main factor why it was chosen as the final TLT installed in the amplifier. The improvised 1:9 TLT has shown decent performance for a narrower spectrum. If the engineer wants to design an amplifier working only in specific bands, this study shows that the utilization of a well characterized transformer with two parallel transmission lines is feasible. However, if broad bandwidth with good characteristics across the whole spectrum is desired, we recommend the use of special single transmission line with a characteristic impedance that is the geometric average of the input and output impedance.

Acknowledgments

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