A Low Noise Preamplifier With Improved Output Reflection For The 2m Band

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The development of LNAs for the frequency range from 1 to 1.7GHz was presented in a lecture at the UKW Conference 2012 in Bensheim, it was subsequently published in [1]. Due to the excellent noise characteristics of the prototype a version for the 70cm band was examined, this gave similar good results as described in [2]. A 145MHz version has also been tested and the results were presented at the UKW conference 2013. The problems that occurred and the solutions found are described in the following text.

1 Introduction

Following the good experience with 1 to 1.7GHz it suggested that development should be focussed on the lower bands. The MM1C properties show that it is for use in the area of 1.9GHz but the noise properties are not documented at lower frequencies. The development for a 70cm amplifier was published in VHF Communications Magazine [2] with the data obtained. It showed that the minimum noise figure was virtually constant at NF 0.3 - 0.4dB even with the reduction in frequency. These results motivated the development of a version of the LNA for the 2m band in the expectation that such values would be measured there as well. There was the following wish list:

- Noise figure: maximum 0.4dB
- (S21) of at least 20dB
- Absolute stability (k greater than 1 up to 10GHz)
- The output reflection S22 that should be possible (Dream value = -20dB at 145MHz)

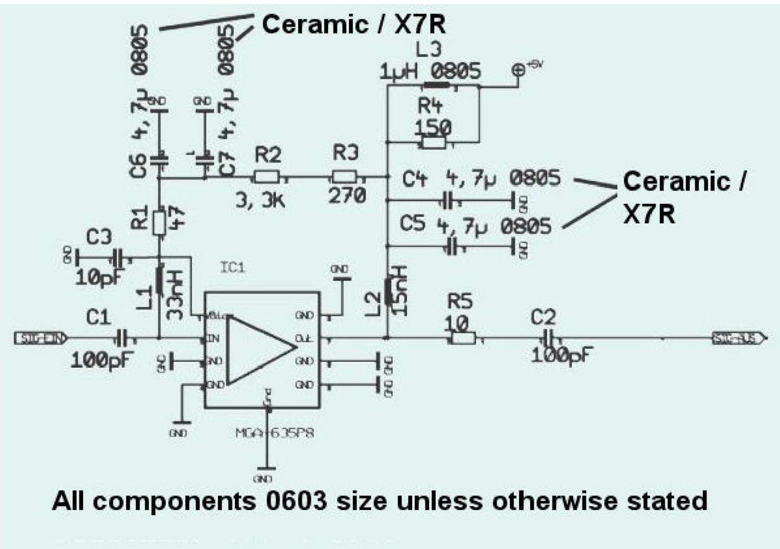


Fig 1: The 1 to 1.7GHz amplifier

Part 1: The development changes

1.1. The amplifier changes

The circuit for the 1 to 1.7GHz LNA is shown in **Fig 1. Fig 2** shows the LNA PCB in its aluminium enclosure, full details are given in [1] and [2]. The most important things are:

It uses a GaAs-pHEMT MMIC cascode amplifier with the “bias” on Pin 1 removed and applied at Pin 2 via an inductor directly to the gate the first

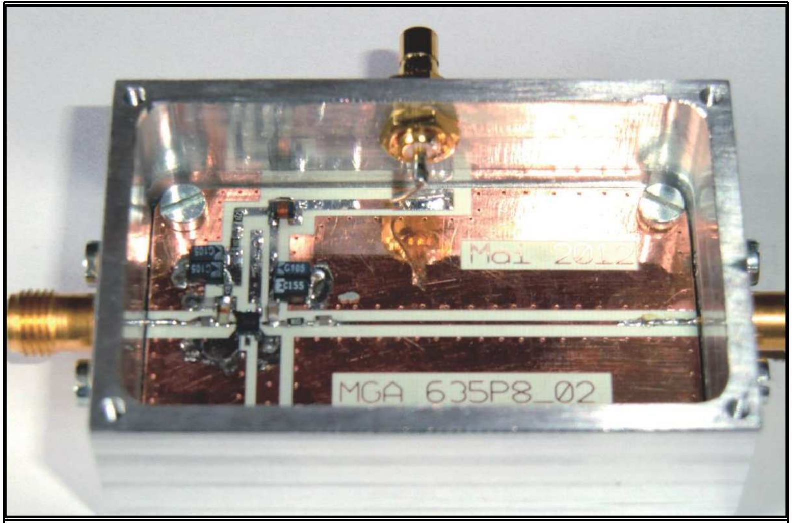


Fig 2: The 1 to 1.7GHz amplifier in its aluminium housing

pHEMT. The other inductor, L2, on Pin 7 provides the supply for the second device. The complete amplifier operates from a single supply voltage of +5V.

A big problem with HEMT devices is stability at lower frequencies giving the tendency to oscillate. I have found a simple trick to solve this: with the decreasing frequency, the resistor R1 becomes more relevant. With approximately 50Ω at the input pin 2 the oscillation is stopped.

The CAD program "ANSOFT Designer SV" [3] was used to determine the noise performance of the circuit shown in **Fig 3** at 145MHz. The changes necessary to the 1GHz version are shown in red on the circuit. This was done by changing the values for L1 and L2 as well as the capacitor C3 and making simulations for the noise figure NF in dB AND continuous control of the stability. An attempt was made to set the minimum noise figure at 145MHz and to optimise noise figure NF. It has sufficient stability to 10GHz resulting from in a small additional 15Ω resistor in the output circuit (but close to the output pin of the MMICs). The grounding coplanar waveguide with a conductor width of 0.59mm, a gap of 1mm on each side and a length of 32.5mm was not missing in the simulation.

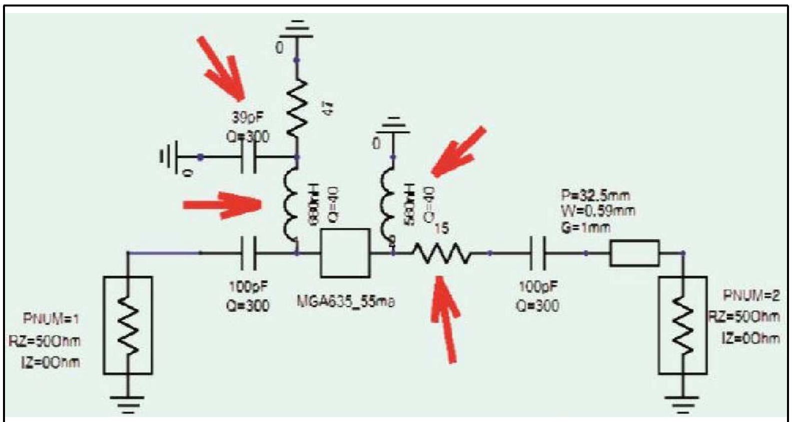


Fig 3: These marked changes make it a 2m LNA

The resulting noise characteristics (see **Fig 4)** are of course a dream! However, one had to check on a prototype whether this is true - after all, no noise parameters are contained in the S parameter file for this frequency range. In these circumstances the simulation program

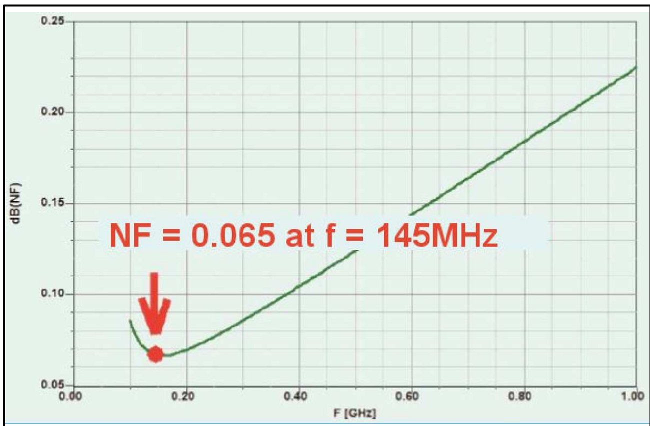


Fig 4: This simulated noise figure can not be believed (see text)

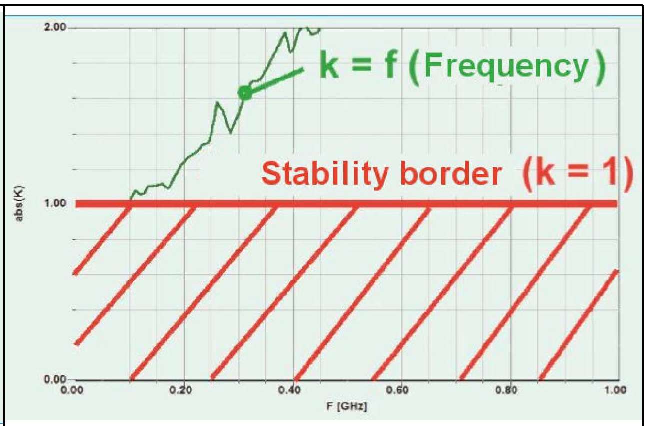


Fig 5: This does not affect stability problems up to 10GHz

proceeds by simply applying a linear decrease of the noise with decreasing frequency!

Also the required stability (k greater than 1 up to 10GHz) is not an issue with the 15Ω resistor in the output circuit - as can be seen in **Fig 5.**

The simulated S-parameters for this frequency range are also of no concern (**Fig 6**).

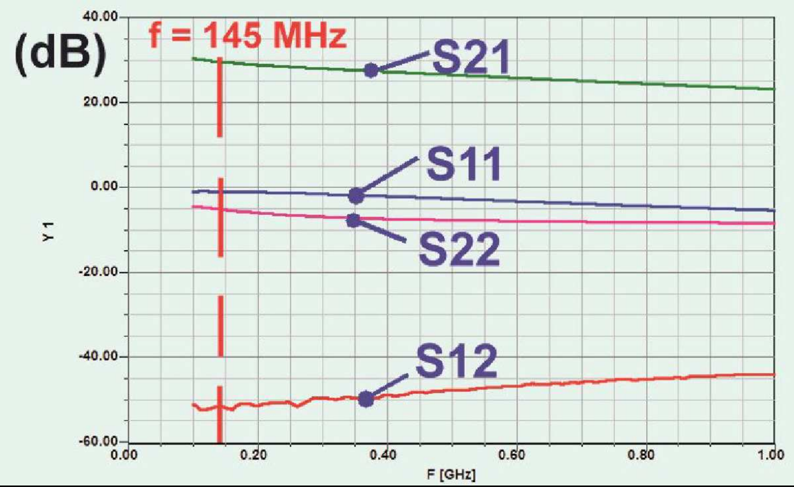


Fig 6: These are the S parameters up to 1GHz that are of interested for operation in the 2m band

1.2. Improving the Output Reflection S22

You cannot change anything at the input of an amplifier after a successful noise adjustment, but the output can be optimised without disturbing the noise characteristics. To do this, simulate the circuit without the 50Ω line at the output in order to directly determine the parameters of the MMIC output pin 7. If we now represent S22 in the Smith chart (**Fig 7**), we see:

At f = 145MHz, S22 results in an output impedance which can be represented by a circuit with R =

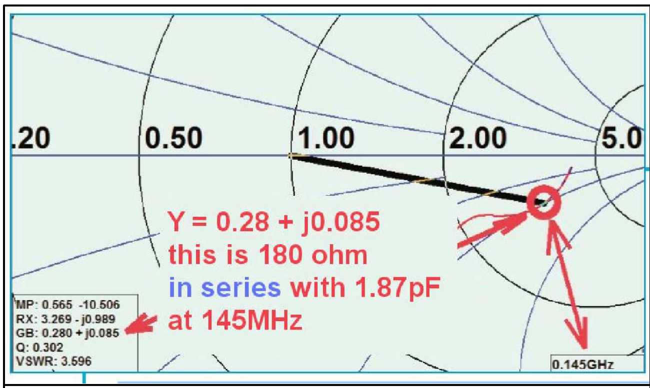


Fig 7: The Smith Chart must be expanded to see and use the S22 details

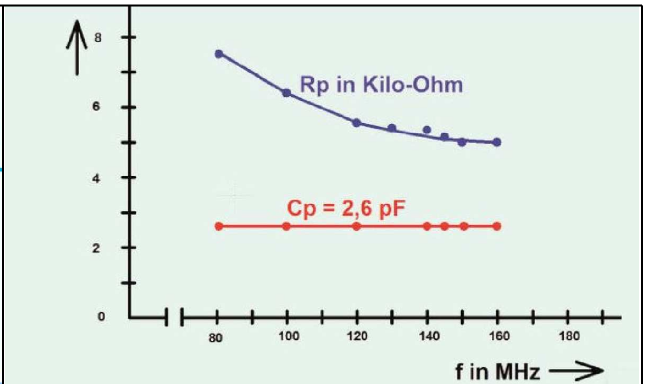


Fig 8: Finally the measurements of the transformer input capacitance and resistance

180Ω in series with C = 1.87 pF. If a transformer with the voltage ratio ü = 2: 1 is inserted there the impedance components change by a factor of 2^2 = 4. The real part of S22 is reduced to 180Ω / ü2 = 45Ω and the capacity naturally increases to C = (1.87pF) x 4 = 7.5pF.

This can be compensated by an additional inductance. The result is a substantially smaller reflection S22. The development of the appropriate transformer is the subject of Part 2.

The output impedance of the first prototype was measured with a BOONTON RX meter in the range from 80MHz to 160MHz. This gave a constant input capacitance of 2.6 pf. In the range of 140 to 160MHz the loss resistance of the ferrite material is a parallel resistance between 5 and 5.5kΩ (**Fig 8**).

This measured input impedance (2.6 pF in parallel with approximately 5.2kΩ) can now be combined with an ideal 2: 1 voltage transformer and used in the simulation since the transmission ratio remains

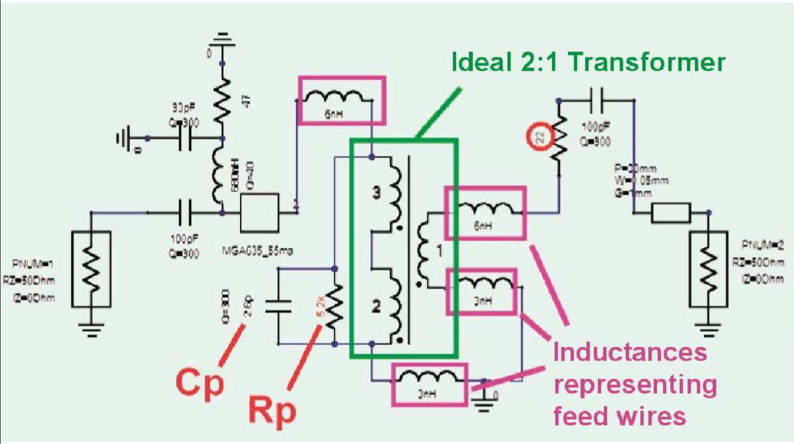


Fig 9: The simulation diagram looks much more complicated with the transformer and associated wiring inductances

constant over a very wide frequency range (see Part 2).

The Simulation circuit diagram was greatly expanded (**Fig 9**). It contains the ideal transformer, combined with the measured input resistance and the input capacitance. The wires to the input and output add additional series inductances. For these thin wires, one must count as 1nH per millimetre of the transformer wires which results in 3nH on each side and 6nH on both ends. In addition the SMD inductance of 560nH at the output has been omitted since the power supply of the MMIC can now be fed via the "transformer".

According to the design, an increase in the "anti-oscillation resistor" at the output of the MMIC from 15Ω to 22Ω does not significantly increase the stability, but does reduce the output reflection S22. Of course the microstrip up to the output should not be left out.

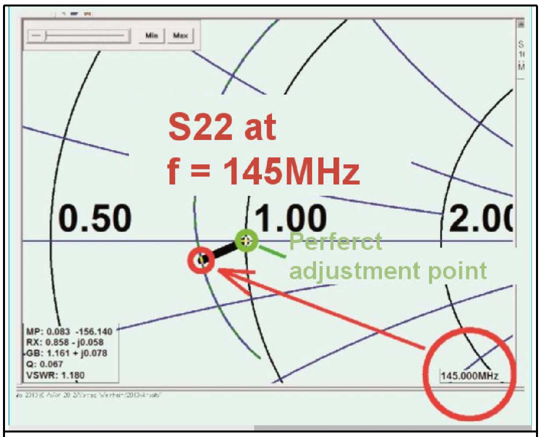


Fig 10: All the work on the transformer has paid off: The S22 curve is now very close to the perfect adjustment point

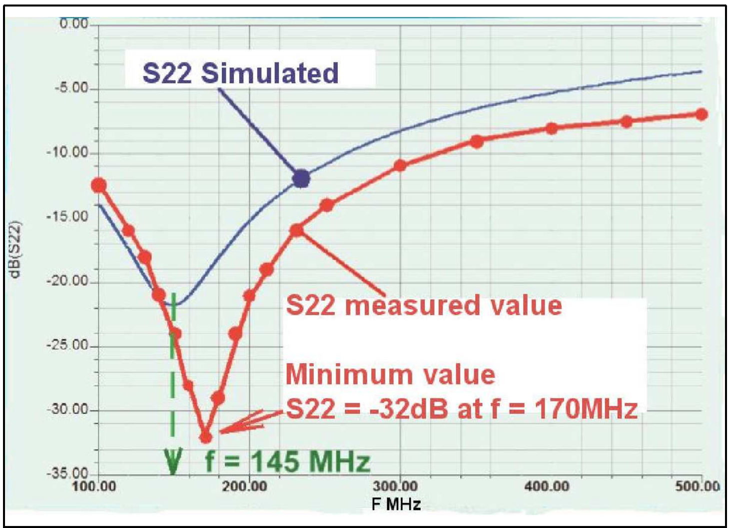


Fig 11: The measurement show a big surprise: the match was much better than expected

The Smith Chart (**Fig 10**) shows the result of this work. A great surprise was the S22 measurement on the prototype board, the result exceeded the greatest expectations (**Fig 11**). The stability factor "k", is above k = 1 up to 10GHz, see **Fig 12**

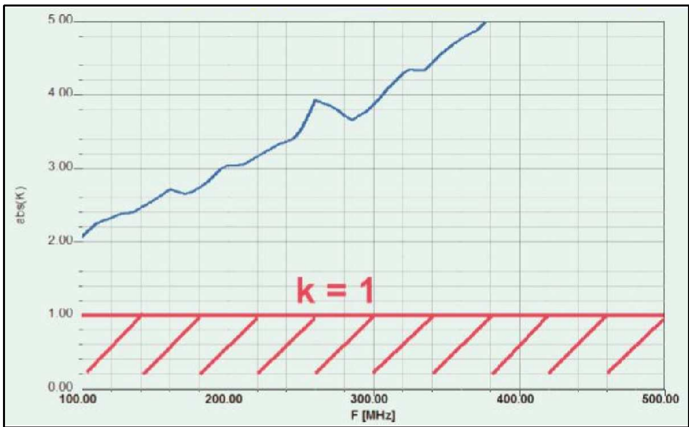


Fig 12: A really satisfying stability curve

The other parameters measured on the prototype are also of interest. First S11 was measured (**Fig 13**): the curve starts with -2dB at 100MHz and drops smoothly to just -4dB at 500MHz. A value of +28dB at f = 145MHz was measured for S21 (**Fig 14**); at 500MHz it is only +24dB.

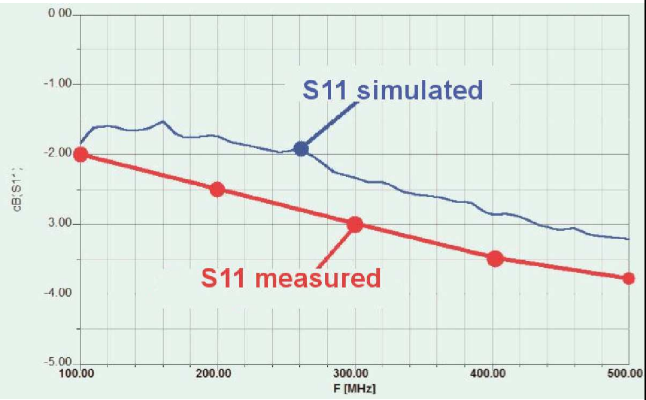


Fig 13: S11 looks good

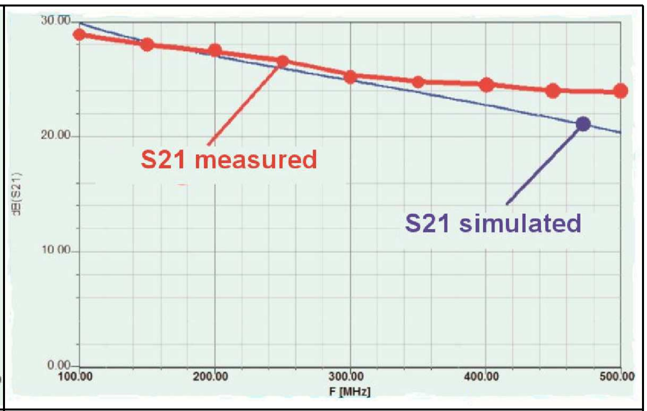


Fig 14: S21 of the finished amplifier is exactly where it should be

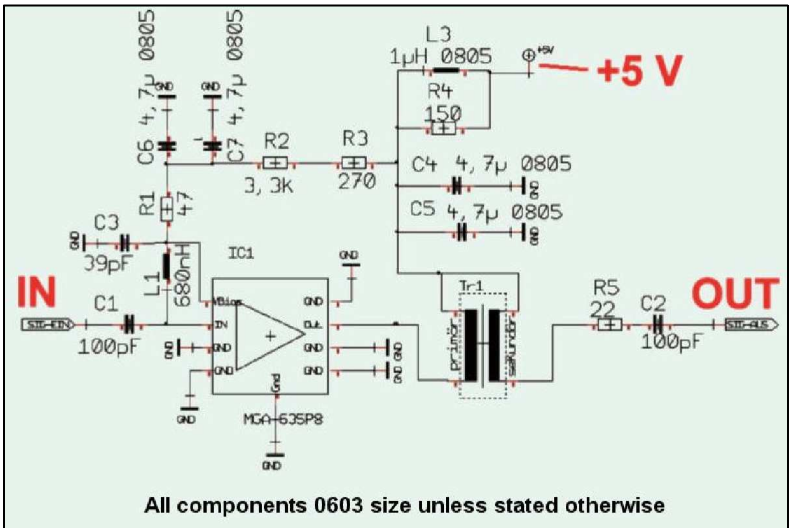


Fig 15: The circuit diagram has a few component changes and an output transformer

(Measurements of S12 are not shown because the values were all below -40dB).

The noise figure also looks good:

The measured noise figure NF starts with 0.35dB at 100MHz, passes through a minimum of 0.32dB at 145MHz and again reaches 0.35dB at 435MHz. Finally, **Fig 15** shows the complete circuit diagram.

Part 2: The output transformer

2.1. Theory

The output transformer consists of coils wound on a ferrite core. If an RF signal is applied to the input and the output is terminated correctly the following effect can be observed if the "coil" has sufficient inductance:

The signal transmitted is not affected by the ferrite core because the magnetisation caused by the current flow in the core balances the return current. On the other hand, the higher the inductance of the coil and therefore its inductive resistance, the better the suppression of "unwanted signals" (this will also determine the lower frequency limit, since more and more current "flows down to ground" through the inductance of the primary side and is therefore missing at the output).

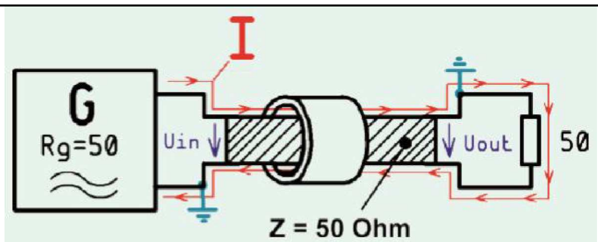


Fig 16: A simple line transformer

This can be seen in the circuit diagram in **Fig 16**:

If the current that flows back and

forth is identical (fundamental) and the upper conductor on the output side is grounded, the output voltage is suddenly antiphase, so you have a "1:1 phase reversal transformer"!

If, on the other hand, the grounding on the output is removed, a "ground-free voltage" is obtained and it becomes a balancing transformer.

2.2. First stage of development: A phase inversion transformer

This stage used the core from an EMI Suppression 6-Hole Ferrite Bead by Wurth Electronics designed for the frequency range of 100MHz to 200MHz. It is intended for broadband chokes and has the following data:

- Order number: Wurth 742 75043
- Material: 3W1 200
- Impedance according to data sheet: 961 ohm at 100MHz

The coil was made from 0.2mm diameter enamelled copper wire with two wires twisted together (with 5 twists per cm). The two wires were 30cm long each and twisted together in a coil winding machine. The winding machine was run for a predetermined "winding period".

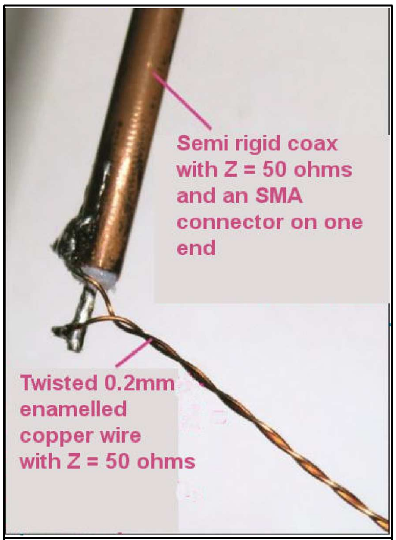


Fig 17: It is easy to measure the 50Ω line with the correct equipment

Reasonable and fast results can obtained with an impulse reflectometer. I used a old professionally instrument from HP that I found in a HAM Radio Flea market! But many radio and TV companies or broadband network installers have something like this that they use for testing and investigating complaints on antennas and broadband cable networks. You may be able to borrow such equipment over a weekend.

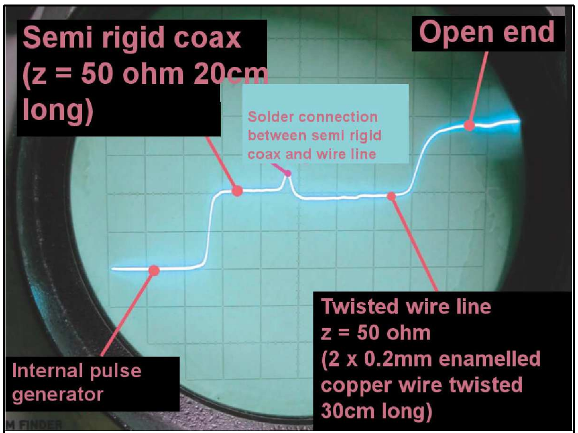


Fig 18: The impulse reflectometer is probably the best measuring instrument for this job

Fig 17 shows how the development was done: take a piece of semi-rigid cable with an SMA connector on one end and open at the other end. The SMA end is connected to the reflectometer and the other end is soldered to the 50Ω cable that was just made then tested.

After several tests of the characteristic impedance with the pulse reflectometer, it was found that 4 to 5 "twists per cm" of 0.2mm enamelled copper wire gave Z = 50Ω. The corresponding display on the reflector probe can be seen in **Fig 18** which is self-explanatory.

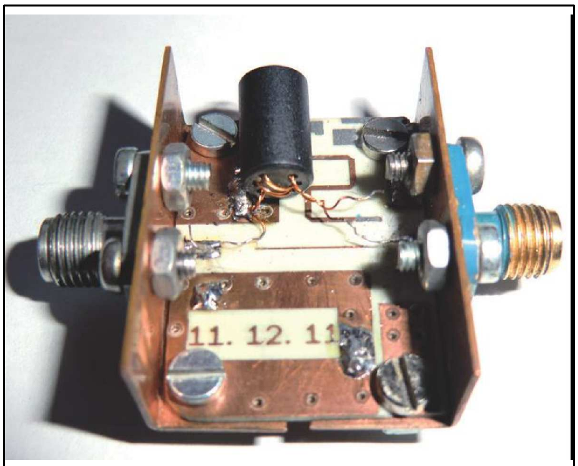
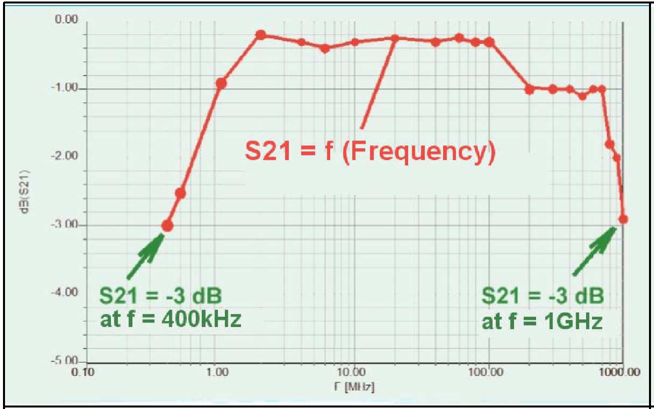


Fig 19: The test setup for the transformer to be tested on the network analyser



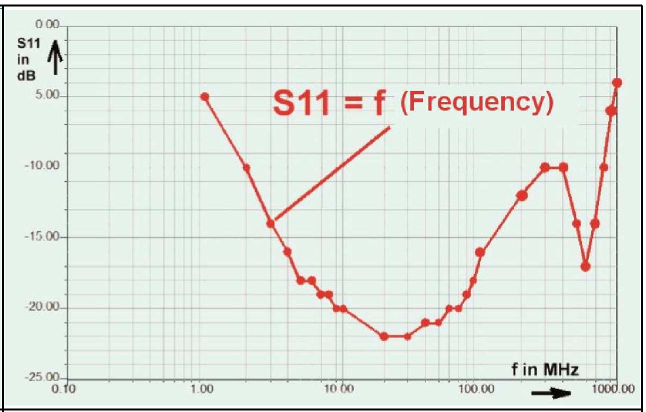


Fig 20: An incredible result: a 3db bandwidth for S21 from 500kHz to 1GHz

Fig 21: The input reflection S11 also shows an interesting curve in this frequency range

Three turns were added to the core using this cable and a simple measuring device with two SMA sockets was assembled from an old RF circuit board.

This arrangement is shown in **Fig 19**; it was connected to the network analyser to determine S11 and S21. The result is shown in **Fig 20** and **Fig 21** and it is quite impressive.

2.3. Second development stage:

The impedance transformer required (with the voltage ratio ü = 2)

This requires two six-hole ferrite cores wound with 100Ω wires. On one side the two wires are then connected in parallel while on the other side they are connected in series. Then:

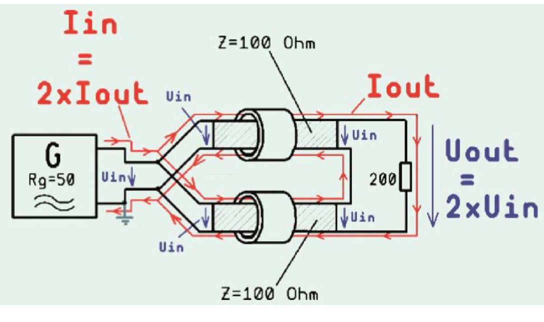


Fig 22: This shows how a 4:1 impedance transformer works

On the right hand side of **Fig 22** (with series connected wires) gives twice the voltage but only half the current compared to the left input side. The result is quadruple the resistance on the right which can be described as follows:

The terminating resistor of 200Ω is transformed down to 50Ω and thus correctly matched to the generator resista-nce.

There are four interesting grounding possibilities on the right side:

- A "ground balanced system" is obtained by grounding at the centre (grounded at the connection point of the two wires) giving two equal phase but equal sized output voltages (each with the amplitude U_{in}).
- If you do NOT ground the centre (NO ground at the connection point of the two wires) you get a "ground free symmetrical system" with two equally large voltages OR twice the output voltage
- Grounding the upper end, the output voltage is twice as high as the input voltage but antiphase.
- Grounding at the lower end, an in-phase output voltage is obtained, also with twice the amplitude of the input signal.

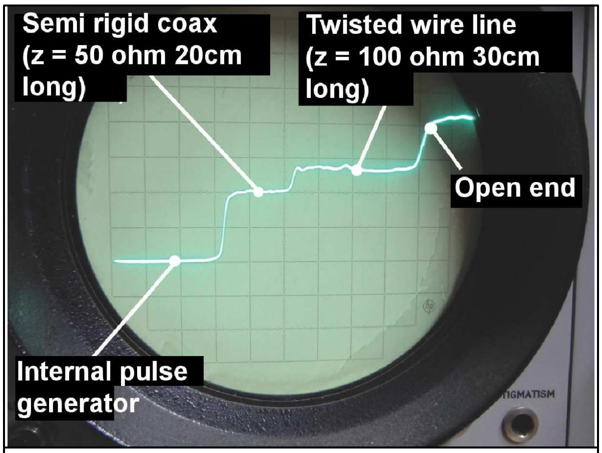


Fig 23: The impulse reflectometer shows the characteristics of the 100Ω line from beginning to end

The biggest problem with this transformer is the availability of a high quality 100Ω line. There is nothing on The Internet because wires with high quality Teflon insulation are required. So there seemed no way forward and that was a big problem. In order to obtain the characteristic resistance, a thicker wire insulation is required to increase the distance between the two twisted wires, thus spoiling the capacitance. The result of many experiments was to use “10 x 0.05” litz wire which gave the desired value of Z = 100Ω with only 4 twist per centimetre. The final state of the development is shown in **Fig 23**. The characteristics of this 100Ω line are shown on the reflectometer.

Two transformers were fabricated and assembled as follows:

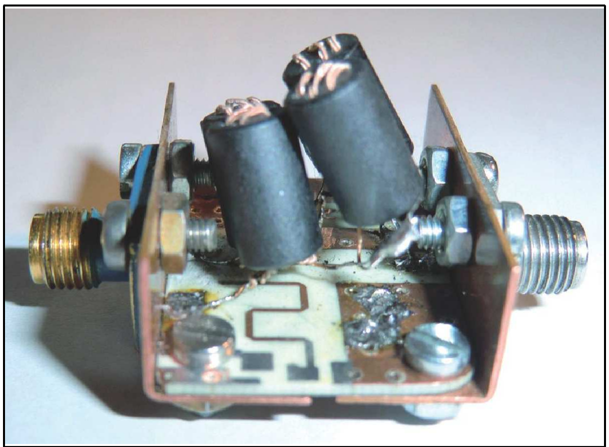


Fig 24: The first test 2:1 transformer

The first transformer transforms from 50Ω to 200Ω and the second is connected to its output as a "mirror image” with its high resistance side. This gets back to 50Ω and the network analyser can be used to determine the properties of the device (**Fig 24**).

The results of the measurements are really interesting because very good values were obtained for S11 and S21 plus a 3dB bandwidth from 500kHz to 1GHz (**Fig 25** and **Fig 26**). This was the same as the phase reversing transformer in the previous development stage

The completed amplifier module is shown in **Fig 27** (compare the output with **Fig 2**). The impedance transformer is glued to the circuit board with epoxy resin (UHU plus).

Here's another important tip:

The track had to be cut at the MM1C output. If this is done with a scalpel it pulls up the track

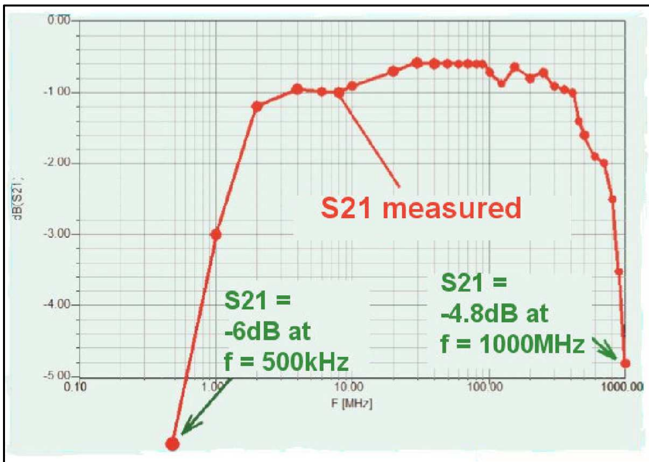


Fig 25: Two transformers connected in series. No more is required for S21

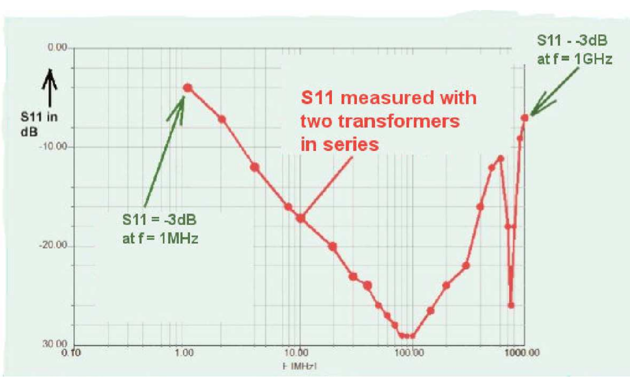


Fig 26: Good values for the input reflection S11 at the required frequency of 145MHz

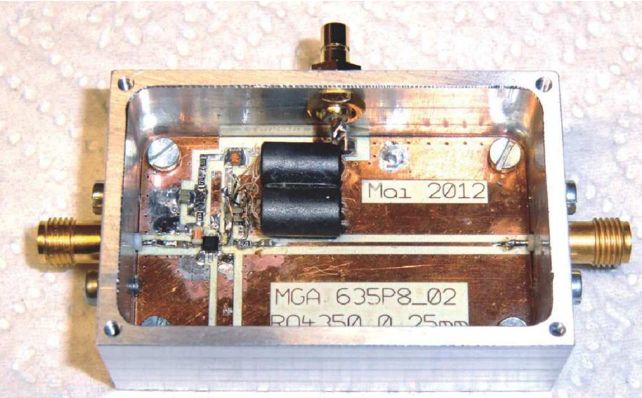


Fig 27: The completed 2m amplifier

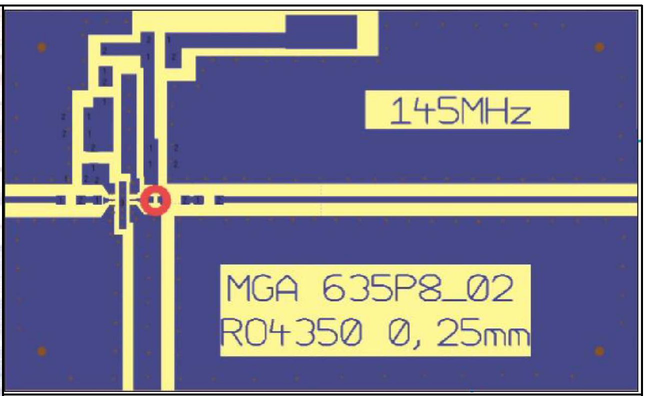


Fig 28: The PCB must be modified by hand as shown for the 145MHz amplifier

(because the copper layer is not very firmly adhered to the R04350 material). This can only be done with a thin diamond cutting disc in the famous small handheld drill called "DremeI”

The same applies to drilling holes: This should be done ONLY with carbide drills and ONLY with the highest speed with the PCB on a hard wood surface. Care should be taken to prevent the drill hitting and damaging the PCB when drilling the large holes for the fastening screws.

The layout of the finished board is shown in **Fig 28**. It is 30 mm x 50 mm and specially prepared for the new module. A board for the 1 GHz amplifier was modified at the marked point by cutting the output track.

Finally finished - this part was so interesting but tiring.

Part 3: What other people say

In addition to acknowledging words, the question often arose:

What does the amplifier support at the input? How is it protected against over voltage or static electricity when connected to an antenna?

These are justifiable concerns and therefore it was necessary to think about this in order to prevent a threat.

3.1. Considerations for the permissible input level

If you take a look at a modern DVT stick (as used for an SDR) you will find a small three leg SMD device in an S0T23 package that contains two back-to-back diodes. When Schottky diodes are used the peak value of the input signal is approximately 0.4V. The recovery times for such modern diodes (usually below 100 picoseconds) works well into the GHz range and the small self-capacitance of each diode (usually approximately 0.5pF) keeps the deterioration of the input reflection within tolerable limits.

The MGA635-P8 has been designed for the "OlP3" (third order output intercept point) as well as the "O1dB" (output 1dB compression point) from 1.9GHz to 3.5GHz

To get to the values for 145MHz consider the following:

At 1.96Hz the O1dB value is about +19.5dBm then for the circuit with an amplification of approximately +19 to 19.5dB (according to the data sheet and my own simulations) this means an input level of approximately zero dBm.

Since the O1dB value changes very little above that frequency (that indicates the start of overloading the final stages) it should also be of the same order of magnitude at 145MHz or only just under it. However, the gain there has risen to +29dB and so the 1dB compression occurs with a 10dB lower input level (-10dBm).

It is assumed that the protection diodes do not cause any distortion to the input at this frequency. The level of -10dBm corresponds to a voltage of 70.8mV or a peak value of almost exactly 100mV. A SPICE simulation was used to determine:

- If the Schottky diodes will cause any problems
- If the higher level and the resulting limiting effect cause additional harmonics in the spectrum.

3.2. The SPICE hour of truth

The first task was to search for an appropriate Schottky diode on The Internet, the BAT17-04 was selected. It is the required back-to-back format with a capacitance of 0.55pF per diode in an S0T23 SMD plastic package. The most important reason for the decision was availability (CONRAD Electronics / 0.22 euro from stock). The next step was the search for a SPICE model for the BAT17, which was quickly found in a large ORCAD collection:

```
.model BAT17 D(Is=3.167n N=1.104 Rs=.7144 lkf=8.133m Xti=3 Eg=1.1 1
+ Cjo=921 .1f M=.3333 Vj=.5 Fc=.5 lsr=50.62n Nr=2 Bv=4 lbv=10u)
* SiEMENS pid=bat17 case=S0T23
* 91-08-29 dsq
*$
```

This short file should be retrieved and copied into a new text file. This must be saved with the correct name and the required extension as a BAT17.mo in the folder "LTsplceIV I lib / sub".

The simple simulations are shown in **Fig 29**. The diodes can be found as "schottky" in the onilne component library, the name must be changed to "BAT17". Our newly created SPICE model is then connected to the directive

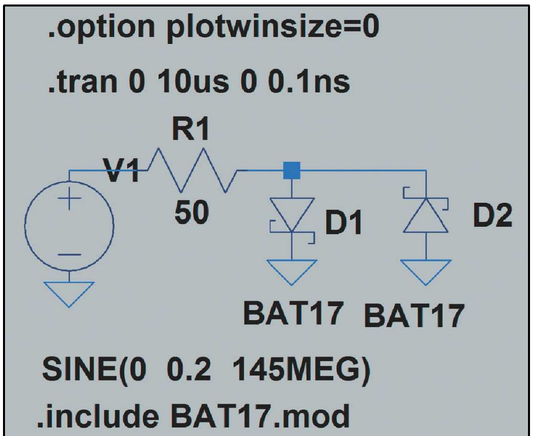


Fig 29: The limiting effect of two back-to-back Schottky diodes can be investigated with LTspice

. Include BAT1 7.mod

And thus made available to the simulation. Do not miss this instruction

.option plotwinsize = 0

For switching off the data compression during storage, this is the only way that a correctly simulated spectrum can be obtained via the FFT (Fast Fourier Transformation).

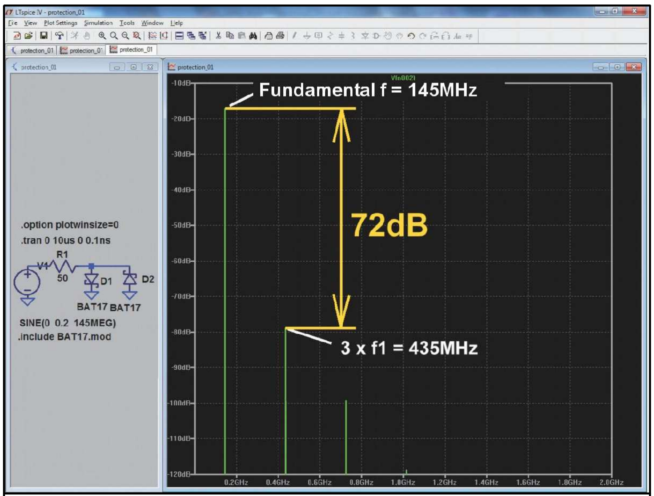


Fig 30: The diodes cannot cause any damage at the 1dB output point

The circuit is fed with a sinusoidal signal with the frequency f = 145MHz. The generator has a peak value of 200mV and a resistance of 50Ω. This result is in an incident wave with amplitude of 100mV. A 10 microsecond simulation was used with a time resolution of 0.1 nanoseconds in the time domain with the following instruction:

.tran 0 10us 0 0.1ns

This achieves 100,000 real computed points and then allows a secure FFT with 65536 samples. **Fig 30** shows the resulting spectrum and in this case only odd numbered harmonics can be found for symmetrical limiting by two back-to-back diodes! An attenuation of 72dB at the third harmonic allows us to breathe a sigh of relief because the diodes cannot cause any degradation at the 1dB compression point; the amplifier alone produces the distortion factor at this point.

It is interesting to compare the additional noise spectra generated by the diodes for increasing input levels in **Fig 31**. The investigation ranges from the peak value = 0.2V through 0.5V and 1V to 2V. This not only shows the level differences of the harmonics but also the effect of distortion are clearly visible on the fundamental. The level increase at f = 145MHz above Uo_peak = 1V is insignificant.

The all clear can be given from **Fig 32** that shows the time domain signal of the limited input voltage with a primary voltage of 2V. It does not go to much more than 0.5V peak and the additional attenuation of the input by the input resistor of the amplifier has not been taken into account.

Finally, for people who like to construct: I have a small number of the gold plated PCBs for the 1 to 1.7GHz version with the MMIC already soldered. Depending on the frequency range required, the necessary board changes and new components can be added.

For those who are interested in such a circuit board with MMIC, please contact me (mail@gunthard-kraus.de). We can also discuss a ready wound 4: 1 impedance transformer for this 2m amplifier version.

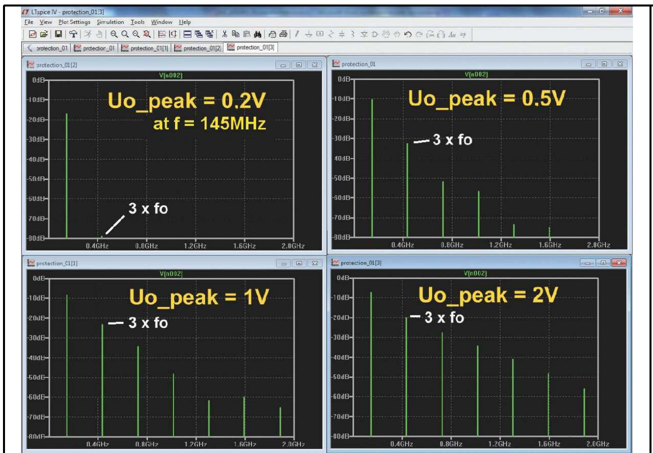


Fig 31: When the primary voltage is increased to 2V many harmonics can be seen plus the limiting effect of the fundamental

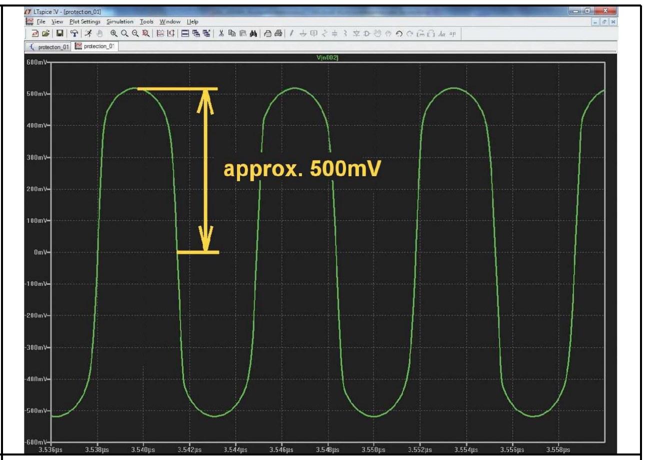


Fig 32: The diodes fulfill their limiting task correctly and do not allow a peak value higher than approximately 500mV

Literature:

[1] Development of a preamplifier from 1 to 1.7GHz with a noise figure of 0.4 dB, Gunthard Kraus, DG8GB. VHF Communications Magazine, 2/2013 pages 90 - 101

[2] A low noise preamplifier for the 70cm band with a gain of 25dB and a noise figure of approximately 0.4dB, Gunthard Kraus, DG8GB, VHF Communications Magazine, 4/2013 pages 201 - 213

[3] Student version of the well-known Microwave CAD software ANSOFT Designer. ANSOFT is officially no longer available but the company has allowed the program to be downloaded from the author's homepage (www.gunthard-kraus.de) (Attention: more than 110 megabytes)