

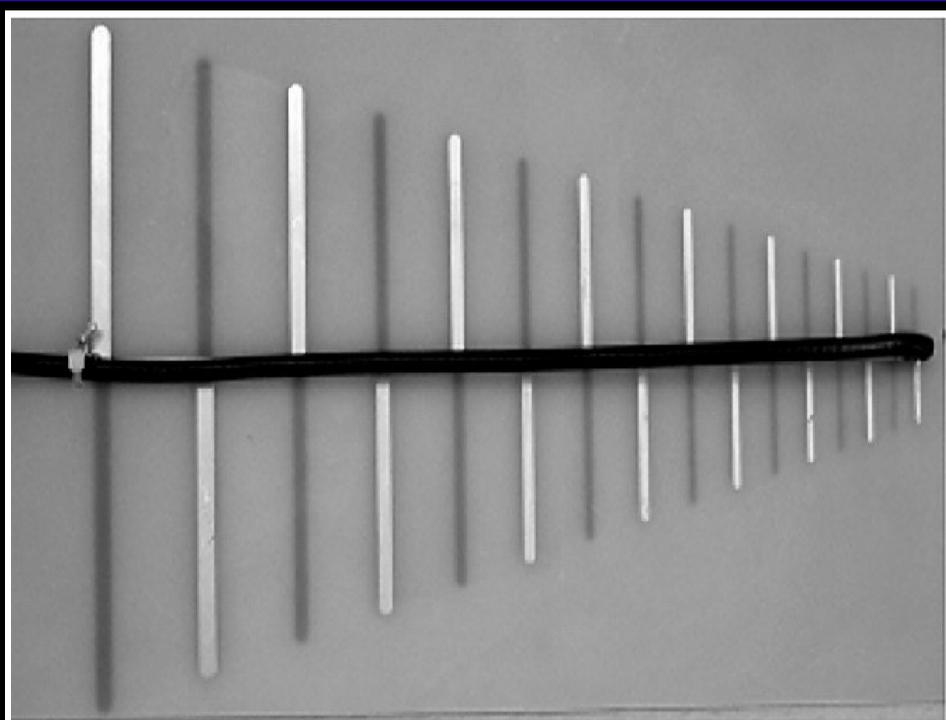


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# **VHF COMMUNICATIONS**

Volume No.38 . Spring . 2006-Q1 . £5.25



**Design for a printed circuit board antenna,  
using a Log Periodic antenna as an example**

*Thomas Bergmann, DG8NTB and Johannes Schad, DG6NDS*

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*Thank you to everyone who has renewed their subscriptions this year. As you will see from the advertisement on page 62 I have now produced two compilation CDs from recent articles, these are now proving quite popular. I am planning to produce some more CDs during 2006, so if you have any suggestions for content, please let me know.*

*This magazine has been printed by a different printer, the previous printer has produced the magazine since 1994 and is now going into semi retirement, I hope the new printer will maintain the same quality and produce the magazine for the next decade*

73s - Andy



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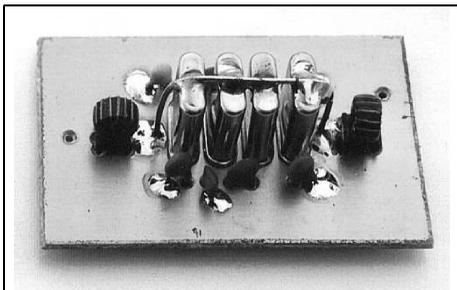
Wolfgang Schneider, DJ8ES

# Design and assembly of a simple 4 pole ladder filter

A 4 pole ladder filter operating at 5MHz with a bandwidth of 400Hz is to be designed for experimental purposes. The input and output impedance is to be 450 $\Omega$ . This can be transformed to 50 $\Omega$  using a simple 1:9 transformer (Fig 1). The complete filter can then be measured using standard equipment.

## 1. Introduction

It is more and more common to use DIY quartz crystal filters for intermediate frequency amplifiers in amateur radio equipment. An argument frequently used in favour of DIY filters is that the costs are low by comparison with commercial



**Fig 1: Prototype crystal filter (ladder filter) with 1:9 transformers fitted to transform the impedance to 50 $\Omega$ .**

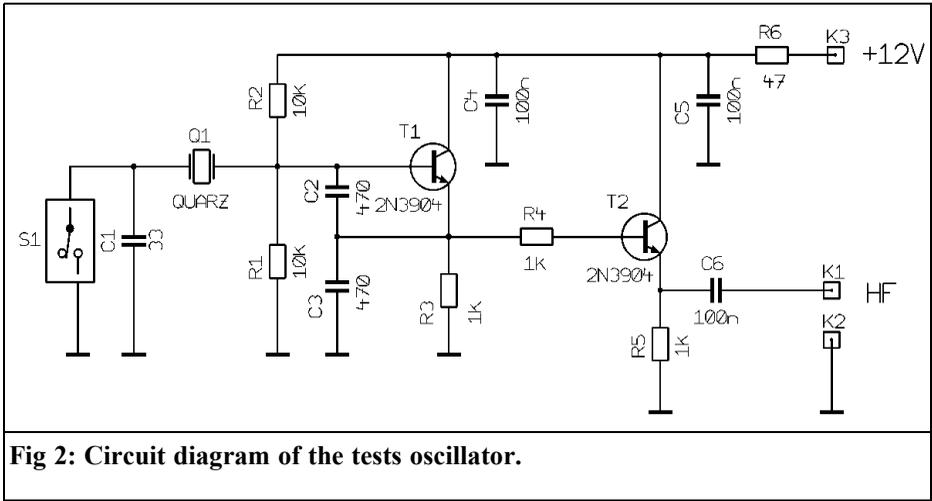
ready-made filters.

However, DIY filters do not come completely free of charge. The number of crystals required must be selected, to meet certain criteria, from those available (or that can be acquired). This puts the emphasis on the measurement of the required crystal parameters.

## 2. The selection of suitable crystals

First of all, a range of identical crystals with the same frequency must be procured. In addition to crystals available on the surplus market, those manufactured for microcontroller applications are also available. These are often available at favourable prices in relatively large quantities. In this example, we shall be using ten 5.068MHz HC49U crystals.

Each crystal is individually measured in a test oscillator designed by G3UUR (Fig 2) and its frequency is noted. For this we need at least an 8 digit frequency counter with a suitable time base, which is connected to the output of the oscillator. The circuit and assembly of the crystal



**Fig 2: Circuit diagram of the tests oscillator.**

oscillator are so simple that no further instructions are needed. Figs 3 – 5 show the PCB layout and completed test oscillator.

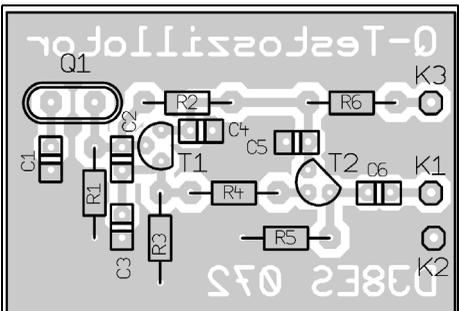
The frequency is measured for each crystal, once with the switch open and once with it closed, and the results entered in Table 1 of the associated Excel program that is available as a download from the VHF Communications web site [1]. The software uses these measurements to calculate the values for  $C_M$ ,  $L_M$  and  $C_0$  that are required later for the actual filter design. The value for  $Q_0$  has only a minor influence on the design and is treated as a constant.

This is not the place for a detailed discussion of the theory or the calculations needed to design a ladder filter. If you want to read about this in greater detail, you are recommended to consult the article in [2].

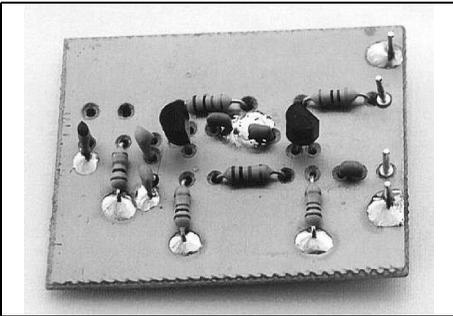
The frequencies of the crystals required for the ladder filter should be very close to one another. This tolerance amounts to maximum of 5 - 10% of the filter bandwidth required. That means, for a bandwidth of  $B = 400\text{Hz}$ , only 20 - 40Hz maximum divergence among them. For the 10 crystals measured, this only applies to the crystals Q3, Q4, Q8 and Q9. The remaining crystals are unsuitable for



**Fig 3: PCB layout for the crystal test oscillator.**



**Fig 4: Component layout for the crystal test oscillator.**



**Fig 5: Picture of the completed crystal test oscillator.**

this project.

### 3. Design and assembly of 4 pole ladder filter

**Step 1**, the measured and subsequently calculated filter parameters are transferred onto the filter design worksheet - i.e. complete lines are copied (see marking on Table 1).

**Step 2**, use the filter design worksheet (Table 2) to work out the nominal filter components. To do this, we need to enter the filter bandwidth required (e.g. 400Hz).

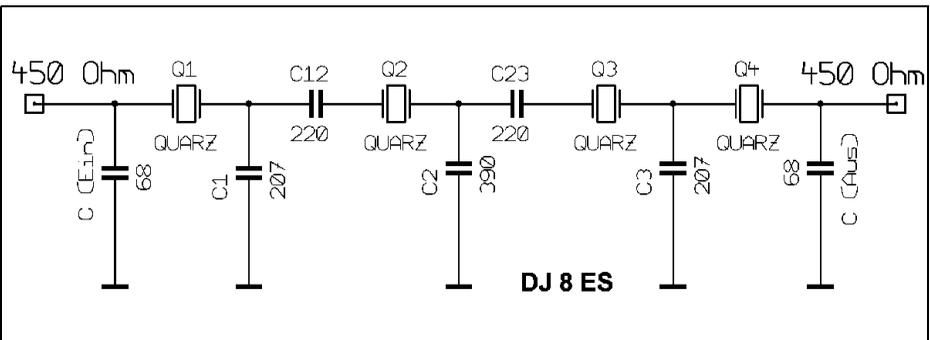
**Step 3**, conversion of these values to the desired input/output impedance of 450Ω.

All the component values required are calculated in this manner. These now have to be standardised to obtain values from the E12 or E24 range. Thus, for example, a calculated value of 206pF can be represented accurately enough using 180pF and 27pF in parallel. The circuit for the 4-pole ladder filters is shown in Fig 6.

### 4. Experiences in the design and assembly of a ladder filter

In each case, the ladder filter was connected using a simple 1:9 transformer at the input and output. This enables us to analyse the DIY filter using the available measuring equipment. The frequency response recorded (Fig 7) confirms the design. The centre frequency is specified as being 5.068040MHz, the 3dB bandwidth is 400Hz (6dB: 500Hz); the insertion attenuation (5.7dB) is also within the normal limits.

As the filter bandwidth increases, however, the design of such filters becomes more critical. In subsequent experiments,



**Fig 6: Circuit diagram of the 5.068Mz, 4 pole ladder filter.**



Microsoft Excel - Ladder-Filter-design-eng.xls

File Edit View Insert Format Tools Data Window Help

B2 = 5068000

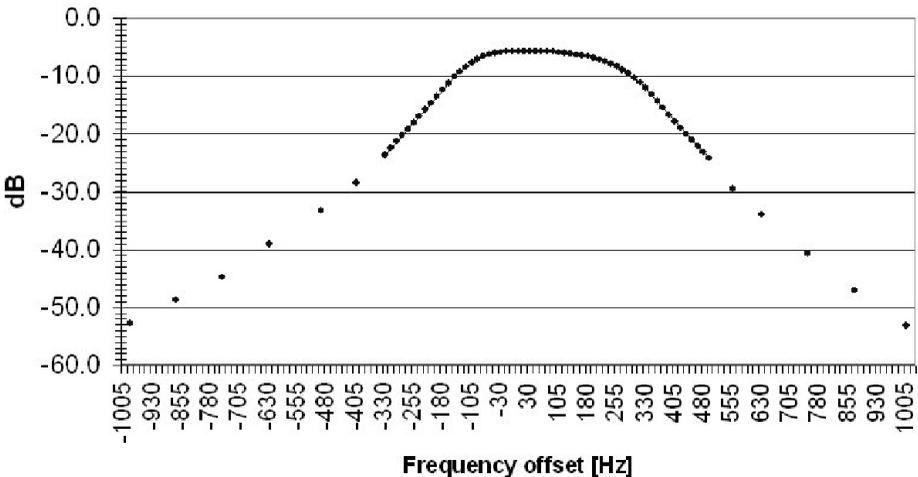
	A	B	C	D	E	F	G	H	I	J	K	L
	Crystal No.	Indicated frequency	Measured frequency with series 33pF C	Measured frequency with no series C	difference in Hz	CM [pF]	LM [n]	C0 [pF]	Q0			
1												
2	1	5,068,000	5,068,795	5,067,822	973	0.01267	0.07791	2,788	200,000			
3	2	5,068,000	5,068,905	5,067,794	1,011	0.01317	0.07498	2,897	200,000			
4	3	5,068,000	5,068,828	5,067,792	1,034	0.01347	0.07331	2,962	200,000			
5	4	5,068,000	5,068,835	5,067,577	1,158	0.01608	0.06646	3,318	200,000			
6	5	5,068,000	5,068,778	5,067,758	1,020	0.01328	0.07432	2,922	200,000			
7	6	5,068,000	5,068,816	5,067,840	976	0.01271	0.07767	2,796	200,000			
8	7	5,068,000	5,068,785	5,067,890	896	0.01167	0.08480	2,567	200,000			
9	8	5,068,000	5,068,840	5,067,810	1,030	0.01341	0.07360	2,951	200,000			
10	9	5,068,000	5,068,848	5,067,862	966	0.01284	0.07688	2,825	200,000			
11	10	5,068,000	5,068,800	5,067,843	957	0.01246	0.07921	2,742	200,000			
12												
13												
14	Data entry field											
15												
16												
17												
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start | ladder filter.doc - Mic... | Microsoft Excel - Lad... | 13:29

**Table 1: The spreadsheet for entering the measurements made on each crystal tested.**

**5.068MHz, 4 pole Ladder filter**



**Fig 7: Frequency response of the 5.068MHz ladder filter.**



Microsoft Excel - Ladder-Filter-design-eng.xls

File Edit View Insert Format Tools Data Window Help

C14      = 400

6	3	5,068,000	5,068,826	5,067,792	1,034	0,01347	0,07331	2,962	200,000	
7	4	5,068,000	5,069,836	5,067,877	1,159	0,01509	0,06646	3,318	200,000	
8	8	5,068,000	5,068,840	5,067,810	1,030	0,01341	0,07360	2,951	200,000	
9	9	5,068,000	5,068,848	5,067,862	986	0,01284	0,07688	2,825	200,000	
10			5,068,837			0,01370	0,07231			

12 2. Calculate the filter components

13

14 Required filter bandwidth [Hz]      400      Data entry field

16	R(in)	C1 [pF]	C2 [pF]	C3 [pF]	R(Ans)
17	237	206	385	206	237

18

19 3. Convert for required input/output impedance

20

21 input/output impedance [Ohm]      450      Data entry field

23	R(in)	C (in) [pF]	C1 [pF]	C12 [pF]	C2 [pF]	C23 [pF]	C3 [pF]	C (out) [pF]	R(out)
24	450	66	206	220	385	220	206	66	450

25

26 Help fields (not for calculation):

27	CE	w <sup>2</sup>	RO <sup>2</sup>	C'
28	6.60842E-11	1.0133E+15	202500	1.39931E-10 [F]
29				139.6306847 [pF]
31		CT	C Mesh	C High
32		219.6567444	134.3616407	83.36097135 [pF]

33

34

35

36

37

38

39

40

Crystal measurements / Crystal selection / Filter design / Frequency response /

Ready

Table 2: The spreadsheet for calculating the filter design.

we also assembled ladder filters with an SSB bandwidth of 2.4kHz. True, with this type of filter the edge steepness is not so good, by comparison with multi-pole SSB filters. However, the deciding factor in the design is the intended application!

The initially proposed number of 10 crystals for measurement is merely a reference value, for the quality of crystals varies a great deal. Various measurements have demonstrated that a batch of 10 crystals had only enough suitable crystals for two filters; on the other hand, 20 measured crystals might also yield only a meagre result, with just four good specimens, which would then be sufficient for only one filter.

## 5. Literature

[1] The Excel spreadsheet can be downloaded from the VHF Communications web site – [www.vhfcomm.co.uk](http://www.vhfcomm.co.uk)

[2] Hayward, Campbell and Larkin: Experimental Methods in RF Design ARRL



*Gunthard Kraus, DG8GB*

# Practical project: Stripline low pass filters for various frequency ranges

Part 3, continued from issue 3/2005

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## 4. (continued): Low pass filters for a cut-off frequency of 500MHz

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As so often, I have to thank one of my fellow radio enthusiasts and VHF readers for the information set out below. His offer to measure this filter with a modern network analyser (WILTRON) with a dynamic range of 100dB in the filter stop band was eagerly accepted. The old HP8410 is not up to this task - after all, it was never built for it. This made the tension all the greater while awaiting the measurement results, firstly this would be an opportunity to find out whether my own measurements were correct and secondly with such a wide measurement range we could get a better idea of where certain effects were coming from (e.g. any kind of collapse in attenuation). Fig 55 gives a very clear idea of the self resonances of the various SMD components (first the capacitors and then the coils). The first resonance clearly defines the slope of the attenuation increase from the pass band into the stop band. The second resonance, just before 4GHz, increases the attenuation, after it has risen again. Note: this is absolutely normal, since above their self resonance the SMD capacitors simply form inductances, thus the reactive impedance becomes higher

as the frequency rises. This is precisely what we don't want for the lower resistor in a voltage divider!

The behaviour of the circuit with the SMD capacitors and the stripline coils can be seen in Fig 56. As with any filter, the thin "coil lines" (or, if you prefer, the "line coils") sometimes no longer form coils, but behave like an RF line with rising frequency; they not only transform, but also change their behaviour as a component completely (you might say they behave like electrical chameleons). This results in the periodically recurring frequency points at which attenuation alternates with transmission.

The best model with regard to the pass band was the version with stripline capacitors and SMD coils shown in Fig 57. Luckily, this remains the same in the filter stop band, S21 never rises above -50dB all the way up to 4GHz.

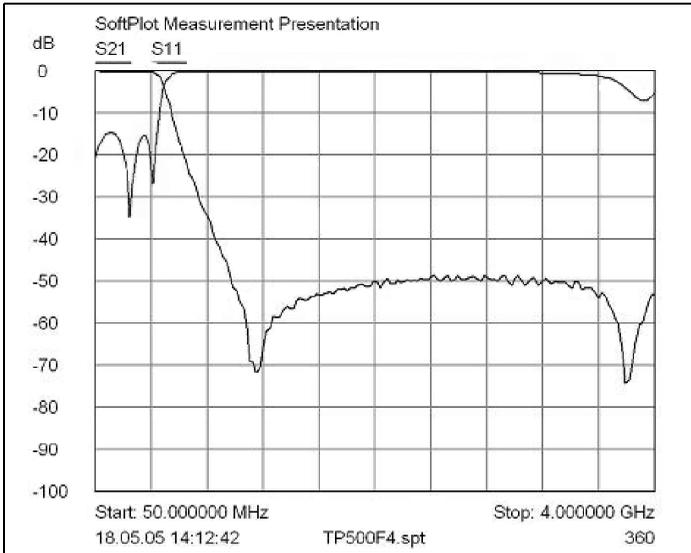
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## 5. Low pass filters for 1GHz cut- off frequency

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### 5.1. Overview

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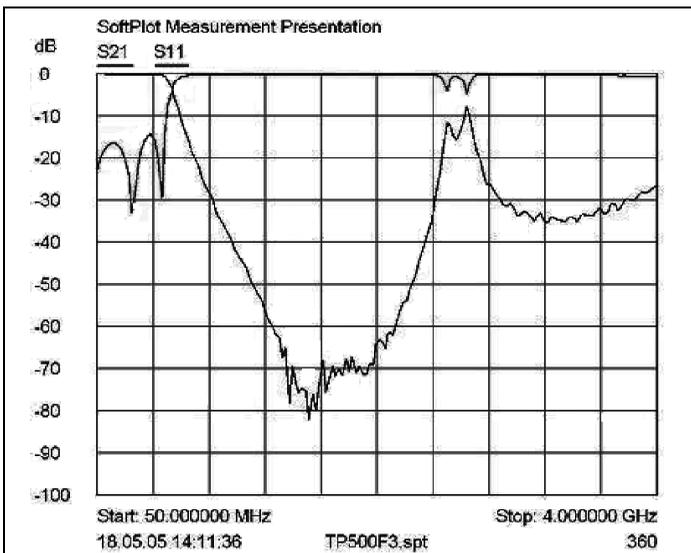


**Fig 55: Frequency response of the SMD version of the 500MHz low pass filter showing the effects of component resonances.**

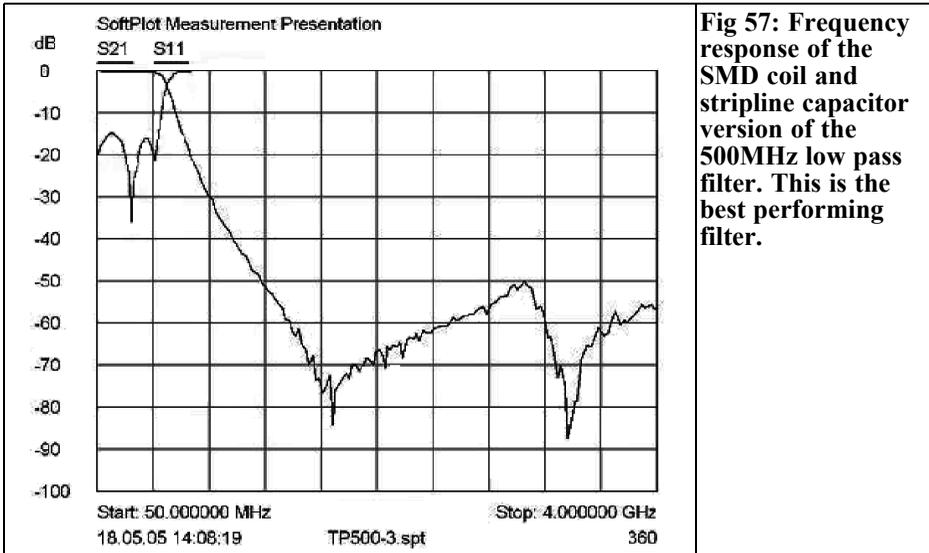
Here we come up against a limit where implementation using discrete components becomes more and more difficult. This is not because it would be insoluble, but because it becomes more and more difficult for the private individual to obtain SMD components and/or to fit them manually because of the values, the quality and also the dimensions involved.

To give a comparison: nowadays mobile telephones are using 0201 SMD resistors and capacitors. That means a size of 0.5mm x 0.25mm and cannot be fitted without using automatic mounting. For this reason, the following three technologies were selected for this project:

- Version 1 as a pure stripline model

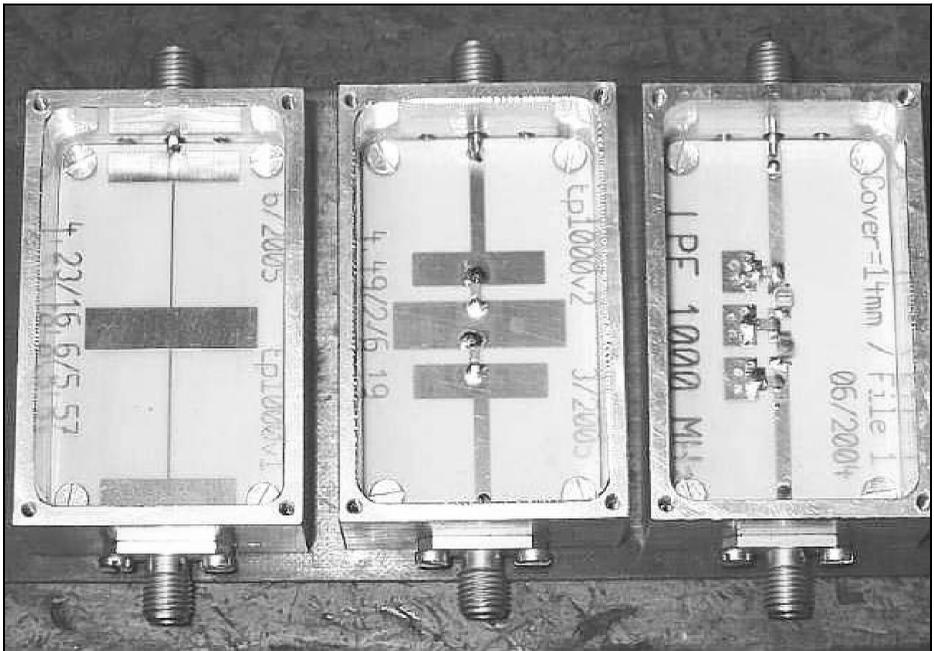


**Fig 56: Frequency response of the stripline coil version of the 500MHz low pass filter showing the behaviour of the "coils" as transmission lines as frequency increases..**

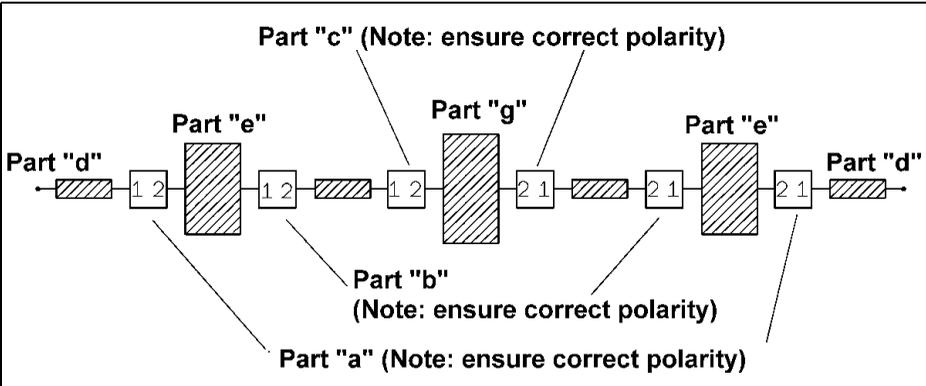


**Fig 57: Frequency response of the SMD coil and stripline capacitor version of the 500MHz low pass filter. This is the best performing filter.**

- Version 2 with stripline capacitors and SMD coils
  - And finally version 3 with SMD components only.
- All three fully assembled filters can be seen in Fig 58.



**Fig 58: Picture of the three types of low pass filter for 1GHz.**



**Fig 59: The start point for the PUFF simulation of the 1GHz low pass filter, type 1.**

- Part "a":** Twoport "s1p83-17.s2p" (Step from 1.83mm to 17mm)
- Part "b":** Twoport "s17-0p25.s2p" (Step from 17mm to 0.25mm)
- Part "c":** Twoport "s0p25-22.s2p" (Step from 0.25mm to 22mm)
- Part "d":** Microstrip 50Ω / 2.56mm long, 1.83mm wide
- Part "e":** Microstrip 8.74Ω / 5.33mm long, 17mm wide
- Part "f":** Microstrip 120Ω / 16.23mm long, 0.25mm wide
- Part "g":** Microstrip 6.91Ω / 7.13mm long, 22mm wide

**5.2. Low pass filter for 1GHz with n = 5; Version 1 as stripline model**

**5.2.1. Design procedure and first printed circuit board**

First we use the "fds" filter program to specify the necessary component values:

$$C1 = C3 = 3.658pF$$

$$L1 = L2 = 10.912nH$$

$$C2 = 6.287pF$$

The stripline widths can be taken from the corresponding 500MHz low pass filter, where we find:

- For C1 and C3: Conductor width 17mm
- For L1 and L2: Conductor width 0.25mm
- For C2: Conductor width 22mm

To determine the line lengths for the capacitors, naturally we use PUFF soft-

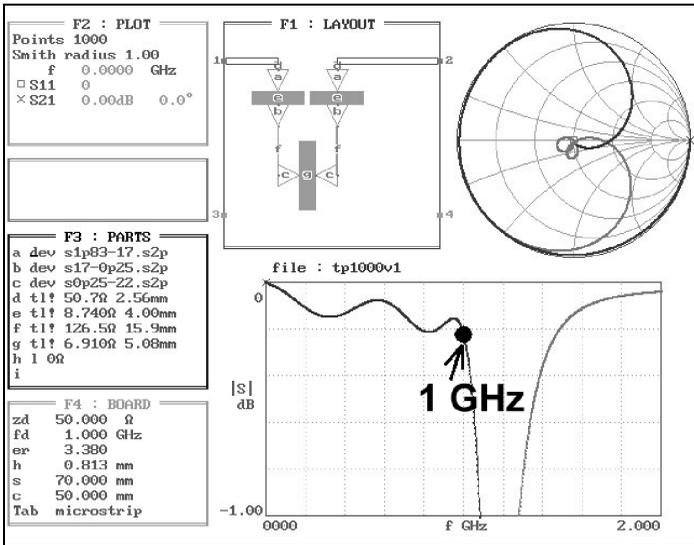
ware for the capacitors as per Fig 46 and Fig 47 in Part 2, simulating open circuit lines at 1GHz. The line length for the coil is a short circuited stripline and has already been determined in Fig 41.

This gives us the following specifications:

- Striplines for C1 and C3: Width 17mm, length 5.33mm
- Striplines for L1 and L2: Width 0.25mm, length 16.23mm
- Stripline for C2: Width 22mm, length 7.13mm

The PUFF simulation circuit diagram can be seen in Fig 59 that look very much like the circuit produced by expensive programs.

The steps between the individual line sections are now indispensable, since they not only describe the transition to the thin line using a "false inductance",



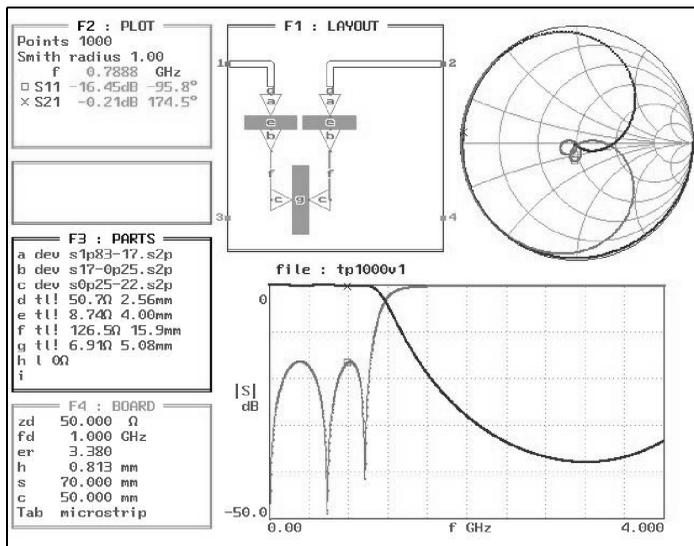
**Fig 60: Simulation of the pass band of the 1GHz low pass filter, type 1, after optimisation.**

but also the “open end fringing” of the wider line section using a “false capacitance”. The student version of APLAC was used to generate the S2P files for the steps, then they were copied into the PUFF directory. This has also been described several times already in Part 1 and Part 2 of this series of articles.

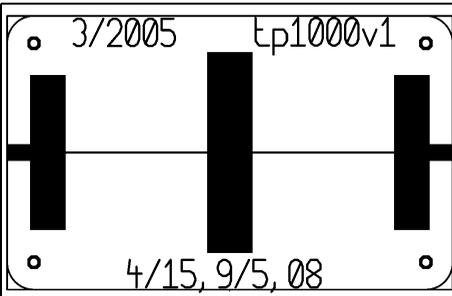
following optimisation, no further length correction is needed for the line sections. The changes that were always needed in the past, shown in the “open end correction diagram” from the PUFF manual, can be omitted from now on!

Thus, by using PUFF, we have taken a giant step forward: With this method,

Thanks to the inclusion of irregularities, the results are now a long way from the specifications, and a comprehensive cor-



**Fig 61: Simulation of the 1GHz low pass filter, type 1, after optimisation showing the response up to 4GHz.**



**Fig 62: PCB layout for the 1GHz low pass filter type 1.**

rection is now unavoidable.

The simulation of the pass band, following successful “manual” optimisation, can be seen in Fig 60. All information concerning behaviour up to 4GHz can be found in Fig 61. A comparison of components list F3 with the simulation specifications of Fig 59 clearly shows the influence of the irregularities and the shortening of the line lengths required. The values involved are as follows:

- L1 goes down from 5.33mm to 4mm
- L2 goes down from 16.23mm to 15.9mm
- L3 goes down from 7.13mm to 5.08mm

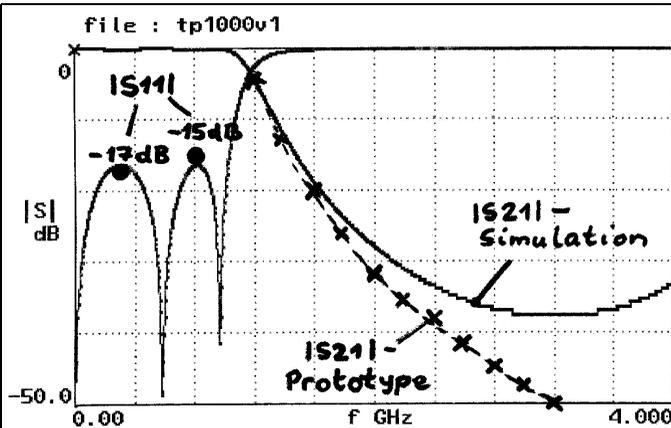
The layout is shown in Fig 62, it can clearly be seen that this time things might be very tight with the housing size, the 50Ω striplines for connecting the SMA

connector are still precisely 2.5mm long.

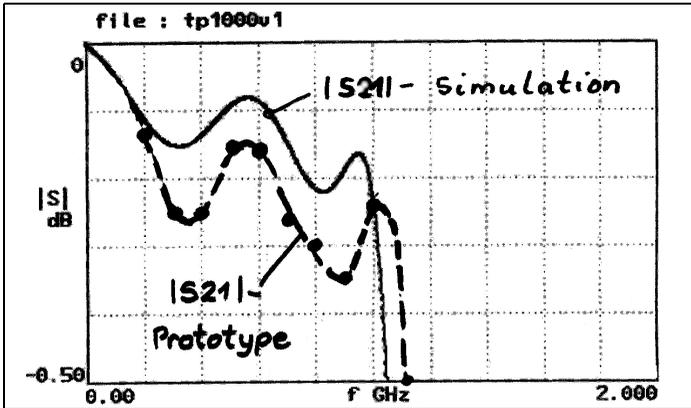
One might get rather tense about the results of the measurements from the prototype. The S11 curve can be seen in Fig 63 and we can see that this has still left us with some work to do, because the two humps are still of different heights, with approximately -17dB and -15dB (the target value for both of them was -16.4dB). It's true that this caused the attenuation to rise in the stop band, but at the cost of a larger ripple in the pass band. If the pass band is extended sufficiently for the S21 curve, we clearly see, in Fig 64, both this ripple and the attenuation, which increases as the frequency rises. We can be satisfied with this overall, but there is still one problem. The cut-off frequency is not at 1,000MHz but is somewhat too high (somewhere around 1,075MHz) and that should be corrected.

**5.2.2. Completion of development**

The correction method below should be demonstrated well on this filter. It has already proved it's worth in various projects, particularly in relation to stripline designs. The variations between the specifications and the second completed circuit board are therefore almost always under 1%! Fanatical DIYers now add another round of corrections, to remove even the last small differences.



**Fig 63: Measured response of the 1GHz low pass filter, type 1, showing that more work is required.**



**Fig 64:** Measured pass band response of the 1GHz low pass filter, type 1.

**The procedure:**

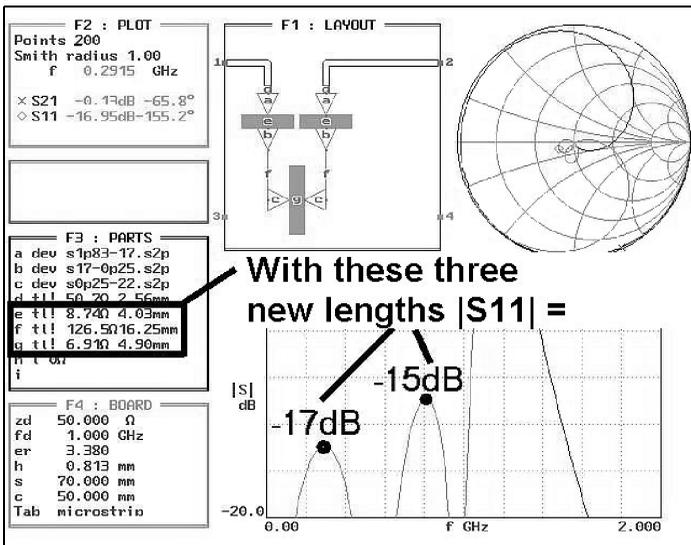
- 1 Keep changing the PUFF simulation circuit until it precisely matches the S-parameters measured on the prototype. Make a note of the component data obtained in this way.
- 2 Now take the data from the first simulation (used to create the layout) and use a pocket calculator to determine, for each component used, the difference between the initial simulation and the measured value simulation.

- 3 Now correct the layout data for the first prototype by these differences. Then set up a new circuit board and hopefully this will have eliminated the discrepancies.

**The individual steps:**

**1<sup>st</sup> step:**

Load the PUFF file (Fig 61) and start varying the lengths of each of the three striplines that form the actual low pass filter in turn. The objective is to obtain a left hand S11 hump at -17dB and a right hand hump at -15dB. For an example of the result of this procedure, see Fig 65.



**Fig 65:** Result of adjusting the line lengths to improve the S11 response.

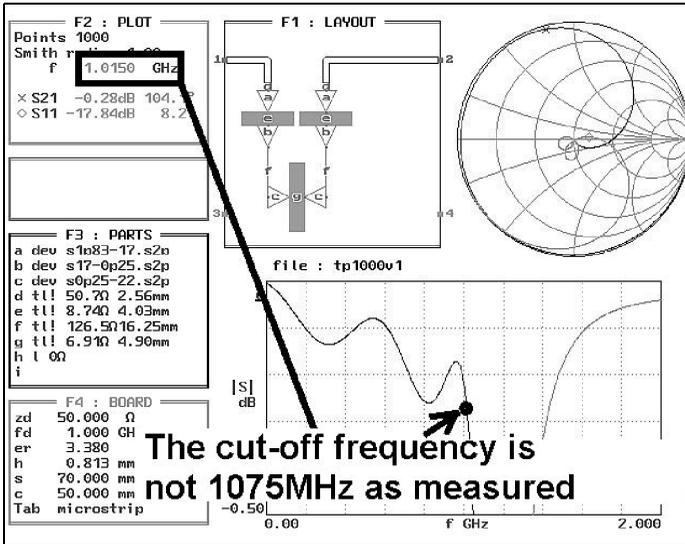


Fig 66: The cut-off frequency is still not correct.

Once you have ticked this point off successfully, you need a pocket calculator because it can be seen in Fig 66 that the cut-off frequency is still not at the measured value of 1075MHz, but only at approximately 1015MHz. So we shorten all three line lengths by the factor  $1015 / 1075 = 0.944186$  and repeat the simulation. This demonstrates that further corrections are still needed: the step irregu-

larities built into the simulation do NOT change their data if the frequency changes, and this makes for small discrepancies. These are no longer all that significant, and thus we eventually arrive at Fig 67 and can go back to the amended line data.

**2<sup>nd</sup> step:**

Now we come to the final evaluation.

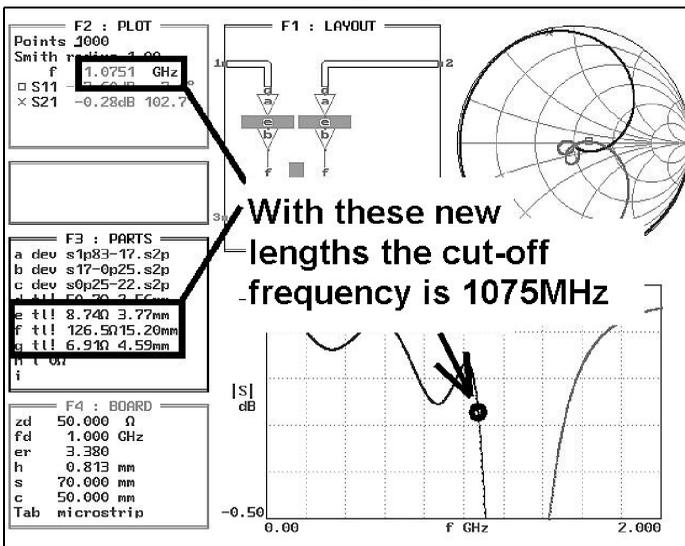
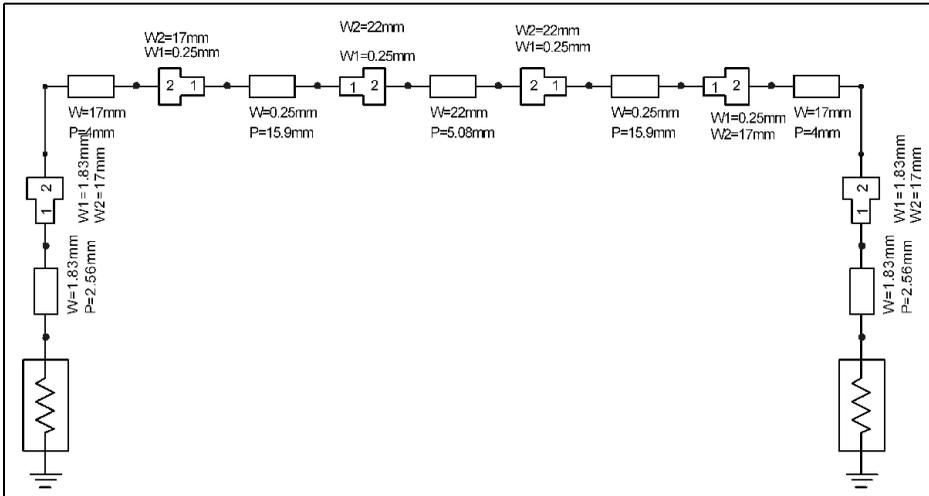


Fig 67: The final simulation showing that the cut-off frequency is as required.



**Fig 68: The input screen used for a simulation of the 1GHz low pass filter, type1, with ANSOFT Designer SV.**

The line length L1 = 4mm has been replaced by 3.77mm; the alteration is 0.23mm

The line length L2 = 15.9mm has been replaced by 15.2mm; the alteration is thus 0.7mm

The line length L3 = 5.08mm has been replaced by 4.59mm; the alteration is thus 0.49mm

**3<sup>rd</sup> step:**

Since the measurement has produced a cut-off frequency that is too high, the line must be lengthened by the differences just determined to give an overall layout correction. Thus we obtain the new line lengths for the next layout:

L1 is lengthened from 4mm to (4mm + 0.23mm) = 4.23mm

L2 is lengthened from 15.9mm to (15.9mm + 0.7mm) = 16.6mm

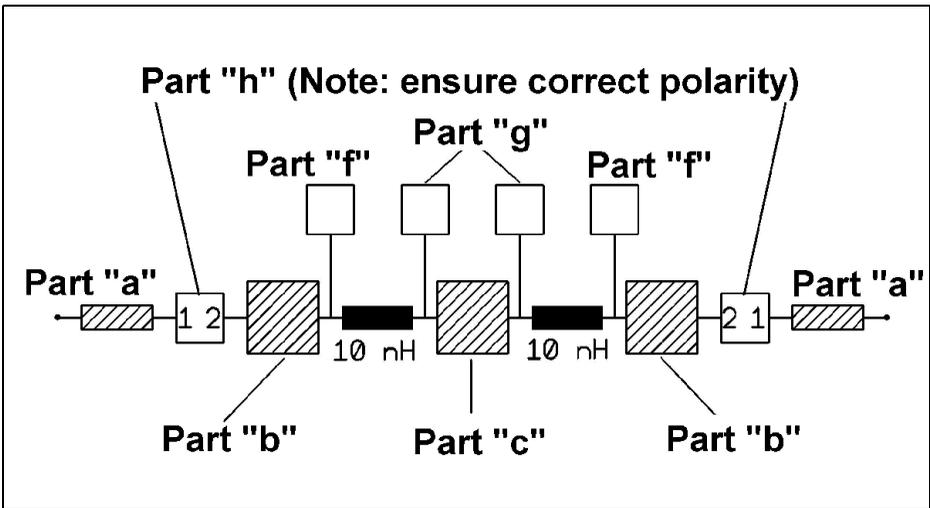
L3 is lengthened from 5.08mm to (5.08mm + 0.49mm) = 5.57mm - and that's it!

the result with the circuit board length, then you'll still have just 1.4mm. left over on each side to connect the SMA sockets.

Now you have to consider whether it wouldn't be better to "fold" the thin "coil lines" into meandering shapes (see the corresponding 500MHz version from Part 2). Apart from the fact that NO solder must flow into the filter structure during soldering (since this makes a noticeable difference to the measured values!), the external, 17mm wide strip-lines are already very near the edge of the circuit board and/or the housing wall, which leads to further unexpected effects. Only big and expensive simulation machines can provide any additional help, but there is one piece of good news: the free student version of the "ANSOFT Designer" is guaranteed to provide models for everything, and this naturally includes striplines running close to the circuit board edge!

Yet you must never lose sight of reality - which here means the circuit board size and/or the housing dimensions! If you just add up the new lengths and compare

In order to keep track of the expenditure on, and the efficiency of, the software from ANSOFT (and out of curiosity as to how precise the PUFF method is with subsequent length correction), a repeat



**Fig 69: The start point for the PUFF simulation of the 1GHz low pass filter, type 2.**

**Part "a":** Microstrip  $50\Omega$  / 15mm long, 1.83mm wide

**Part "b":** Microstrip  $8.74\Omega$  / 5.33mm long, 17mm wide

**Part "c":** Microstrip  $6.9\Omega$  / 7.13mm long, 22mm wide

**Part "f":** Oneport "17mm.sip" Open end extension of  $8.74\Omega$  line

**Part "g":** Oneport "22mm.sip" Open end extension of  $6.94\Omega$  line

**Part "h":** Twoport "s50-8p7.s2p" step from  $50\Omega$  to  $8.74\Omega$

simulation as an ANSOFT project was carried out to design, build a prototype, measure the results and subsequent optimisation using the three point method above. The ANSOFT entry screen required for this can be seen in Fig 68; even as a PUFF user, you'll be at home here straight away. But the operating procedure is a bit of a challenge.

The result of this procedure and the ANSOFT proposals on the new layout are naturally of great interest. They are as follows:

- L1 = 4.25mm
- L2 = 16.67mm
- L3 = 5.59mm

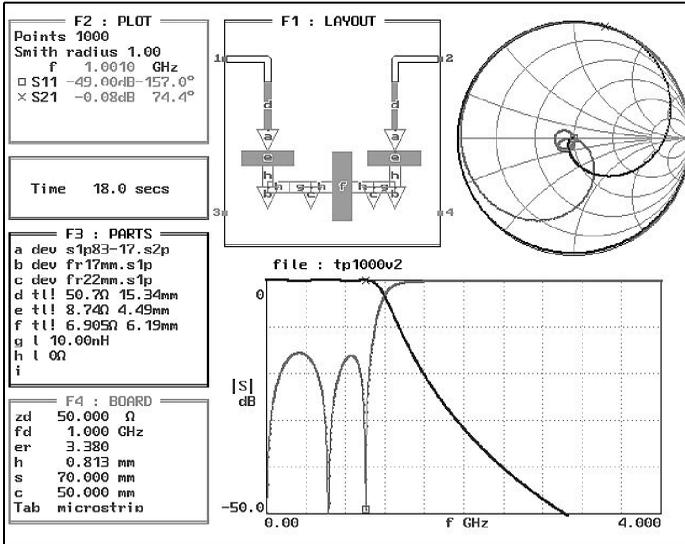
In comparison with the new values previously obtained from PUFF, we can only say: all the best!

### 5.3. Low pass filter for 1GHz with $n = 5$ ; Version 2 with SMD inductances and stripline capacitances

#### 5.3.1. Layout development

As with the 500MHz low pass filter in the previous section, we now attempt to get by again using commercially available SMD coils from the standard range; value of 10.9nH is required. So a value of 10nH is proposed, and we can use this to re-create the simulation screen (Fig 69).

The stripline data for the "capacitors" can simply be taken from version 1, which we have just looked at. A subsequent optimisation is essential (with an appropriate compromise due to the fact that the inductance value is not exact). Instead of making use of the steps between the individually linked together line sections, we have to work with "open extensions"



**Fig 70: Optimised simulation of the frequency response for the 1GHz low pass filter, type 2.**

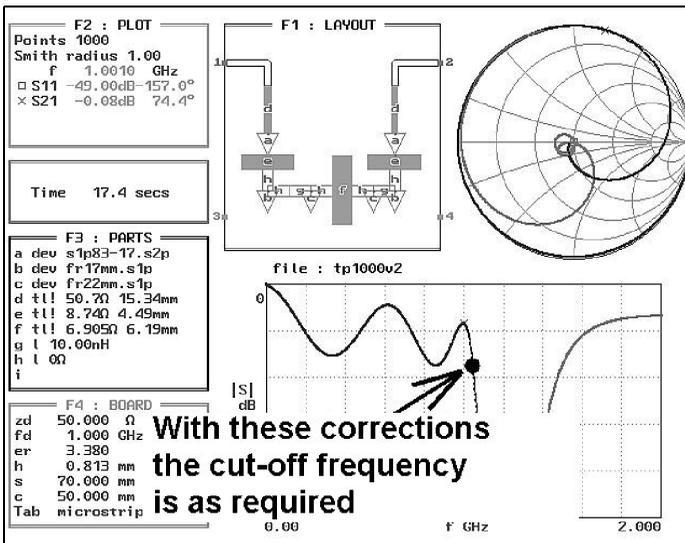
again (see 500MHz low pass filter, version 2), which are taken into the F3 component list as APLAC S1P files.

measurements will supply regarding the losses actually present and the attenuation this causes.

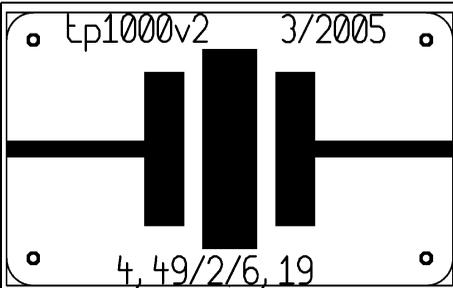
The optimisation and simulation results for loss-free coils can be seen in Fig 70 and Fig 71. Fig 72 shows that the layout becomes more and more delicate as the frequencies rise. It is certainly of interest to see what information the prototype

**5.3.2. Measurement results**

We can see from Fig 73 that the previous simulations were definitely not all that far from reality. The simulation curve applies to an assumed loss resistance of



**Fig 71: Optimised simulation for the pass band of the 1GHz low pass filter, type 2.**



**Fig 72: PCB layout for the 1GHz low pass filter type 2.**

1Ω (i.e. a coil Q of approximately 60 at 1GHz) and the measurement values are better below 700MHz, whereas above this value they are slightly worse than the simulation. This is plainly and simply because in the CAD program we have to work with a loss resistance which remains constant throughout the frequency range. In reality, the coil Q is constant over wide ranges and consequently the loss resistance falls as the frequency drops - which the measurements confirm.

One small cosmetic fault is also present, in the shape of the measured cut-off frequency, which is slightly too low, whereas the maximum S11 values in the pass band are fairly precise at -16dB (instead of -16.4dB) and so no more correction is necessary.

Since the S21 curve in the stop band does not differ essentially from the simulation as per Fig 70 so there is no need to describe it. Since all these discrepancies can be balanced out, we have shown just version 1 of the filter (and once again the old rule applies that the first circuit board version can never be fully satisfactory in all ranges).

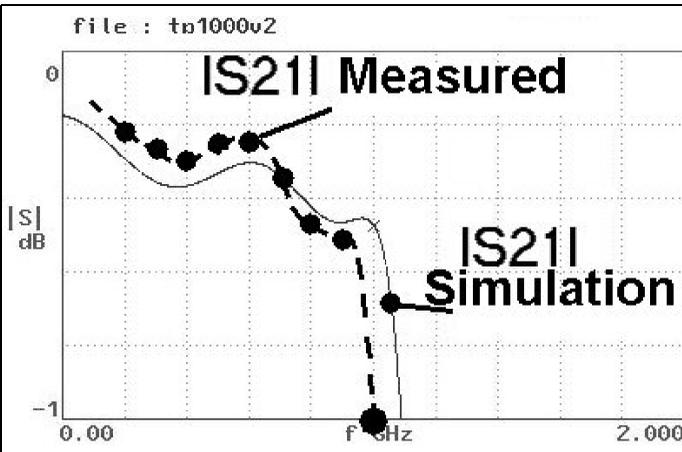
**5.4. Low pass filter for 1GHz with n = 5; Version 3 with SMD components**

**5.4.1. Layout development**

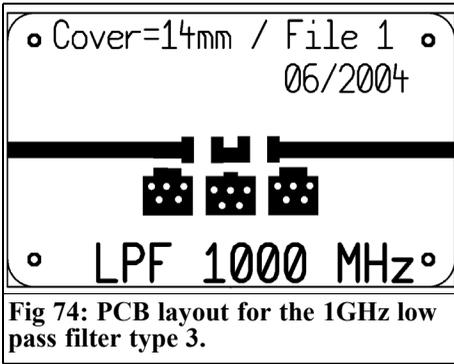
For what is probably the last time (because the cut-off frequency keeps going higher and higher), let's try and assemble the low pass filter using discrete components only. The advantages hoped for from this at 1000MHz are exclusively in the filter stop band, unless any housing resonances are added, the circuit would have to work as desired and block neatly up to about 4GHz. In the pass band, using components from standard ranges causes problems in maintaining the ripple values or the cut off frequency.

But initially we need the component values from Section 5.2.1. as our starting point:

$$C1 = C3 = 3.658\text{pF}$$



**Fig 73: Measured response for the 1GHz low pas filter type 2.**



**Fig 74: PCB layout for the 1GHz low pass filter type 3.**

$L1 = L2 = 10.912\text{nH}$   
 $C2 = 6.287\text{pF}$

A glance at the circuit board (Fig 74) permits us to suspect that things won't remain like that, for both the SMD pads and the connection lines between the components bring additional capacitances into play which cannot be ignored. So we reach for the standard simulation screen (from Part 2) with the appropriate SMD version of the 500MHz low pass filter and adapt it for the new circumstances.

The correct initial parameters for the PUFF simulation is provided by Fig 75 and after only a little optimisation work we obtain a thoroughly suitable result (Fig 76 and Fig 77). Here we were

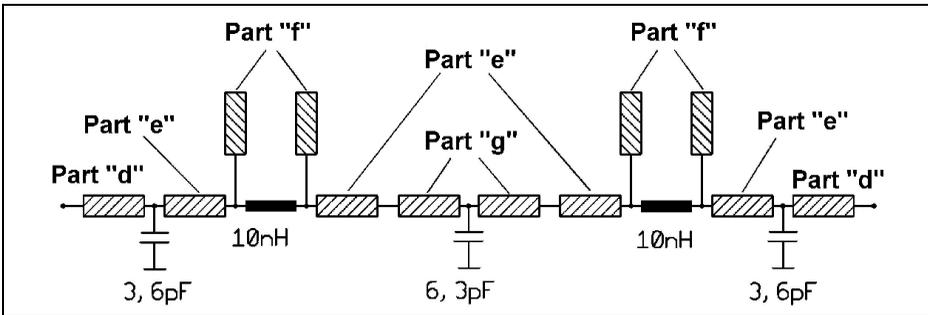
working with a series loss resistance of just  $1\Omega$ , in order to take account of the coil Q of 60 at 1,000MHz.

Things are going very well with components in accordance with the standard range (3.9pF and 6.2pF for capacitors, 10nH for coils). The only cosmetic problems are, firstly, a cut-off frequency that is somewhat too low (which is giving us an attenuation of 1dB at 1,000MHz) and secondly a somewhat poorer reflection (left hand hump = -14dB, right hand hump = -15dB, instead of both S21's = -16.4dB. The rather higher ripple in the transmission range due to this reflection was accepted.

**5.4.2. Measurement results**

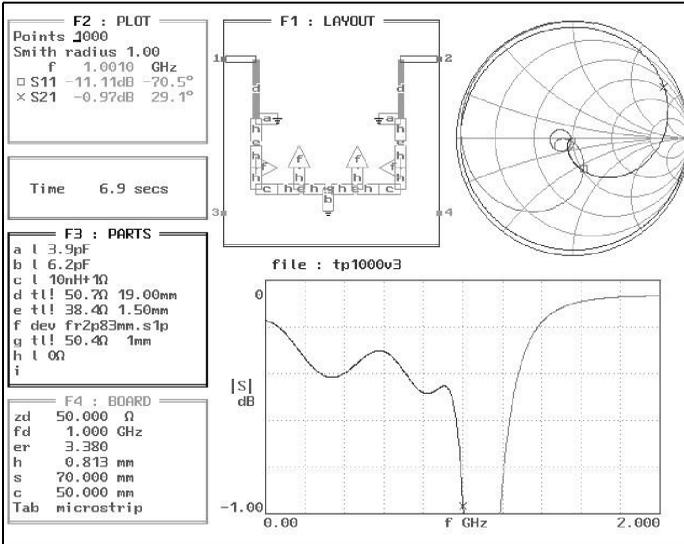
A lot of information can be obtained from Fig 78 and careful analysis alone can supply additional important predictions:

The measured ripple is lower than the value in the simulation - so we should expect a lower reflection value in the pass band. The cut-off frequency is likewise lower than expected; as the users we must decide whether this is tolerable. On the other hand, the abrupt steepness of the transition from the pass band into the



**Fig 75: Start point for PUFF simulation of the 1GHz low pass filter type 3.**

- Part "d":** Microstrip  $50\Omega$  / 19mm long, 1.83mm wide
- Part "e":** SMD-pad Microstrip  $38.4\Omega$  / 1.5mm long, 2.83mm wide
- Part "f":** Open end extension for SMD-pad "fr-2p83.s1p"
- Part "d":** Microstrip  $50\Omega$  / 1mm long, 1.83mm wide

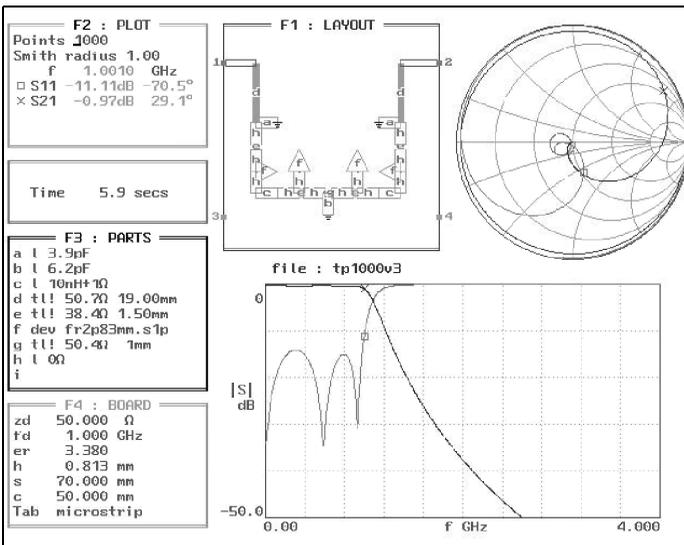


**Fig 76: Simulation of the frequency response for the 1GHz low pass filter type 3.**

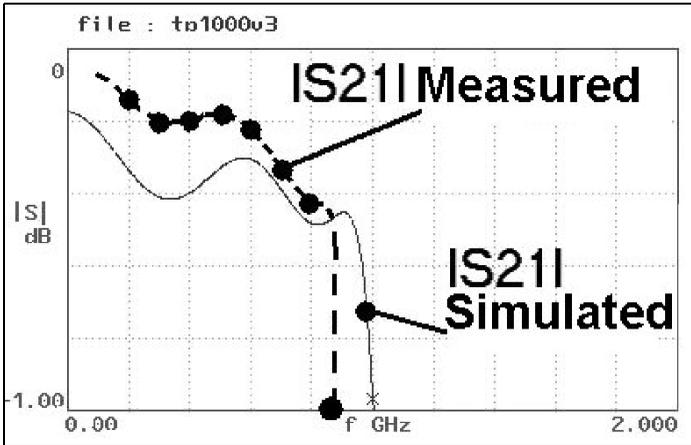
stop band is very suspicious, since this does not correspond to the simulation at all.

Fig 79 confirms our observations, this also explains why there is a steep rise in attenuation; the self resonances of the capacitors (in each case, there was only one example soldered in of 3.9pF or 6.2pF) accounts for this surprise! Above

the self resonance the effect of the capacitors starts to make its appearance: they act as inductances, their reactive impedance increases with the frequency, and so S21 is above -50dB at 3GHz. The only way to remedy this is with considerably smaller SMD versions, or with a parallel circuit containing several capacitors to increase the self resonance.



**Fig 77: Simulation of the frequency response for the 1GHz low pass filter type 3.**



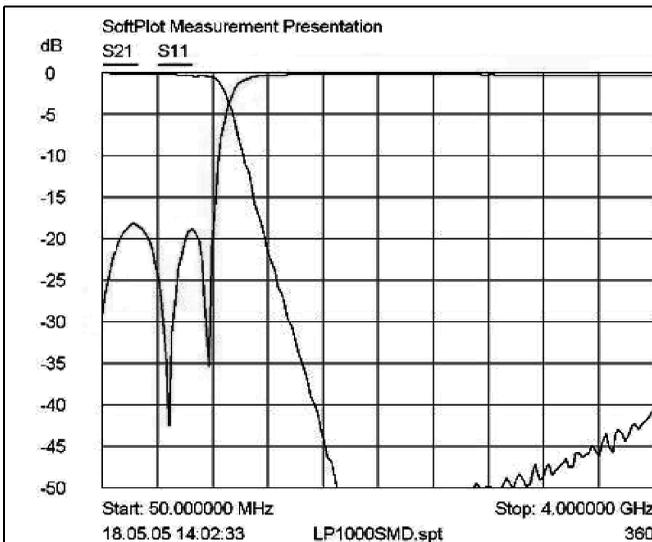
**Fig 78: Measured frequency response for the 1GHz low pass filter type 3.**

## 6. Stripline low pass filters for the range between 1 and 2GHz

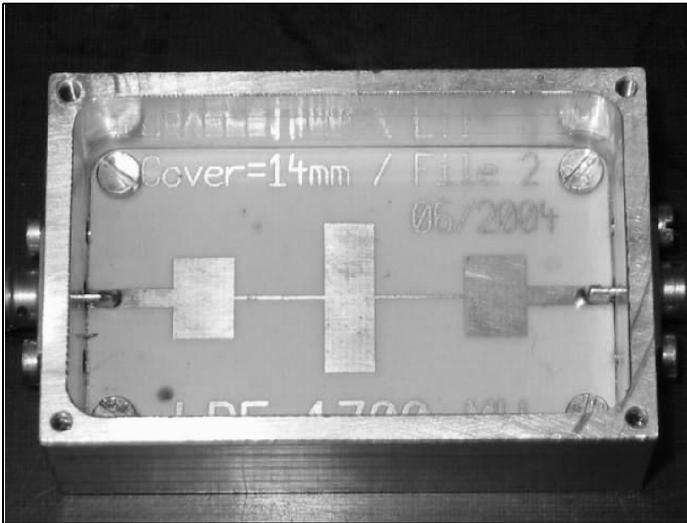
### 6.1. Initial comments

From now on, development is restricted to pure stripline versions, since they are the easiest to create and can be copied most easily. But before making a start let's define a few points:

- The widths of the line sections used as capacitors and coils should be kept the same between 1 and 2GHz. The conversion to a different cut-off frequency is made by simply changing the line lengths.
- The width of the line sections acting as capacitors must be reduced from those used for the 1GHz version. As the frequency increases everything gets shorter so the ratio of width to length becomes unfavourable for the capacitors. The consequence of this is that the simulation becomes more



**Fig 79: The frequency response of the 1GHz low pass filter, type 3, showing the effects of self resonance above 3GHz.**

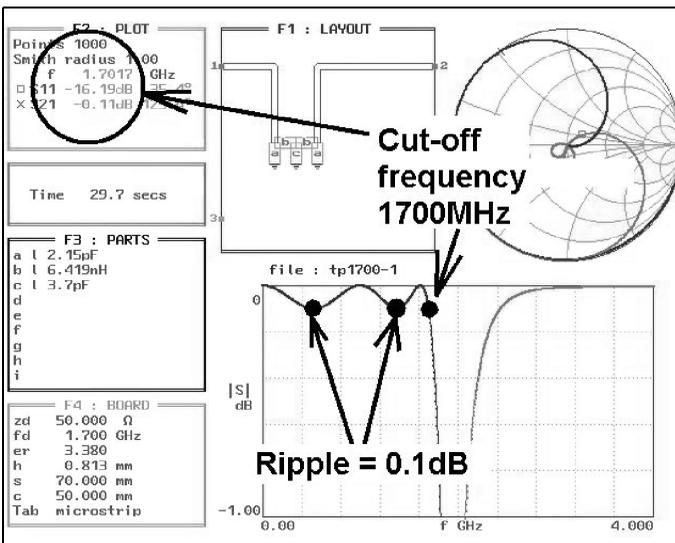


**Fig 80: Picture of the 1700MHz low pass filter.**

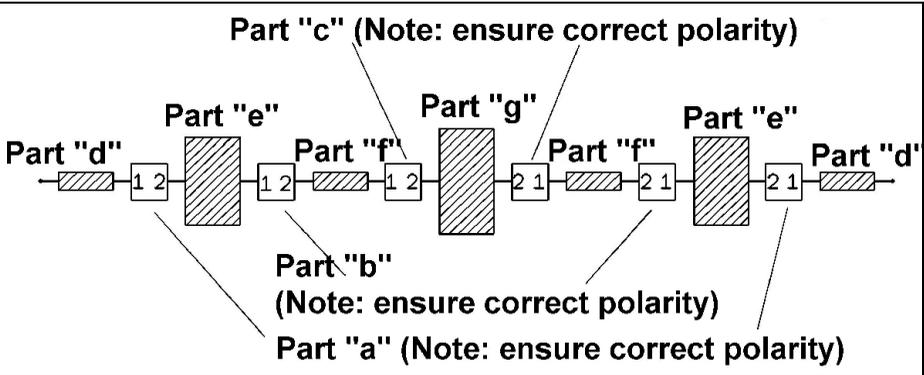
and more imprecise for line models and we really have to reach for EM simulators (such as SONNET, mstrip40, etc.). Experience tells us that this upper limit lies at a ratio of approximately 5:1. Above this value the line model simulation with an S-parameter program slowly but more and more clearly deviate from the measurement results on the prototype.

For this reason and after a few test simulations, the following values were selected:

The “external capacitors” of the circuit consist of stripline sections with a characteristic impedance of 17Ω. With the selected R04003 material and a circuit board thickness of 0.813mm, this gives a conductor width of 7.93mm.



**Fig 81: Simulation of the 1700MHz low pass filter.**



**Fig 82: Start point of PUFF simulation for 1 - 2GHz low pass filter.**

- Part "a":** Twoport "st50-17.s2p" step from 1.8mm to 7.93mm
- Part "b":** Twoport "st17-120.s2p" step from 7.93mm to 0.25mm
- Part "c":** Twoport "st120-10.s2p" step from 0.25mm to 14.53mm
- Part "d":** Microstrip 50Ω / 8.35mm long, 1.83mm wide
- Part "e":** Microstrip 17Ω / 6.06mm long, 7.93mm wide
- Part "f":** Microstrip 120Ω / 9.35mm long, 0.25mm wide
- Part "g":** Microstrip 10Ω / 6.03mm long, 14.53mm wide

A line section with a characteristic impedance of  $Z = 10\Omega$

acts as the "central capacitor". It thus has a width of 14.53mm. The value of 0.25mm, selected as the minimum on the basis of the manufacturing options, is retained as the minimum width for the "coils".

A 1300MHz low pass filter was assembled for the 23 cm band, together with a 1700MHz low pass filter for GPS and Meteosat applications. Since the design procedures remain the same, let's look at the 1700MHz version below.

**6.2. Stripline low pass filter with a cut-off frequency of 1,700MHz**

**6.2.1. Design with PUFF**

The completed prototype built into its standard aluminium housing can be seen in Fig 80.

To conclude this series of articles, let's

briefly run through the individual design stages again:

**1<sup>st</sup> stage: Determining component values**

Either we use the filter program "fds", or we simply go back to the initial values for the 1000MHz low pass filter from the previous section. If we multiply by the factor

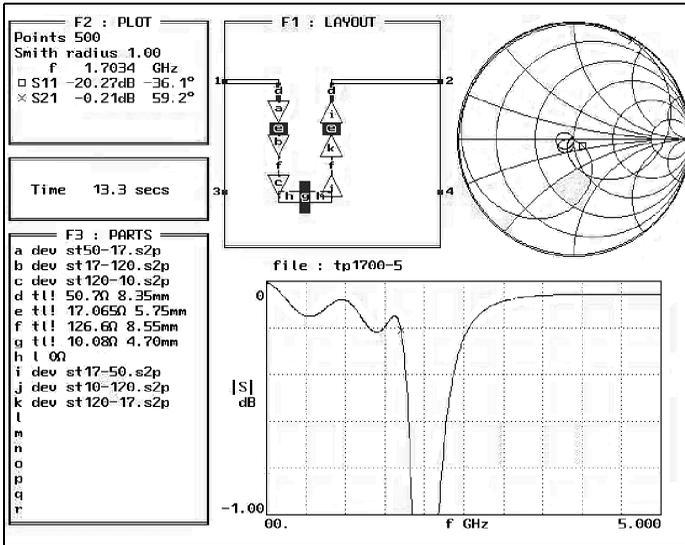
$$1000\text{MHz} / 1700\text{MHz} = 0.5882352$$

we obtain the following result:

$$\begin{aligned} C1 = C3 &= 2.15\text{pF} \\ L1 = L2 &= 6.415\text{nH} \\ C3 &= 3.7\text{pF} \end{aligned}$$

**2<sup>nd</sup> stage: Check simulation of ideal filter**

To do this, we quickly enter the filter circuit into the set-up file for PUFF, and the S21 range from 0 to -0.5dB for the frequency range from 0 to 5GHz. We can very easily tell whether both the ripple



**Fig 83: Final simulation of the pass band for the 1700MHz low pass filter.**

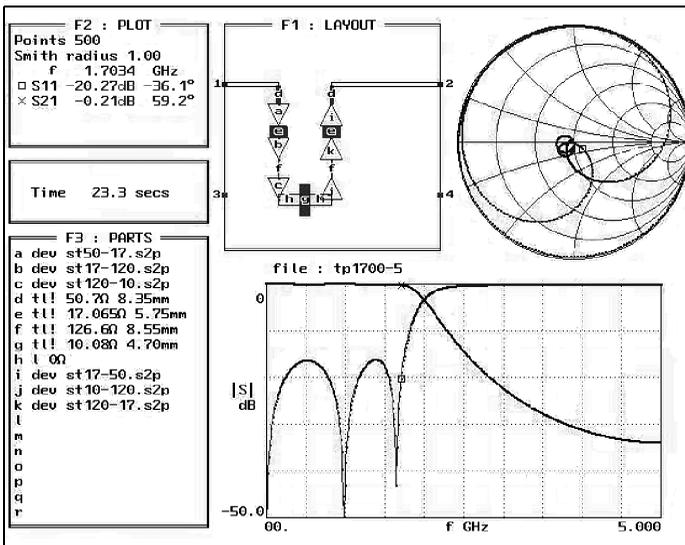
(0.1dB) and the cut-off frequency (1,700MHz) are correct; Fig 81 confirms that.

**3<sup>rd</sup> stage: Determining “capacitor line lengths”**

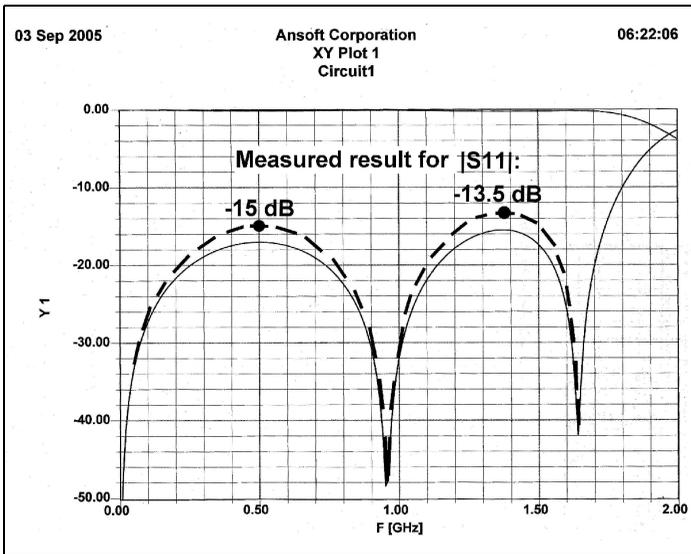
Striplines are now used for the two external capacitors, with  $Z = 17\Omega$  and a width of 7.93mm. Using PUFF, we simu-

late an open circuit line and then increase the capacitance (by pressing the "Equals" key, when the cursor is on "S11"). We rapidly learn from this that a capacitance of 2.15pF is associated with a line section with a length of  $L1 = 6.06\text{mm}$ .

We do the same thing for the centre capacitor. Here a line with  $Z = 10\Omega$ , a width of 14.53mm and a length of  $L3 =$



**Fig 84: Final simulation of the frequency response for the 1700MHz low pass filter up to 5GHz.**



**Fig 85: Simulation results for the 1700MHz low pass filter using ANSOFT Designer SV versus measured results.**

6.03mm is selected, which gives precisely the 3.7pF required.

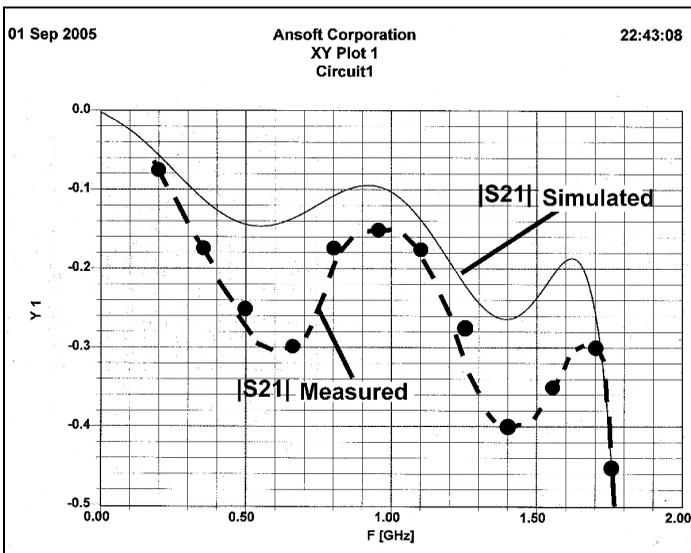
**4<sup>th</sup> stage: Determining of “coil line lengths”**

Instead of an open circuit stripline, we now simulate a short circuited line, that will behave like an inductance when line lengths are short. With a characteristic

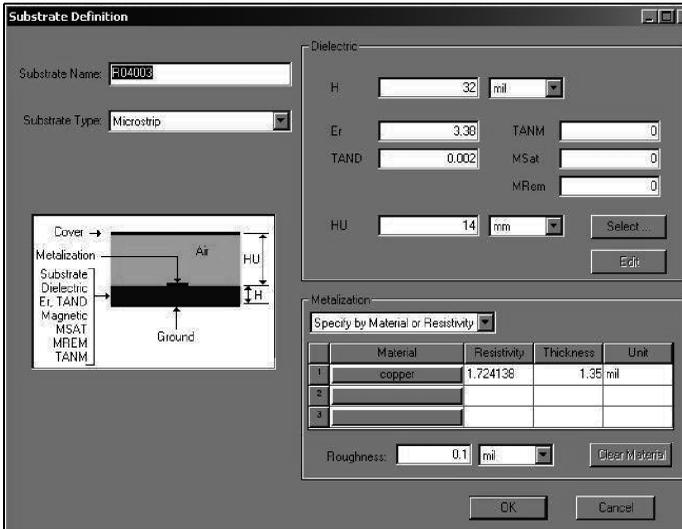
impedance of  $120\Omega$  (corresponding to a conductor width of 0.25mm), for the 6.415nH required, we need a length of precisely  $L_2 = 9.53\text{mm}$ .

**5<sup>th</sup> stage: Simulation screen**

As for earlier projects, we should first sketch the simulation screen to be entered into PUFF. This initial position can be



**Fig 86: Simulation results of the pass band for the 1700MHz low pass filter using ANSOFT Designer SV.**



**Fig 87: Circuit board and housing data entered into ANSOFT Designer SV.**

seen in Fig 82.

### **6<sup>th</sup> stage: the “Steps”**

Each change in the width of the striplines creates irregularities, which are considered as “steps” in the simulation. In this case, we reach for APLAC and there we generate, by brief simulations (each using line sections 1 micrometer long), the S-parameter files required for the three width changes required: from the 50Ω feed to the first “capacitor”, from the “first capacitor” to the “coil” and finally from the “coil” to the “centre capacitor”. These three files are copied into the PUFF work directory.

### **7<sup>th</sup> stage: PUFF simulation and optimisation**

We now key the simulation screen from Fig 5 into PUFF, paying particular attention to the correct “polarity” of the “steps” in the right hand half of the filter (reversed direction of assembly!). We can then optimise some of the “steps” introduced, reducing the line section lengths considerably. Fig 83 shows the final condition being aimed for in the pass band. But Fig 84 demonstrates that

the reflection gives no grounds for criticism either.

### **8<sup>th</sup> stage: Printed circuit board layout and prototype production**

This just requires some work - there are no problems. Looking at Fig 80 at the start of this section provides enough information. The only important preparation is to provide a correct listing of the required line sections for the work in the CAD program:

Feeders: Width 1.83mm, Length 8.35mm

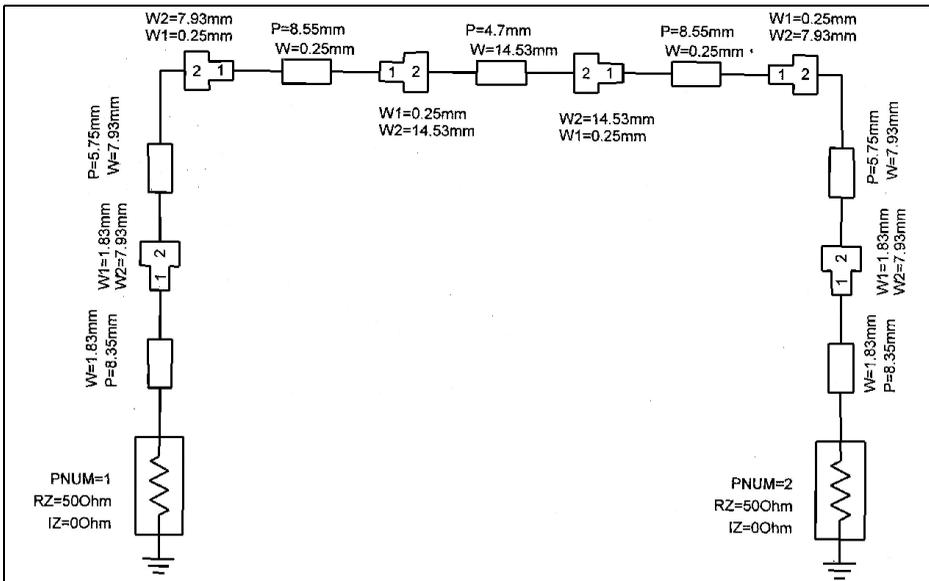
First and third capacitors: Width 7.93mm, Length 5.75mm

Both coils: Width 0.25mm, Length 8.55mm

Middle capacitor: Width 14.53mm, Length 4.70mm

### **6.2.2. Measurement results and check simulation using ANSOFT Designer SV**

Following some very promising initial experiences using the new free student version of the professional program AN-



**Fig 88: Circuit diagram used with ANSOFT Designer SV for the 1700MHz low pass filter.**

SOFT Designer, this specimen circuit board was analysed immediately and the measurement results were included in the simulation results.

An assessment of Fig 85 shows the following: the Designer calculates in advance that the two S11 humps are not identical - and the measurement confirms that. The maximum value for both humps is in each case 2dB higher than expected, and the effect can be seen immediately in Fig 86. This gives rise to an increase in the S21 ripple. The basic attenuation also rises more than had been expected; we are talking about a deterioration of about 0.1dB at 1,600MHz, as against the simulation - and that's nothing too awful.

In conclusion, we should compare the necessary expenditure for the user on the Designer, as against the PUFF simulation. Fig 87 shows how all circuit board and housing data is entered into ANSOFT, this is really well arranged.

The circuit diagram of the low pass filter, as it must be set up by the operator with the help of the Editor supplied, can be seen in Fig 88. All steps are derived from the component library supplied. If required, the dimensions of the striplines come from the "Line-Calculator". It has already been described as "TRL85" [1] and is now integrated into the program package, together with a "Filter-Calculator". The operator just has to program the sweep and the results output in order to obtain Figs 85 and 86.

In a forthcoming edition of VHF Communications there will be a description of how a fine practical project was incorporated into the Designer SV.

## 7. Conclusion

From low frequencies to high frequencies



- all my own work! The motto of this low pass filter series would run something like that. And if you have read this series of articles attentively, it will now be clear to you how the last circuit board development could be finished or converted to higher frequencies.

idea or other that they put aside owing to the frighteningly high development costs involved. The modern aids referred to can certainly make design much easier in some cases!

But time doesn't stand still and, as can be seen from the example of the ANSOFT Designer SV, programs are continually becoming more precise in their predictions and more comprehensive in the options they offer. Because in this case there's not even any additional expenditure involved, all developers should roll their sleeves up and dig out some old

**8. References**

[1] An Interesting Program, TRL85.exe, VHF Communications 4/2001, pp 199 - 205

**Correction to the article "An interesting program: Circuit simulation using PSPICE" by Gunthard Kraus, DG8GB in issue 4/2005**

On page 225, there were two mistakes in the paragraph at the top of the right hand column:

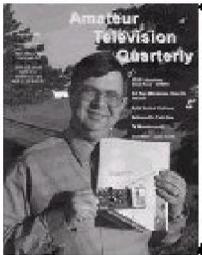
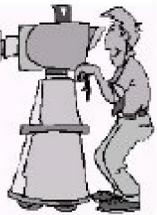
- The value of the resistor from the connection point between the two capacitors to ground should be between 10 and 100MΩ.
- The theory on inductances should have read "a DC voltage" being applied.

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*Gunthard Kraus, DG8GB*

# Internet Treasure Trove

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## Spice Simulators

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Who would like to download the most popular SPICE of programs on the market and compare them? Here is the correct address.

Address: <http://www.web-ee.com/Downloads/simulator/simulation.htm>

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## HF-fax

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That sounds very simply, but behind it is an extensive homepage with unbelievable amounts of information. The whole weather satellite community seems to have concentrated here. Everything is discussed here and made available from antennas to the technology.

Address: <http://www.hffax.de/>

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## Linear Technology

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The PSPICE article in the last issue generated a large amount of interest and resulted in some mail contacts and/or references. One was particularly interesting: Linear Technology offer a complete

Spice machine called "SWITCHER CAD" for investigating of DC to DC converter circuits. Fortunately every other circuit can be simulated and the company say it has "the most precise and most stable PSPICE version available free of charge". Looking at their web site showed that there is a lot of other interesting things to download:

"Switcher CAD" (PSPICE simulator), "Noise" (Program for the investigation of the noise from OP amps), "BodeCAD" (program for the production of Bode diagrams), "FilterCAD" (program for the investigation active filters), "Spice Models" (collection of various Spice Models), "The Configurator" (amplifier Design Tool), "LT1568 filter Design Guide" (active filters using the LT1568)

Address: <http://www.linear.com/company/software.jsp>

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## EmWONDER / S-parameter Resources

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In this harmless sounding homepage you will find not only S-parameters available on the Internet, but also the links for SPICE and S-parameters. It goes into more detail with a reference to an S-



parameter to Spice Model converter.

Address:  
<http://www.emwonder.com/sparameters/>

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## Cougar

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This company was called “Teledyne” and known as a microwave specialist. The new name brings a new side to the company with much to discover. Beside application notes and data of sheets there is an interesting conversion table..

Address:  
<http://www.cougarcorp.com/index1.asp>

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## Watkins Johnson

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It is important to look at active companies web sites regularly to see what new information is available Watkins Johnson have many application notes as well as first class technical notes. In addition there is also an extended technical archive, here there are nearly all the older, but outstanding, documents made available in scanned form (unbelievably back to 1974!).

Address:  
<http://www.wjcommunications.com/technotes/index.asp>

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## Mentor Graphics / IC Nanometer Design

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This company is normally mentally linked with the computer engineering, not for the WLAN and Bluetooth standards plus the RF technology available. Under the column “Physical verification” there is a whole collection of good simulators. Well hidden, but worthwhile investigating...

Address:  
[http://www.mentor.com/products/ic\\_nanometer\\_design/techpubs/index\\_noflash.cfm?v=google&p=adwords\\_broad\\_rf%20transistor&s=1x1&g=ic&c=ic\\_ocid\\_263\\_cfp\\_67\\_ceid\\_3-1x1](http://www.mentor.com/products/ic_nanometer_design/techpubs/index_noflash.cfm?v=google&p=adwords_broad_rf%20transistor&s=1x1&g=ic&c=ic_ocid_263_cfp_67_ceid_3-1x1)

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## Usefull links for electronics

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This Hungarian homepage contains a lot of interesting information for electronics. Take if there is anything of interest for you.

Address:  
<http://www.pollak.sulinet.hu/www/elek.html>

**Note:** Due to the fast movement of Internet content it is impossible to represent the up to date position. We apologise if the Internet addresses are no longer available or are changed at short notice by their operators.

Neither the authors nor the publisher can accept responsibility for the correctness of the content of the web sites mentioned.



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1999	Y	Y	Y	Y
1998	Y	Y	Y	Y
1997	Y	Y	Y	Y
1996	Y	Y	Y	Y
1995	Y	Y	Y	Y
1994	Y	Y	Y	Y
1993	Y	P	P	P
1992	Y	Y	Y	Y
1991	P	P	P	P
1990	P	P	L	P
1989	P	P	P	P
1988	P	P	P	L
1987	P	P	P	P
1986	L	Y	Y	Y
1985	Y	Y	Y	Y
1984	Y	Y	Y	Y
1983	P	Y	Y	Y
1982	P	P	P	P
1981	P	L	P	P
1980	P	P	P	P
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Carl G. Lodström, SM6MOM & KQ6AX

# The Noble Art of Measuring Optical Power.

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## 1.

### Background

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I often read about the mWs of optical output power from pointers and lasers. Browse the auctions on eBay and you will be surprised how many are just 5mW! One may be led to believe that the manufacturers manage to get just this level of output power with great precision.

In reality it is the maximum allowed power for a device to be rated as a “Class IIIa” device. More than 5mW and the risk of damaging an eye becomes considerable. From [1] we can get the following definition:

#### *Definitions of Laser Classes*

##### *Class II laser products*

*Lasers categorized as Class II cannot harm the retina because the human blink reflex is sufficient to provide protection. The power output is less than 1 mW. They are generally in the wavelength range of 630-680 nm.*

##### *Class IIIa laser products*

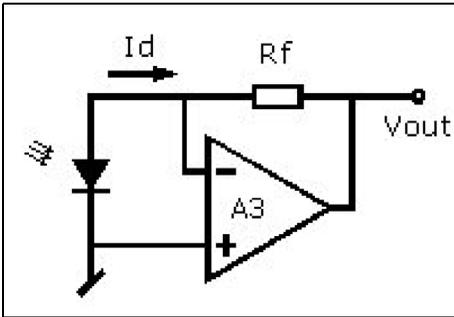
*Class IIIa lasers, which can be safe for momentary viewing, are a recognized eye hazard if viewed through optics (telescopes, magnifiers). The power*

*output is between 1-5 mW. They are generally in the wavelength range of 630-680 nm.*

So, how to determine how much we really get out of these devices or how good is an optical filter, a window, a lens and a mirror? For the amateur, experimenting with lasers and optics, it is useful to possess the means of measuring optical power.

From a previous article, Alexander Meier, DG6RBP, had a description of an optical output power meter [2]. He covers the issues well, although I would disagree with the practice of using the frequency of various lights. It is almost impossible to measure (directly) whereas the wavelength is what matters in gratings, diffraction, particle absorption of the light, scattering... Besides, filters and lasers are always sold by the nanometer (nm,  $10^{-9}$ m), not by frequency. The need for a monochromator is also discussed. Fortunately I described how to build a simple but effective one in my Optical Communications article, 2<sup>nd</sup> part [3]. The power meter that Alexander describes has trimmers for setting the calibration coefficient for various wavelengths. One would need the means to determine the wavelength to select the proper one.

Even though Alexander has covered the subject well, I have some additional points of view that may be useful.



**Fig 1: Schematic of a T-I amplifier.**

I had almost finished the photo diode part of this article when I remembered one more method, using a device that I thought up and built ten years ago! It is such a good one that we will save it for last!

## 2.

### Photo Diode Detector

The best detector for this purpose is the silicon photo diode. The current from it (in a short circuit, or at least into the low resistance of a good mA meter) is proportional to the optical power it is exposed to, but for a given wavelength. So a photo diode and a  $\mu\text{A}$  meter will do well. Chose the meter range for what you want to measure and the detector size. If the detector is large, and you hold it in full sunlight, maybe you will get 50mA! A narrow laser beam on the same detector may only give 1mA.

This will be hard to read with any kind of resolution, on the same meter. A smaller detector may give 3mA in sunlight and still 1mA from the laser (provided that the detector is not smaller than the beam) so now we can read both on one and the same (5mA?) meter movement. One can of course start out with a more sensitive movement (say  $50\mu\text{A}$ ) and shunt it to various ranges, selectable by a switch.

This will work just as well as any electronic circuitry (usually a Trans-Impedance Amplifier, a “TI-amp” is used for this), although currents smaller than  $1\mu\text{A}$  are difficult to measure without an amplifier of sorts.

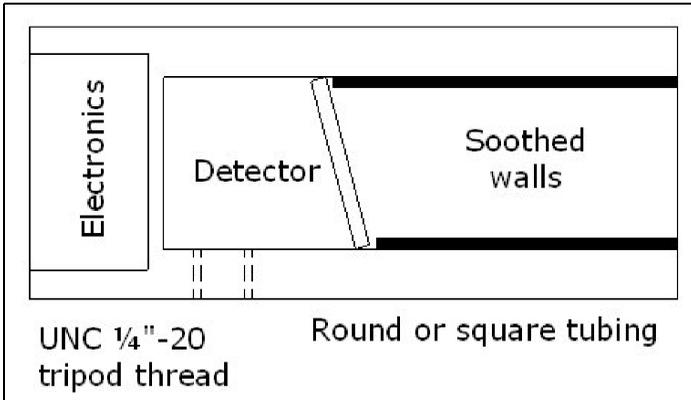
Fig 1 shows the typical T-I Amplifier. Input current,  $I_d$ , from the diode is opposed by an equal and opposite current from the output through  $R_f$ . An output voltage of  $-I_d \times R_f$  results. With  $R_f = 1\text{M}$  the output will of course be  $-1\text{V}/\mu\text{A}$  input current. Because  $I_d$  is opposed by an equal current from the output through  $R_f$ , there is no change of voltage at the node of the inverting input of the amplifier. This is the same as presenting a short circuit to the detector diode. This is very good when we want to measure the current from a photo diode.

#### 2.1 The detectors

The dark current of the detector diode and the bias current of the amplifier determine the lower limit for detection. It is not difficult to find these components with less than  $100\text{pA}$  respective, so very low light levels can be detected. It is advantageous (from this point of view) to select a “PN”, not a “PIN”, diode. They have a much lower dark current. Let us compare some data of two otherwise similar diodes by Siemens:

	BPW 33	BPW 34
Dark Current	20 pA.	2 nA
$I_d@1000 \text{ lux}$	$78 \mu\text{A}$	$80 \mu\text{A}$
Rise time	$1.5 \mu\text{s}$	$0.35 \mu\text{s}$
Type	PN	PIN

“PIN” means that it has an Intrinsic layer between the P and N layers. This makes a faster detector, but with the larger dark current, thus it is less suited for *very* low light levels. For the power meters we are dealing with here, one may select the faster -34 anyway. Especially since one never know when it will be of interest to study fast modulation of a light source. But, as we will see later, the -33 has its place too, so get a couple of each while you are at it! The -34 is “more IR”,



**Fig 2: The slanted detector mount in a tube.**

peaking at about 900nm, thus a little less favourable for 650nm measurements. One BPW 34 of mine gave 2.5mA in the Californian sun.

Both diodes have an area of 7mm<sup>2</sup> resulting in the listed responses to a 1000lux illumination from a reference incandescent lamp with a 2856K filament temperature. This determines the “colour temperature” from the lamp, and the detector is very colour sensitive, favouring red and IR light.

The more modern laser pointers have a diode emitting at 635nm, making them appear much brighter, as 670nm is the edge of visibility. It may be pointed out that although the 635nm one looks brighter, the 650nm laser will register as “higher power” when measured with a Silicon detector because 670nm is closer to the detector maximum. This will be true for all unfiltered Si and Ge Detector diodes. An exception is the photoptically filtered detectors such as BPW 21. They have a maximum in yellow-green, mimicking the sensitivity of the eye. This is more useful for exposure meters than in laser communication. Unless you use a green laser.

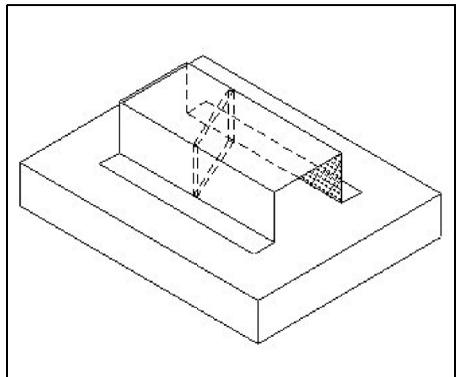
For communication by light, IR is preferred as it matches the detectors and penetrates smoggy air better over long distances. It may also have less diffraction from the emitting area in the diode, making collimation of the beam easier.

The only drawback is that it is mostly invisible.

You will soon find that even an incandescent lamp registers very well on a Silicon Photo Detector! Lamps have a lot of IR and small lamps can even be modulated directly by audio frequencies for communication.

**2.2 Mechanical solution**

The detector can be enclosed in a tube with sooted inside walls. If the detector is mounted at an angle to the tube, light is not reflected back to the source (Fig 2). The detector may not have to be tilted more than shown (15°) or so. Anyhow, the reflection will hit the blackened wall of the tube and will have almost no effect on the measurement and no effect on the



**Fig 3: Diagram of a simple detector housing.**

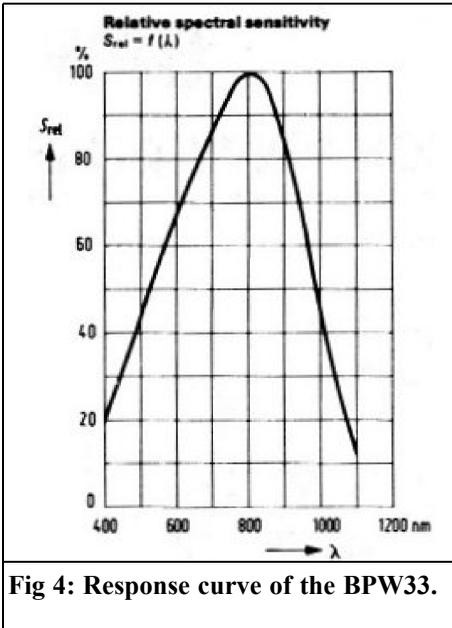


Fig 4: Response curve of the BPW33.

laser. In mounting the detector one may consider putting it deep enough that all light always reflects into the sooted wall. One may also consider mounting it deep enough that the moderate tilt allows the cosine correction for various input angles to be small and remain about the same for varying input angles.

It is a good idea to base this design on a detector diode that is not encapsulated, and without a window, since reflections (single and multiple) and additional losses will complicate measurements. The detector must of course be large enough for to receive the entire beam. Study the surplus market, and eBay, for some good size detectors! Fig 3 shows a very simple detector housing! The inside of which is sooted with a cigarette lighter.

### 2.3 Calibration

Studying a data sheet, like the one for BPW 33 [4] it seems that Osram has taken over this product These days, when MBA's are running engineering companies, it is hard to keep up with the many

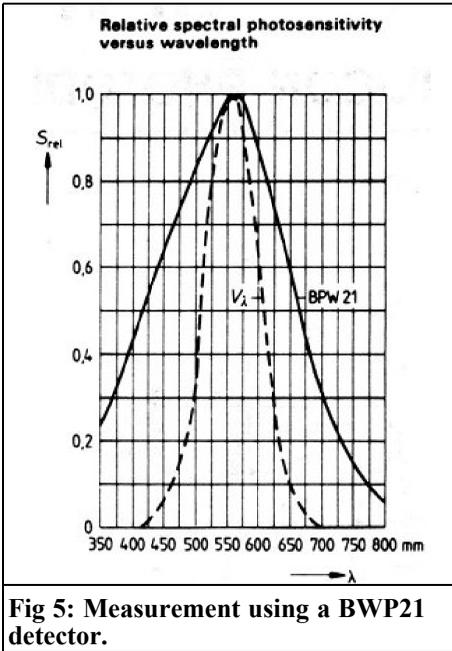
clever mergers and fancy name changes.

Anyhow, we can see that at 650nm the sensitivity is down to about 80% from the 800nm maximum. At 850nm the "spectral sensitivity" is given as 0.59A/W. The sensitivity curve (Fig 4) shows about 95% there. So it is reasonable to expect  $0.8/0.95 \times 0.59 = 0.5A/W$  for a 650nm source. 5mW would then make 2.5mA. With a 15° angle the cosine correction would be 0.966 resulting in a 2.4mA current. Assuming the use of an analogue meter movement, a 50µA instrument we will be able to read levels down to  $50E-6/0.59 = 85µW$ , full scale. Just make sure that you do not drop more than some 100mV in the meter. R = 2k or so.

With a TI Amplifier we can easily read currents of 0.1nA and less. 0.1nA would correspond to 17nW of light or a 1/300,000 of 5mW. This will probably cover all needs. For an output of, for example, 1V/mW, with the same detector and light wavelength, Rf should be  $1000/(0.5 \times 0.966) = 2.07k$ . I would make it variable from 1.5 to 2.5 or so, and have a knob with a calibration coefficient dial. One could even have wavelengths marked on the knob so it can be set to a matching calibration coefficient.

Possible opportunities for calibrating it against known beams should be welcomed. So it makes sense to put a battery pack together, and visit that Laser Exhibition when it comes to a town near you! Or bring your laser pointers and get them calibrated on the exhibitors meters! At the same time you can ask them if their power meter is corrected for various wavelengths and come across as a real pro!

If you do not have any data on the detector that you found in the Junk Box, it is probably safe to assume about 0.6A/W. For detectors like BPW21, filtered to be similar to the sensitivity of the eye, the value is lower, 0.34A/W at 550nm (yellow-green light). Almost all of this is due to the fall-off @ 550nm for



**Fig 5: Measurement using a BWP21 detector.**

a silicon detector. It reads precisely 1mA in the bright California Sun. It is listed as  $10\mu\text{A}$  at 1000lux. As sunlight is supposed to be 100,000lux, this agrees well. But we have no use for this filtering when measuring on IR and red sources.

I made a few measurements on an unknown, not very large, HeNe laser (632.8nm) and on a diode laser pointer, the 650nm variety. I got 0.36mA with a BPW 21. We can see its response curve in Fig 5 (and  $V$  is the response of the eye) at some 0.7 for 633nm. At the peak it has 0.34A/W so with  $0.34 \times 0.7 = 0.24\text{A/W}$  0.36mA should mean 1.5mW from the laser. This may very well be the case.

The 650nm laser pointer gets a coefficient of 0.6 with this detector.  $0.34 \times 0.6 = 0.204\text{A/W}$ . The current I read, 0.4mA, then corresponds to 2mW from the laser pointer. This may be true, it was not advertised as a 5mW device either. The output from a laser diode without optics is more complicated to measure. It really requires an Integrating Sphere. That is a

sphere, painted white inside, where the light bounces off the walls, so that no direct light reaches the detector. All sources can be measured in such a device, but making one is not justified.

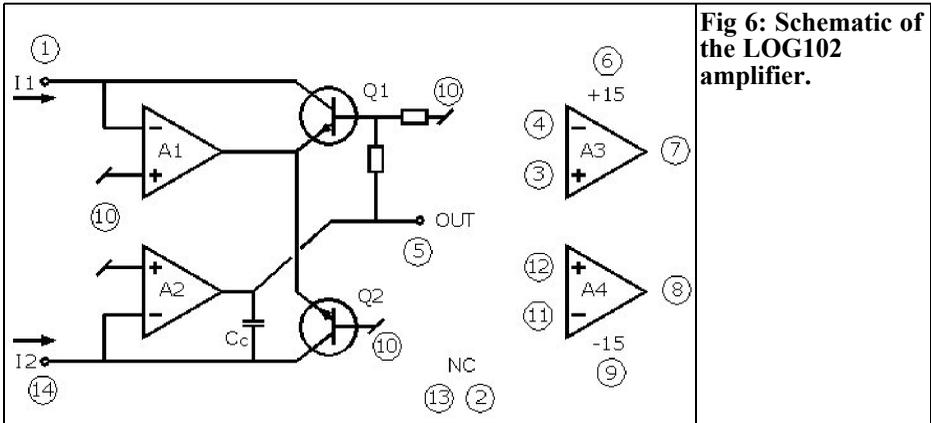
The laser diode output can spread quite a bit, especially in the plane perpendicular to the emitting aperture, so holding it very near a detector diode, with a large enough surface, is one way, although the measurement will be flawed by the outer rays impinging on the detector at an angle. One could have three or more diodes (in parallel) set up as a rounded wall, as three sides in a hexagon, covering most of the angle in one plane. As the shape of the output is elliptic, this may not be a bad idea. The cosine error will be small. Especially "stackable" detectors ought to do well, like the BPY 64P that is 6x6mm with a minimal dead area on the edges.

Another way to capture the entire beam is to shine it sideways, from the focal point, onto one half of a parabolic reflector from a flashlight. The reflector has a very large Numerical Aperture (NA) and will easily capture and re-direct all the light from the laser diode. It can be focused on even a small detector diode as much as a metre away. The efficiency of a good reflector is likely over 80% but not over 95%, so a good estimate of the output power can be made. No cosine correction will be needed because the entire beam will reach the detector from within a few degrees of the normal. This may be the best method for measuring a bare diode, without its optical assembly.

### 3.

## Making the invisible visible, IR viewers

For situations like this, using infrared, it is beneficial to have means of observing the beam focused on a screen. With the



**Fig 6: Schematic of the LOG102 amplifier.**

laser diode and reflector mentioned above, the intensity will be pretty high, so a regular digital camera, or a video camera, will probably see it even though they have filters to block IR. In a dark room one can perhaps see it with the naked eye, but do not look into the beam, especially not when focused.

One interesting device, recently available, is the NexImage camera from Celestron. It is based on the popular PC cam by Philips, but comes with fantastic software for astronomy. It is very sensitive, gives real time video and connects to the USB port of a PC, which also supplies the power. It needs a lens, but it is easy to rig a magnifying glass or an old film camera lens. It has a photoptic filter (greenish) that lets quite a lot of IR through. I have used it, with a mirror telescope, to focus the IR beam on the shopping centre, used in my article about Optical Communication. At the time I was unaware of this camera. Now I achieved a round 5cm spot at 200m from an IR laser diode, and a projector lens 25/1.4 for 8mm film. This is described in the addendum to the article on the VHF Communications web site [6]. Other PC cameras will probably do as well for making IR visible.

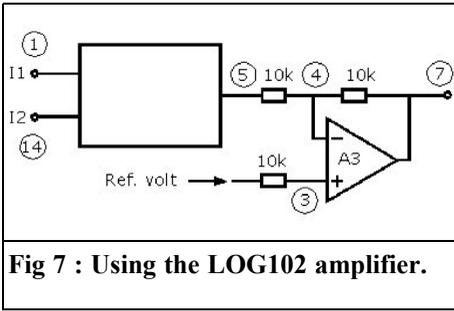
#### 4.

### Wide dynamic range measuring amplifier

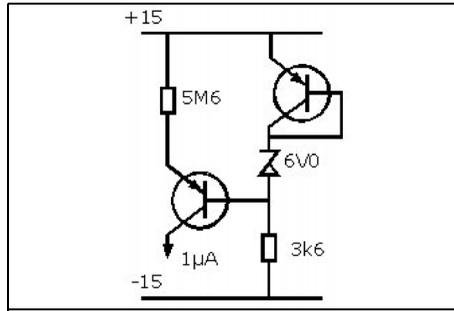
Just as it is nice to have a power meter for RF calibrated in dBm, a logarithmic scale, it is just as nice a feature for light meters! Fortunately there is an IC that will do this for us! It used to be seriously pricey, I remember when it was way over \$20, but now I see it at \$7.25, not much for such an agile device. The device is LOG 102 [7] by Burr Brown but Texas Instrument seems to have bought them, chips and all. There is a “Request Samples” area that I would visit it first.

So what is so great with this device? What is in it? It has current inputs (“short circuits” to the source) capable of handling the range of 1nA to 1mA, 6 decades, 120dB of light as light power is converted to current in the detector(s). There are two current inputs and two uncommitted OP amps (Fig 6). Each decade of input current imbalance results in one volt on the output. Using a BPW 21, and 1 $\mu$ A on the other input, this would be the response from full sunlight to the light from a half moon!

The A<sub>1</sub> and A<sub>2</sub> amplifiers have a typical input bias current of  $\pm 5$ pA, and as the



**Fig 7 : Using the LOG102 amplifier.**



**Fig 8: Circuit diagram of the current generator.**

measurement range of the device goes down to 1nA, it is obvious that a BPW 34 or BPW 21, with a dark current of 2nA, are not the optimal detectors if we want to reach the lowest levels. Although I could easily detect the half moon with the LOG 102 and a BPW 21, it would have been a better choice to use BPW 33 with its dark current of about 20pA. If we are only going to measure output levels from lasers we will be closer to the 1mA end of the range, so the dark current is of no consequence. Well, how about speed then?

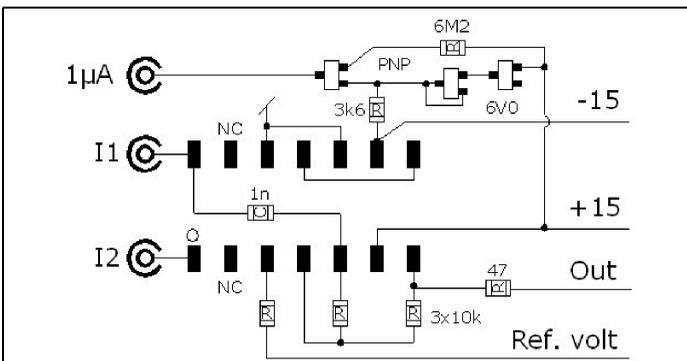
The LOG 102 is no speed demon, the darker, the slower. Reducing the value of  $C_c$  will increase the speed, but also the noise level at low input levels. I use 1nF in mine, but could probably have done well with 100pF. Experiment if speed is of the essence!

Looking into the LOG 102, it is based on the logarithmic relation between  $I_b$  and  $V_{be}$  in a transistor. The circuit is actually comparing two such currents, with two

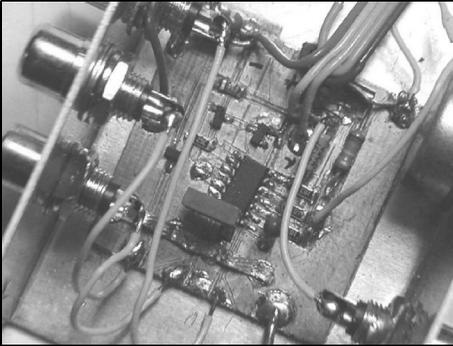
transistors, something that in my opinion makes it even more useful, although maybe not for our power measurements. Remember: "If you can measure something in a bridge, do it!" A dual photo diode could feed each input and the *difference* in illumination becomes an output signal. It can be very useful for position sensors, transmission and reflection measurements, colour temperature and spectroscopy, providing one reference channel and one measurement channel.

Anyhow, for our power meter, we will measure the current from one detector only (or with both the halves in parallel) so the other input will have to be satisfied from some constant current source, as zero input current falls outside the operational range (Fig 7).

Of course, I cannot describe a 14 page data sheet here, so I urge you to down-



**Fig 9: Layout for the LOG102 amplifier using SMD components.**



**Fig 10: Picture of the LOG102 amplifier circuit board.**



**Fig 11: Picture of the LOG102 amplifier front panel.**

load it [7] and study it, but I will supply enough information to make the thing work.

#### 4.1 The input range.

The input currents  $I_1$  and  $I_2$  can each range from 1nA to 1mA. Output voltage is the logarithm of the *ratio* between them, plus the offset. The full output range of 6V can only be obtained if one of the inputs is “halfway”, at 1 $\mu$ A. As the other input goes from 1nA to 1mA, the output goes from -3 to +3 V (or the other way around). By shifting the output by 3V positive, with the Reference Input, the output range becomes 0 to 6V, which I generally find more useful. With a pot, fed with 5V from a stable 78L05 regulator, the zero crossing can be set anywhere from 0 - 5V.



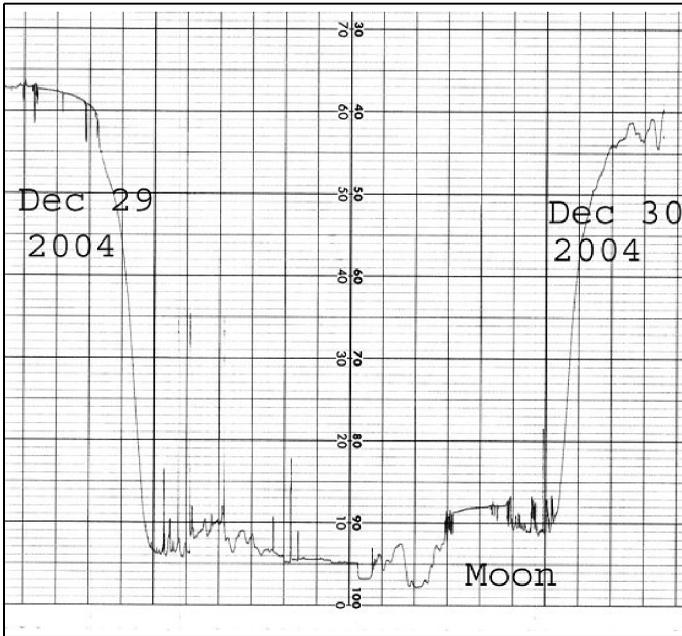
**Fig 12: Picture of the LOG102 amplifier rear panel.**

#### 4.2 The construction:

A resistor could generate the 1 $\mu$ A for “the other” input, but they advise against it in the data sheet, as the resistor noise would make the bottom of the range unusable. I took their advice and did not even try it. They present a simple Constant Current generator with two PNP transistors and a Zener diode (Fig 8). I built it and it works fine. It is reasonable stable. At +18°C I measured 986nA and when warmed up to about 45°C it gave 990nA. I used 6M but 5M6 may have been a better choice. The Zener voltage is important because between 6.0V or 6.2V is the range where  $V_z$  is temperature stable.

The power levels are all very small so I elected to build this with surface mount components. If you have a stereomicroscope, it makes life a lot easier. I used a sharp pocket knife (later confiscated on a flight as I forgot to put it in the checked luggage) and a piece of board laminate and started cutting, Figs 9 and 10 show the result, it is not beautiful, but it works!

The inputs and the 1 $\mu$ A output have one connector each, as do the signal output on the other side. I decided on phono plugs. A short cable, with a plug on each end, connects the 1 $\mu$ A current to the desired input. Or, differential diodes connected to both inputs can be used. Figs 11 and 12 show the front and rear panels.



**Fig 13: An 18 hour recording of a near full moon and the sun on the two adjacent days. X axis: one div/hr. Y axis 2 div/current decade.**

On the rear panel it says “0-1 $\mu$ A” I wonder what I thought of? It is wrong! Input range is 1nA ~ 1mA.

As usual for all my  $\pm 15$ V projects I employ a 5 pin, 240 $^\circ$ , DIN connector. The remaining two pins become inputs or outputs, as required. Here one of them is an output. My power supplies have two BNC connectors connected to these two pins, making hookups of projects like this, to other instruments, meters or recorders quick and simple. I even built a 12V  $\Rightarrow$   $\pm 15$ V switch mode supply, with this output connector, for mobile use!

Fig 13 shows a recording of a near full Moon and the days before and after. Above the M in “Moon” is an undisturbed hour of moonlight. In the early part of the evening, passing cars shone spikes of light on the detector. Both the Sun and the Moon recordings have brief dips from clouds passing by. We may also notice that the recording dropped as deep as 2cm below what was recorded for full moonlight. This could indicate that it got as dark as 1% of full Moon (scale is 10x per division) but I suspect

phenomena could begin to affect measurements of 0.01 Moon levels. Possibly not exactly a true registration, but the Moon light clearly registered. 12 divisions below the 1mA from the BPW 21 in full sunlight is  $10^{-15}$ A. Not likely a true registration, but the moonlight clearly registered, with some  $10^{-13}$ A.

## 5.

### The other method

Using a bolometer is a fundamentally non-electric method. A body heated by the irradiation with the temperature of the body measured electrically, although in principal a sooted thermometer will work.

From July 16 through July 22, 1994, pieces of an object designated as Comet P/Shoemaker-Levy 9 collided with Jupiter. (<http://www2.jpl.nasa.gov/s19/>)

At the time I was observing, it was clear



that I probably would not get to see much of it. It would not be totally dark and the impacts were destined to take place on the backside of Jupiter, just before coming up over the horizon. So what is a young man to do? The impacts must release a lot of heat on an otherwise cold planet. Maybe I could detect the heat pulses? How to make a really sensitive detector for radiated heat?

I decided to try a thermoelectric cooler (TE cooler), a small 12V device with 127 junctions. They also work as thermal generators. It should work if I could keep one side cool on something with a great thermal mass, soot the other side and irradiate it! Amplify the resulting DC voltage, which ought to be proportional to the power of the radiating source. One heat source => so much output voltage, two such heat sources => twice the output voltage I assumed. For our very low temperature differences, loss from re-radiation ought to play no role. I built it and it seemed to work fine. The sensitivity was very good.

The comet impact drew closer, I set up my detector but I could not even detect Jupiter against the even colder skies! This was not promising! Finally I realised that the corrector plate on my Schmidt Cassegrain telescope was the temperature I "saw", regular outdoor temperature. The glass plate is probably opaque to heat radiation at several  $\mu\text{m}$ . And if nothing else, it is warmer than space or Jupiter, so it "shines" its own 300K radiation onto the detector. If I had used a Newtonian telescope, with mirror surfaces only, it could have worked. Fortunately the "seeing" was excellent so I rigged the telescope with a video camera and got a nice video of the black dots from the impacts as they came up over the horizon.

### 5.1 Construction

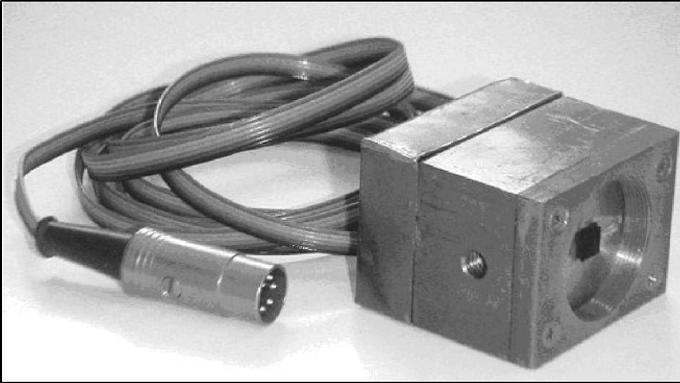
The TE cooler I used is size 6.0 x 6.2mm (by Melcor?) on the top plate. It is glued with RTV (silicone rubber) to a block of copper, about 350g in weight. The con-

ductors from the cooler go through the copper block, in drilled holes, to the other side, where an Instrumentation Amplifier ("IN Amp") is located under a cover, preventing thermal EMF from air currents. The gain of this excellent IN Amp, an AD524 from Analog Devices [8], is set to a fixed 1000x, but it is pin selectable to 100, 10 and 1x. With the  $R_G$  resistor the gain can be set to any value one may desire, but 1000x is fine for this purpose. It is not cheap, but it is very good. There are other kinds that are not bad either, but I think this is the one with the lowest noise.

The DIN connector system I use, described earlier, carries the supply voltages and the output signal.

Letting the Sun shine on the sooted side, with the copper block at room temperature equilibrium, generates an un-amplified signal of 16mV. The Solar constant is 1.37kW/m<sup>2</sup> in space and ~ 1kW/m<sup>2</sup> where I live. I found a number of 950 ~ 1100W/m<sup>2</sup> for Germany. Here in California it may be a bit more, but it will vary with time of year everywhere.

As the element is  $6 \times 6.2 =$  or  $37.2E-6\text{m}^2$  the power from the Sun is about 37mW. The un-amplified output signal is thus 2.313mW/mV. This can be detected with an ordinary DMM, with a resolution of  $100\mu\text{V}$ ,  $1/10\text{mV} = 0.2313\text{mW}$ . The added AD 524 improves measurements. A stable 1000x gain is very useable, but one has to be careful with circulating air currents, thermal EMF and heat radiation leaking in! The copper block must not be touched for a while before the measurement, or it should be thermally insulated. Amplified output signals of as little as 1mV (thermally generated:  $1\mu\text{V}$ ) are good, steady and useable, and I have not even low pass filtered the output! The bandwidth at a gain of 1000 is some 25kHz so the DC voltages we are looking at could be amplified a lot more with low pass filtering. 1mV amplified output should correspond to  $2.313\mu\text{W}$ . Compared to the photo diode detector with



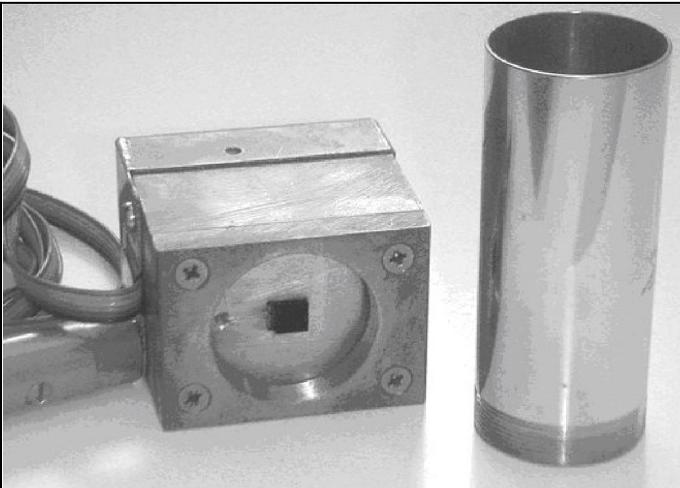
**Fig 14: Bolometer showing tripod mount.**

0.5A/W,  $2.313\mu\text{W}$  will generate  $1.16\mu\text{A}$ . With  $R_f = 865\Omega$  such a detector also gives 1mV out. The optical detector can be at least 1000 times more sensitive, although a lower practical limit of  $2\mu\text{W}$  is probably enough for measuring laser beams.

Photo diodes are good, but I believe this bolometric detector has the great advantage of a flat frequency response. The same for IR, visible or UV! It will depend on how black the soot is. I made fine soot with a normal butane cigarette lighter. Somebody has investigated this with a spectrograph and found the soot from bensol, burned in a normal wick lighter, to be especially black. I have not confirmed this.

## 5.2 Mechanical construction

You need a good size piece of copper. Mine weighs 333g and it seems to be sufficient for stability both thermal and mechanical. Maybe a much smaller block would be sufficient, I have not tried. Make it too small and it would start to warm up from the beam. There has to be one good flat surface to glue the TE cooler to. Use glue that will not get brittle, such as RTV (Room Temperature Vulcanising) silicone rubber or contact cement. Glue the side with the leads to the copper because is usually a little longer. Make sure to press it while the glue cures because a thin layer of glue is beneficial. Mask it all with tape and soot the outer side of the TE cooler. Be



**Fig 15 : Bolometer showing telescope mounting tube.**



careful not to heat it too much, at the most some 130°C, or the junctions in the TE cooler will melt and you have to start again.

It may be good to have a protruding collar, and/or a cover, to protect the TE cooler. I built mine for fitting to the telescope, I found a piece of tube used for bathroom sinks and I cut a thread for it in the copper block (Fig 14). It had the correct OD, 1¼", the same as the eye-pieces. The tube is made of brass, shiny on the inside. In my application it did not matter but for measuring lasers it may or may not be so good. If you need a tube, make it shorter and put a few dark baffles in it so light does not reflect from the walls onto the TE cooler. Sooting all the insides will make the instrument sensitive to the heat from touching the tube but less sensitive to stray heat radiation. The chrome plated outside of my particular tube was probably a good idea. Black baffles inside the tube will radiate heat and that will be detected. With the polished inside, I can handle the tube and get no reading from the heating of it. In Fig 15 you can see a tripod thread, UNC 1¼-20 in the block.

## 6.

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and  
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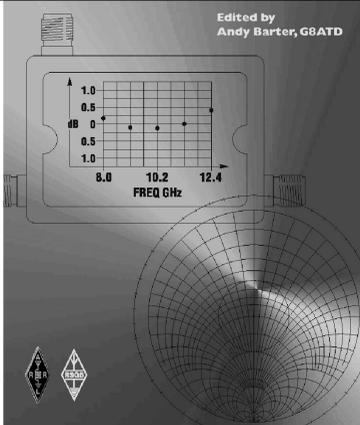
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# The International Microwave Handbook

## INTERNATIONAL MICROWAVE HANDBOOK

Edited by  
Andy Barter, G8ATD

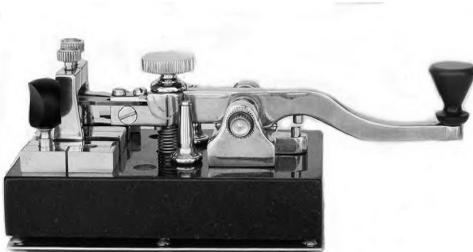


The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area, along with excellent designs from around the world to fire the imagination. Material has been drawn from many sources including the RSGB journal RadCom and the ARRL publications QST & QEX. Alongside this material a truly international range of sources have been used including items from Germany, Denmark, New Zealand, Slovenian and many more.

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Thomas Bergmann DG8NTB and Johannes Schad DG6NDS

# Design for a printed circuit board antenna, using a Log Periodic antenna as an example

**As part of the work for a dissertation at the Coburg Polytechnic, a broadband directional antenna had to be developed and manufactured. The following article is concerned with the problems encountered during this project. Printed circuit board antennas - why they don't always work properly, and how things can go well, in spite of everything, using a Logarithmic Periodic Dipole Antenna (LPDA) as an example!**

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## 1. Introduction

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### 1.1. Why LPDA

In contrast to all other types of antenna, the LPDA has a very wide operating range. The antenna length alone determines the bandwidth of the antenna. Its gain lies between 6 and 10dBi, depending on the model. Another welcome factor is that the electrical characteristics of the antenna display almost no change over the entire operating range. In the aerospace sector, the advantages of such antennas were well known in the early days enabling the number of individual antennas on aircraft to be greatly reduced.

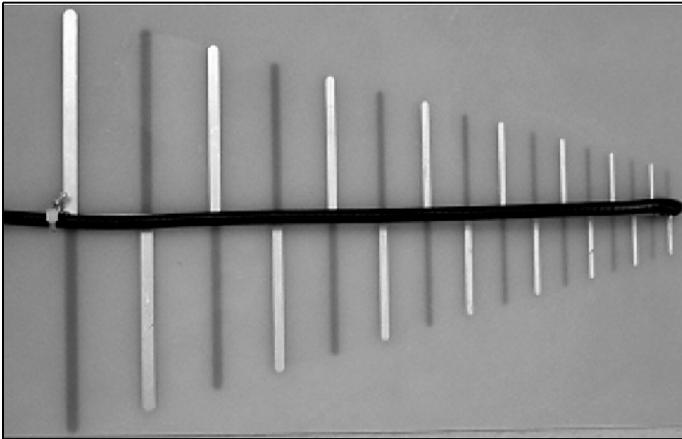
In addition to the standard 2.4GHz WLAN band, additional WLAN and WiMAX frequencies are being released in the near future. The LPDA can thus be used as a replacement for several individual antennas. A particular challenge is posed by development for relatively high frequencies, most assembly instructions relate to LPDA's up to 500MHz.

For frequencies above 1GHz, it is no longer possible to assemble a conventional LPDA using air spaced round metal elements without problems. For this reason an attempt was made to create an LPDA using stripline techniques and normal printed circuit board material for a frequency range of 1 to 3.3GHz.

Since the subject of LPDA's is very wide-ranging, only the most important characteristics and insights will be considered here. This article should be understood as a basis for further research in this area.

### 1.2. About the LPDA

In the 12th, updated edition of Rothammel's antenna book [1], in Chapter 27, we find the story of the origin and basic principles of such antennas. These basic principles are of no importance to the calculation, copying and results comparison of logarithmic periodical antennas. We are not going to go into the theoretic-



**Fig 1: A basic logarithmic periodic dipole antenna using stripline techniques.**

cal principles behind logarithmic periodic antennas.

Of all the logarithmic periodic forms of antenna, the LPDA is the most widely used. This is because it is the only one for which the calculations required can be completed at an acceptable cost. The gain, as already described in the introduction, lies between 6 and 10dBi. So from the point of view of gain alone, this antenna is no better than a waveguide antenna. Compared with high gain antennas such as yagi or quagi systems it is certainly inferior, but the LPDA beats these forms of antenna for its broad band performance. The front-to-back ratio of the LPDA, at 15 to 25dB, is superior to most antennas. The first logarithmic periodic antenna was developed and publicised by D. E. Isbell in 1960 [2].

The basic LPDA created using stripline techniques is shown in Fig 1. As can be seen, the antenna consists of several dipoles with varying lengths, thicknesses and intervals. The connections between the dipoles are crossed giving the phase changes necessary for operation. The LPDA is essentially connected in front of the shortest dipole, i.e. to the front in the direction of radiation. A load resistor, fitted behind the longest dipole, can just be seen in Fig 1.

## 2. Operating principle

A TEM wave runs from the feed point (in front of the shortest dipole) along the stripline and excites the dipole. Since the phase of the currents lag in the direction towards the feed point, the radiation is directed towards the feed point, so that the TEM mode runs contrary to the feeder. This phase lag generated by the phase reversal from element to element on the feeder. The major part of the energy is radiated through each dipole with a length corresponding to half a wavelength. These dipoles form the “active zone”  $B_{ar}$ . Only a small part of the energy reaches the longer dipoles lying behind them. In common with the dipoles that are too short, they essentially make no contribution to the far field.

As the main radiating section of the antenna, the “active zone” is responsible for the far field characteristics of the LPDA. As the frequency increases, the active zone moves forward in the direction of the feed point. If the frequency has changed by the factor  $\tau$  then this moves forward one dipole. The TEM mode is attenuated so much in the active zone by the energy radiation that it becomes insignificantly small behind this



zone. This means that the section behind the active zone no longer has any influence on the electrical characteristics of the antenna. Thus the type of load (stub or dummy load resistor) has no influence on the operating frequencies.

If these antennas are operating below their operating frequencies, then the TEM moves along the stripline and cannot be radiated due to a lack of resonance. Without a load resistance, this means that the entire RF power is reflected back to the transmitter, which results in a very poor VSWR. If there is a relatively large mismatch, this can lead to the destruction of the transmitter. For this reason, the LPDA was fitted with a load resistor.

The basic principles for calculating the LPDA using stripline techniques are contained in the "Rothammel" antenna book [1] and the "High-frequency Engineering Pocketbook" by Meinke, Gundlach [3].

### 3. Typical calculation of an LPDA from 1 to 3.3GHz

#### 3.1. Bandwidth B

The bandwidth B of the antenna is determined as follows by establishing the upper and lower limiting frequencies,  $f_o$  and  $f_u$ :

$$B = \frac{f_o}{f_u - f_u \cdot 0.05} = \frac{\lambda_u + \lambda_u \cdot 0.05}{\lambda_o}$$

$$B = \frac{3300\text{MHz}}{1000\text{MHz} - 50\text{MHz}} = 3.47$$

To prevent matching and the front-to-back ratio deteriorating abruptly at the lower frequency limit, this should be corrected down by approximately 5%. In the formula above, this factor has already

been taken into consideration.

#### 3.2. Grading factor and interval factor

The grading factor,  $\tau$  and the interval factor  $\sigma$ , influences the LPDA apex angle. In general, we can say that the gain increases as the apex angle, decreases, and as the grading factor approaches the value of 0.98.

$$\sigma = 0.25 \cdot (1 - \tau) \cot \frac{\alpha}{2}$$

or

$$\cot \frac{\alpha}{2} = \frac{4\sigma}{1 - \tau}$$

We can also say that an increase in  $\sigma$  leads to an increasing number of elements.

$$\tau = \frac{1_{n+1}}{1_n} = \frac{R_{n+1}}{R_n} = \frac{S_{n+1}}{S_n} < 1$$

A reduction in the grading factor leads to an increase in the antenna length,  $A$ . The interval factor,  $\sigma$  is also designated as the relative dipole interval and can be selected freely, like the grading factor.

$$\sigma = \frac{d_n}{\lambda_n} = \frac{d_n}{2 \cdot l_n}$$

We can find optimal values for both variables with Fig 2. The following inter-relationship applies for finding the best value for  $\sigma$  for a given grading factor:

$$\sigma_{opt} = 0.258 \cdot \tau - 0.066$$

If we select  $\sigma < \sigma_{opt}$  the gain is reduced, while on the other hand with  $\sigma > \sigma_{opt}$  the radiation pattern becomes worse (emergence of side lobes). Depending on the grading factor, the values for the optimal interval factor lie between 0.12 and 0.19.

From the diagram we selected  $\sigma = 0.09$

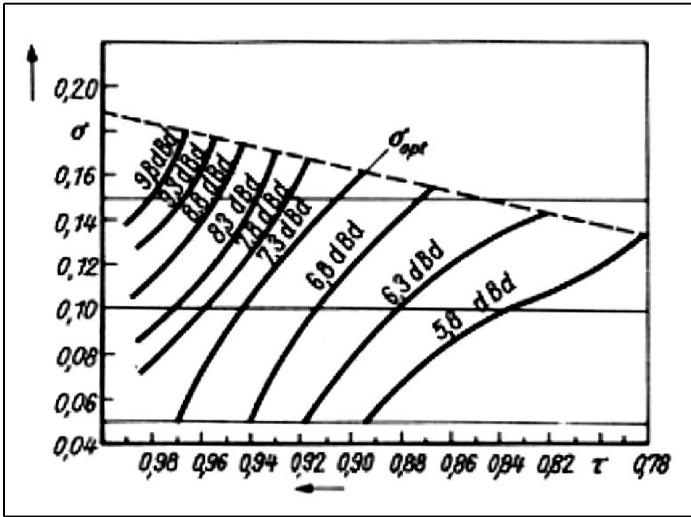


Fig 2: Chart of grading factor against interval factor.

and  $\tau = 0.9$  for this antenna. Thus the apex angle becomes:

$$\cot \frac{\alpha}{2} = \frac{4 \cdot 0.09}{1 - 0.9} = 3.6 \Rightarrow \alpha = 31.05^\circ$$

$$\sigma_{opt} = 0.258 \cdot \tau - 0.06 = 0.258 \cdot 0.9 - 0.06 = 0.17$$

With  $\sigma = 0.09$  and  $\tau = 0.9$ , the gain to be expected from the diagram is approximately 6.3dBd or 8.45dBi. When the values for  $\tau$  and  $\sigma$  have been established, we can quickly obtain a general idea of the size of the antenna.

### 3.3. Structure size $B_s$ and number of elements $E_n$

Using the operating range  $B$  and the active zone  $B_{ar}$ , we can now determine the structure size  $B_s$  of the antenna.  $B_s$  is the length ratio of the longest element to the shortest. For the active zone  $B_{ar}$  we obtain:

$$B_{ar} = 1.1 + [3.08 \cdot (1 - \tau) \cdot \sigma]$$

$$B_{ar} = 1.1 + [3.08 \cdot (1 - 0.9) \cdot 0.09] = 1.38$$

Using the value for the active zone  $B_{ar}$

and the operating range  $B$ , we can now determine the structure size  $B_s$  of the antenna:

$$B_s = B \cdot B_{ar}$$

$$B_s = 3.47 \cdot 1.38 = 4.79$$

Using  $B_s$ , we can now calculate the number of elements,  $E_n$ , required:

$$E_n = 1 + \frac{\lg B_s}{\lg \left( \frac{1}{\tau} \right)}$$

$$E_n = 1 + \frac{\lg 4.79}{\lg \left( \frac{1}{0.9} \right)} = 15.9$$

The result for  $E_n$  must always be rounded to a whole number. We thus obtain a total of 16 elements for the present antenna. We can also use the structure size to calculate the antenna length  $A$  in accordance with the following formula:

$$A = \left( 1 - \frac{1}{B_n} \right) \cdot \left( \frac{4 \cdot \sigma}{1 - \tau} \right) \cdot \frac{\lambda_{max}}{4}$$

$$A = \left( 1 - \frac{1}{4.79} \right) \cdot \left( \frac{4 \cdot 0.09}{1 - 0.9} \right) \cdot \frac{0.316m}{4} = 0.22m$$



The formula referred to on Page 642 of the 12<sup>th</sup> edition of Rothammel's antenna book for the antenna length, A (27.4.12) is incorrect. The formulae referred to above must be used instead.  $\lambda_{\max}$  in the formula relates to the lower limiting frequency corrected by 5% (950MHz).

### 3.4. Element lengths, thickness and intervals

Since the LPDA is to be assembled on circuit board material using stripline techniques, all the considerations and calculations change from the procedure referred to in [2]. In [2] the shortening factor taken as insignificant but this must be taken into account. It has in fact been demonstrated that in an LPDA design the shortening factor causes a deviation of approximately 15%! We can also use non-homogenous track widths to create constant impedances,  $Z_a$ , for all dipoles, which represents a powerful advantage for subsequent calculation.

The lengths of the individual elements can be calculated in relation to the longest dipole from the shortening factor or from the grading factor,  $\tau$ . The longest half-wave dipole is calculated as:

$$l_1 = \frac{\lambda_{\max}}{2} \cdot V$$

First we have to consider how thick the longest element actually may or should be. Since we could produce the antenna with the help of a milling plotter, we had the option of giving all the elements different widths. For normally assembled LPDAs that cannot be done because it is almost impossible to obtain aluminium rod in the various thicknesses required; the expenditure would be far too great. So all known calculation models start from an identical element diameter.

For this design, we can make use of this advantage and make all the elements to give a constant shortening factor i.e. the

elements are made varying widths, corresponding to their lengths. Since we do not know how wide the last element is, there is an optimising problem. Assuming we select a width of 4.8mm for the longest element, then after a long time spent in calculations we obtain a width of just 0.99mm for the shortest element.

#### Proof:

$$l_1 = \frac{\lambda_{\max}}{2} \cdot V = \frac{0.316m}{2} \cdot 0.89 = 140.52mm$$

$$d = 4.80mm$$

The longest element is the half wave dipole for 950MHz. The corresponding wavelength we obtain is approximately 31.6cm. If we select an element thickness of 4.8mm, then the shortening factor of the half wave dipole based on the ratio  $\lambda/d = 65.79$ . We obtain a shortening factor of precisely 0.89. This again means that in practice the element must be made smaller by the factor 0.89. Thus we obtain a length of 140.52mm for the longest element. Since all subsequent element lengths are dependent on  $l_1$ , we obtain the lengths in Table 1.

So that we can obtain the same impedance level,  $Z_a$ , and the same shortening factor for all the dipoles, all thicknesses are selected in accordance with the same  $l/d$  ratio. Corresponding to the longest element, we obtain:

$$v = l/d = 140.52mm / 4.8mm = 29.3$$

Since the shortest element is still only 28.93 mm long, we obtain an element width of:

$$d = l/v = 28.93mm / 29.3 = 0.99mm$$

As can be seen, the optimising problem lies in the fact that we cannot know in advance how wide the smallest element is. So that we do not have to calculate all 16 elements every time and then to our



**Table 1: List showing the dimensions for the LPDA. L, element length, d element width, S element spacing.**

$I_2 = I_1 \cdot \tau = 140.52\text{mm} \cdot 0.9 = 126.47\text{mm}$	$d = 4.32\text{mm}$	$S_1 = 28.40\text{mm}$
$I_3 = I_2 \cdot \tau = 126.47\text{mm} \cdot 0.9 = 113.82\text{mm}$	$d = 3.88\text{mm}$	$S_2 = 25.56\text{mm}$
$I_4 = I_3 \cdot \tau = 113.82\text{mm} \cdot 0.9 = 102.44\text{mm}$	$d = 3.49\text{mm}$	$S_3 = 23.00\text{mm}$
$I_5 = 92.42\text{mm}$	$d = 3.14\text{mm}$	$S_4 = 20.70\text{mm}$
$I_6 = 82.98\text{mm}$	$d = 2.83\text{mm}$	$S_5 = 18.64\text{mm}$
$I_7 = 74.68\text{mm}$	$d = 2.55\text{mm}$	$S_6 = 16.78\text{mm}$
$I_8 = 67.21\text{mm}$	$d = 2.29\text{mm}$	$S_7 = 15.10\text{mm}$
$I_9 = 60.49\text{mm}$	$d = 2.06\text{mm}$	$S_8 = 13.59\text{mm}$
$I_{10} = 54.44\text{mm}$	$d = 1.85\text{mm}$	$S_9 = 12.23\text{mm}$
$I_{11} = 49.00\text{mm}$	$d = 1.67\text{mm}$	$S_{10} = 11.01\text{mm}$
$I_{12} = 44.10\text{mm}$	$d = 1.50\text{mm}$	$S_{11} = 9.90\text{mm}$
$I_{13} = 39.69\text{mm}$	$d = 1.35\text{mm}$	$S_{12} = 8.91\text{mm}$
$I_{14} = 35.72\text{mm}$	$d = 1.22\text{mm}$	$S_{13} = 8.02\text{mm}$
$I_{15} = 32.15\text{mm}$	$d = 1.09\text{mm}$	$S_{14} = 7.22\text{mm}$
$I_{16} = 28.93\text{mm}$	$d = 0.99\text{mm}$	$S_{15} = 6.50\text{mm}$

disappointment discover that the last element has become too small, an optimising program has been written. Using this program, all elements can be calculated on the basis of the longest element, the shortening factor and the initial thickness. We can then decide quickly which values are still acceptable and usable.

The element intervals are less critical and can be calculated in accordance with the following formula:

$$S_1 = \sigma \cdot \lambda_{\max} \quad \text{or}$$

$$S_1 = [0.5 \cdot (I_1 - I_2) \cdot \cot\alpha/2]$$

$$S_1 = \sigma \cdot \lambda_{\max} = 0.09 \cdot 31.6\text{cm} = 28.40\text{mm}$$

$$S_2 = S_1 \cdot \tau = 28.40\text{mm} \cdot 0.9 = 25.56\text{mm}$$

$$S_3 = S_1 \cdot \tau^2 = 28.40\text{cm} \cdot 0.9^2 = 23.00\text{mm}$$

$$S_4 = S_1 \cdot \tau^3 = 28.40\text{cm} \cdot 0.9^3 = 20.70\text{mm}$$

...

If we add up all the individual intervals, we arrive at an antenna length, A, of 22.56cm. Further back, using a formula for A, we obtained a value of 22cm and this shows that the rough formula for estimating the antenna length is reasonably reliable.

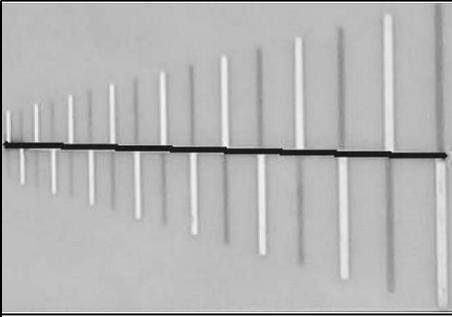
### 3.5. Assembly and feeding the LPDA

The assembly and feeding of the LPDA using stripline techniques requires considerably more than the process described by all antenna books for current LPDAs and has shown itself to be a really tough nut in research. But, with the help of a self-written calculation program, we were in a position to get a handle on this problem.

#### The problem:

Once we have calculated all the element lengths, intervals and thicknesses, we have to start thinking about the transmission line. Since we had a double-sided printed circuit board made from FR4 material, our first thought was to have the corresponding dipole elements mounted on one side. A stripline was to act as a waveguide between the two sides. This even had the advantage that by using a suitable stripline width we could have matched the impedance of the dipole to the 50Ω coaxial cable.

The dark line shown in Fig 3 represents this stripline. As can be seen, the corresponding dipole elements and the second part of the stripline are on the opposite



**Fig 3: Picture of the antenna showing the stripline.**

side of the FR4 board. It goes without saying that you must pay the greatest attention to precision when milling the structure.

Once all data needed for assembly had been worked out, the LPDA was digitally displayed on the computer with the help of the EAGLE circuit board layout program. Once the antenna had been milled, it became clear when the LPDA was measured that the ripple, between 1 and 3.3GHz was decidedly high. There were four VSWR oscillations from absolute matching up to 1:2.5.

Due to the fact that the stripline was matched in relation to both the impedance of the individual dipole and the impedance of the coaxial cable, an optimal matching could be expected. Moreover, the radiation diagram of the antenna was more than remarkable; it was difficult to find an unambiguous lobe. The antenna had formed several lobes at some frequencies (double lobes) at +/-45 degrees to the actual main direction of radiation. Following much checking and re-examination of all the formulae, it became clear that some errors had crept into the chapter on LPDA's in Rothammel's antenna book [1]. No matter how many times the antenna was revised, no significant improvement was achieved. The validity of the formulae used had been confirmed with the help of several textbooks.

After making yet another copy of an LPDA on a different printed circuit board material, RO4003, it became clear that this antenna was different from the LPDA on the FR4 material (the stripline was correspondingly matched to the thickness of RO4003).

The difference in the two circuit board materials lay in the fact that RO4003 is rather thinner than FR4 and also has a lower dielectric constant. Moreover the dielectric loss factor is smaller than for FR4 by a power of ten.

The breakthrough idea came during the lunch break. The problem lay in the stripline, which was completely filled with epoxy material. Since FR4 has a dielectric constant of about 4.3, the wave propagates in the stripline at a reduced speed. We know from coaxial cables that the wave velocity in a dielectric is always lower than in free space:

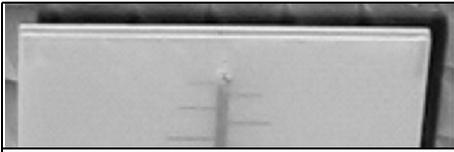
$$v_R = \frac{c}{\sqrt{\epsilon_r}} = \frac{300000 \text{ km/s}}{\sqrt{4.3}} = 1.45 \cdot 10^5 \text{ km/s}$$

$$\frac{v_R}{c_0} \cdot 100 = \frac{100}{\sqrt{4.3}} = 48$$

[% of  $c_0$ ]

In plain language, this means that the wave coupled into the stripline cannot propagate itself in accordance with the phase condition for logarithmic periodic operation. The waves radiated from the dipoles into space overtake the wave in the stripline and reach adjacent dipoles faster. The effect resulting from this can hardly be detected mathematically, but it is all the more measurable. Since RO4003 has a dielectric constant of approximately 3.3, this effect was less pronounced. That was also the reason for the small but still noticeable difference in the VSWR.

In order to pursue this suspicion, each half of the antenna was milled on a



**Fig 4: Picture showing the antenna made from two PCBs.**

separate FR4 circuit board. The two halves were then placed on one another and fastened with spacers in such a manner that the dielectric of the stripline consisted exclusively of air ( $\epsilon_r = 1$ ). Measurements on the antenna confirmed the suspicion. In addition to a VSWR of better than 1:1.5 over the entire operating range, a considerably more favourable radiation diagram was measured.

The LPDA halves can be seen placed one on top of another in Fig 4. Both boards do influence the wave field, but since the majority of the field is outside the epoxy material, the error this causes is very slight.

### 3.6 Feed and matching

The dipoles of the LPDA were calculated in such a way that each dipole has the same impedance level. A 50Ω coaxial cable (type RG58) was used to connect the antenna. We can now match the input impedance,  $Z_a$ , of the dipole to the 50Ω cable. The impedance level,  $Z_a$ , is calculated as:

$$Z_a = 120\Omega \cdot \left[ \left( \ln \frac{1}{d} \right) - 2.25 \right]$$

$$v = \frac{l}{d} = \frac{140.52mm}{4.8mm} = 29.275$$

which gives us an impedance level for the dipole of:

$$Z_a = 120\Omega [(\ln 29.275) - 2.25] = 135.21\Omega$$

To match the radiation resistance of approximately 135Ω to the coaxial cable,

we calculate the stripline in such a manner that it acts as a transformation line:

$$Z_0 = \frac{R_e^2}{8 \cdot \sigma^1 \cdot Z_a} + \left[ R_e \cdot \sqrt{\left( \frac{R_e}{8 \cdot \sigma^1 \cdot Z_a} \right)^2 + 1} \right]$$

$\sigma^1$  is the average interval factor and is calculated from:

$$\sigma^1 = \frac{\sigma}{\sqrt{\tau}} = \frac{0.09}{\sqrt{0.9}} = 0.0949$$

where  $R_e$  represents the impedance of the 50Ω coaxial cable used

$$Z_0 = \frac{(50\Omega)^2}{8 \cdot 0.0949 \cdot 135.21\Omega} + \left[ 50\Omega \cdot \sqrt{\left( \frac{50\Omega}{8 \cdot 0.0949 \cdot 135.21\Omega} \right)^2 + 1} \right] = 79.97\Omega$$

The stripline's impedance is therefore 79.97Ω. In order to be able to determine the dimensions of the stripline, we need to know that this is independent of various factors. At this point, we have to go into the following optimisation problem:

The most precise calculation for the width of the stripline is provided by the effective strip width,  $w_{eff}$ . In addition to the interval,  $a$  or  $h$ , and the line width,  $b$  or  $w$ , it also takes into account the influence exerted by the thickness,  $d$ , of the conductor strip. But since the interval,  $w$ , cannot be predicted automatically, the following formula from [Erich Pehl, P.35] with  $w = b$  serves as an approximation. We therefore need the value from the following equation merely as an initial starting point, so that we can come up with more precise values later.

$$Z_0 = Z_L = 120\Omega \cdot \pi \cdot \frac{1}{\sqrt{\epsilon_r}} \cdot \frac{a}{a+b}$$

For:  $d \rightarrow 0$  and  $\frac{b}{a} > d$

$$(a+b) \cdot Z_L = 120\Omega \cdot \pi \cdot \frac{1}{\sqrt{\epsilon_r}} \cdot a \Rightarrow b = \frac{120\Omega \cdot \pi \cdot a - Z_L \cdot \sqrt{\epsilon_r}}{Z_L \cdot \sqrt{\epsilon_r}}$$

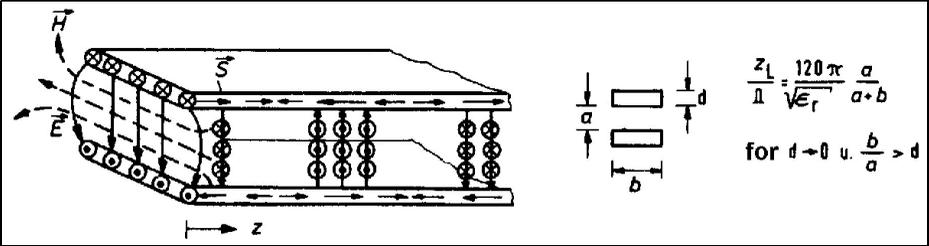


Fig 5: Diagram showing the relationship between E and H fields.

FR4 material was used as spacers for the two LPDA halves. Excluding the copper coating, this is 1.53mm thick. Since the dielectric in the stripline is air, we can calculate the width of the stripline,  $b = w$ , after converting the above formula.

for the stripline width,  $w_{eff}$  a value is required for  $w$ . In the simulation, it became clear that the formula set out above provides relatively good values for  $w$ . So we propose the following method in order to optimise the stripline:

$$w = b = \frac{120\Omega \cdot \pi \cdot a - Z_L \cdot a \cdot \sqrt{\epsilon_r}}{Z_L \cdot \sqrt{\epsilon_r}}$$

$$= \frac{120\Omega \cdot \pi \cdot 1.53mm - 79.97\Omega \cdot 1.53mm \cdot \sqrt{1}}{79.97\Omega \cdot \sqrt{1}} = 5.68mm$$

Use the stripline formula to calculate an approximation for the stripline width,  $w$ . Then insert this value into the precise formulae for the microstrip and use it to calculate the effective strip width,  $w_{eff}$ , the effective dielectric constant,  $\epsilon_{eff}$  and the associated impedance level,  $Z_{L'}^{eff}$ . The impedance level thus obtained,  $Z_{L'}^{eff}$ , must correlate as well as possible with the calculated value for  $Z_0$ . If this is not the case, alter the values for  $w$  by hand, and try them out until  $Z_{L'}^{eff} = Z_0$  is attained. Since this manual optimisation is almost uncontrollable due to its complexity, this optimising aid was used in our LPDA program.

Previous simulations and calculation models have shown that the formula set out in the book “Microwave Technology” [4] by Erich Pehl is not suitable for creating the optimum stripline. There is also a discrepancy in the conditions for using the formula. The reason for this is that the ratio arising from  $b$  and  $a$  gives a unit-less value. But for the use of this formula the author requires that this value should be bigger than a length expressed in units. Maybe he just forgot to make the corresponding value here unit-less as well. Unfortunately, Rothammel's antenna book [1] cannot confirm this, as the formula it sets out (5.7.1) for the stripline in its 12th edition is unusable. The pocket book on high-frequency techniques from Meinke, Gundlach [3] merely indicates the fact that the stripline has double the value of a microstrip with the same dimensions.

From the simple formula for the stripline, based on a height of  $h = 1.5mm$ , we obtain a width of  $b = w = 5.68mm$ . We use this value to calculate the effective strip width in accordance with the equations for the microstrip. At this point, we should point out once again that these formulae have been matched to the stripline. To make the results valid for the stripline, instead of the height,  $h$ , we now have to use  $h/2$ :

But for us to be able to use the microstrip formula and thus attain the optimal value

$$(A) \quad w_{eff} = w + 1.25 \cdot \frac{d}{\pi} \cdot \left( 1 + \ln \frac{h}{d} \right)$$



For  $2 \cdot w/h \geq 1/2\pi$

$$(B) \quad w_{eff} = w + 1.25 \cdot \frac{d}{\pi} \cdot \left( 1 + \ln \frac{4 \cdot \pi \cdot w}{d} \right)$$

For  $2 \cdot w/h < 1/2\pi$

Since  $2 \cdot w/h \geq 1/2\pi$ , formula (A) applies for the calculation. As can be seen from Fig 5,  $d$  is the thickness of the copper coating; for FR4 this is 35µm thick. For  $w_{eff}$  this gives us:

$$w_{eff} = 5.68mm + 1.25 \cdot \frac{0.035mm}{\pi} \cdot \left( 1 + \ln \frac{1.53mm}{0.035mm} \right) = 5.75mm$$

To calculate the impedance level,  $Z_L$ , we also need the value for the effective dielectric constant,  $\epsilon_{eff}$ :

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \cdot \left( \frac{1}{\sqrt{1 + 12 \cdot \frac{h}{2 \cdot w_{eff}}}} + 0.04 \cdot \left( 1 - \frac{2 \cdot w_{eff}}{h} \right)^2 \right)$$

Since  $2 \cdot w_{eff}/h > 1$ , the first part of the formula now applies. For  $\epsilon_{eff}$  this gives us:

$$\epsilon_{eff} = \frac{1+1}{2} + \frac{1-1}{2} \cdot \frac{1}{\sqrt{1+6 \cdot \frac{1.53mm}{5.75mm}}} = 1$$

The following calculation of the impedance level will show how close the approximation for  $b = w = 5.68mm$  has come to the value  $Z_0$  obtained of :

$$(A) \quad Z_L = 2 \cdot \frac{Z_0}{2 \cdot \pi \cdot \sqrt{\epsilon_{eff}}} \cdot \ln \left( \frac{4 \cdot h}{w_{eff}} + \frac{w_{eff}}{2 \cdot h} \right)$$

For  $2 \cdot w_{eff}/h < 1$

(B)

$$Z_L = 2 \cdot \frac{Z_0}{\sqrt{\epsilon_{eff}}} \cdot \frac{1}{\frac{2 \cdot w_{eff}}{h} + 1.393 + 0.667 \cdot \ln \left( 1.444 + \frac{2 \cdot w_{eff}}{h} \right)}$$

in all other cases

Since  $2 \cdot w_{eff}/h > 1$ , the formula (B) applies for the calculation. For  $Z_L$  this gives us:

$$Z_L = 2 \cdot \frac{120\Omega \cdot \pi}{\sqrt{1}}$$

$$\frac{1}{\frac{2 \cdot 5.75mm}{1.53mm} + 1.393 + 0.667 \cdot \ln \left( 1.444 + \frac{2 \cdot 5.75mm}{1.53mm} \right)} = 72.69\Omega$$

As can be seen, a stripline with a width of  $w = 5.68mm$  would have an impedance level of app.  $Z_L = 72.69\Omega$ . So we are only just below the required value of  $Z_0 = 79.97\Omega$ . By means of skilful testing, with the help of the optimisation program, we can arrive at the best matching of the point with a stripline width of  $w = 5.01mm$ . The effective strip width is then  $w_{eff} = 5.08mm$ .

Since we now have all the data needed for assembly, the dimensions of the antenna are transferred into the circuit board layout program EAGLE. The two LPDA halves are then milled out and assembled.

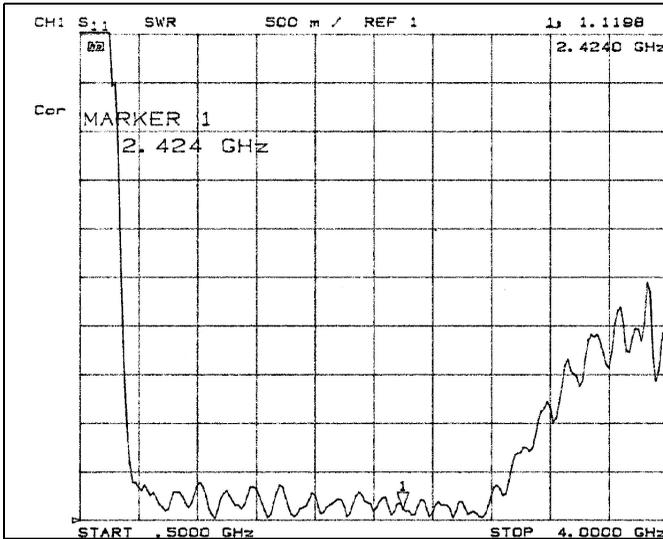
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## 4. Measurement results and discussion

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First the VSWR behaviour of the LPDA antenna was measured using a network analyser. Since all dipoles in the system had the same l/d ratio and thus the same impedance level,  $Z_0$  and the stripline was matched, it should give an optimal VSWR.

The measured VSWR diagram of the LPDA is shown in Fig 6. At 800MHz the VSWR is already better than 1:1.5. It is only from 3200MHz onwards that the VSWR exceeds this value again. This



**Fig 6: Measured VSWR of the LPDA antenna from 500MHz to 4GHz.**

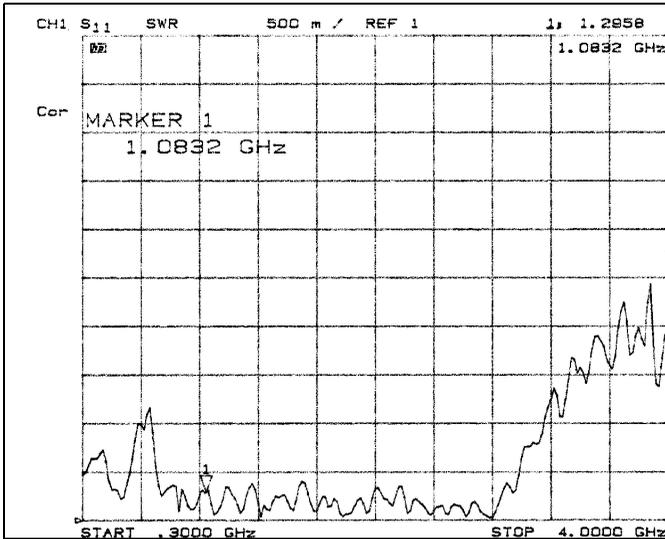
behaviour confirms the validity of the formulae used.

It can be seen that there is an apparently recurring ripple in the matching area. This ripple arises due to the matching in the individual dipoles. It has already been mentioned that the grading factor, provides evidence relating to the number of dipole elements. The higher the value of  $\gamma$ , the more elements the antenna possesses. Since each element is in resonance only for a restricted frequency range, this ripple occurs. For example, if the frequency increases, the active zone (the elements matched) migrates towards the feed point. If we select a frequency at which the corresponding dipole is just no longer matched, but the area for the best matching of the next dipole is not reached, then the VSWR increases, due to the mismatch.

It can also be seen from Fig 6 that above the operating range the VSWR increases slowly but steadily. Below the operating range it increases steeply up to some very high values. The behaviour above 3.3GHz can be explained by the fact that although there are no longer any elements in resonance, the longer elements

are now operated at double the basic frequency. Thus, for example, the half-wave dipole for 2GHz acts as a full wave dipole for 4GHz. According to Rothammel [1], this also changes the input impedance, which becomes more higher. Since the matching to the entire system now no longer applies, the VSWR increases.

Below the operating range, practically all the elements are too short for a resonant radiation. The input wave runs along the stripline and cannot be radiated due to the lack of matched elements. The wave is reflected back to the transmitter and leads to an extremely bad VSWR. Under certain circumstances, this can even lead to the destruction of the transmitter. For this reason, a load was inserted at the end of the antenna (behind the longest element). The value of the load must be calculated for the impedance level of the stripline, something that tests have confirmed. The best results were attained using an 82 $\Omega$  chip resistor. The effect of the load is impressively demonstrated in Fig 7. As was to be expected, this did not alter the behaviour of the system in any other way. You must use loads that are suitable for RF work!



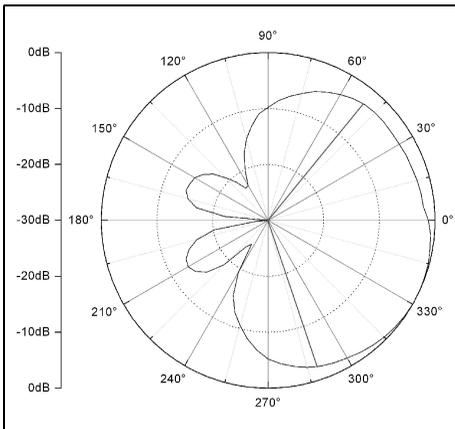
**Fig 7: Effect of load resistor used with the LPDA antenna.**

**4.1 Radiation diagrams for 1 - 3.3GHz LPDA**

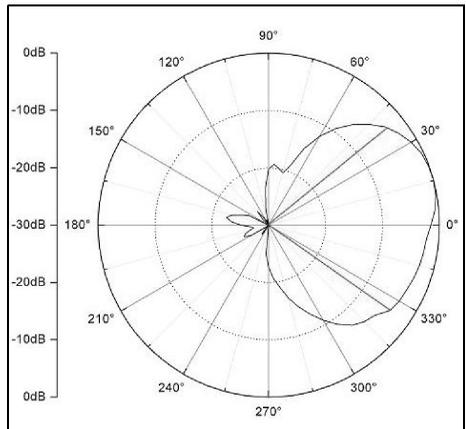
In terms of pure theory, the radiation diagrams of a logarithmic periodic antenna should not alter. But since in practice tolerances and inaccuracies frequently occur, there will also be negligible changes in the antenna diagrams. Only some of the radiation diagrams

plotted are discussed below.

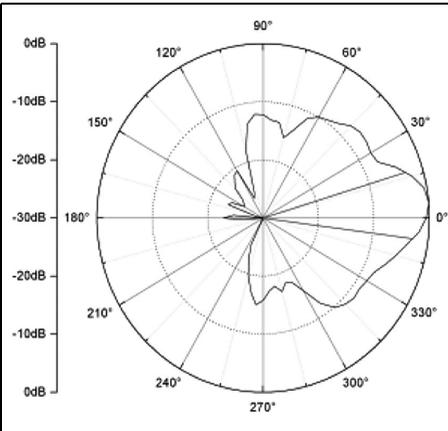
The radiation diagram of the LPDA at an operating frequency of 1GHz can be seen in Fig 8. The striking features are a double lobe in the reverse direction and a rather large apex angle. The gain of 4.9dBi was determined using the short-circuit measurement. A measurement dipole for 900MHz was used as the trans-



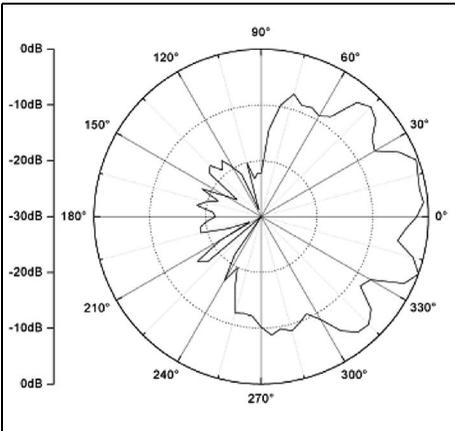
**Fig 8: Radiation diagram for the LPDA antenna at 1GHz: Apex angle 120°, Front/back 15.1dB, Front/sidelobe 13.4dB, Gain 4.9dBi.**



**Fig 9: Radiation diagram for the LPDA antenna at 1.7GHz: Apex angle 70°, Front/back 22.6dB, Front/sidelobe 19.2dB, Gain.**



**Fig 10: Radiation diagram for the LPDA antenna at 2.4GHz: Apex angle 25°, Front/back 22.8dB, Front/sidelobe 12.1dB, Gain 8.5dBi.**



**Fig 11: Radiation diagram for the LPDA antenna at 3.3GHz: Front/back 18.3dB.**

mitter antenna. The very slight gain at the lower limiting frequency can be traced back to the fact that there is no longer a reflecting element behind the last radiating element. Part of the energy is thus concealed in the two side lobes in the reverse direction.

The radiation diagram of the LPDA for an operating frequency of 1.7GHz is shown in Fig 9. The apex angle has now been reduced to approximately 70 degrees. The front-to-back ratio has likewise become considerably better and is now 22.5dB. The antenna gain could not be determined, since no suitable transmitter antenna was available. Though, because of the smaller apex angle and the improved front-to-back ratio, we can assume that the gain is markedly higher than the 4.9dBi from the previous measurement.

At 2.4GHz, the 3dB apex angle has now been markedly reduced, to 25 degrees. The front-to-back ratio has scarcely altered from 1.7GHz (Fig 10). A gain of 8.5dBi was measured using short-circuit measurement.

If we take the frequency higher, we find that numerous side lobes have formed at the top end of the operating range. We could no longer discern a main radiation direction (Fig 11). What had triggered this effect could not be explained. Though the possibility exists that the radiation diagram has become so badly marred due to reflections. But the front-to-back ratio, at 18.3dB, continues to be very good.

However, a comparison with the results of an LPDA measurement in the book "Broadband antennas" by Dubost and Zisler confirms that LPDA's have a decidedly high front-to-back ratio and that the apex angles are generally bigger than those for Yagi antennas.

## 5.

### Matching and balancing of circuits

The LPDA implemented here was calculated and assembled in such a way that



we can directly connect a  $50\Omega$  coaxial cable. Since normally a transformer is also required to feed the antenna, it is possible that the antenna will have a higher ripple value,  $s$ , in the operating range. In principle, there should be no problem in assembling a BALUN (Balanced/Unbalanced), though we decided not to do this.

Naturally the possibility also exists of feeding the antenna using a symmetrical cable. Using suitable matching, with the help of the stripline, we could also use a symmetrical  $200\Omega$  or  $240\Omega$  cable.

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## 6.

### Design Program

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The program compiled for the antenna design described above is not exactly user-friendly and, as has become clear that it contains small errors. So the little program will not be described in greater detail here or published. Our thanks go to the editors for their comments on the existing design or calculation programs, or those they have created themselves.

Those wishing to carry out their own experiments on LPDA's can, for example, insert the formulae given in the article into a table calculation.

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## 7.

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[2] Publication from D.E. Isbell relating to logarithmic-periodic dipole antennas

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[5] Broadband Antennas - Theory and Practise, by Dubost Zisler, Oldenburg-Verlag

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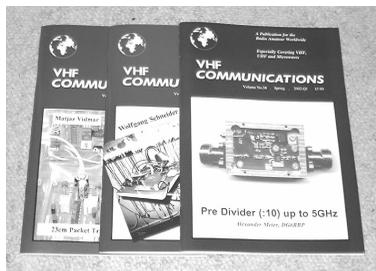
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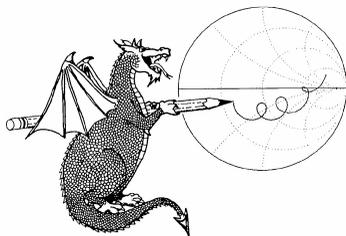


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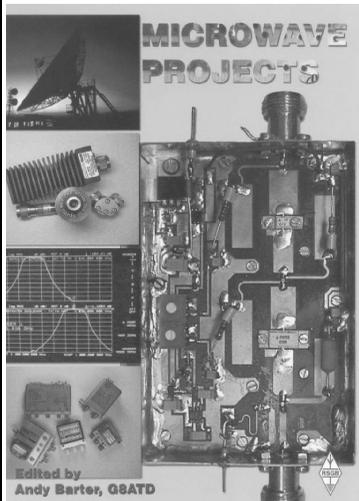
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# Microwave Projects



Microwave Projects is aimed at those who are interested in building equipment for the amateur radio microwave bands.

Packed full of ideas from around the world this book covers the subject with a variety of projects. The book has many contributors who have a wealth of experience in this area and they have produced many projects, design ideas, complete designs and modifications of commercial equipment, for the book.

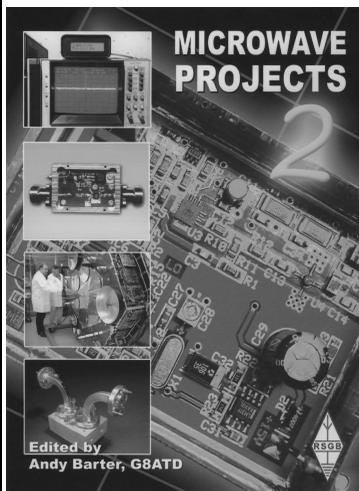
This title provides much useful information as to what can be achieved effectively and economically. Aimed at both the relative novice and the "old hand" the book also covers useful theory of designing microwave circuit and test equipment for the projects. The book includes chapters covering:

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- Transverters and transmitters
- Receive amplifiers
- Power amplifiers
- Filters
- Miscellaneous, including rare articles on using YIG oscillators and microwave absorbers.

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**If the kit or PCB is not in this list please contact K. M. Publications**

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DG6RBP	12GHz Divide by 1000 prescaler	4/03		£ 160.00
DG6RBP	12GHz Divide by 10 prescaler	1/04		£ 160.00
DG6RBP	Universal measuring amplifier for low dc voltages	2/05		£ 130.00
DG6RBP	Universal PLL oscillator	3/05		£ 118.00
DG6RBP	ATV transmitter with PLL for 10GHz	4/05		£ 193.00

KIT	DESCRIPTION	ISSUE	No.	PRICE
DJ8ES-019	Transverter 144/28MHz	4/93	06385	£ 120.00
DJ8ES-019/50	Transverter 50/28MHz	2/95	06392	£ 120.00
DJ8ES-047	Log Amplifier up to 500MHz with AD8307	2/00	06571	£ 42.00
DG6RBP-002	Pre Divider (:10) up to 5GHz	1/02		£ 115.00
DB6NT-Rotor	Simple Speed Control for Rotators	2/02	06533	£ 35.00

PCB	DESCRIPTION	ISSUE	No.	PRICE
DJ8ES-019	Transverter 144/28MHz or 50/28MHz	4/93	06384	£ 10.00
S53MV	Set of PCBs for Matjaz Vidmar Spectrum Analyser	4/98-4/99	S53MV	£ 65.00
DJ8ES-047	Log Amplifier up to 500MHz using AD8307	2/00	06569	£ 6.50
DG6RBP-002	Pre Divider (:10) up to 5GHz	1/02		£ 18.00
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DJ8ES-054	Frequency Generator to 4GHz - oscillator	1/02		£ 9.50
DJ8ES-054	Frequency Generator to 4GHz - microcontroller	1/02		£ 15.50
DB6NT-Rotor	Simple Speed Control for Rotators	2/02	06532	£ 5.50
DB6RBP	ATV Squelch	4/02	06542	£ 3.50
DG4RBF	Set of 4 PCBs for FC4000 Frequency counter	1/00		£ 20.00
DG6RBP	Universal measuring amplifier for low dc voltages	2/05		£ 16.50

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DG4RBF	EPROM for FC4000 Frequency counter	1/00		£ 10.00

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**Minimum shipping charge £5.00**

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