

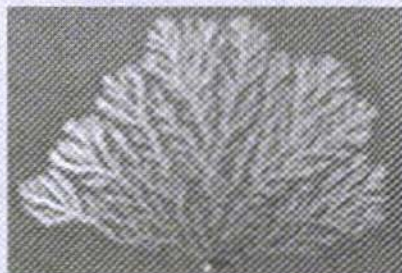
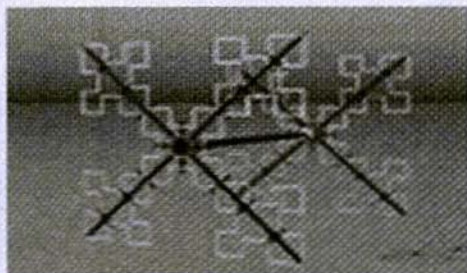


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VHF COMMUNICATIONS

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The Fractal Antenna

Angel Vilaseca, HB9SLV

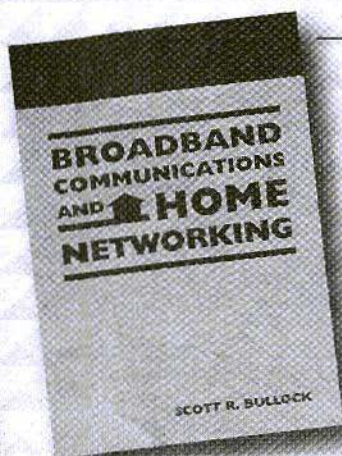
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Apologies for some omissions in issue 3/2001, I was using a new all electronic method for production. Hopefully the teething problems are sorted out for this issue.

A lot of interesting articles again, in particular the article from Joop van Sundert, PD1APO on the practical use of the YGO3 antenna optimiser discussed in issue 3/2001. If there are any other readers who have an article based on a subject covered in VHF Communications I am always pleased to publish them.

I have received information from The California Institute Of Technology that they are no longer supplying PUFF. It is, however, bundled with two books available from Amazon.com:

Complete Wireless Design by Cotter W Sayre, current price \$69.00

The Electronics of Radio by David B Rutledge, current price \$44.95

73s Andy



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Wolfgang Schneider, DJ8ES

70 MHz Preamplifier for Frequency Counters

Many frequency counter modules, both industrial products and D.I.Y. equipment, require a "respectable" TTL level at the input. A suitable preamplifier must therefore be used for level adjustment to enable practical frequency measurements.

The preamplifier described here offers a high degree of input sensitivity in the frequency range from 1 Hz to 70 MHz and supplies the standard level for TTL modules at the output, with an input impedance of more than 1 M Ω .

1. Circuit description

The input impedance of the amplifier is very high, due to the FET J310 (T1) in the input stage, being more than 1 M Ω . The two back to back 1N4148 diodes protect the input from saturation. The BF506 transistor (T2) is the first amplifier stage. The circuit diagram can be seen in Fig. 1.

Most of the signal gain is provided by a Motorola line receiver - type MC10H116 (IC1). It is usually used to process weak signals in digital communications and is specified for frequencies up to 100 MHz and above.

The following BF506 (T3) transistor acts as a switching stage. With appropriate

design it could provide TTL levels directly at the output for low frequencies (< 10 MHz), but not for a frequency up to 70 MHz! The problem is to find appropriate switching transistors with a correspondingly high transit frequency for processing TTL signals. Transistors are required with a limiting frequency of approximately 10 to 20 times the operating frequency.

As the measurement frequency rises, the amplitude of the square wave signal at T3 falls. The LT1016(IC3) comparator switches between low and high levels controlled with its threshold voltage which is set with a R15 (22 k high-precision potentiometer). The desired TTL level is thus guaranteed over the entire frequency range from 1 Hz right up to 70 MHz.

2. Assembly instructions

The amplifier is assembled on a two-sided copper-coated epoxy printed circuit board with the dimensions 40 mm x 80 mm. The layout can be seen in Fig. 2. The components side of the board is also the earth surface.

The board is first drilled, using a 0.8 mm drill for the components and a 1 mm drill for the soldering studs. The mounting

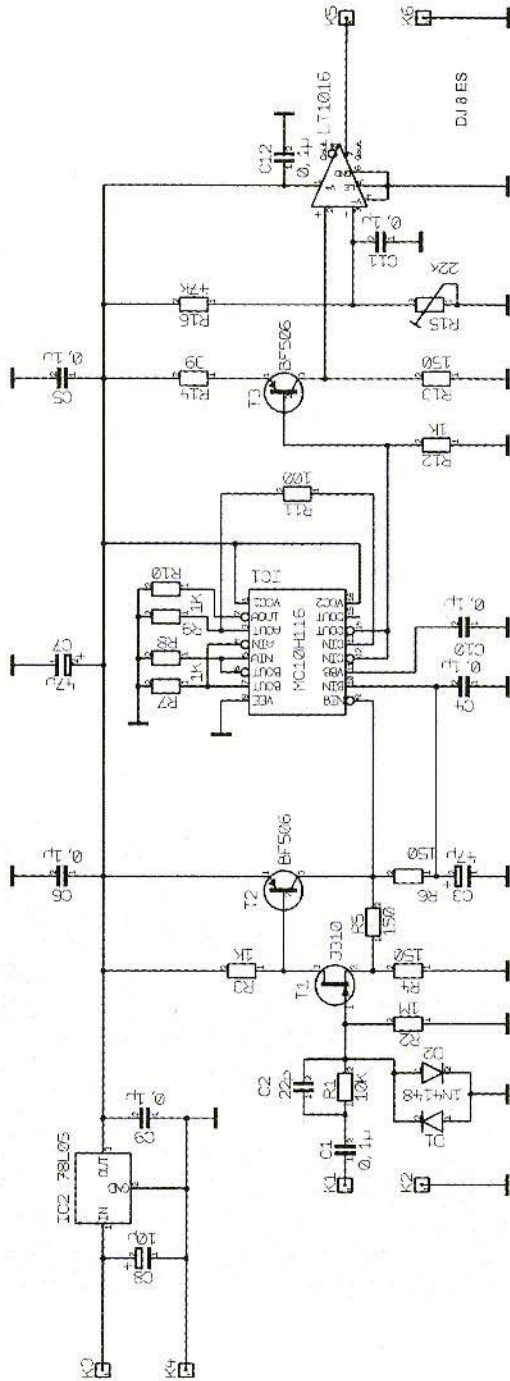


Fig 1: Circuit Diagram of Preamplifier

holes should have a diameter of 3 mm. In the second step, all holes for connections, excluding earth connections, should be counter sunk using a 2.5 mm drill.

The printed circuit board is populated, in no particular order, in accordance with the components drawing in Fig. 3, including down to the two integrated circuits MC10H116 (IC1) and LT1016 (IC3), and the components are soldered into position. Earth connections are soldered on both sides here, so that good earthing is achieved.

The two remaining ICs should not be inserted until the operating voltage of +5 V has been checked at the output of the 78L05 voltage regulator (IC2).

3. Components list

- 1 x MC10H116, line receiver IC, Motorola
- 1 x IC LT1016, comparator IC, Linear Techn.

- 1 x 78L05, voltage regulator IC
- 1 x J310, FET
- 2 x BF506, transistor
- 2 x 1N4148, diode
- 1 x electrolytic capacitor, 10 μ F/ 25 V, RM 2,54 mm
- 2 x electrolytic capacitor, 47 μ F/ 16 V, RM 2,54 mm
- 1 x high-precision potentiometer, 22 k Ω , RM 2,54 mm
- 1 x printed circuit board, DJ8ES 051
- 5 x soldering stud, 1 mm
- 1 x soldering lug, 3 mm

Resistors, 1/8 watt, RM 10 mm:

- 1 x 39 Ω
- 1 x 100 Ω
- 4 x 150 Ω
- 6 x 1 k Ω
- 1 x 10 k Ω
- 1 x 47 k Ω
- 1 x 1 M Ω

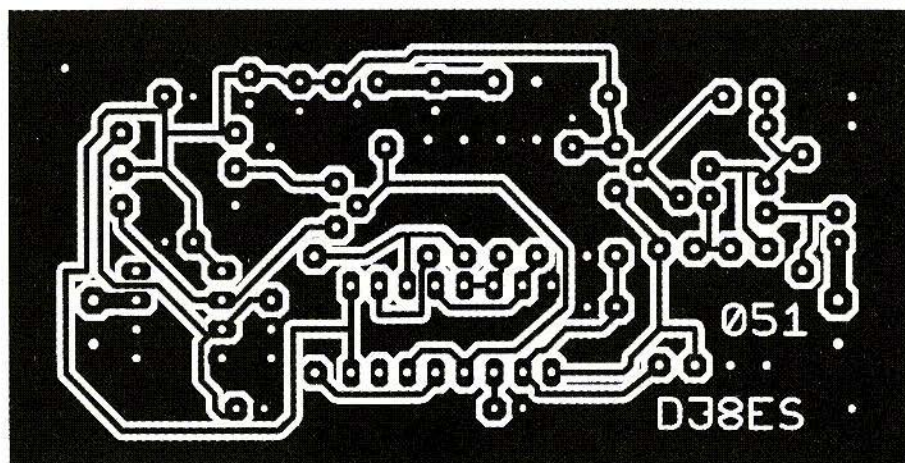


Fig 2: PCB Layout DJ8ES-051

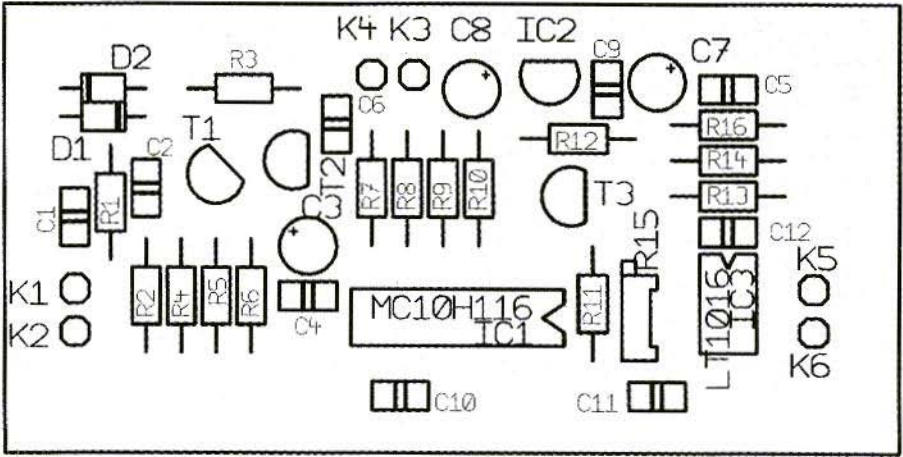


Fig 3: Component Layout

Ceramic capacitors, RM 2,54 mm:

- 1 x 22 pF
- 8 x 100 nF

4.

Putting into operation

Once the supply voltage has been applied, the amplifier is ready for operation.

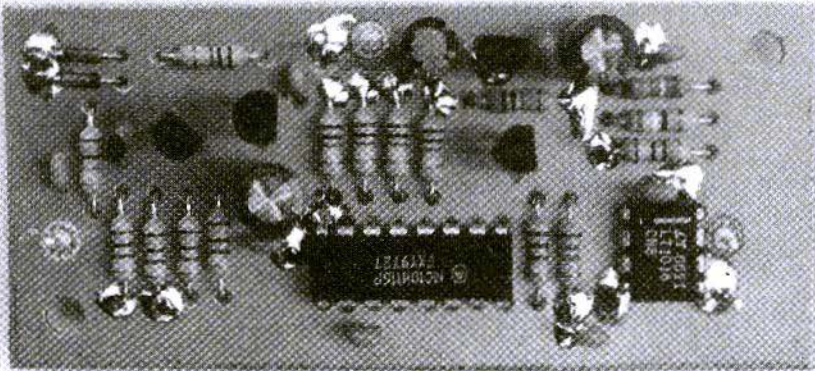


Fig 4: Photograph of Completed 70MHz Preamplifier

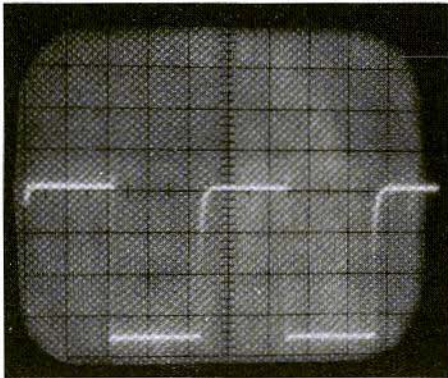


Fig 5: Square Wave Signal Output at TTL Levels

The current consumption is approximately 75 mA at an operational voltage of +12 V.

To test the equipment first set the switching threshold of the LT1016 comparator

(IC3) using the R15 high-precision potentiometer to the mean value of the voltage level at the output of T3. Then testing is simply carried by connecting to a frequency counter set at maximum input sensitivity and at maximum measurement frequency. If you do not have a suitable standard signal generator, this can also be done, with the help of an oscilloscope using a lower frequency.

The amplifier (Fig. 4) operates over a broad frequency range, from less than 1 Hz right up to 70 MHz. The output always offers a square wave signal with a TTL level. This can be seen using an oscilloscope with an appropriate band width (Fig. 5). The input sensitivity is high impedance over the frequency range ($>1\text{ M}\Omega$) right up to 30 mV. Only close to the upper or lower limiting frequency does the sensitivity fall back somewhat.

In conclusion, I would like to express my heartfelt thanks to OM Frank Peter Richter (DL5HAT) for his active support.



Gunthard Kraus, DG8GB

An Interesting Program, TRL85.exe for Synthesis and Analysis of Microwave Problems

1. Brief description

Name: TRL85.exe

Area of application:

Precise synthesis and analysis of

- Microstrips
- Edge coupled microstrips
- Striplines
- Edge coupled Striplines
- Coaxial Cables

File type and file size:

TRL85.EXE-File with Windows user interface and additional ONLINE help. File size approximately 2 Mb

No problems were detected during operation with WIN 95/98/ME and with a fast computer.

Bugs or serious computing errors detected:

Left-hand edge of results report is often displaced too far to the left on screen and can no longer be brought back.

User friendliness:

The program is easy to understand and it is relatively easy to familiarise yourself with it. Uses metric or Imperial units as desired. Very practical structure.

Online help:

Yes, high-quality, provided the results are within the scope.

Procurement options:

Download from <http://www.ansoft.com>

2. Description of TRL85

The ANSOFT company has prepared a free, limited student version of its very expensive and comprehensive microwave CAD program "Serenade" for downloading from the Internet.

Address: <http://www.ansoft.com>.

It comes accompanied by the TRL85 stripline calculator, the "TRL85.exe-file" which can be copied into another directory at any time once Serenade has been installed, together with the online help, or can be transferred to another computer.

TRL85 is based on the same computing algorithms as PUFF or PCAAD, but is markedly more powerful.

Apart from the well-known WINDOWS tools, you will appreciate, for example, the option which allows you to take into

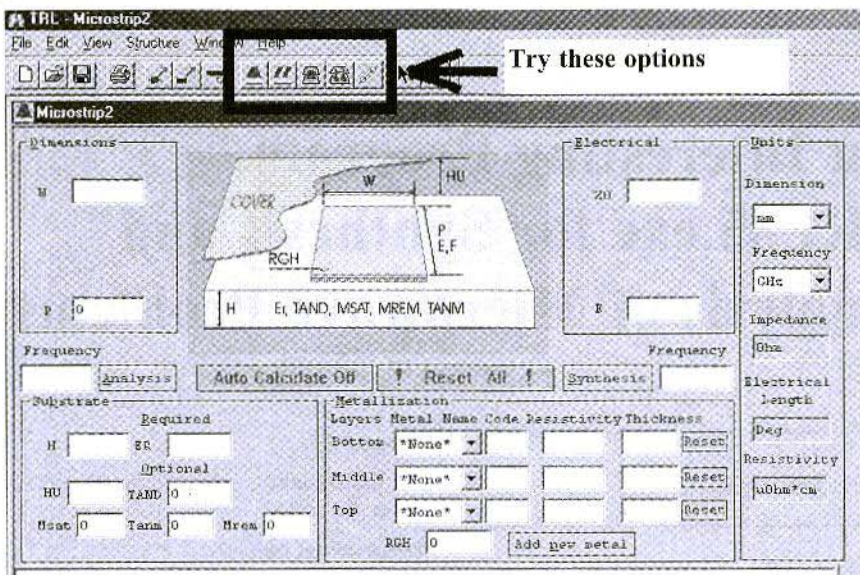


Fig 1: TRL5 screen display showing Stripline Calculation option

account the incorporation of the circuit into a housing using "distance to cover". In addition, the relationships between losses and frequencies are carefully and separately logged, broken down into power losses and insulator losses, etc..

2.1. Working with TRL5

The rest of this article assumes that a TRL5 directory is now available on the computer in use.

For that, you can set up a start button for the TRL5 program on the WINDOWS screen or simply move to the directory for TRL5 and click on the file "TRL5.exe".

On the screen then displayed (Fig. 1) you should first click on each of the symbols marked there in succession. In this way you can see the basic structure of the various analysable modes of conductivity for yourself, together with the screen

menus offered.

The areas covered are:

- Microstrip
- Edge coupled Microstrip
- Stripline
- Edge coupled Stripline
- Coaxial cable

The following examples can be used to familiarise yourself with the program on a step-by-step basis.

2.2. Typical application:

Analysis and synthesis of a 50Ω microstrip supply line

For the well-known printed circuit board material Rogers RO 4003 with the following data:

- $\epsilon_r (\epsilon_r) = 3.38$
- Board thickness $H = 0.813$ mm

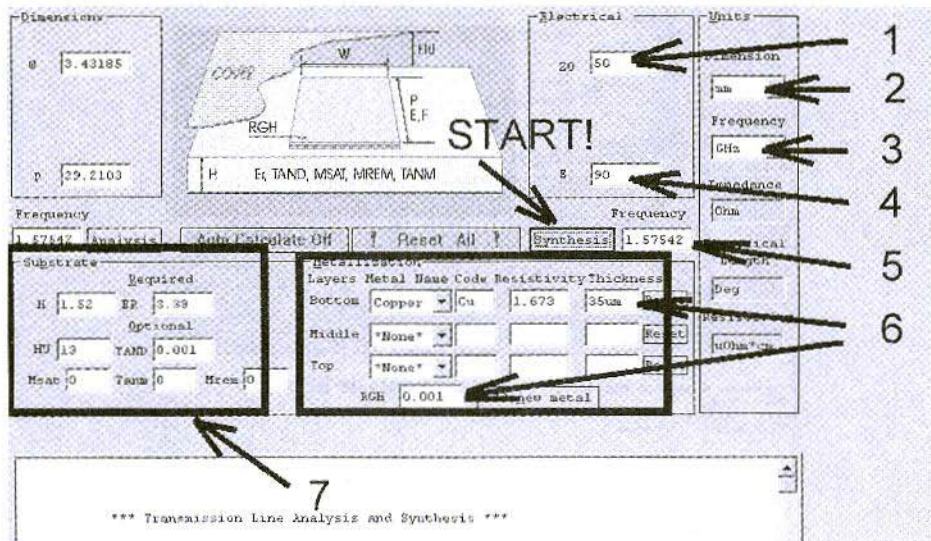


Fig 2: Steps for Microstrip simulation

- Copper coating on both sides 35 μm
- Dielectric loss factor TAND = 0.001 at 1.6 GHz
- Screening plate 13 mm above the board

A quarter-wave version is to be investigated, with an electrical length of 90 degrees at the GPS frequency $f = 1575.42$ MHz.

We therefore select the “Microstrip” option button and we first examine the screen in detail:

In the right-hand half of the menu is the “Synthesis” button. This is used if a specific impedance level and a specific electrical circuit length (in degrees) have been selected, at the operating frequency required. As a result, the physical (= mechanical) length and width of the circuit are obtained for the board data entered at the bottom left-hand corner and the values for the track material entered at the bottom right-hand corner.

In the left-hand half of the menu is the “Analysis” button. This can be used to specify (from the mechanical dimensions) the electrical characteristics (the impedance level, the electrical length in degrees, the losses in dB / mm, even separated into dielectric and conductor losses, etc.).

The following steps should now be carefully carried out one after another (Fig. 2):

- (1) Please enter a 50 Ω impedance level here.
- (2) Is the dimensional unit the mm?
- (3) Is the frequency already being measured in GHz?
- (4) Let the electrical length be 90 (the associated dimensional unit has already been set to degrees).
- (5) Enter frequency 1.57542 (GHz) correctly.
- (6) The specifications for the track material [copper / thickness = 0.035 (mm) / RGH = roughness = 5 μm] must go in this field. Under

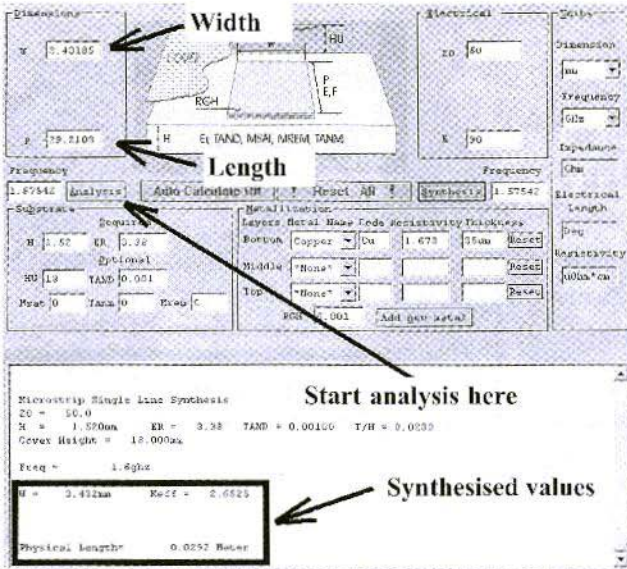


Fig 3: Results of Synthesis on Microstrip Line

“Bottom” select the option “Copper” and for “RGH” enter the proposed 5 μm .

(7) And here finally all data concerning the printed circuit board (thickness = 0.813 (mm) / ER = 3.38 / cover distance from board = 13 (mm) / TAND = 0.001).

When everything has been done, press the “Synthesis” button and the corresponding result appears on the screen (Fig. 3).

In the top left-hand corner of the menu, the two boxes for W (= width) and P (= physical length) are suddenly filled with the calculated values.

Underneath the menu area, in a large separate field, a results report is displayed. In the marked frame, here too we have the values for the length and width of the stripline, which are:

- Width = 1.83 mm
- Length = 29.24 mm
- for $Z = 50\Omega$

Now press the “Analysis” button again and see what new options are offered:

In the menu not much has changed: it is just that the rounding errors can be seen for the impedance level and the electrical length, under “Reverse Calculation”.

In the report field (Fig. 4) on the other hand, we now also have the attenuation value at 1.6 GHz, and this divided into “D” (= dielectric loss), “C” (= conductor loss) and “T” (= total loss).

If desired, TRL85 saves all results in a separate file this is an additional option you are given when leaving the program.

2.3. Typical application:

Coupled line pair

With bandpass filters made from coupled line pairs, once the filter design is complete we obtain the values for the EVEN and ODD resistances. They must first be converted into mechanical data before the printed circuit board is designed. “PUFF”

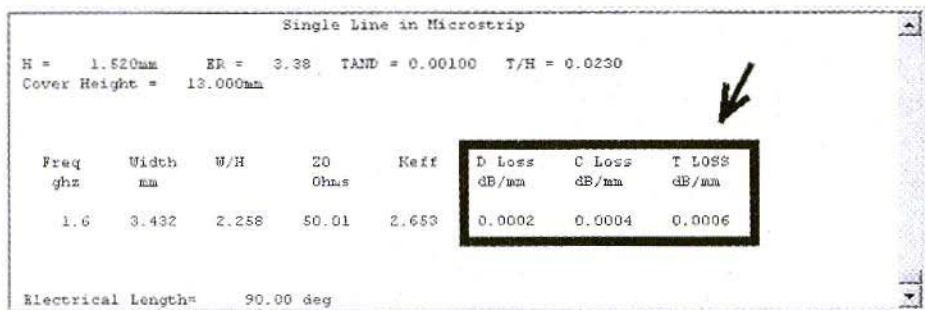


Fig 4: Analysis Report for Microstrip Line Showing Attenuation Values

can certainly do this too, but with "TRL85" we can also just additionally calculate the incorporation into the screening housing with the cover distance of, for example, 13 mm.

The following data are assumed for the line pair in such a filter:

- ZEVEN = 65Ω
- ZODD = 40.8Ω
- Electrical length = 90 degrees

The same printed circuit board made of

RO4003 is used again and the data to be entered for it are:

- Relative permittivity, ER = 3.38 / Printed circuit board thickness, H = 0.813 mm /
- Copper coating on both sides 35 μm / Dielectric loss factor TAND = 0.001 at 1.6 GHz / Screening plate 13 mm above printed circuit board.

Now use the second button from the left

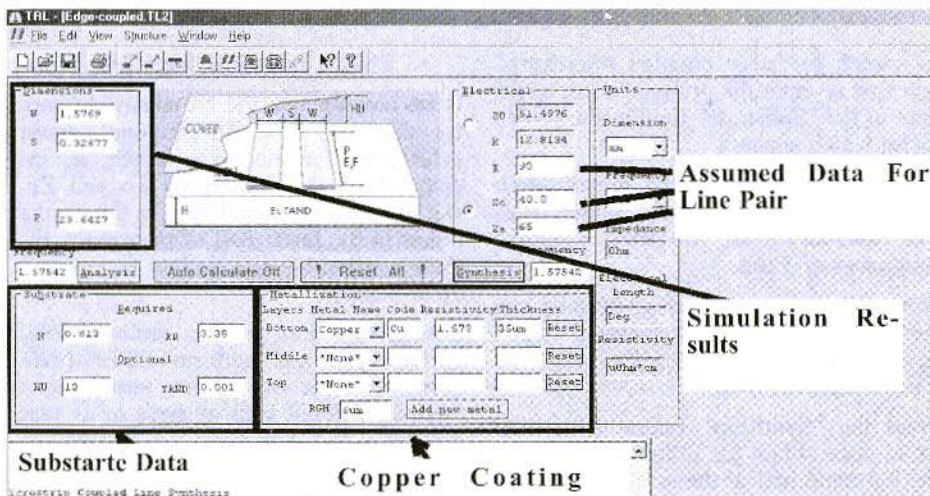


Fig 5: Data Setup For Simulation Of Coupled Line Pair

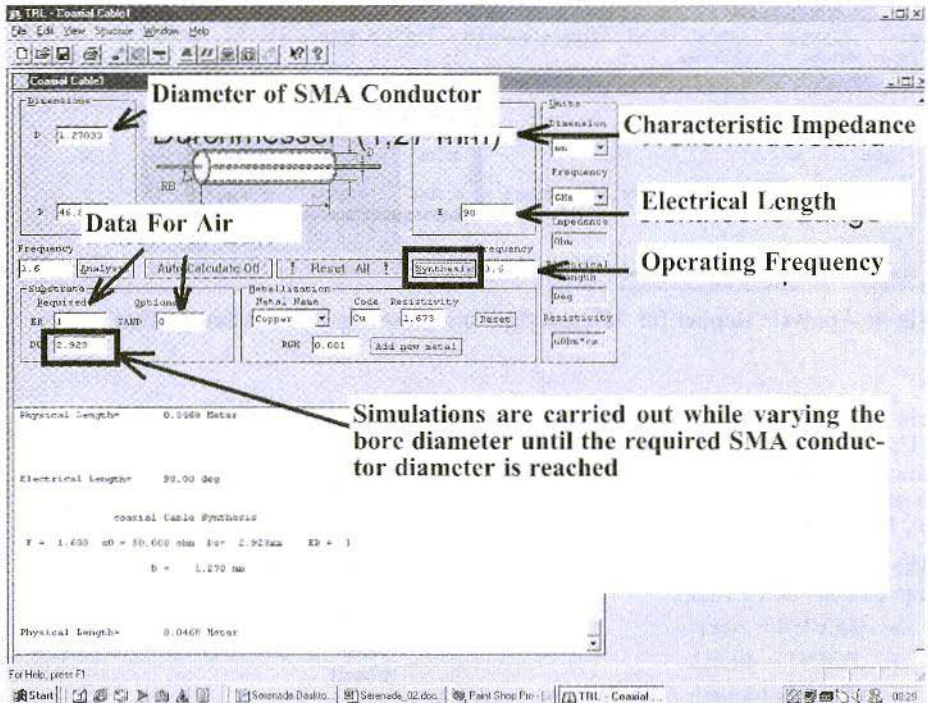


Fig 6: Using TRL85 for Simulating a Coaxial Line Problem

to switch to “edge coupled microstrip” and first re-enter the printed circuit board data in the “Substrate” field in the screen menu which appears.

We also need the data on metallisation again. So we select “copper” under “Bottom” and for “RGH” we select a surface roughness of $5 \mu\text{m}$.

Now things get interesting, because we can now finally enter the electrical length, E , the ODD resistance, Z_o , and the EVEN resistance, Z_e .

Press the “Synthesis” button once, and we at once obtain the following values in the left-hand half of the screen (Fig. 5):

- Conductor width $W = 1.58 \text{ mm}$
- Interaction gap $S = 0.33 \text{ mm}$

- Phys. circuit length $P = 29.65 \text{ mm}$

If we now operate the “Analysis” button, we can see how TRL85 once again determines the electrical length, E , together with the resistances Z_o and Z_e . But the most important thing is the results report in the lower half of the screen, for there we find the precise loss factors and other details.

But please don't forget: unfortunately, TRL85 can not perform OPEN-END corrections for us on line terminations working without load we have to do that ourselves and determine the shortenings required using, for example, the well-known diagrams in the “PUFF Manual”!



2.4. Typical application:

Coaxial line

Here we are not trying to go over the data for a coaxial cable but to solve the following problem:

The internal conductor of an SMA flanged bush with a diameter of 1.27 mm is fed through the wall of an aluminium housing to the printed circuit board. This housing wall is 3 mm thick and the continuous bore should be selected in such a way that the impedance level, even within the wall, is 50Ω . What drill diameter should be selected? The solution is shown in Fig. 6.

Call up "TRL85", select the "Coaxial cable" menu and enter an impedance level of 50 (Ohms), an assumed electrical length of 90 (degrees), an operating frequency of 1.6 (GHz) and the data for air ($ER = 1$, $TAND = 0$). The program then also needs the external diameter and calculates the internal diameter for this. Here we can start with any value as the external diameter. Finally, press the "Synthesis" button.

We alter this external diameter and con-

tinue to create simulations until the internal diameter reaches the pre-set value of 1.27 mm. For this, we need a bore with a diameter of 2.9 mm in the housing wall.

3.

Summary and assessment

The "TRL 85" stripline calculator provides an outstanding supplement to simulation using "PUFF" and is indispensable once the housing has to be included as well.

In other respects, "PUFF" and "TRL85" supply practically identical results. Its ease of operation makes it a joy to use, and unfortunately this is not the only way in which "PUFF" is put in the shade. There is also the direct modelling of all data (impedance level, splitting loss fractions, etc.) at the pre-set design frequency, together with the WINDOWS interface.

Note :

You can look forward to further presentations of interesting software here.



Joop van Sundert PD1APO

Optimising yagi antennas with YGO3 Genetic optimiser

1.

Introduction

For 8 years I have been involved in the design and optimisation of Yagi antennas. I have used programs such as Yagicad and YOC for the design of such antennas, but the results did not give me a satisfied feeling as to the performance; the theoretical results appeared to be very promising, but in practice the results were not as good as one might expect. Lately I have been using the NEC2D code, and since then results have improved. Personally I had some problems with the NEC input files, simply because I was so used to enter data according to the requirements of Yagicad and YOC. In this phase of the project I was assisted by Dr. Gerry van Dooren PA3EWA, who has professional experience in the use of NEC.

2.

Yagi design and optimisation

I am using the following procedure to design and optimise a Yagi antenna. The first order design is obtained using Yagimax 3.11. The latter program also has an optimiser which is used to obtain the rough geometry of the yagi antenna.

Since the user interface of this program is very good (compliments to the author of Yagimax!), it is quite easy to obtain a reasonably performing yagi antenna.

After the optimisation in Yagimax, the numerical design is saved, and this geometrical data is converted into a NEC input file. This program was supplied by PA3EWA, and was specifically written for this purpose. This input file is used as input to NEC2D-500k, after which I look at the results, and apply some minor manual corrections in order to obtain a real impedance such that the run time in YGO is (hopefully) reduced. The comparison of results given by Yagimax and NEC2D reveals that for the identical design some considerable differences can be obtained. As an example, I have applied the above procedure to a 23 element yagi antenna. The data Yagimax supplied was a gain of 18.5 dBi, front-to-back ratio (F/B) of 38 dB, and an impedance Z of $(49+j1.8)$. The corresponding results obtained when using NEC2D were: gain = 17.9 dBi, F/B = 21 dB, and $Z=(56+j4.5)$.

I have built antennas of various design and by mounting them at the same height on the mast, I could make good comparisons of gain, F/B, side lobe level and radiation pattern. I have done comparisons for around 10 antennas, and the preliminary conclusion for these is that the results as predicted by NEC2D come

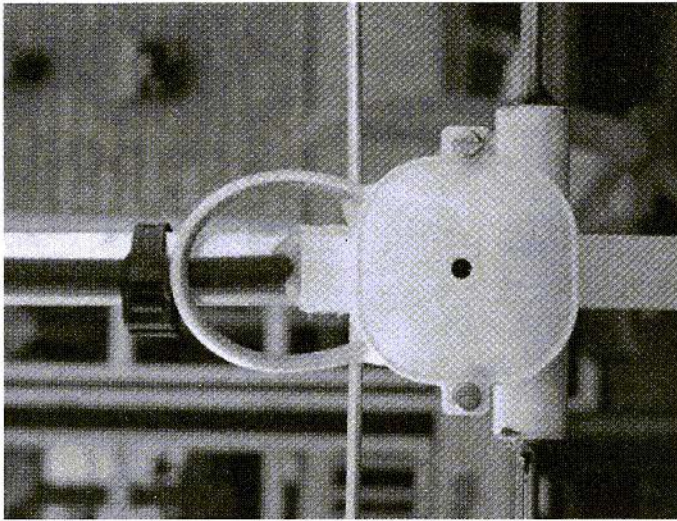


Fig 1: Top view of Balun

closest to the data as measured in practice.

3.

Further optimisation of the yagi antennas

Recently, I have started working with the program YGO3, that I obtained from the unofficial NEC archive as maintained by Ray Anderson. This code uses genetic algorithms to optimise antenna geometry's for NEC. It produces NEC input files, runs NEC and optimises the NEC input data based on the results from a previous run. The basic principle is that the fittest antenna, i.e. the antenna with the optimal performance will survive.

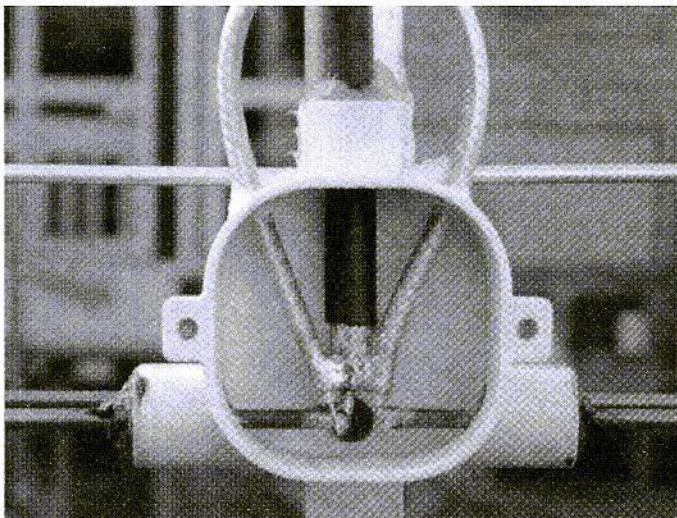


Fig 2: Internal view of Balun

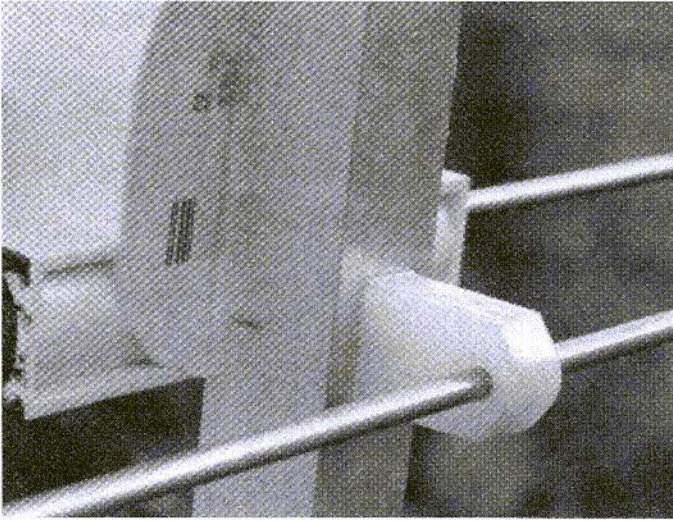


Fig 3 : View of Dipole Fixing

Please refer to dedicated antenna journals for details on the theory of this method. After some e-mail correspondence with Ray to solve some problems with the seed files, I managed to get the program running. The theoretical results as obtained appear to be very promising. The disadvantage being that it can take a *very* long time to obtain an optimum result.

To begin with, I started using YGO with

a seed file directly. This appeared to be quite cumbersome, and I decided to try something else. Now I write the seed file myself using data from Yagimax and NEC2D. After having defined the seed file normalised to the wave length, the number of runs in YGO is limited, because the design obtained using Yagimax, with its data then converted into a basic NEC input, is already a fairly acceptable design. To view the designs, it appears to

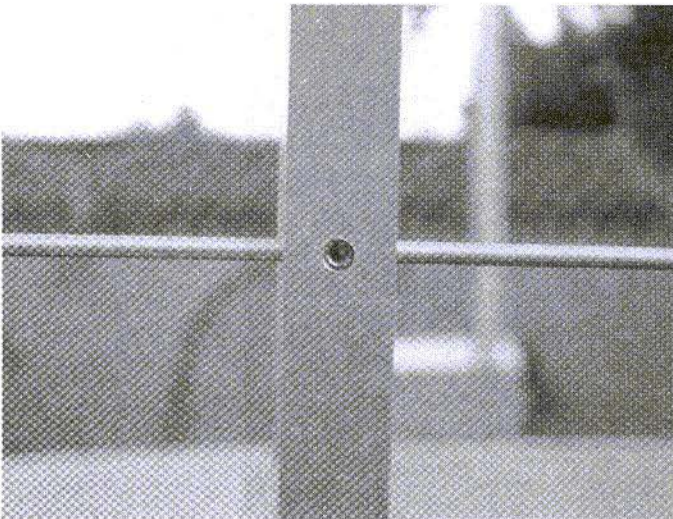


Fig 4: View of Element fixing

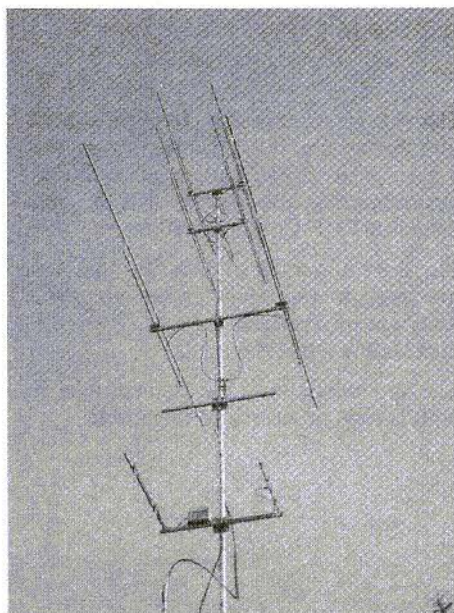


Fig 5: Picture of the aerials in Position

be favourable to enter a seed file name in YGO3, choose a short run. Note that I normally put foms to 000.00. Also make sure to start from generation 1!

Some further info on manually altering input files is the following. After you have looked at a plot in YGO3, you can open the file YGO1.1 and you can make changes as desired. After doing this and saving the file, one can immediately view the changes by starting YGO PLOT. After this, you open the file GEO1.1. At the bottom of this file, the geometry in wavelengths is shown. By putting this geometry into the seed file, one obtains a manually altered seed file, and it is unnecessary to make long runs. In this way I have designed and optimised various antennas. I also added the configuration file used, but there remains the need to experiment with several runs to obtain the result as desired. By opening e.g. GEO1.80 you can compare the geometry of the antenna with the file 25el_7.dat (added). The geometry in GEO1.80 is the

optimised design with starting point 25el_7.dat. Another file, 26el_4.dat is a design employing 2 reflectors, that at the moment is used in a horizontal stacked array of two yagis. See the photographs and the figure, where the pattern for the 25 element design is shown with a solid line, and the design with the 26 element is in dashed pattern.

4.

Practical antenna building

I have added seed files 25el_7.dat and 26el_4.dat. Both of these files have been optimised with YGO3. Both designs have been built and tested by undersigned. The 25el_7 is a design for 432.1MHz, while the 26el_4 is a design for 486.5MHz. Obviously, other frequencies can be used to design antennas for. When applying frequency scaling it is very important to pay attention to the director and reflector element radii, because in the frequency scaling step, e.g. to 1296MHz, the radii obtained can give rise to dimensions with inferior and impractical mechanical characteristics. Typically I try to use element diameters of 5mm. This causes the antenna to appear very slender, but yet is mechanically strong.

When I build the antennas, I have mounted the elements *through* the boom, in galvanic contact with it. I have added various photographs. The balun is made from RG142U with a transformation ratio of 4:1 because a folded dipole is used as radiating element. In the theoretical design an element correction of 11mm was added, depending on frequency used and diameter of the boom.

5.

Note From Editor

Conclusions

I have explained how YGO is used using Yagimax and NEC2D. Some results have been shown, and personally I have the opinion that the comparison between theory and practice is very reasonable.

I wish all amateur designers good luck with using YGO3. If there are any tips and/or suggestions with respect to YGO, I am happy to hear these.

73 de Joop PDI APO in The Netherlands
e-mail: PD1.APO@12move.nl

Thank you very much to Joop for submitting this article. It is always very interesting to hear from readers who use the information in the articles that are published in VHF Communications.

There are a number of data files that were too large to reproduce in the magazine. They are all available from the web site - <http://www.vhfcomm.co.uk>

The appendix shows the element data for both aerials described in the article.

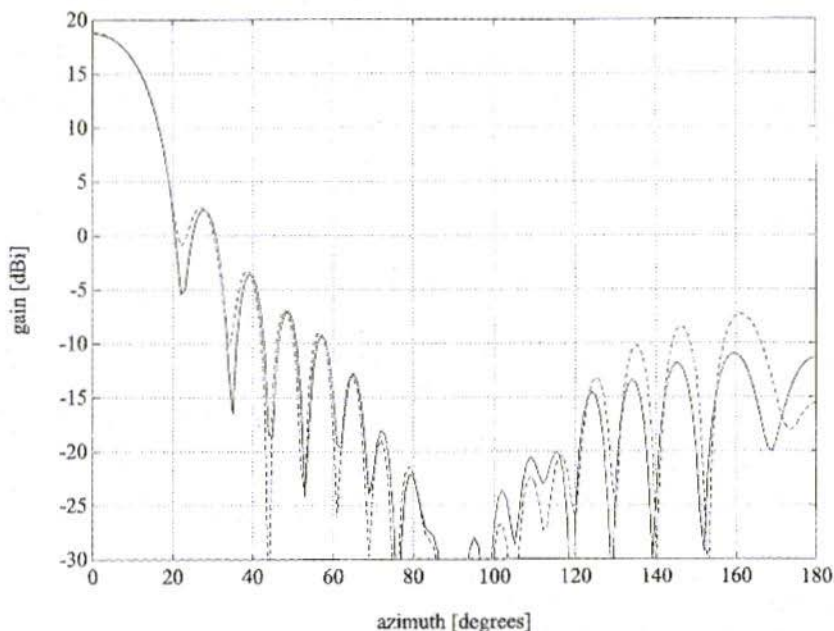


Fig 6: Graph showing the Gain Plot of The Antennas



Appendix 1

Data For 25 Element Aerial

Run ID: 05-17-2001, 19:16:01

Figure of Merit = 18.13, best chromo in gen #40

Coefficients: a= 150 b= 10 c= 10 d= 10 e= 10 f= 10

This line and the next two are reserved for comments.

All three comment lines must be present.

Do not erase any of the header or comments lines!

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Appendix 1

Data For 26 Element Aerial

Run ID: 05-17-2001, 19:16:01

Figure of Merit = 18.13, best chromo in gen #40

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This line and the next two are reserved for comments.

All three comment lines must be present.

Do not erase any of the header or comments lines!

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Angel Vilaseca, HB9SLV

The Fractal Antenna

A Revolutionary Idea

Fractals are mathematical functions. This concept is normally connected with computer-generated images. Fractals became extremely popular in the eighties for two reasons: the first factor was the meteoric development of micro-informatics and the rapid increase in the computational power of micro-computers! The second contributing factor was the work of the world-renowned French mathematician, Benoît Mandelbrot, who had made his name working in the field of fractals.

1. Order and chaos

Fractals were originally just a mathematical concept. Since then, they have become a tool which allows splendid images to be generated by computers. But that's not all they also offer a general concept of the universe, a unifying principle of science, thanks to Benoît Mandelbrot and numerous other researchers.

Fractals are linked to chaos theory. In 1988, a book was published called "CHAOS" by James Gleick, and became very popular. It includes several astonishing propositions which have since become common knowledge; like the one about the butterfly in China which flaps its wings and, through a chain of chaotic

events, subsequently causes a tornado in the Caribbean.

Fractals combine order and disorder in a unique way! If we look attentively at one of the illustrations in Fig. 1 above, we can see that they are disordered within a small frame and yet ordered on a large scale. A fractal can be chaotic, as in these diagrams, or deterministic. "Deterministic" here means that they are composed on the basis of a single pattern (for example a triangle or a rectangle), which is called the generator, and that they are based on successively changing scales (which are called "repetitions"). The number of repetitions can be infinite. You need at least two of them to be able to speak of self-similarity. Examples of fractals are the Koch curve, Fig. 2 on the left, the Sierpinski triangle, or the "cylinder head gasket" developed by Mandelbrot, Fig. 2 below.

The term "fractal" was coined by Benoît Mandelbrot. It means that it can be proven that fractal images have no complete dimensions i.e. they are incomplete. This does not happen in Euclidian geometry! In classical geometry, a point is defined as having no dimensions. A line has one dimension, a plane has two and finally a volume has three dimensions. So far so good. It can be demonstrated that a given fractal, for example

$\log 4 / \log 3 = 1.2618$ dimensions
or again

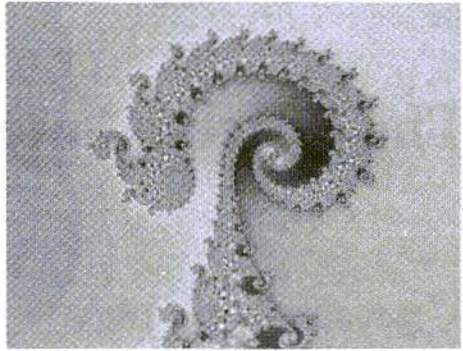


Fig 1: Two examples of Fractals

$\log 2 / \log 3 = 0.6309$ dimensions! This point can be understood intuitively. If, for example, we take a Sierpinski triangle, this is an area, so it has two dimensions. For the first repetition, a triangle is removed from the whole and here again we find once more three smaller triangles, but these also still have two dimensions. If we repeat this procedure in stages, the areas become smaller and smaller. After an infinite number of repetitions, the triangle will be made up of an innumerable number of areas which will be infinitely small. The entire area will be approaching zero, but nevertheless the Sierpinski triangle appears to possess an area. It can be mathematically demonstrated that the number of dimensions is less than two.

2.

From theory to practise

The subject of this article is the application of the theory of fractals to the design of antennas. The objective is not simply to discuss a modern subject but to demonstrate that there really is some potential here. Researchers have already investigated this path extensively and have established that the theory of fractals makes it possible to create much more

powerful antennas (with regard to the size, radiation efficiency, antenna efficiency and band width or slight side lobes). There are some antennas which can just not be designed without fractals.

Fractal antennas are a new area of development only 4 or 5 years old; for this reason, there are very few descriptions to be found. Fig. 3 shows two industrial developments which have already been produced.

This area of development looks very promising against the background of an enormous expansion in wireless communication. The developers are working away at it intensively and protecting their own developments through patents. However, little by little it is becoming possible to get at the secrets of even this new type of antenna. This article is already a beginning.

3.

Characteristics of a fractal

To obtain some idea of what a fractal is, try and answer the following question: how long is the coast of Corsica? That seems to be an easy question. All you need to do is find a map of Corsica and pick up a ruler - right? Wrong!

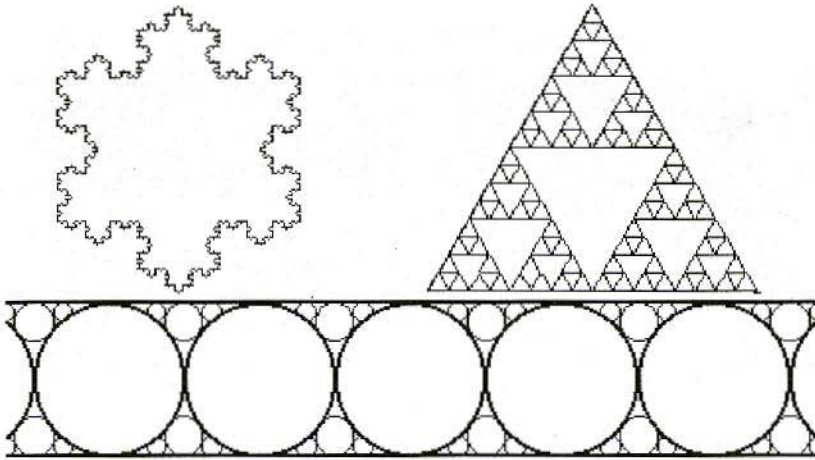


Fig 2 : Examples of the Koch Curve, the Sierpinski Triangle and the Cylinder Head Gasket developed by Mandelbrot

You take a large-scale map and a pair of compasses. Working from the scale of the map, set the points, e.g. to 50 km and then measure, approximately, the length of the Corsican coastline. Now repeat the process once more, but this time set the compass points to 5 km. Since the coastline is very jagged and consists of numerous promontories and bays, the value now obtained will be greater than the first result. If we use more and more detailed maps and smaller measurement steps, the value at each stage will be greater than the preceding value. We would thus finally arrive, in theory, at an

infinite length.

It could be said that this theoretical contour of Corsica is a fractal diagram; it gives us a coastline with an infinite number of indentations. They all look the same, irrespective of whether they extend 1 km, or 10 or 100 km. Incidentally, the same also applies to a cloud. There are also computer programs which apply fractal theory to the synthetic generation of landscapes. And take it from me, I can confidently assert that these synthetic landscapes (Fig. 4) look more "realistic" than reality!

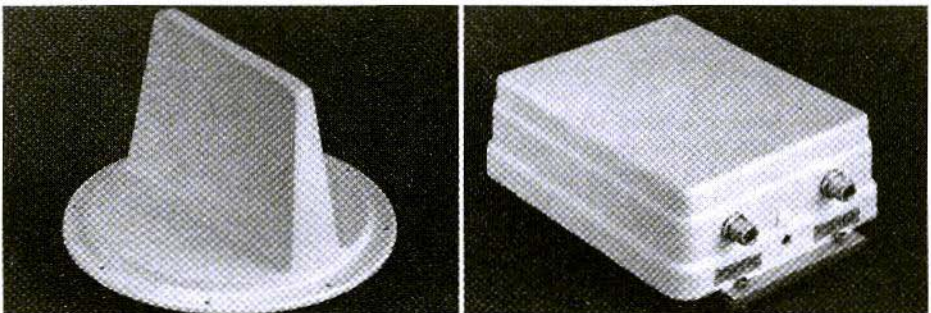


Fig 3 : Examples of industrial developments of Fractal Antennas, from Sistemas Radiantes S.A.



Vanilpporant.com

El Lago Paris

R./vanderlippe

Fig 4: A Fractal Landscape, produced with software called "Bryce" from R. Vanderlippe

And so we come to a characteristic which can be extremely valuable for an antenna it always behaves the same way with waves of varying wavelengths, irrespective of whether the wavelength is 10 cm, 1 m or 10 m. In other words, a wide band antenna.

If we use a dipole which has a fractal form instead of the standard rectilinear form, we can establish that it is made up of a large number of sections, each of different length, and that each will radiate in a certain part of the spectrum as well as possible (Fig. 5).

On the other hand, everyone knows that the shortest distance between two points is a straight line. But it has now likewise been recognised that the longest distance is the fractal line! In this way, we obtain curved radiating elements which result in very much more compact antennas with less losses and a higher degree of efficiency. Fig. 6

Another possible way of using the theory of fractals is the log periodic antenna; it is well-known for its extensive band width. In a log periodic antenna, the theory of fractals is applied, not just to the individual elements but to the entire architecture of the antenna. Although it had already been invented many years

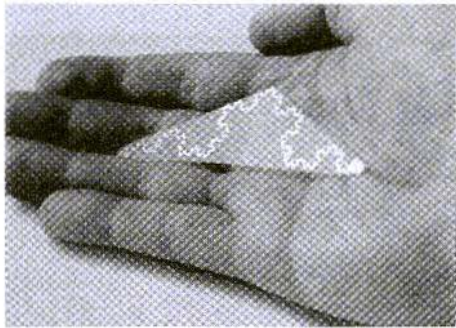


Fig 5: A Fractal dipole developed using the Koch curve by Nathan Cohen - NIIR

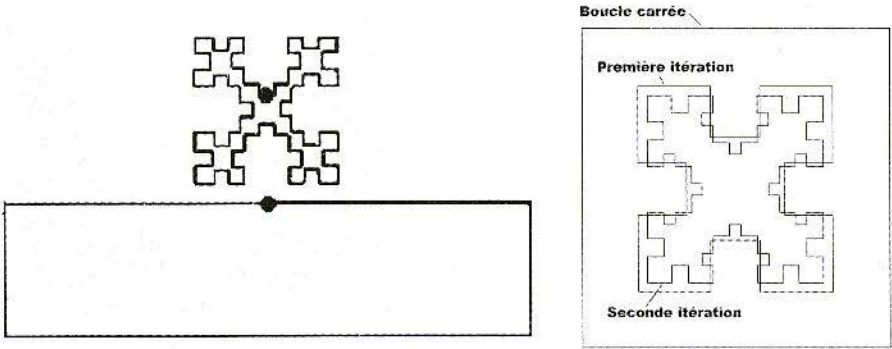


Fig 6: The development of the Fractal form of a rectangle. The right hand side shows the iteration process.

ago, before there was any talk of fractals, it can be recognised that it has one of their essential characteristics: self-similarity. This is one characteristic property which immediately strikes anyone who sees a fractal image. What this observer sees is one element repeated infinitely, from the tiniest detail right up to the overall view. Like the marmelade jar on which there is a label with a photo of a little girl holding a marmelade jar, on which there is a photo of a little girl - etc..

It can be demonstrated mathematically

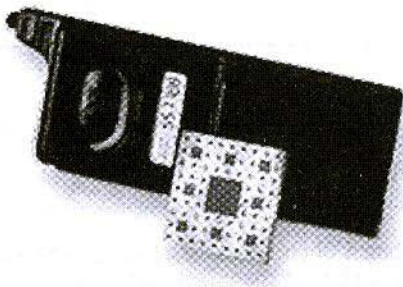


Fig 7: A Fractal antenna developed for a mobile phone by Nathan Cohen - NIIR

that, for an antenna to have good broadband functioning, it must have a point of symmetry and it must be self-similar. I.e., its appearance must always be the same, irrespective of the scale. In other words, it must obey the law of fractals.

Various well-known fractal representations, such as the Sierpinski triangle, have been put forward as antennas. The common point of all these structures is their good behaviour in relation to their reactivity, which usually means we can do without any matching. This implies a simplification of design, greater reliability and lower losses.

It is possible to create radiating fractal structures for the UHF range as printed circuits. In the ultra-short wave range as well, thanks to the good ratio of the wavelength to the mechanical length with this approach, printed circuit structures can be used (Fig. 7).

There are structures drafted in one piece, but there are also antennas consisting of a metallic plane, in which holes have been stamped in accordance with fractal distribution.

The concept of fractal antennas can be applied to individual antennas or to antenna arrays. In the case of arrays, the

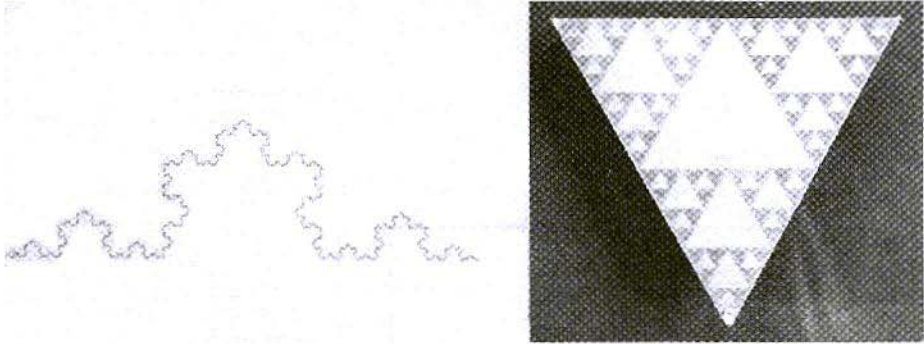


Fig 8: Right - a Koch curve, Left - A Sierpinski triangle developed as a printed circuit by The Carles Puente Polytechnic at Barcelona University

distribution of elements can be regular, as in certain radar antennas, or irregular, as in certain radio telescope arrays, which were built by linking together already existing instruments, which are distributed all over the world.

A fractal array can combine the robustness of an irregular array with the efficiency of a regular array, and this with only a quarter of the elements.

In the case of individual antennas we can, as an example, take the Koch curve or the Sierpinski triangle (Fig. 8).

These representations, e.g. of the coiled shape, combine inductances and capacitances, which removes the need for any matching circuit and extends the transmission band, in that it improves the gain. Antennas have already been put on the market which can advantageously replace the rubber antennas of portable equipment. Some have been created using small printed circuits which are integrated into the housing of a mobile phone.

It is true that fractal antennas allow miniaturisation, but we should not go beyond a reduction in size by a factor of 2 to 4. Otherwise there is a danger of sacrificing the yield. And, just as with all other antennas, it will naturally not be possible to have small dimensions, large

band widths and high gain simultaneously. But fractal antennas can come closer to this than other types and thoroughly successful compromises can be obtained between these three characteristics.

Any type of antenna can be fractalised: monopole, dipole, helix, printed antennas, etc.. The elements are formed in accordance with a fractal representation, or gaps are introduced into the antenna.

In the case of flat antennas such as printed circuits, which are made up of several elements, the coupling between them is generally the main factor which restricts their power i.e. gain and their impedance matching. This problem is avoided in the case of the fractal antennas.

4. Experimental fractalisation of a loop

Nathan Cohen, N11R, had the following experience: he manufactured a frame antenna with an edge length of 15 centimetres, in the form of a rectangular printed circuit. He then manufactured three other fractalised antennas with the same dimen-

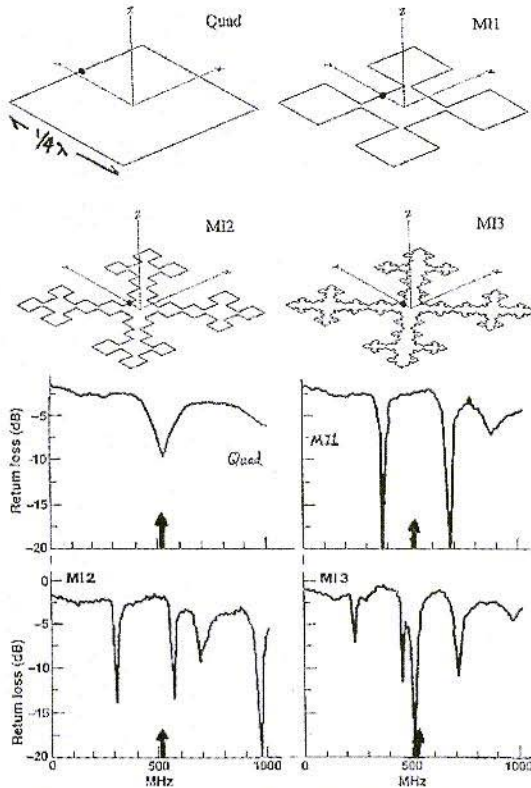


Fig 9: Experimental results on a Quad Loop by Nathan Cohen - NIIR

sions, but each time adding a repetition in accordance with the Koch curve. Naturally the size of the loop went up each time.

Compare this with the problem of the Corsican coast which we mentioned earlier. He then measured the reflected power (proportional to the SWR) of each of these antennas. This told him that the frequencies at which the SWR was lowest (which means that the impedance here was approximately 50 Ohms) kept changing, and that the more peaks there were on the frequency axis the lower these were (Fig. 9).

For antennas with three repetitions, for example, we have a usable peak at one-third of the frequency of the non-fractal-

ised loop. It should be borne in mind that the peaks are not harmonics of the start frequency. It can be determined subsequently that the peaks also correspond to zones in which the impedance is real i.e. not reactive. If the radiation resistance is at 50 Ohms and is not reactive, that means that no additional matching is necessary.

5. Fractalisation of a dipole

In this example we are not dealing with experiments but with the results which were calculated by Nathan Cohen, NIIR,

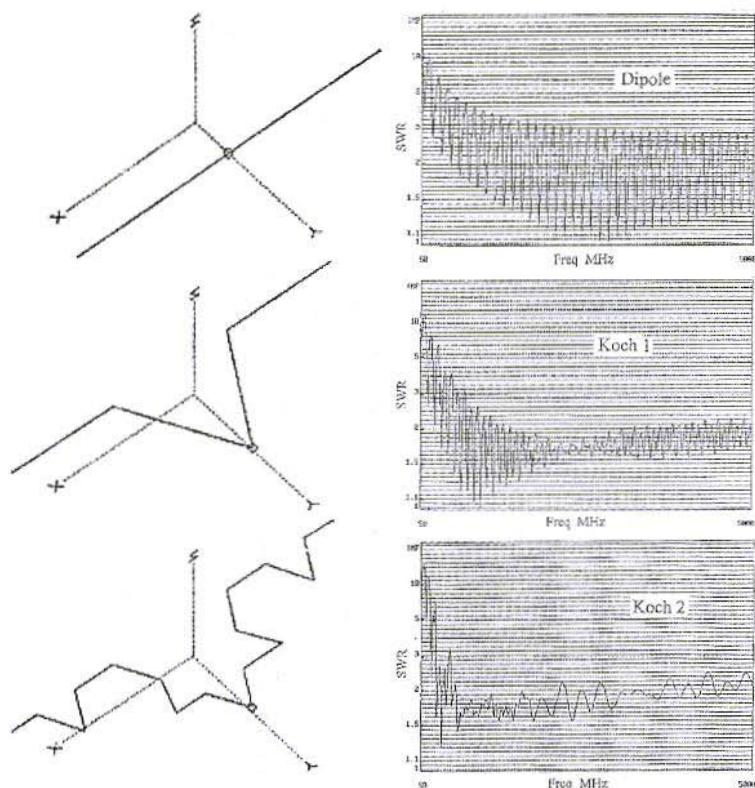


Fig 10: Fractal Dipoles by Nathan Cohen - NIIR

with the help of “NEC4” software (Fig. 10).

Here, instead of a loop, we take a dipole dimensioned for 65 MHz, to which we apply two repetitions of the Koch pattern in succession. But this time we sweep a much bigger range, going right up to 5 GHz. With a rectilinear dipole we now measure a host of points with a low SWR for all harmonics of 65 MHz.

On the other hand, the fractalised dipoles demonstrate expansion and lower troughs. The optimal impedance i.e., for which the trough is weakest lies around 350 Ohms.

Thus we can see that fractalisation can be used to create antennas with an extremely

wide transmission band.

6. Practical Applications

A practical application of the fractal theory has been put forward by Nathan Cohen, NIIR: a quad with two elements for the 10-m-Band. Fig. 11

It measures 1.5 m x 1.5 m without any efficiency losses as against the unfractalised version. It does not need a matching circuit, and its impedance, Z , is 50 Ohms. It can thus be energised directly through a coaxial cable. Fig 12.

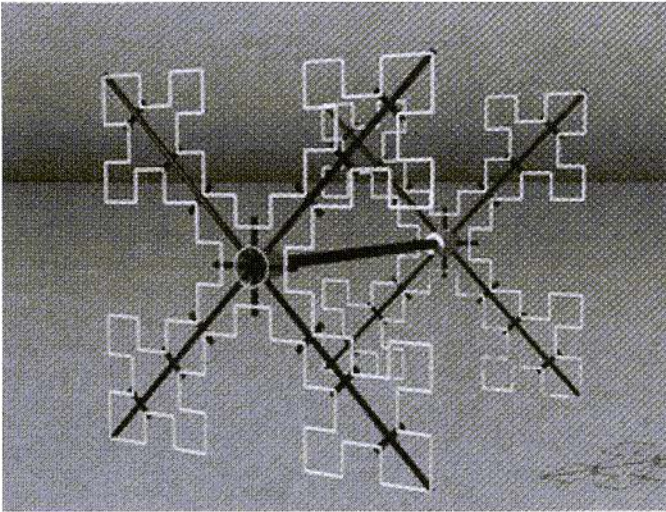


Fig 11: A Two Element Quad for the 10 m band by Nathan Cohen - N1IR

Using this antenna, radio contact could be made with European locations at 1 Watt and with the Pacific region at 2 Watts. A further version was measured for the 2-m band, with a gain of 4 dB and a front-to-back ratio of at least 15 dB. The 2-m version has a transmission band width of 500 kHz with an SWR below 2.

Fig. 14 shows a general view of the antennas. The two elements have the same dimensions. They can be manufactured using copper wire (1.5 mm. or more). The simplest method is to use a template on a wooden board on which nails are knocked in at the bending points, 1 to 26, in accordance with the pattern, and the copper wire can thus be very precisely bent. Each of the four sides of an element is bent on the template in succession in this way.

Fig. 13 shows a quad element for 10 m with corresponding dimensional specifications. The framework of the quad can easily be manufactured using PVC tubing. The same material can be used for the transverse struts. The typical impedance of the antenna can be increased if the length of the middle transverse struts is increased.

Instead of the coax cable, a coil with two

turns is connected to the distributing point of the reflector, so that its frequency of resonance can be decreased by app. 600 kHz. The distance between the active element and the reflector is 170 cm.

Like all fractal antennas, this antenna is also resonant in more than one band. N1IR detected resonances at 52, 97, 125 and 141 MHz. At 125 MHz, the front-to-back ratio is 10 dB and the gain is likewise 10 dB, which is a lot for an antenna which is contained in a cube with a wavelength of only 0.6.

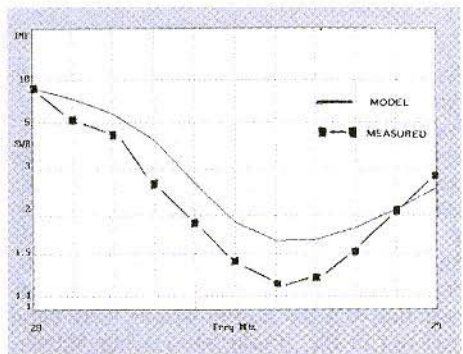


Fig 12: The measured and theoretical SWR for the Two Element Quad

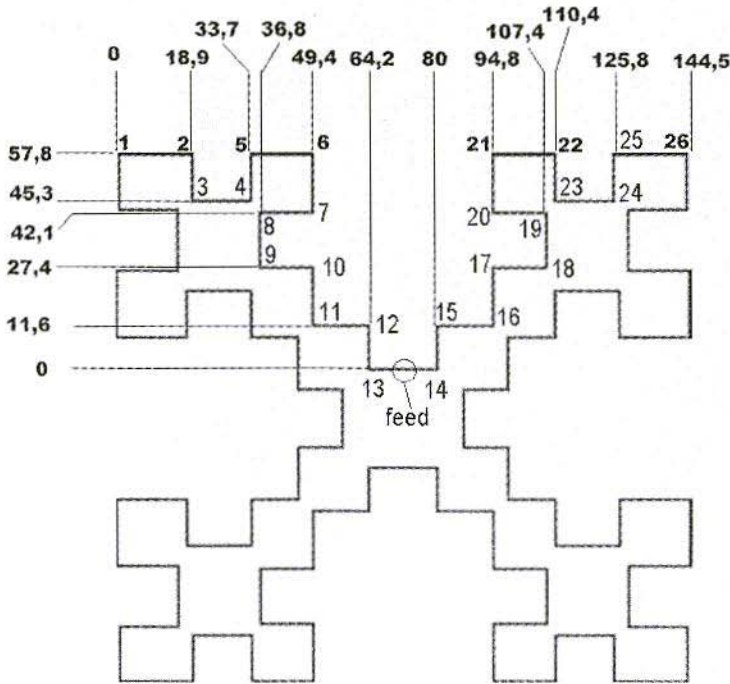


Fig 13: Dimensions for the Two Element Quad

6.1. Two fractal antennas for the 70-cm band

Another practical implementation for the American 70-cm band is proposed by Richard Kutter of the University of Dayton. To demonstrate the validity of the concept of the fractal antenna, he first examined a dipole and an antenna proved a tempting control. He then compared them with two fractalised loops, one with one repetition, the other with two. The radiation and SWR simulations were carried out in the 70-cm band using MININEC Pro. To ensure that the fractalised antennas remained in the desired band,

their dimensions were matched. He simulated the properties of two quad antennas with two, then with three loops (Fig 15 and 16). Unfortunately he gives no details of the precise dimensions of his antennas.

6.2. Fractal antennas for microwaves

The higher the frequency involved, the fewer details we are given by authors. It is, in fact, in the ranges of mobile telephony and microwaves that the applications are most interesting from the commercial point of view...

Here are some examples of ultra-high frequency antennas, on which numerous



Fig 14: A picture of the Two Element Quad antenna for the 10 m band

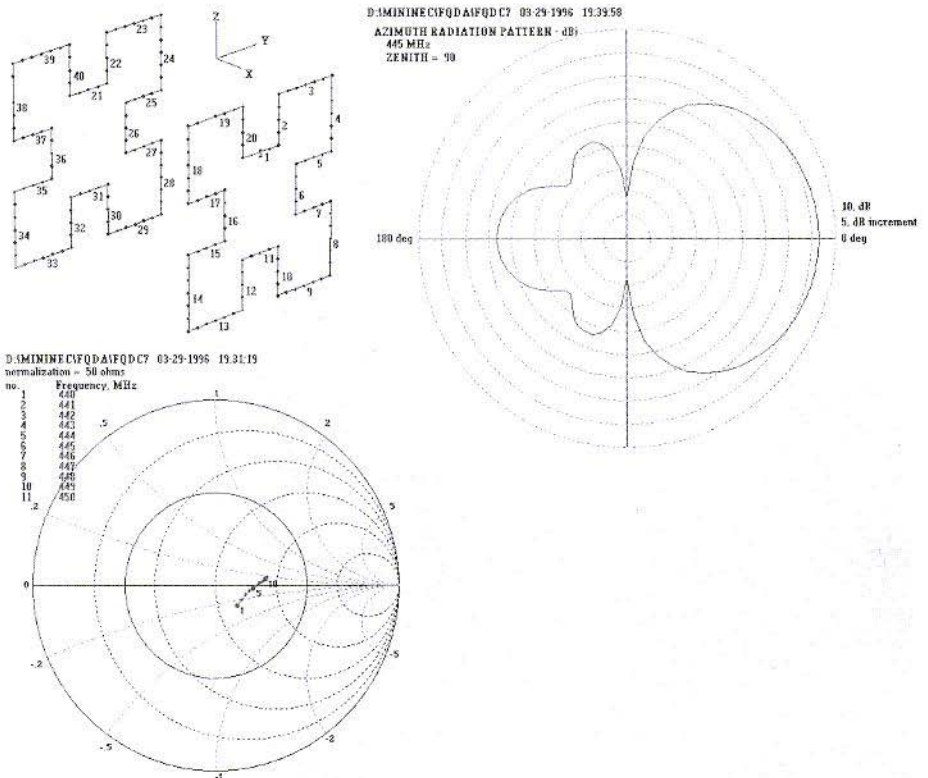
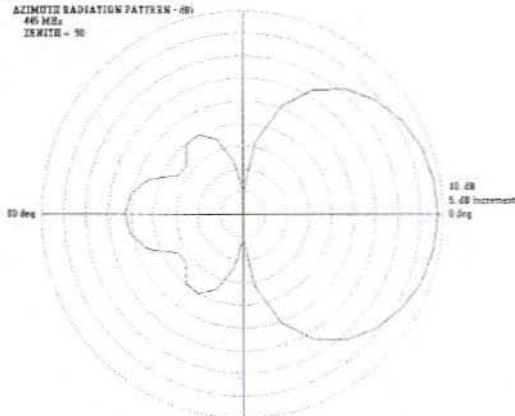
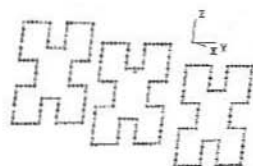


Fig 15: Simulation of a Two Element Quad for the 70 cm band



D:\MINKI\CF3QD\AIF3QDD 04-11-1996 09:10:43
 AZIMUTH RADIATION PATTERN - dB
 440 MHz
 ZSMITH = 50



D:\MINKI\CF3QD\AIF3QDD 04-11-1996 09:13:29
 normalization = 50 dBm

Frequency, MHz
 1 240
 2 241
 3 242
 4 243
 5 244
 6 245
 7 246
 8 247
 9 248
 10 249
 11 250

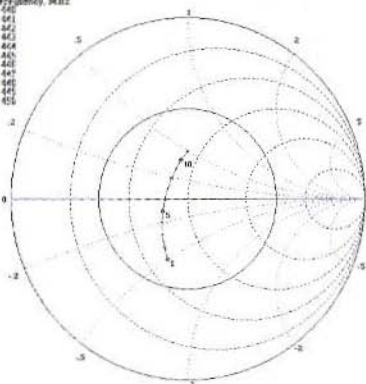


Fig 16: Simulation of a Two Element Quad for the 70 cm band

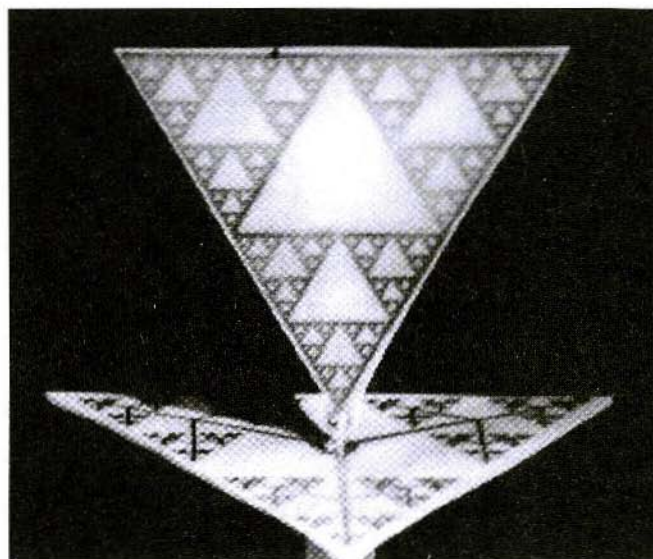


Fig 17: The Sierpinski Triangle used at microwave frequencies

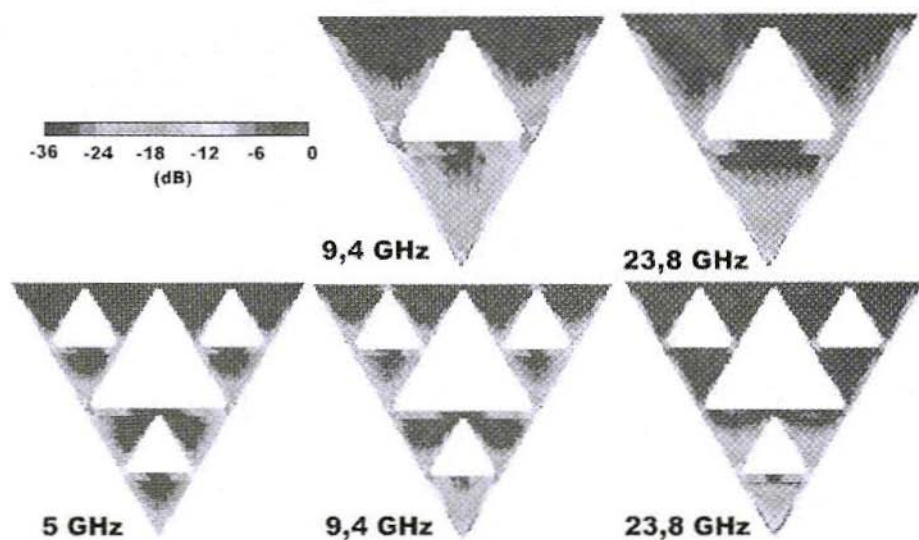


Fig 18: Examples of antennas scaled for various microwave frequencies

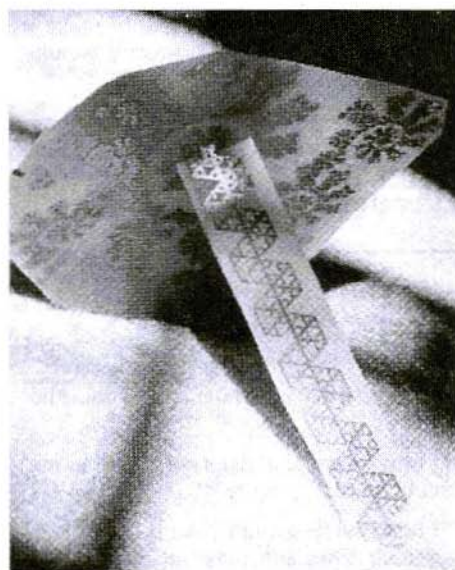


Fig 19: A wideband antenna developed by Nathan Cohen - NIIR



Fig 20: A Fractal version of the helix antenna by Nathan Cohen - NIIR

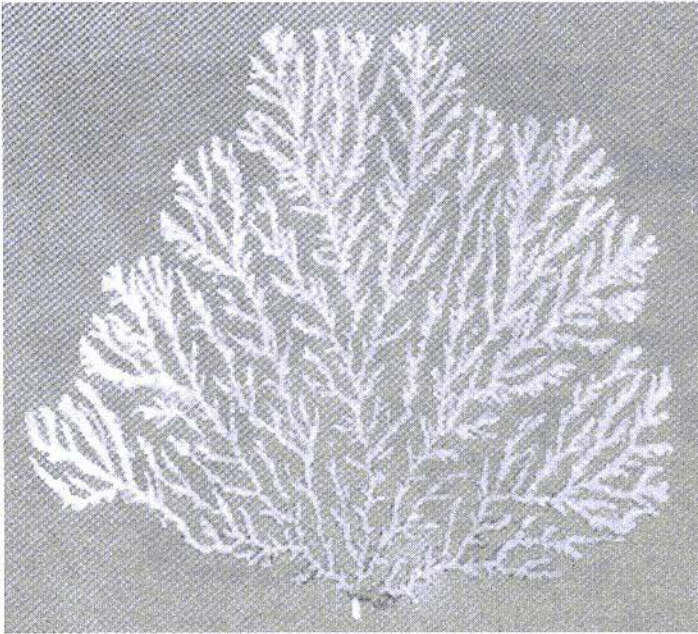


Fig 21: My Favourite, a beam antenna developed by The Carles Puente Polytechnic at Barcelona University

teams are currently at work all over the world. But, who knows? Perhaps, once again, another important discovery will emerge from the amateur radio community?

6.3. The Sierpinski triangle

One structure often used in ultra-high frequency engineering is the Sierpinski triangle. It is powered through one of the corners, in that it is connected up to the internal conductor of a coaxial line. The screening is connected to the earth surface, which can be both a closed surface and also a fractal figure. The behaviour of such an antenna can be analysed with the help of software based on the moments method, such as, for example, EmSight. The program has already been introduced here. The said method makes it possible to calculate the intensity of the currents which prevail in the conductors.

The point of optimal matching for patch

antennas is generally empirically determined. However, a start has been made on using mathematical models to develop simulations, which will certainly make it possible to create more powerful simulations in the coming years.

7. Literature

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- [1] "CHAOS" by Gleick, 1987, ISBN 0 7493 8606 1
 - [2] <http://fractenna.com/>
 - [3] <http://www.crhc.uiuc.edu/~kutter/the-sis/>
 - [4] <http://www.antlab.ee.ucla.edu/~johng/fractal.html>
 - [5] http://www-tsc.upc.es/ccf/research_lines/antennas/fractals/fractal_antennas.html



PUFF version 2.1

Microwave CAD Software

Unfortunately this product is no longer available by itself. It is now part of two paperback books :

- Complete Wireless Design by Cotter W Sayer
 - The Electronics Of Radio by David B Rutledge
- Both books are available from Amazon.com

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Gunthard Kraus, DG9GB

Modern Design for Band Pass Filters Made From Coupled Lines

Nowadays there are many aids available to any electronics developer. Even for development work in the area of high-frequency engineering, there is some very powerful software in existence, some of which is available without charge on the Internet.

The use of modern design and simulation tools is described below by means of examples.

1. Foreword

Some years have passed since the series of articles entitled "Design and realisation of microwave circuits" in [1], in which this subject was dealt with comprehensively. In the intervening period, the options for procuring information and for circuit simulation have multiplied greatly. In addition, the analysis options are more precise, thanks to continuous improvements in the CAD field.

A "test version" or "student version" of almost any modern CAD or simulation programs can now be obtained from the Internet, including the original manuals, which are usually complete textbooks in themselves - and mountains of application notes on almost any subject. The real problem thus becomes how to make a suitable choice. "Know-where and know-

how" are also important, for all test versions of what are usually very expensive programs have some kind of limitations. And there's nothing more irritating than to slog away at familiarising yourself with a new program and then suddenly realise that the program available just can't go any further with the problem you're working on.

So the idea here is to demonstrate the correct and successful design of stripline band-pass filters, together with their implementation in practise. We shall compare not only the procedures but also the degrees of success, using the tried and tested CAD program "PUFF" (Version 2.1) and the ultra-modern student version of ANSOFT Serenade.

2. A glance at the technology

Band-pass filters serve to "separate out" a specific desired frequency range, while simultaneously suppressing, as far as possible, all undesirable signals outside this range. The following filter models can be considered for the microwave range in this context:

- Waveguide filters (for very high power levels)
- Coaxial filters



- Helix filters
- Filters made from ceramic resonators
- SAW filters
- Inter-digital band-pass filters
- Stripline filters with coplanar structures
- Microstrip filters made from coupled lines
- Hairpin filters, etc.

If we also lay down additional requirements, such as

- DIY manufacture as simple printed circuit board at lowest possible cost
- easily convertible to other frequencies without high costs or problems
- no smoothing
- absolute reproducibility

then the two last types are usually given preference. In this context, hairpin filters represent a modified version of the standard stripline filter for shortening the construction length and increasing the edge steepness. The disadvantages of larger dimensions must be taken into account here.

3.

Principles of stripline band pass filters made from coupled lines

Here we are using “coupled lines”, i.e. two striplines which are running in parallel and close together. Due to this running together closely in parallel, we obtain not only a capacitive coupling (via the electrical field) from one line to the other, but also a magnetic coupling. The magnetic field of one line induces an electrical voltage in the second line and thus transfers electrical power. The remarkable thing here is that the different waves triggered through this coupling from one line onto the other are added together in only one direction. But in the other direction they are in antiphase and try to cancel each other.

This is precisely the behaviour of a directional coupler and it is also the main applications area for this line structure. This behaviour can be used to separate forward and return waves in a system!

However, the description of such a component for simulation can be expanded further:

Due to the fact that some of the waves

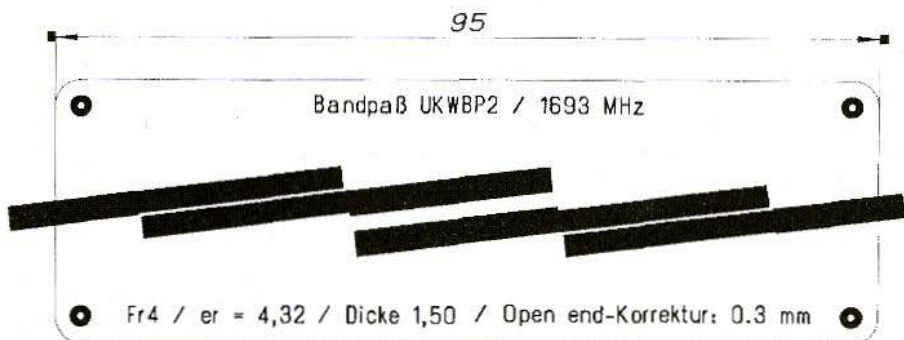


Fig 1: PCB layout for 16943 MHz Bandpass Filter

triggered in the second line are in-phase and some are anti-phase, it is necessary to specify two different impedance levels, namely

- a) the EVEN impedance (or: in-phase impedance) and
- b) the ODD impedance (or: anti-phase impedance).

The EVEN impedance level is always higher than the system impedance (usually 50 Ω), whereas the ODD impedance is essentially lower than the system impedance used.

The relationship of the three impedance levels to one another always depends on the formula

$$Z^2 = Z_{EVEN} \cdot Z_{ODD}$$

Note:

As soon as even one impedance is specified, many CAD programs react in the following way: if the impedance exceeds the system impedance, Z , then it is indicated as the EVEN impedance, and the missing ODD value is calculated in accordance with the above formula - and vice versa.

3.1 Now for stripline band-pass filters

If several such coupled line pairs, with 90 degrees of electrical length, are connected together in series, the line sections act as resonators and "the input signal is transmitted smoothly from input to output, only in the range around this frequency" in this way, the desired band-pass behaviour is obtained. Unfortunately, this is repeated at the odd multiples, i.e. for example the triple frequency, etc...

For the practical implementation, see Fig. 1.

It can easily be seen that, in addition to the three coupled line pairs, the 50- Ω striplines are also used as connection to

the SMA sockets. The underside of the printed circuit board is a continuous earth surface.

4.

The design path: from the standard low-pass to the stripline band-pass filter

4.1 Preliminary work

The circuit developer is initially faced with the following decisions:

Which type of filter is the correct one?

The choice will fall, for example, on Bessel, Butterworth, or Chebyshev filters. Basically:

Chebyshev filters display ripple in the transmission range, but as against that they can offer good edge steepness for the transition into the filter attenuation band.

If, in contrast, we need better group delay behaviour and no ripple in the transmission range, we go for Butterworth filters, though their edge steepness in the filter attenuation band is markedly lower than that of the Chebyshev type.

If the filter has to remain absolutely gentle and as smooth as possible at all points, that leaves only the Bessel filter. Mind you, we pay for this "gentle" behaviour with a very "tired" transition from the transmission range to the filter attenuation band (in order to keep phase distortion as low as possible). Thus, there is scarcely any "edge steepness" to speak of in the filter attenuation band.

Then comes the question of the degree of filtration, N , which for normal low-pass filters directly corresponds to the number of components needed. A greater degree of filtration brings about steeper edges in the filter attenuation band, but in practise



the attenuation in the transmission range is also increased, due to the greater number of components and their losses.

In practise, the type of filter which is very frequently used is the Chebyshev, with N between 3 and 5. For this reason, a filter from this group is taken as an example here.

The next decision relates to the system impedance (usually 50 Ω). Moreover, especially for Chebyshev filters, the maximum passband ripple, the reflection factor, etc., must be determined.

It should be borne in mind that the variables:

- passband ripple (oscillations of S21 and / or the transmission loss)
- reflection factor r
- voltage standing wave ratio VSWR
- S11
- reflection attenuation ar

are inseparably associated with each other in the Chebyshev type! The following relationships apply here:

- a) Between the reflection factor r and the voltage standing wave ratio VSWR:

$$r = \frac{VSWR - 1}{VSWR + 1}$$

- b) Between the reflection factor r and the passband ripple (maximum transmission loss in dB)

$$a_{max} = 10 \cdot \log \frac{1}{1 - |r|^2}$$

- c) Between S11 and / or S22, the reflection factor r and the reflection attenuation ar:

With correct matching, S11 and / or S22 correspond precisely to the reflection factor of the filter, but are normally specified in dB:

$$S11 = S22 = 20 \cdot \log |r|$$

The reflection attenuation is then simply the “negative dB value of S11 or S22”! Correctly:

$$a_r = 20 \cdot \log \frac{1}{|r|}$$

The following summary table (drawn up in accordance with the above formula) is intended to serve as a small design aid:

Reflection factor r	Reflection attenuation ar	S11 or S22	Chebyshev Ripple of Trans loss
50 %	6dB	6dB	1.25dB
20 %	14dB	14dB	0.177dB
10 %	20dB	-20dB	0.0436dB
5 %	26dB	-26dB	0.01dB
2 %	34dB	-34dB	0.0017dB
1 %	40dB	-40dB	0.00043dB
0.5 %	46dB	-46dB	0.0001dB

(In practise, a maximum reflection factor between 5 and 10 % is usually sufficient...)

4.2. Designing a GPS bandpass filter

A bandpass for GPS with the following data is intended to serve as a design example:

Filter type	Chebyshev
Mean frequency	f ₀ = 1575 MHz
Lower limiting frequency	f _{Min} = 1550 MHz ripple limiting frequency
Upper limiting frequency	f _{Max} = 1600 MHz ripple limiting frequency
Degree of filtration	N = 3
System impedance	Z = 50 Ω
Max. reflection factor	r _i = 10 %
Max. ripple	a _{rmax} = 0.0436 dB
Reflection attenuation in transmission range	ar = 20 dB

S11 in transmission range S11 = -20 dB
if at all possible, less than

The selected reflection factor $r = 10\%$ gives a maximum ripple of 0.0436 dB in the transmission range.

This means that S21 can fall as low as -0.0436 dB, whilst S11 and S22 never exceed 20 dB.

Note:

The following calculation path is taken from the book "Microwave Engineering" by David Pozar [5], Page 484.

Additional note:

The degree of filtration, N , should always be selected to be odd (i.e. 3, 5, 7...), because only then are the source resistance and the load resistance identical. Apart from this, make sure that the number of coupled line pairs is always 1 more than the selected degree of filtration. For $N=3$, there must thus be four line pairs in the layout.

And now to the individual design steps:

1st step:

First we need the filter coefficients of a single low-pass filter for this case. For this we can, for example, use the "faisyn" program (obtainable, for example, from <http://www.rfglobalnet.com>).

The above filter data are entered in succession when the program makes the corresponding requests, and the option "Parallel Capacitor" is selected. Thus the following table is finally obtained, with the 4 coefficients required for the calculation (Fig. 2):

$$g_1 = \text{cap1} = 0.8532$$

$$g_2 = \text{ind1} = 1.1038$$

$$g_3 = g_1 = \text{cap} = 0.8532$$

$$g_4 = \text{normalized load resistance} = 1.0$$

2nd step:

Specification of fractional bandwidth of pass:

$$\Delta = \frac{f_{\max} - f_{\min}}{f_{\text{center}}} = \frac{1600 - 1550}{1575} = \frac{50}{1575} = 0.031746$$

3rd step:

Now we come to the admittance inverter constants for the four line pairs:

1st line pair:

$$Z_0 \cdot J_1 = \frac{\sqrt{\pi \cdot \Delta}}{\sqrt{2 \cdot g_1}} = \frac{\sqrt{\pi \cdot 0.031746}}{\sqrt{2 \cdot 0.8532}} = 0.24175$$

2nd line pair:

$$Z_0 \cdot J_2 = \frac{\pi \cdot \Delta}{2 \cdot \sqrt{g_1 \cdot g_2}} = \frac{\pi \cdot 0.031746}{2 \cdot \sqrt{0.8523 \cdot 1.1038}} = 0.05138$$

3rd line pair:

$$Z_0 \cdot J_3 = \frac{\pi \cdot \Delta}{2 \cdot \sqrt{g_2 \cdot g_3}} = \frac{\pi \cdot 0.031746}{2 \cdot \sqrt{1.1038 \cdot 0.8532}} = Z_0 \cdot J_2 = 0.05138$$

Normalized Lowpass Prototype Filter Components:

Wc=1 rad/sec, Normalized Source Resistance=1

CAP 1 0 C= 0.8532

IND 1 2 L= 1.1038

CAP 2 0 C= 0.8532

The Normalized Load Resistance= 1

Press any key to continue...

Fig 2: Filter Coefficients fFrom "Faisyn"

4th line pair:

$$Z_0 \cdot J_4 = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_3 \cdot g_4}} = \sqrt{\frac{\pi \cdot 0.031746}{2 \cdot 0.8532 \cdot 1.00}} = Z_0 \cdot J_1 = 0.24175$$

4th step:

The EVEN and ODD impedances of a line pair are generally specified in accordance with the following formulae:

$$Z_{EVEN} = 50\Omega \cdot [1 + Z_0 \cdot J_N + (Z_0 \cdot J_N)^2]$$

$$Z_{ODD} = 50\Omega \cdot [1 - Z_0 \cdot J_N + (Z_0 \cdot J_N)^2]$$

For the first and fourth line pairs we obtain:

$$Z_{EVEN} = 50\Omega \cdot [1 + 0.24175 + 0.24175^2] = 65\Omega$$

$$Z_{ODD} = 50\Omega \cdot [1 - 0.24175 + 0.24175^2] = 40.8\Omega$$

For the second and third line pairs the values are:

$$Z_{EVEN} = 50\Omega \cdot [1 + 0.05138 + 0.05138^2] = 52.7\Omega$$

$$Z_{ODD} = 50\Omega \cdot [1 - 0.05138 + 0.05138^2] = 47.56\Omega$$

5.

Use of PUFF

5.1. Simulation of ideal circuit using PUFF

First start "PUFF" And load the SETUP file. Then press the "F4" key and enter the following values for the Rogers material R04003:

with thickness 0.032

for impedance level $z_d = 50 \Omega$

the design frequency $f_d = 1575$ MHz

the dielectric constant $\epsilon_r = 3.38$

board thickness $h = 0.813$ mm.

F4 : BOARD	
zd	50.000 Ω
fd	1.575 GHz
er	3.380
h	0.813 mm
s	200.000 mm
c	50.000 mm
Tab	microstrip

Fig 3: Starting Parameters

The printed circuit boards size "s" should be 200 mm, and 50 mm. is a sufficient distance between the connections (Fig. 3).

Then move into field "F3" and successively enter there the data for the two coupled line pairs required. Please transfer them precisely as shown in Fig. 4!

F3 : PARTS	
a c	65.00 Ω 40.80 Ω 90.0°
b c	52.70 Ω 47.56 Ω 90.0°
c	

Fig 4: Data For Successive Line Pairs

Press the "F1" key to make the layout window appear (Fig. 5). And now please pay attention, first move the cursor as far to the left as you can. Then press the shift key for upper case lettering and keep pressing the "Cursor Left" key until you get to the desired location. Press "I" and port 1 is connected immediately.

Press letter "a" on the keyboard, followed by "Cursor Right". This positions the first line pair. Then press "b" and next "Cursor down", which connects up line pair "b".

Now press "Cursor down" again and the third pair is already sitting there on the screen. Finally press "a", "Cursor down"

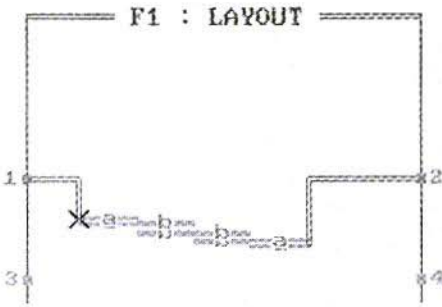


Fig 5: PUFF Layout Window

again and then "2" and port 2 is likewise connected to the exit of the circuit.

Use "F2" to go to the simulation window. Using "Cursor Up2 or "Cursor Down", you can move, not just in the top left-hand "Plot window", but also along the

axes of a diagram in the bottom right-hand corner ("linear plot") (Fig. 6).

Here you pre-set, for example:

500 simulation points

Smith radius = 1

Representation of S11 and S22

Horizontal scales in the ratio: 1.5
...1.7 GHz

Vertical scales in the ratio: -1.00...0
dB

And now please press "p", and you can watch the computer at work. If you think that's too slow, press "Q", and then all the calculations are done and the image is built up off-screen, and things go considerably faster!

If you look at the result now, you'll undoubtedly be a little disappointed: it's nowhere near a ripple with maximum 0.0436 dB; it's bigger by a factor of more

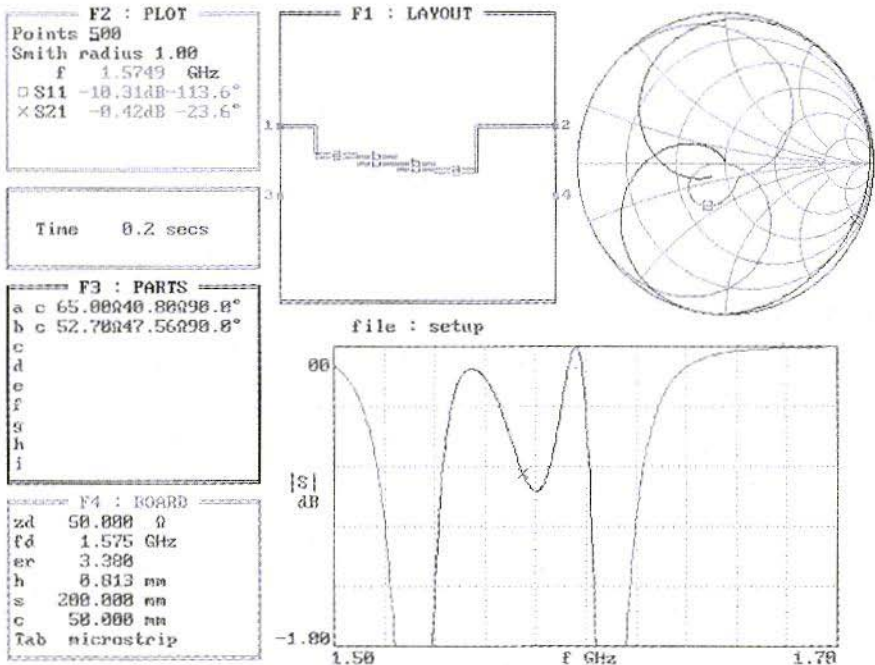


Fig 6: PUFF Simulation Window

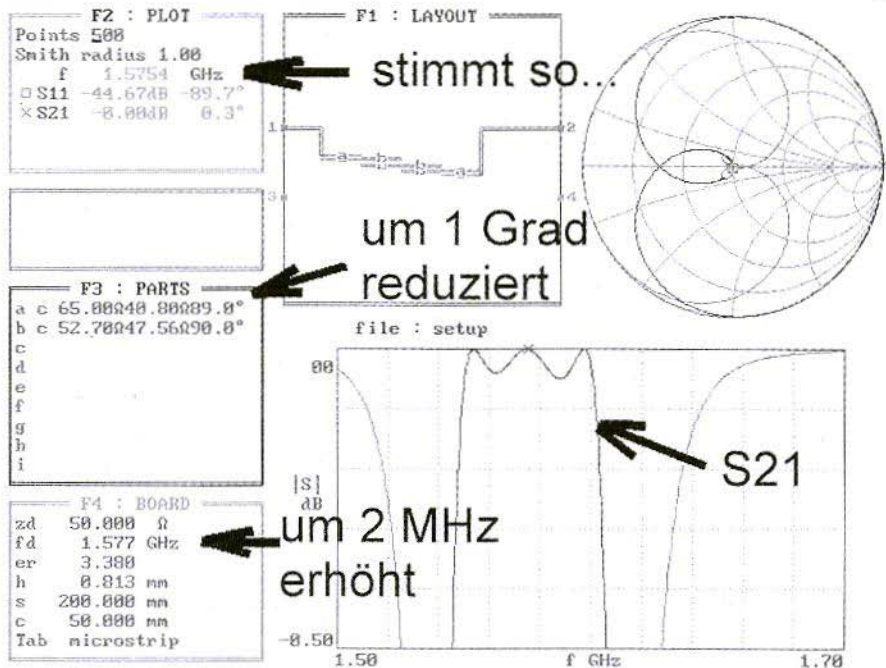


Fig 7: PUFF Simulation After Slight Corrections

than 10 and the value is 0.5 dB. But don't worry, we can still get round that. We just need to make some very slight corrections to the line data!

We need only reduce the electrical length of the first and fourth line pairs (part "a" in the parts list) by approximately 1 degree and increase the design frequency by 2 MHz in order to obtain the theoretical curve (Fig. 7)!

5.2. Simulation of the real circuit with PUFF

First call up the "SETUP.PUF" file from the PUFF directory into a text editor and then enter the remaining printed circuit board data (thickness of copper layer = 35 micrometres and surface roughness for

a strip conductor polished until it gleams with grits and grinds or scouring powder approximately 2 micrometres. Loss factor "lt" of board material R04003 at this frequency max. 0.001):

```

....
zd 50,000 Ohms {normalizing impedance.
0<zd}
fd 1.575 GHz {design frequency. 0<fd}
er 3.380 {dielectric constant. er>0}
h 0.813 mm {dielectric thickness. h>0}
s 200.000 mm {circuit-board side length.
s>0}
c 100.000 mm {connector separation.
c>=0}
r 0.010 mm {circuit resolution, r>0, use
Um for micrometres}
a 0.000 mm {artwork width correction.}
mt 0.035 mm {metal thickness, use Um
for micrometres.}
sr 2.000 Um {metal surface roughness,
use Um for micrometres.}
lt 1.0E-0003 {dielectric loss tangent.}
    
```


....

The amended setup file is loaded back into PUFF and then the exclamation mark for each pair of coupled lines is entered in field "F3". This switches to "realistic modeling with all side-effects" (Fig. 8).

If we now place the cursor in field "F3" on part "a" and enter the equals sign, the actual data of the coupled lines are immediately faded into the dialogue field (Fig. 9).

Now we have to keep changing the values entered under "a" until the data determined in the preceding chapter are displayed in the dialogue field:

$$Z_e = 65 \Omega$$

F3 : PARTS	
a	c!65.00Ω40.80Ω90.0°
b	c!52.70Ω47.56Ω90.0°

Fig 8: Switch To Realistic Modeling

$$Z_o = 40.8 \Omega$$

electrical length $l = 89$ degrees

It can be seen that to bring this about the entry for "a" has to be

Ze:	65.006Ω
Zo:	40.796Ω
l :	89.032°

Teil "a"

F3 : PARTS	
a	c!66.10Ω43.56Ω89.6°

Fig 9: Data For Part "a"

Ze:	52.702Ω
Zo:	47.542Ω
l :	90.011°

Teil "b"

F3 : PARTS	
a	c!66.10Ω43.56Ω89.6°
b	c!55.65Ω49.26Ω90.7°

file : setup

Fig 10: Data For Part "b"

$$c! 66.1 \Omega 43.65 \Omega 49.6^\circ$$

This procedure must be repeated for part "b" (Fig. 10). The target is to obtain this display in the dialogue field:

$$Z_e = 52.7 \Omega$$

$$Z_o = 47.56 \Omega$$

$$l = 90 \text{ Grad}$$

For this, finally, we need the entry:

$$c! 55.6 \Omega 49.26 \Omega 90.7^\circ$$

Fig. 11 shows the result of the circuit simulation if the losses are taken into account.

The design frequency continues to remain at 1577 MHz, but following the simulation use <Page Down> to move the display cursor to 1575 MHz. We now have a transmission loss of approximately 2.5 dB.

If we correspondingly switch the value range in the two axes of the lower diagram, we can take another look at the long-range selection, i.e. the behaviour in the range between 1 and 2 GHz (Fig. 12).

5.3. Determination of mechanical, uncorrected line data with PUFF

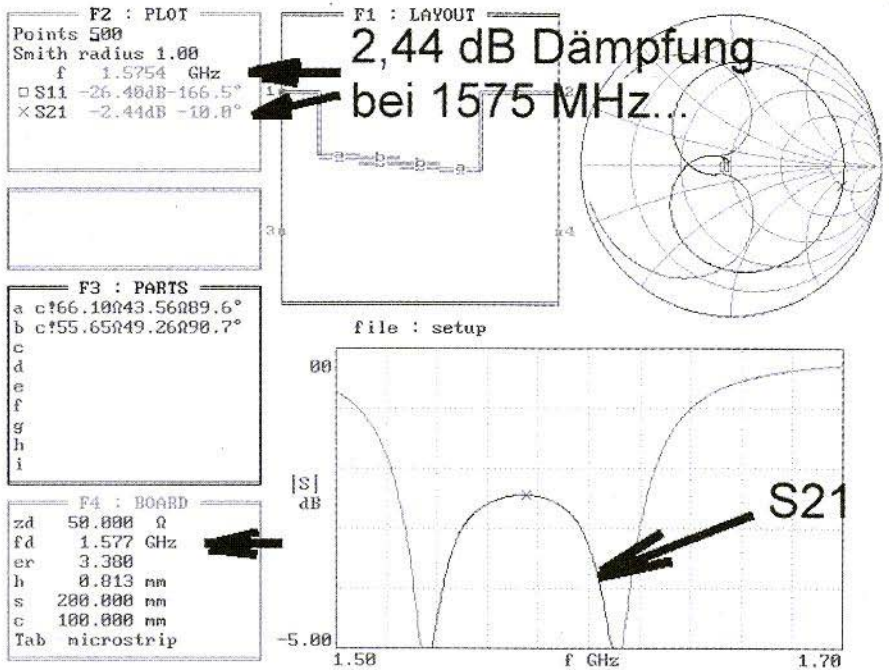


Fig 11: Results With Losses Taken Into Account

In order to get at the dimensions of the coupled lines, we move back again into field "F3" and delete the call-up sign at part "a" behind the letter "c" (for coupled lines). As soon as we key in the equals sign behind here, we obtain the desired values in the dialogue field (Fig. 13):

- Length l = 29.34 mm
- Width w = 1.59 mm
- Interaction gap s = 0.31 mm

Repeat this for part "b", i.e. the two central pairs of coupled lines, and we correspondingly obtain Fig. 14:

- Length l = 29.15 mm
- Width w = 1.82 mm
- Interaction gap s = 1.84 mm

Then, as a preliminary to the board design, determine the width of the 50 Ω feed likewise in the same way. It is modelled as "lossy transmission line with

90 degree length" and, as part "c", supplies a required width, w = 1.84 mm, following the deletion of the exclamation mark (Fig 15).

5.4. Necessary layout corrections

Here we are dealing either with striplines open at the end or with the meeting of two striplines which are of different widths. In both cases, the familiar OPEN-END correction is required, due to fringing, but one peculiarity should be taken into account here at the open ends of coupled lines:

The two line pairs are coupled to each other both electrically and magnetically. It is certainly true that the electrical field lines project beyond the open ends of the lines (so we need to do some shorten-

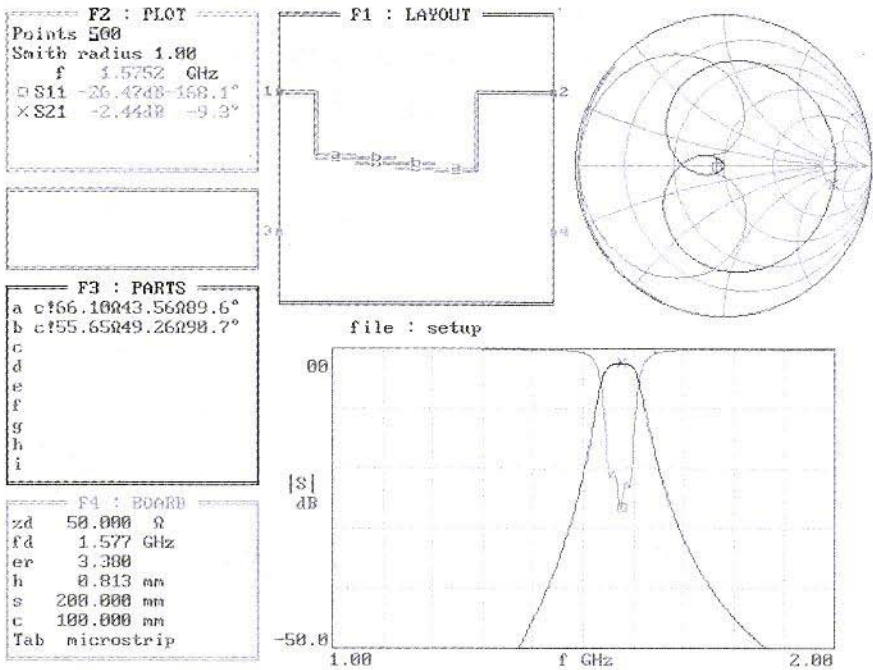


Fig 12: Simulation Results In The Range 1 - 2 GHz

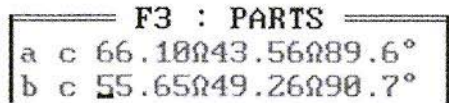
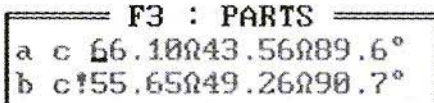
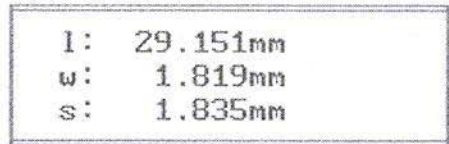
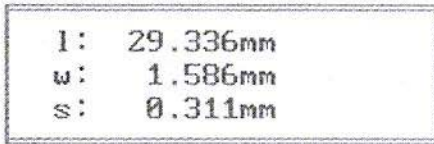


Fig 13: Results For Line Pair "a"

Fig 14: Results For Line Pair "b"

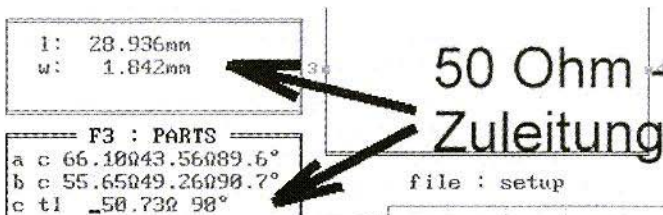


Fig 15: Determining The Width Of 50Ω Line



ing...), but the magnetic coupling decreases linearly in this area right down to zero.

For this reason, calculate in only half the "OPEN-END-EXTENSION" which would otherwise be normal and shorten the line correspondingly!

Apart from this, we now need several tools to create the layout:

- a) The well-known, tried and trusted diagram for determining the OPEN-END-EXTENSION from the PUFF manual (Fig. 16).
- b) A simple hand-drawn sketch (Fig. 17) with the electrical data of the individual line pairs already-determined. Enter the necessary corrections.
- c) A printed circuit board CAD program, which simultaneously makes it possible to solve tricky construction problems (e.g. GEDDY-CAD, tried and trusted for such microwave tasks for many years).

Procedure

1st step:

The first and fourth line pairs consist of two striplines each with a width of 1.59 mm. With the help of PUFF we obtain the impedance level for the pre-set printed circuit board data:

The result gives us: $Z = 54.6 \Omega$

So we go to the above diagram from the PUFF manual. For this task, it supplies (with $\epsilon_r = 3.38$) an open-end extension $\Delta l/h$ of approximately 0.45. So these section pieces must be reduced by half of $0.45 \times 0.813 \text{ mm} = 0.18 \text{ mm}$ at all open ends!

2nd step:

At the start and end of the bandpass, the 50- Ω feed is connected up with a width of 1.84 mm and turns into the (narrower!) stripline with a width of 1.59 mm. So the narrower line must be extended by a little piece measuring $(1.59 \text{ mm}/1.84 \text{ mm}) \times 0.45 \times 0.813 \text{ mm} = 0.05 \text{ mm}$.

3rd step:

The two central line pairs have conductor widths of 1.82 mm. The associated impedance level (according to PUFF) is 50.4 Ω and at $\epsilon_r = 3.38$ requires an

Fig 16 Diagram For Determining The Open End Extension From The PUFF Manual

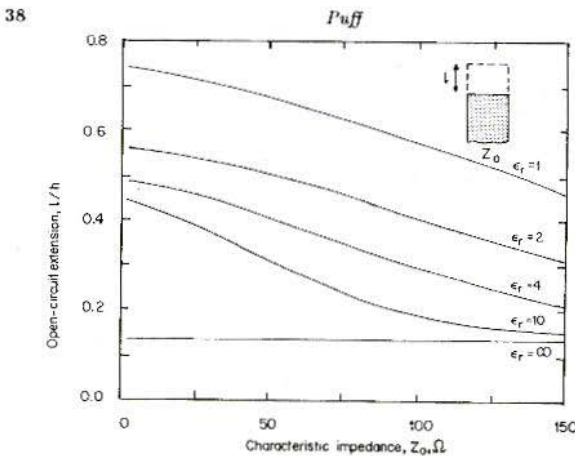


Figure 7.2 The open-circuit end correction in microstrip, plotted from (7.2). The artwork length correction in a parts list should be *negative*.

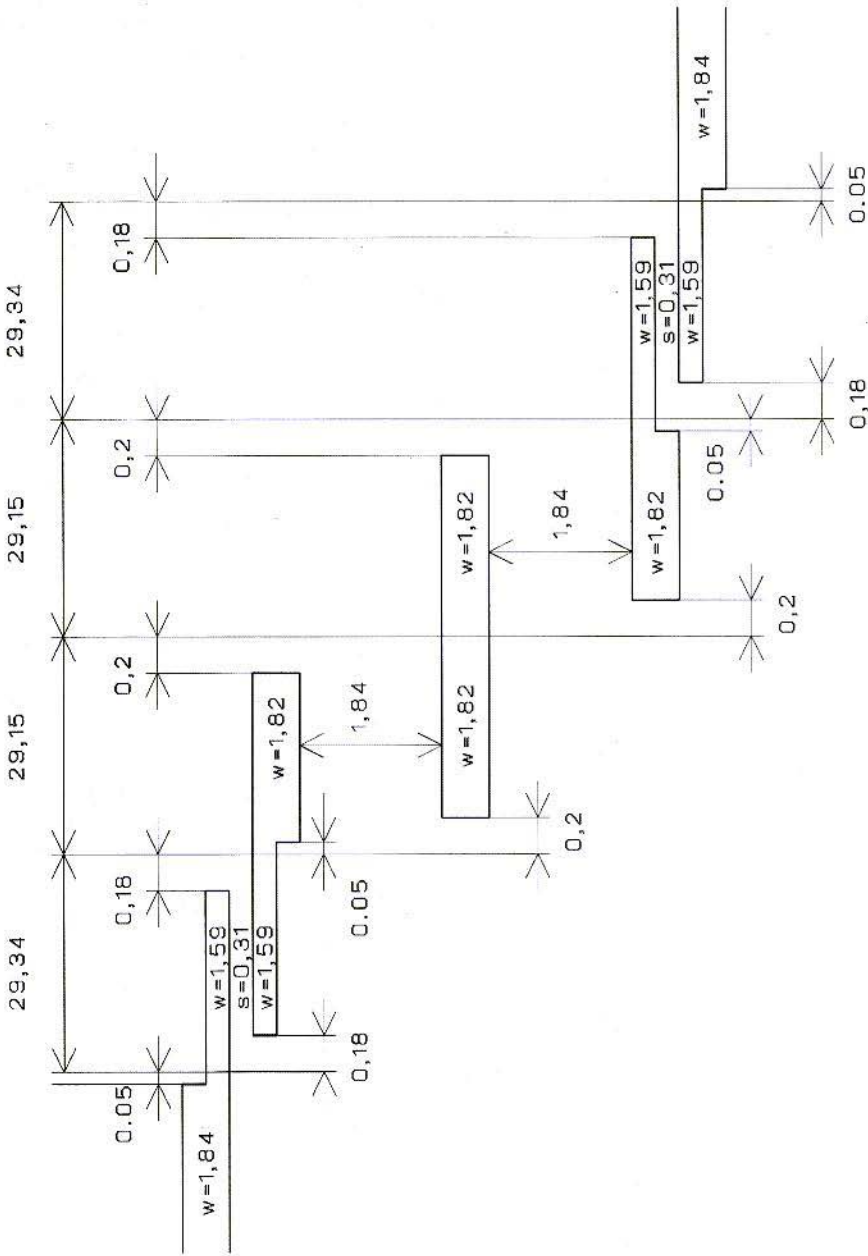


Fig 17: Sketch For The Filter

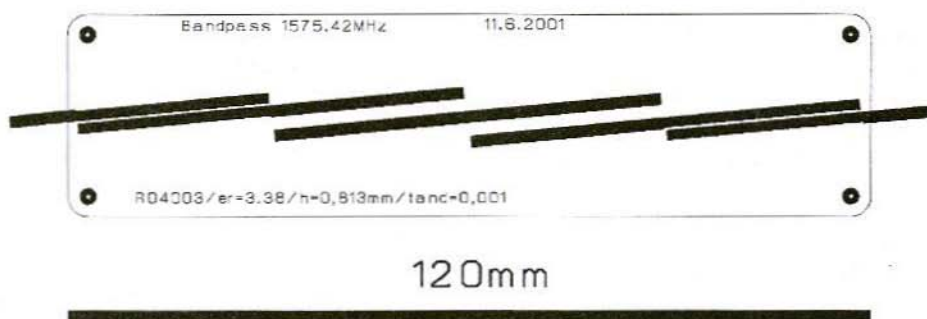


Fig 18: The Finished Layout

OPEN- END correction of 0.48×0.813 mm = 0.39 mm.

Again, only half of this value, i.e. approximately 0.2 mm, needs to be cut off from the two ends.

4th step:

When the first line pair meets the second and the third meets the fourth, there is a correction of

$(1.59 \text{ mm} / 1.82 \text{ mm}) \times 0.48 \times 0.813$ mm = 0.05 mm.

The wider line must be shortened and the narrower line must be lengthened by this amount.

You should never omit entering all these details on a sketch out of laziness. It is an obligatory stage in the layout design (Fig. 17) and you need to take considerable trouble just to produce the drawing. But this is as nothing in comparison with the time and expense which will be wasted if the behaviour of the prototype produced inexplicably displays big departures from the simulation and you laboriously have to check every individual dimension on the completed printed circuit board. Its really very annoying if a gross error in the board layout turns out to be the reason for this.

Normally if you've followed all the instructions in this article the divergences

between the simulation and prototype are max. 1 - 2%.

The finished layout, prepared for incorporation into a milled aluminium trough, the internal dimensions of which are 30 x 120 mm, is shown in Fig. 18.

One more tip for those who don't know what to do with the thick 120 mm long line under the printed circuit board: we need this to assemble the board equipment, unless there is a photo-plotter available, which can be used to bring us back to the correct dimensions by means of photography. Only in this way will we again find the correct scale and be able to handle the manufacturing problem posed by the narrow interaction gap for line pairs 1 and 4.

5.5. Use of TRL85 Stripline Calculator

To determine the data for the incorporation of the circuit into a screening housing, the "TRL85" stripline calculator from Ansoft is used.

Ansoft are known for supplying very high quality and expensive HF-CAD programs, but have always also had their eye on instruction and training! So on their Homepage on the Internet () we find a

student version of the microwave CAD package "Serenade" which can be downloaded free.

Following installation it will be recognised that a very good stripline calculator has been filed in a separate directory as "TRL85.EXE-file" together with excellent ONLINE help. It can be used separately at any time, copied down and transferred to other computers. You will very soon learn to appreciate its WINDOWS user interface.

If we now compare the simulation results from "TRL85" with the values from PUFF, we can make the following statements:

- a) In normal cases, the data determined by PUFF and TRL85 for single and coupled striplines are practically identical.
- b) In addition to this, TRL85 offers the advantage that screening can be brought into the simulation in the form of the "Cover Height" (the distance between the cover and the board).
- c) With TRL85 all data which is of interest (impedance level, losses,

broken down into dielectric loss, conductor loss and total loss... etc.) can be determined directly for a specific design frequency and given out. Unfortunately, PUFF can't provide for that, as there, though you certainly have to carry out calculations using these values, they are not displayed.

The "TRL85" program is presented in this issue in the section headed "The interesting program".

In Table 1 below, we now contrast the TRL85 microstrip simulations for operation without and with screening.

A comparison with the values determined by PUFF and used in preceding chapters shows that it is high time for a printed circuit board and correct dimensions.

Another tip:

TRL85, unfortunately, won't automatically apply the OPEN-END correction either. So again we have to fall back on the use of the diagram from the PUFF manual when a line end is hanging in the air somewhere.

Bandpass Specification	TRL85 without screening	TRL85 with screening Distance = 13mm
50 Ω Microstrip	Width = 1.84 mm	Width = 1.84 mm
First Line Pair	Ladder Width = 1.58321 mm	Ladder Width = 1.5769 mm
E=89, 1575.42 MHz	Spacing = 0.33225 mm	Spacing = 0.32877 mm
Z _c =65 Ω , Z ₀ =40.8 Ω	Length = 29.2968 mm	Length = 29.3133 mm
Second Line Pair	Ladder Width = 1.8216 mm	Ladder Width = 1.81525 mm
Z _c =52.7 Ω , Z ₀ =47.56 Ω	Spacing = 1.87272 mm	Spacing = 1.82327 mm
E=90, 1575.42 MHz	Length = 29.186 mm	Length = 29.2009 mm

Table 1: TRL85 Simulations With and Without Screening



6. Repetition of design using ANSOFT-SERENADE

6.1. Simulation of ideal electrical circuit

As "SERENADE" is already installed on the PC, on account of the "TRL85" stripline calculator, we can carry out a repeat operation and see how such a circuit is investigated by this very modern program. Naturally, we are interested, above all, in what improvements it can give us, in terms of ease of operation or precision.

Procedure

Launch the SERENADE software and start a new project (e.g. "BP1575_1").

Then look for the "ideal coupled line" (Fig. 19), position it on the screen four times and each time (see preceding chapter!) enter the EVEN and ODD impedances, the electrical length of 90 degrees and the operating frequency applying for this, 1575.42 MHz, in the "Property Editor". When the component is positioned, the editor opens automatically. If it does not, just double click on the left-hand mouse button on the circuit icon in the wiring diagram. Our ports are connected up at the input and output, but that's not enough yet!

The "Harmonica" circuit simulator is the problem now that two of the four connections have simply remained open for each coupled line pair, this is prohibited.

We could now apply a very high-ohmic resistance (e. g. 10 Mega-Ohms) to such an open connection. However "ideal line pieces working without load at the end (Stubs)" are considerably better and have lower losses, with an electrical length of zero at 1575 MHz and with the impedance level $Z = 50 \Omega$. They cause no additional losses. Nor do they alter the data for the coupled lines.

Then the frequency block is re-set and the range between 1 GHz and 2 GHz is represented in 1 MHz steps. Rogers R04003 material is again used as substrate, with a thickness of only 32 MIL (0.813 mm), as this gives the filter structure smaller dimensions. The other data are as follows:

Dielectric constant	$\epsilon_R = 3.38$
Metallisation Met1	Copper with thickness 35 μm
Surface roughness	RGH = 2 μm
Loss factor	TAND 0.001.

The entire circuit, as used for simulation, is shown in Fig. 20. Even for those who don't know the program yet, the components just produced can easily be identified. Frequency and substrate control blocks certainly need no further explanation.

Following a precise check of the circuit, the simulation can begin (button with gears) and the Report Editor can be activated (turquoise / grey button). Se-

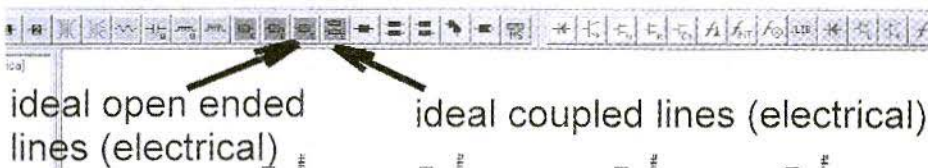


Fig 19: Toolbar For Serenade

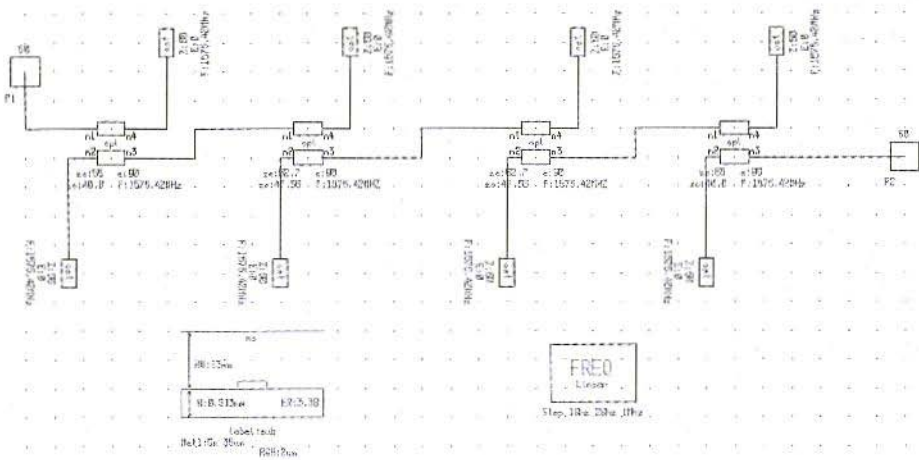


Fig 20: Circuit Used For Serenade Simulation

lect, for example, S11 and S21 in dB representation and examine the result (Fig. 21).

The result looks promising and S11 is never worse than the intended value in the passband 20 dB.

Use the right-hand mouse button and click on “ZOOM IN” repeatedly to bring out the precise sequence of S21 in the range from 0 to 0.1 dB between 1550 and 1600 MHz.

Only in this way can we assess whether the design path from the preceding chapter really supplies the correct values desired.

Fig. 22 shows a perfect and well-formed filter curve.

In practice, both the mean frequency (1575 MHz) and the minimum ripple value (0.041 dB) are in accordance with the design pre-settings.

6.2. Simulation of physical circuit using HARMONICA

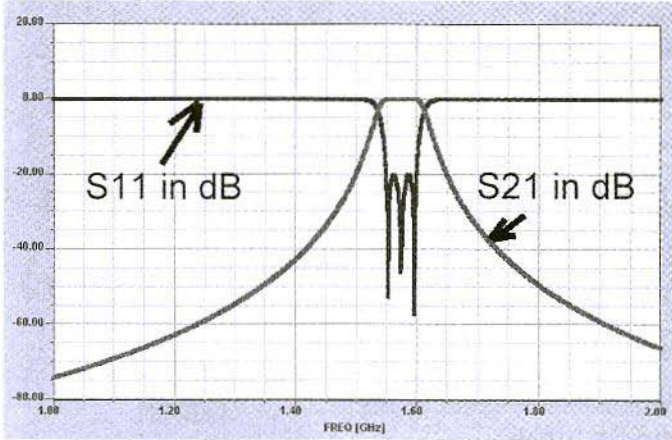
Apart from simulating the bandpass with “electrical components”, HARMONICA also offers the option of a structure made up of “physical circuits”. This requires the conductor width, interaction gap and conductor length to be entered, the dimensional unit being “mm”. And these values can be obtained only through the TRL85 stripline calculator, which can even be called up from the operating screen by pressing a button.

Here only the values for the first and fourth line pairs ($Z_c = 65 \Omega$, $Z_o = 40.8 \Omega$, $E = 90$ degrees) need be entered, with the track data (copper with $35 \mu\text{m}$ thickness and a roughness of $2 \mu\text{m}$) and the printed circuit board and housing data (board thickness $H = 0.813\text{mm}$, $ER = 3.38$, cover height above board, $IIU = 13\text{mm}$, $TAND = 0.001$).

If you then press the “Synthesis” button,



Fig 21: S11 and S21 Curves



you obtain a representation corresponding to Fig. 23, You obtain:

- Conductor width $W = 1.58$ mm
- Interaction gap width $S = 0.3$ mm
- Circuit length $P = 29.61$ mm

This procedure is repeated for the second and third line pairs ($Z_e = 52.7 \Omega$, $Z_o = 47.56 \Omega$, $E = 90$ degrees).

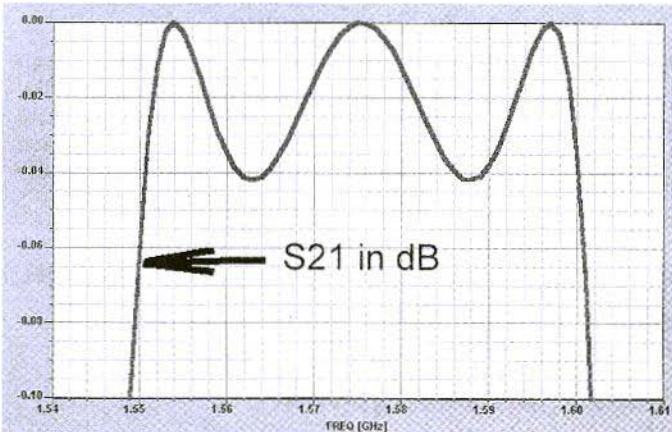
Then we obtain:

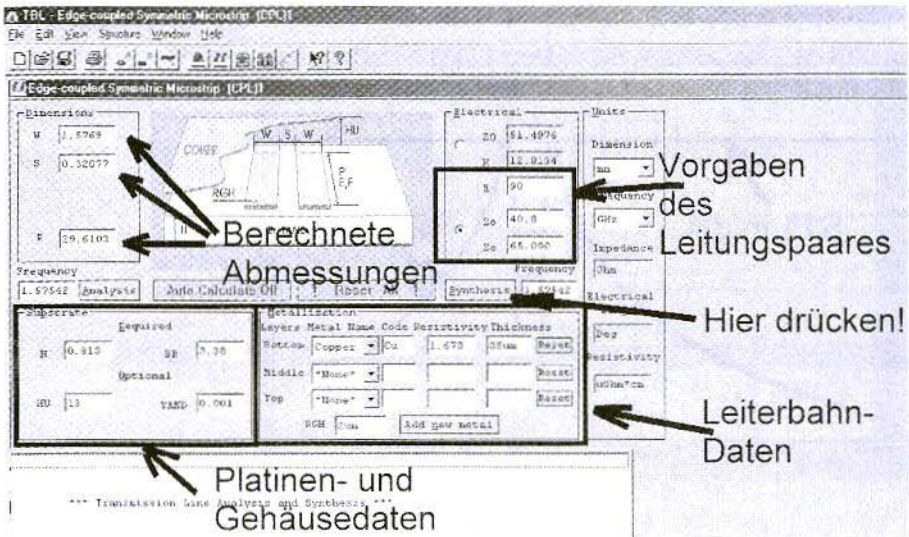
- $W = 1.82$ mm
- $S = 1.82$ mm
- $P = 29.20$.

It is naturally interesting to compare this with PUFF, although it should be remembered that the part involving the housing and the distance of 13 mm between the board and the cover is not included in this calculation!

Moreover, for simulation using PUFF,

Fig 22: Cloese Up Of S21 Curve





the electrical length for the first and last line pairs is presumed to be 89 degrees, whereas for SERENADE it is 90 degrees. The uncorrected PUFF values are as follows:

Line pairs 1 + 4:
 $W = 1.58 \text{ mm}$
 $S = 0.31 \text{ mm}$
 Line length $P = 29.34 \text{ mm}$;

Line pairs 2 + 3:
 $W = 1.82 \text{ mm}$
 $S = 1.84 \text{ mm}$
 Line length $P = 29.15 \text{ mm}$;

It can be seen that the differences between the two simulations are not so devastating that one of them immediately appears as completely unusable.

But let us simulate the bandpass with the "physical TRL85 values" again, consider the result and ponder on:

- where the differences with the simulation using PUFF come from and
- how can we arrive at the correct values.

To do this, though, we must draw a new circuit diagram, and first we must delete the old one completely.

Even when we create the new circuit diagram, things move forward splendidly. The coupled line pair is actually there with the two open ends as a completed component (Fig. 24). This naturally makes the work considerably easier.

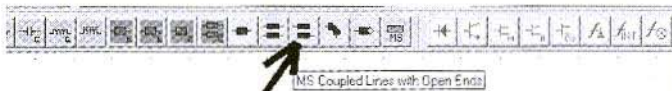


Fig 24: Component For Coupled Line Pair With Open Ends

Gekoppeltes Microstrip-Leitungspaar mit offenen Enden

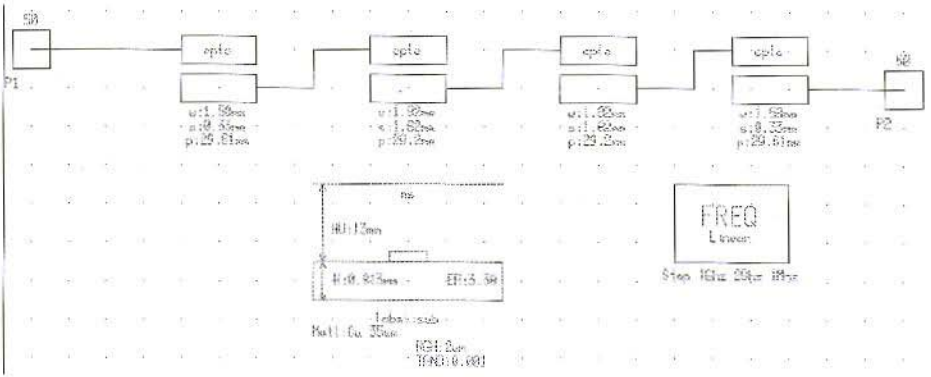


Fig 25: Revised Circuit

The screen operates in a much tidier way, even when the necessary data are entered (Fig. 25).

This is how the S-parameters look after simulation using HARMONICA (Fig. 26).

If we zoom into the representation of the

passband (Fig. 27), then several points strike us:

- a) The lowest transmission loss is predicted to be similar by both programmes (PUFF: approximately 2.5 dB, HARMONICA approximately 2.8 dB).

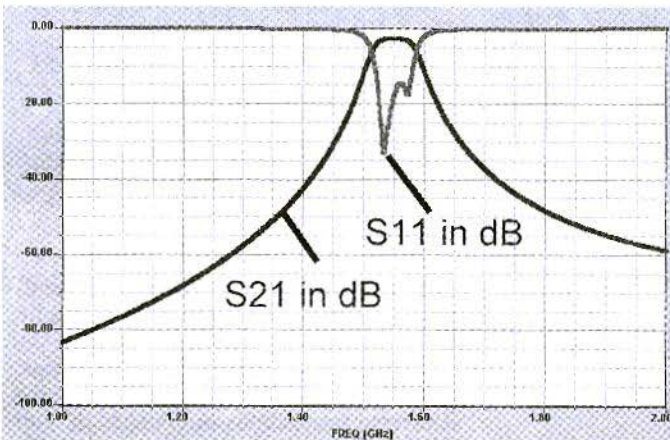
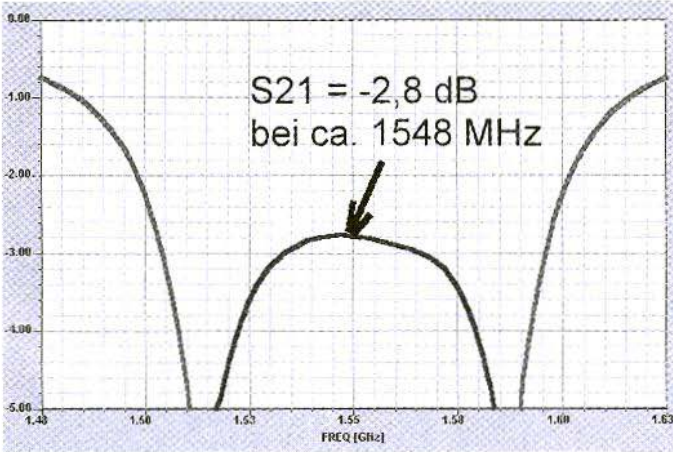


Fig 26: S Parameters From Harmonica Simulation



Fig 27: Close Up of S21 Curve



- b) The mean frequency of the passband according to the HARMONICA simulation reveals a divergence of 30 MHz (approximately 2 %) and is clearly too low. The program therefore does not make automatic OPEN-END corrections!

Luckily, if we use Ansoft, we don't need to carry out the same actions on the diagram as are listed in the PUFF manual to make corrections here we have something which is extremely useful.

What we actually do is to take specific values for the line pair through variables, pre-set maximum and minimum values for the S-parameters at specific frequencies, and then let the optimiser do the job of reconciling all these wishes.

Here we have the following steps:

1st step:

In the first and fourth line pairs, the physical length, P, is replaced by a variable, P1. We correspondingly use variable P2 for the second and third line pairs (Fig. 28).

2nd step:

Call up a "variable control block" (Menu

path: "Parts/Control Blocks/Variables") and enter the permissible variation range for P1 and P2 between question marks. In the middle is the original initial value (Fig. 29).

3rd step:

We now have to formulate the optimisation goals. There is an original button for this, in the form of a yellow and red practice target. In the student version only a maximum of 3 optimisation goals are permissible, but this should just be enough at first.

The optimisation goals here are:

- In the "Frangel" range, from 1.55 to 1.6 GHz, S21 should not fall below 3.3 dB (Goal1)
- In the "Frangle2" range, S11 should be lower than 20 dB (Goal2) (Fig. 30).

4th step:

Only now should you press the "Optimisation" button. The program normally indicates that it must first analyse the circuit, and asks for permission to do this. Grant it permission and also finally give it precise instructions in connection with the number of optimisation searches, the

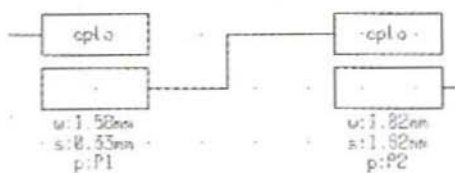


Fig 28: Variables P1 and P2

representation, etc. (Fig. 31).

5th step:

Now please follow the sequence exactly:

- 1) Pre-set, for example, 2000 searches;
- 2) Select "RANDOM" as optimisation type;
- 3) This tick has to go;
- 4) Now press "Optimise" and wait until the program has found the best approximation to the pre-settings.
- 5) Now start another circuit analysis, as this is the only way to update the results diagram.
- 6) Now close this menu and obtain the diagrams with S11 and / or S21 in the foreground (Fig. 32).

The assessment can be found in Fig. 33, and the result looks very satisfactory.

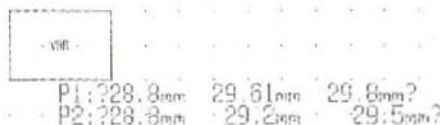


Fig 29: Entry Of Variables P1 and P2

The data within the passband have changed in accordance with Fig. 34; the result is useable.

The only question remaining is: where have the accompanying new circuit data got to, with which the optimiser has produced the curves above?

Its very simple: you'll find them in the variable block instead of the initial values (Fig. 35)!

Finally, assemble all data for the layout preparation in accordance with "AN-SOFTs physical variants".

First and fourth line pairs:

Conductor width $W = 1.58$ mm
Interaction gap $S = 0.33$ mm
Line length $P = 29.04$ mm

Second and third line pairs:

Conductor width $W = 1.82$ mm
Interaction gap $S = 1.82$ mm
Line length $P = 28.80$ mm

Now only one question remains: "Which of the two programs is really right"?

Theres only one way to find the answer to this question: produce another printed circuit board with just these dimensions, measure it under exactly the same conditions as for the "PUFF product" using the network analyser, and then cold-blood-



Frangle1:1.55GHz 1.6GHz
Frangle2:1.55GHz 1.6GHz

Goals1:MS21 -3.3dB GT
Goals2:MS11 -20dB LT

Fig 30: Goals For Optimisation

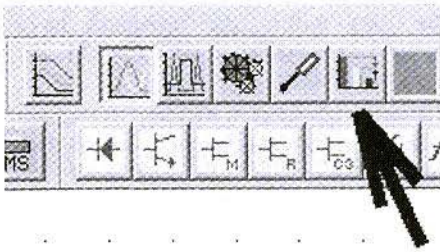


Fig 31: Starting Optimisation

edly analyse the results and compare them.

7. Literature

[1] Design and realisation of microwave circuits, Gunthard Kraus, VHF Communications, from issue 4/96, P. 244 - 250, at irregular intervals to issue 2/99.

[2] PUFF manual, original English version

[3] Ansoft-Serenade manuals (supplied when program is downloaded)

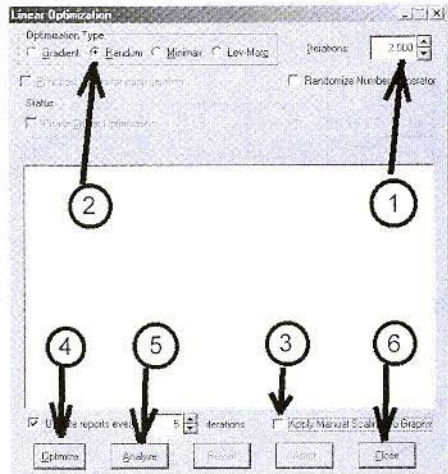


Fig 32: Closing Menu

[4] APLAC manuals (supplied when program is downloaded)

[5] Microwave Engineering by David Pozar (John Wiley & Sons, New York, ISBN 0-471-17096-8).

[6] Microwave Filters, Impedance-Matching Networks, and Coupling Structures by G. Mattiacci, L. Young and E. M. T. Jones. (Artech House Publishers,

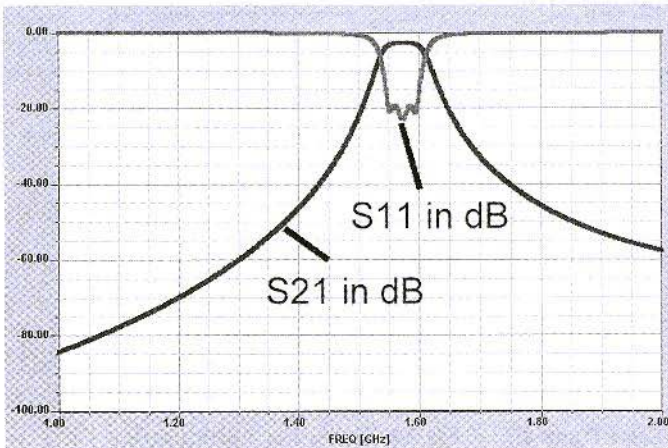
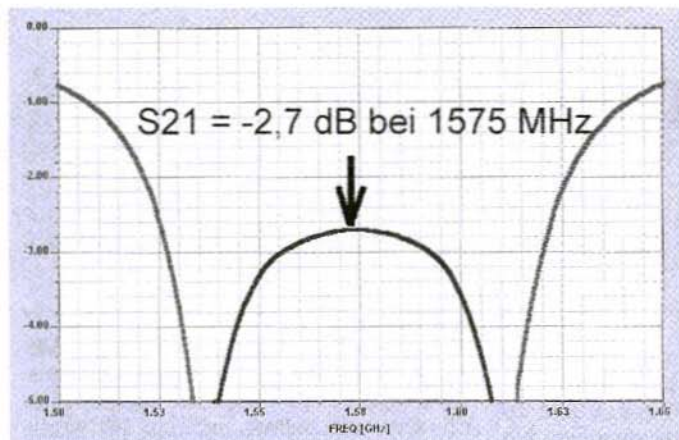


Fig 33: Simulation Results



Fig 34: Simulation Results



ISBN 0-89006-099-1).

[7] Microstrip Lines and Slotlines by K: C. Gupta, Ramesh Garg, Inder Bahl and Prakash Bhartia. (Artech House Publishers, ISBN 0-89006-766-X).

Software, manuals and tutorial are combined on an "ANSOFT-CD" and can be obtained from the author, provided the costs are reimbursed. Please E-mail me at: krausg@elektronischule.de.

Fig 35: New Circuit Data



Gunthard Kraus, DG8GB

Internet Treasure Trove

SONNET Lite 7.0

The manufacturer of the well-known EM simulator, which has already been introduced here in Part 1 of the article on modern designs for patch antennas, has improved the free "Lite Version". SonnetLite 7.0 now offers expanded simulation options. Thus, for example, there are now more ports available, the online help has been rewritten, pared-down algorithms allow more extensive simulations to be carried out with the same main memory restrictions as previously, etc..

In addition, a version is now on sale with max. 32 MB main memory, but for \$ 500.

Address: <http://www.sonnetusa.com>

Applied Microwave and Wireless

The well-known American technical review has now archived almost all technical articles up to 1989 like the "Microwave Journal" and they can be accessed through the homepage. Some highly interesting subjects are covered, although even here the conspicuous copyright stamp must be respected, including the ban on reproduction. Copies of material can be obtained without charge following registration.

Address: <http://www.amwireless.com>

MINI-CIRCUITS

A well-known name for a very wide range of HF and microwave components (couplers, mixers, amplifier components, dividers, attenuators, etc.....). Its worth visiting their homepage, as there you will find not only numerous documents for downloading but also an endless succession of new ideas. The most recent products are passive mixers with fabulous IP3 values going up to +38 dBm, or the interesting and reasonably priced "Blue-cell Mixer".

Address: <http://www.minicircuits.com>

APLAC

The Finns in the far North are also hard at work: now APLAC 7.61 rolls up as a free student version with the usual advance publicity and improvements.

The offer of a 45-day full test version is very advantageous. As soon as you are actually up to speed with the student version and are coming up against restrictions in using it, due to the size of the project you're tackling, you apply for a license and continue to work on the problem with the full version. Although the computer used must form part of a network!

Address: <http://www.aplac.com>



PCAAD40

If this heading sounds somehow familiar, there's a good reason why: version 2.1 was recently introduced in VHF Communications in the section entitled "An interesting program".

The author, David Pozar (who has also been responsible for some very good technical books on microwave techniques) has developed the program quite considerably. There is now a large WINDOWS 32-bit version, with over 30 different models of antennas alone for investigation and analysis. Unfortunately though, it is no longer free, but costs \$ 390.

Address:

<http://www.antennadesignassociates.com>

VECTRON

A manufacturer which has specialised in SAW filters, and particularly in crystals and crystal oscillators of all kinds (TCXOs, VCXOs, OCXOs, etc...) for all conceivable frequencies.

The data sheets for each product are in pdf format. They are very attractive, thorough and informative, as are also the large number of application notes available for downloading on almost every subject in this area.

Address:

<http://www.vectron.com>

TACTRON

This firm has for years been a supplier of American microwave components and simulation software such as the MICRO-WAVE OFFICE, etc.

On the homepage you can naturally find documents covering the complete product

range, but there are also some very interesting new items:

Under the heading "myTactron" an IIF electronics "flea market" has been set up, where anyone can offer surplus components, new or used equipment, stickers, etc. at no charge.

Address:

<http://www.tactron.de/mytactron>

ARCOM

Anyone wanting to become more closely acquainted with dwarf wave techniques will find a manufacturer of power amplifiers, low noise amplifiers, down and up converters, transmitters and receivers here. Not only is it interesting to glance at the many data sheets available, but it is also worth keeping an eye on the new "Education Page". This is intended to become a forum in which like-minded people can exchange and extend their knowledge.

Address:

<http://www.arcom-inc.com>

IONBEAMMILLING

You read it right: not only is that the firm's name, but they "really do carry out milling with ion beams".

Here all possible microwave and millimetre structures (couplers, attenuators, filters, etc.) are marked out, for example on gold-coated ceramic plates, using the "atomic sand blasting process" (actual quotation from firm!). The way the procedure works is also precisely described. Truly the most modern high-tech manufacture - genuine information for those interested.

Address:

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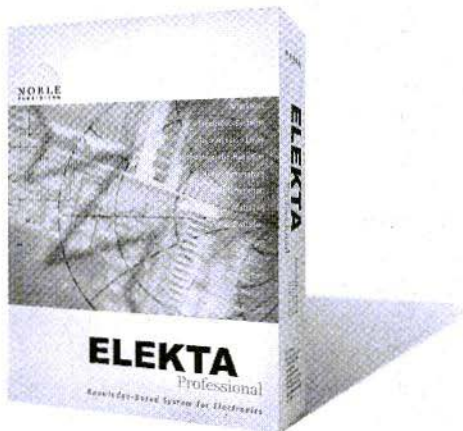
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