

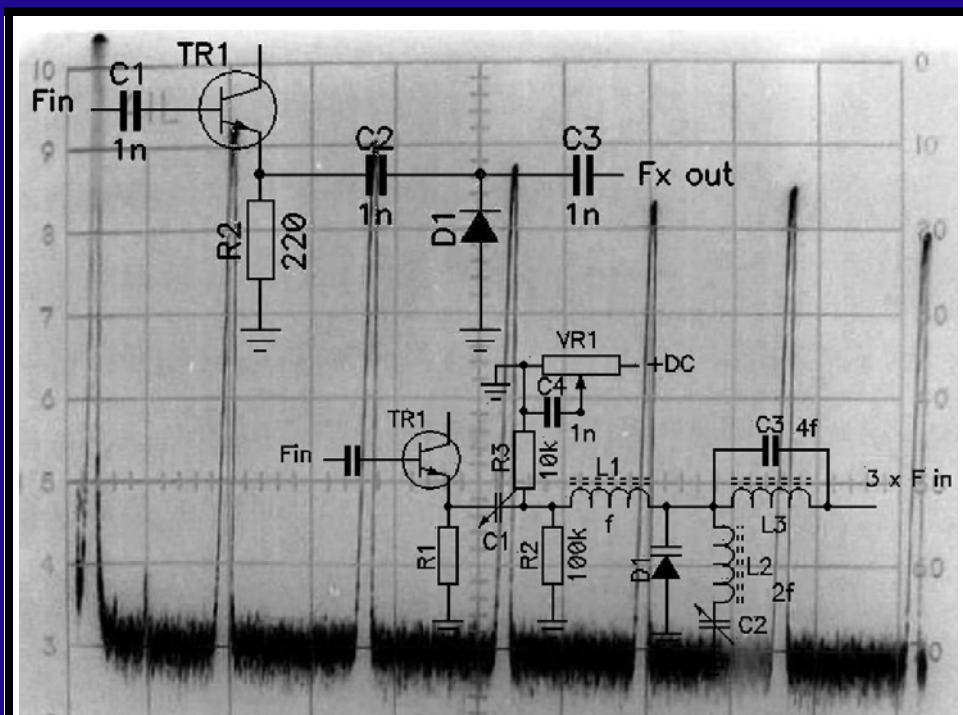


*A Publication for the
Radio Amateur Worldwide*

*Especially Covering VHF,
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Diode Multipliers

John Fielding ZS5JF

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John Fielding is a new author to VHF Communications Magazine, he lived and worked in the UK until he moved to South Africa and has a wide experience in amateur radio. He sent me two interesting articles on multipliers, the first one is in this issue and the second on transistor multipliers will be in issue 4/2006.

I had the article by Dragoslav Dobricic on Ultra linear low noise preamplifiers translated to go in the two RSGB publications; Microwave Projects 2 and The RSGB Handbook. I thought I should publish the article in this magazine as well, the full article for all bands plus PCB layouts can be found on the web site shown in the references.

73s - Andy



K M Publications, 63 Ringwood Road Luton, Beds, LU2 7BG, UK

Telephone / Fax +44 (0)1582 581051, email : andy@vhfcomm.co.uk

web : <http://www.vhfcomm.co.uk>



Gunthard Kraus, DG8GB

An interesting program: ANSOFT designer SV 2.2

The filter calculator in ANSOFT designer SV (student version) was used in my article in the 1/2005 and 3/2005 editions of VHF Communications, it only showed a small sample of its possibilities. The complete software is a free, capacity limited, version of the professional and expensive microwave CAD program ANSOFT DESIGNER, but not for commercial purposes. In the following article I will describe the effort required to use this software and the rewards, demonstrated using a practical project.

1.

Overview

Characteristics: Linear circuit simulator with the emphasis on stripline circuits. It is a filter program, a Line calculator (you still remember TRL85) and starting from version 2.2 also a simple EM simulator with integrated close and far field option.

Best features:

- Good prompting with very comprehensive menu structure as well as outstanding on-line help. The complete manual of the full version can be

downloaded from the ANSOFT web site.

- There are no limits on the circuit size, the circuit type or the occupied main memory, so you can “play without limitations”!
- The filter calculator is very useful, it works precisely so it is unnecessary to use any other free filter programs such as fds, faisyn, filtry etc.
- The Line calculator is a development of TRL85. For example “Coplanar microstrip waveguides” are supported. The results of computation can be transferred immediately to the attributes of the line concerned in the circuit.
- Hundreds of different constructional models are available with no restrictions, none of the familiar stripline parts are missing.
- The simulation accuracy is very high.
- Extensive collection of advice, Application Notes, questions and answers (FAQ), technical information etc. in a free of charge “Knowledge Base” on the ANSOFT web site. This is for the full version, but much is directly usable with this free version.

Bad features:

- The nonlinear simulation is disabled, likewise the automatic tuning and optimisation option.



- The Gerber Plot option to produce a file for PCB manufacture is switched off.
- The EM simulator has close and far field options, however because the number of cells it is very limited and is only applicable for small structures.
- Operating system: at least Windows 2000 or XP

Bugs or calculation errors: The author discovered a bug in the simulation of strip line low pass filters. ANSOFT confirmed this when contacted and promised to correct it in the next version (as a comfort: testing the full version showed the same bug). In such a complex piece of software it is unavoidable but not tragic.

Desirable Improvements: A special request for ANSOFT: please extend the range of the EM simulator to something like the power of SonnetLite (the free version with 16 megabytes of main memories). The close and far field option for antenna simulation are missing there, so ANSOFT would then have an advantage.

Download and file size:

- Over 100Mb as a packed file.
- Application Notes, technical information, FAQs etc. from the Knowledge Base must be downloaded separately and stored separately.
- Internet address:
<http://www.ansoft.com>

2.

Introduction

All of your experience does not help: anyone who has only worked with PUFF so far will get a shock when they start

their first design because you are confronted with a typically complex Windows screen with some windows and many buttons. Unfortunately the “starting procedure” is almost as complex for a circuit with a capacitor and a coil as a complicated microwave circuit. As soon as you get used to the controls things get better and better. You can get fast results to many difficult problems with precision simulations.

3.

Development of a 1575MHz stripline band pass filter using coupled lines (edge coupled type)

3.1. Project design

This type of filter construction is frequently used because striplines are easily reproduced and economical to manufacture. The mechanical size reduces with rising dielectric constant and frequency, but they are well suited to circuit simulation and can be manufactured accurately. The initial results for this of the project are shown as a PCB layout in Fig 1, but some explanation is required:

The direct output to a PCB layout program is disabled in the student version of the program. Therefore the author works with the free demonstration version of TARGET3001, which is well suited for these draft layouts and has sufficient capacity. There is a trick to tilt the lines, they are first drawn accurately with a CAD program with straight lines as they would be on the top surface of the PCB. This is then copied onto an empty PCB layout, rotating it slightly until the 50Ω striplines at the input and output are at the centre of the left and right hand edges of the PCB.

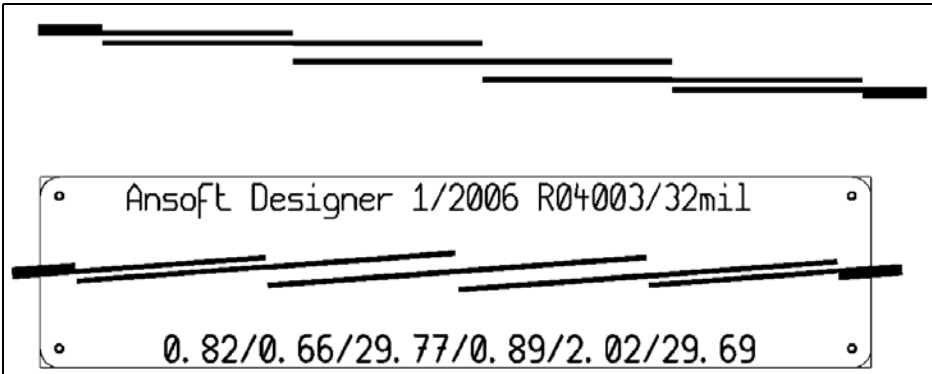


Fig 1: PCB layout for an Edge Coupled Microstrip Bandpass Filter for the GPS frequency of 1575MHz. See text to see how to produce this layout.

But now to the filter. The defaults for the filter circuit (Chebyshev type) are:

- Input and output impedance: $Z = 50\Omega$
- S21 Ripple in the pass band: 0.1dB (result in $S11 = -16.4\text{dB}$)
- Filter degree: 3
- Centre frequency: 1575MHz (GPS receive frequency)
- Bandwidth: 50MHz
- Printed circuit board; Material: Rogers R04003, Thickness: 32 mil (0.813mm), Dielectric constant: 3.38, Copper layer (thickness): 35 μm

(1.35mil or 1oz), Surface roughness: 2 μm (0.08mil), Cover height: 14mm

3.2. Draft using the filter designer

First SV designer must be downloaded, installed and started. Once started the screen shown in Fig 2 is displayed, the different windows are for different tasks, please look at these and remember the different meanings.

First, open the Project pulldown menu from the upper menu bar and click on "INSERT filter Design". Fig 3 shows the filter design wizard, now carefully select

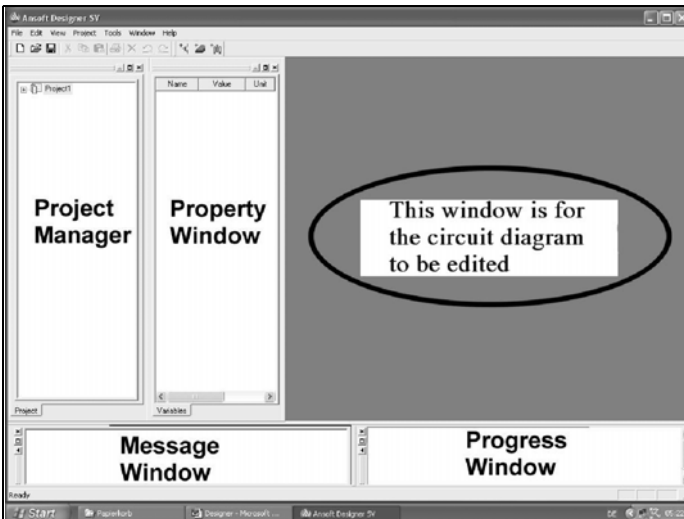


Fig 2: Most of the information is shown in these five windows, make sure you know their significance.

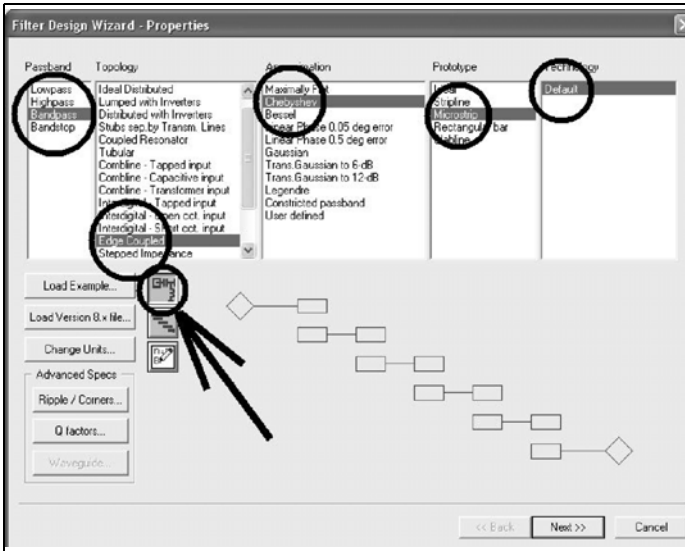


Fig 3: Use the mouse to select the filter type required with coil and capacitor.

the following options from left to right:

- Band-pass filter
- Edge coupled
- Chebyshev
- Microstrip
- Default
- Make sure that the key with the coil and capacitor is pressed and click on

NEXT

This takes you to the next screen shown in Fig 4 that displays the specifications:

- Filter degree = filter order: 3
- Ripples = Ripple [dB]: 0.1
- Centre frequency = f_0 [GHz]: 1.575
- Range = BW (bandwidth) [GHz]: 0.05

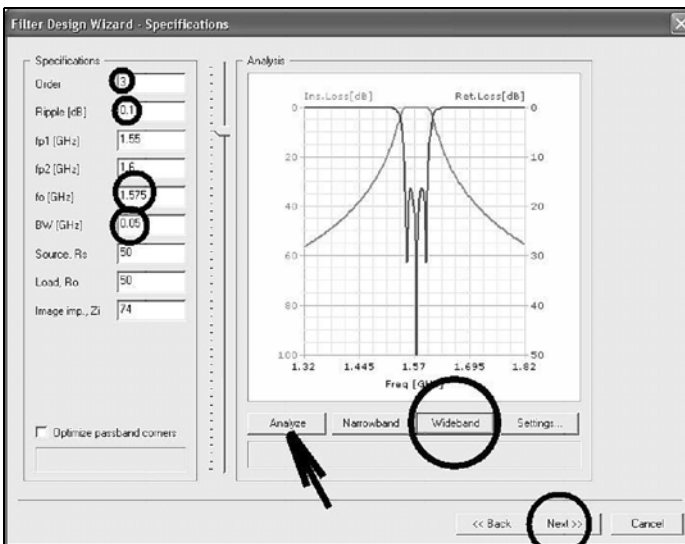


Fig 4: The filter data is displayed and can be viewed as required.



Fig 5: The data for an ideal filter is already entered, the PCB data must be amended.

Clicking on the Analyze button completes all missing values (e.g. the lower and upper critical frequency). Please try it out!

The lower three lines are left unchanged, the input and load resistance are 50Ω, the Image Impedance (Z_i) is set carefully because it is crucial to the layout of the final product:

- If one selects a very small value (e.g. 10Ω), the lines are shapeless and more than 15mm wide.
- If one selects a very high value (e.g. more than 100Ω), both the feeder widths become tiny (under 0.01mm) so cannot be manufactured at a reasonable cost.
- From experience the author has found that the proposed value of 74Ω is a reasonable compromise.

The right hand half of the screen shows the expected curves for S11 and S21. There is a wonderful facility to switch between Narrowband and Wideband. Selecting Wideband immediately shows a larger frequency range. Each filter design is simulated completely with all un-

wanted effects, it is possible to view many different parameters, just click on different settings and test the effect – Unbelievable! After each press of OK the result of the analysis appear immediately!

Pressing NEXT displays the next input sheet shown in Fig 5. The upper windows are for entering the PCB material, here it is R04003 material:

- Substrate dielectric constant, ϵ : 3.38
- Substrates height, h [mm]: 0.813
- Conductor thickness, t [mm]: 0.035
- Cover height, ch [mm]: 14

The last entry “frequency, f [GHz]” is automatically set to 1.5748016, based on the selected centre frequency of 1575MHz, by the program and cannot be changed.

Practice: By clicking with the mouse in each input field you can examine whether it makes a difference to the simulated data. When you have calmed down from the practice, check the correct list of variables is:

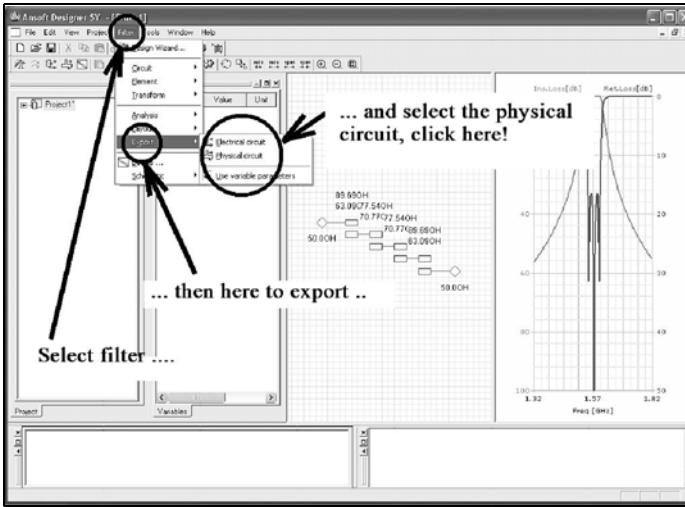


Fig 6: The ideal filter has been simulated, this is how to realise the physical circuit.

Variables:

- w1 = 0.8175mm
- s1 = 0.678mm
- p1 = 30.22mm
- w2 = 0.8874mm
- s2 = 2.015mm
- p2 = 29.97mm
- w3 = 0.0

The meaning of the designations is not difficult to understand:

- “w” = feed conductor for the concerned coupled line,

- “s” = gap width
- “p” = the physical conductor length.

If NEXT is pressed and then FINNISH, a picture of the filter is displayed on the screen including Even and Odd resistances as well as the transmission and reflection characteristics.

However the design is not yet completely finished, because this shows the ideal solution and in practice there is more to be added. Fig 6 shows what must be done to get to the “Physical Circuit”. The benefit of doing this is shown in the design shown in Fig 7.

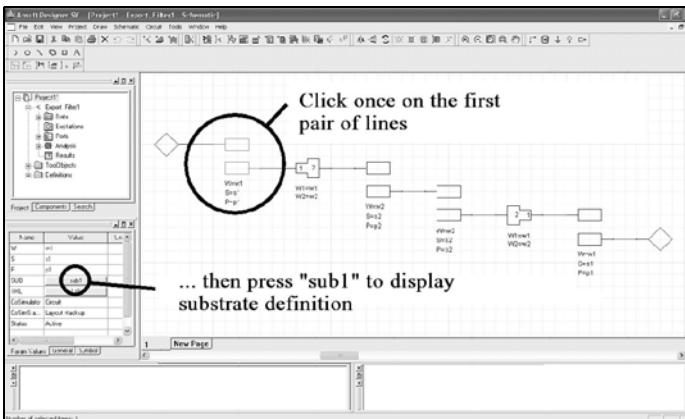


Fig 7: This is the simulation data, every part of the design can be modified.

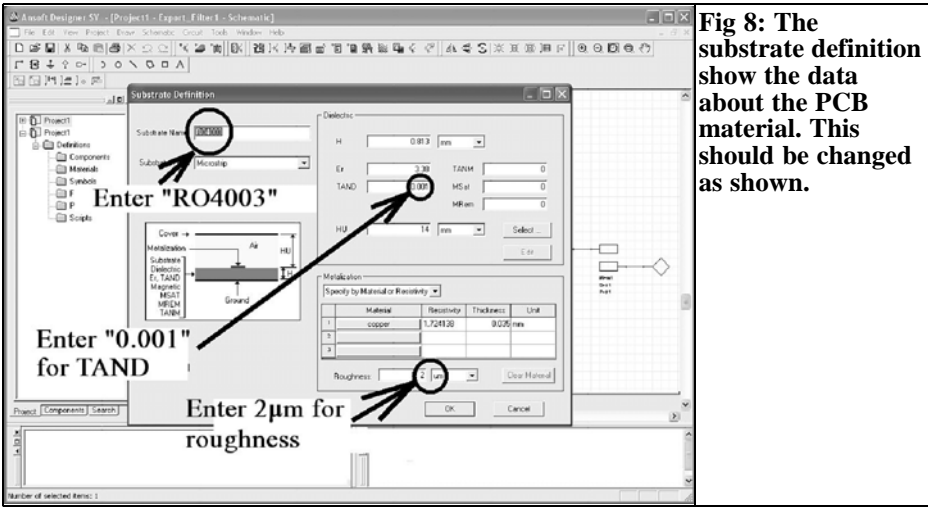


Fig 8: The substrate definition show the data about the PCB material. This should be changed as shown.

Suddenly the filter circuit is complete with all necessary practical additions necessary for a realistic simulation! But before this simulation the following measures are necessary:

Click the first coupled pair of lines ONCE which opens it's attribute menu in the "Property Window". Now press the "sub1" (substrate) key once to go to the menu for the substrate (SELECT substrate). Click on Edit to get the complete input window for the PCB. Fig 8 shows how all entries should read, most are already set correctly but some must be set correctly (i.e surface roughness = 2µm, substrate name "R04003" or dissipation factor TAND = 0.001). Check to make sure that everything is correct and accept the changes by pressing OK.

By the way: There is a substrate list in the designer library that contains the R04003 material. However the entry shows TAND = 0.0027, this value applies to 10GHz but the filter is working at 1.5GHz. TAND changes with the square root of frequency so the correct value must be used.

$$TAND(1500MHz) = 0.0027 \cdot \sqrt{\frac{1.5GHz}{10GHz}} = 0.001$$

Everything clear? Once each entry has been made and the substrate has been given the name "R04003" press OK to save the settings. If the material specification or the steps between lines are not set up the simulation will abort.

Now for the actual analysis: Press the "Analyze" button (Fig 9), if there is a problem with the set up you will be notified immediately. If the simulation runs without problems you can look at the "Messages" and/or the "Progress Window". However, after the simulation is complete there are no results displayed because the program wants to know how

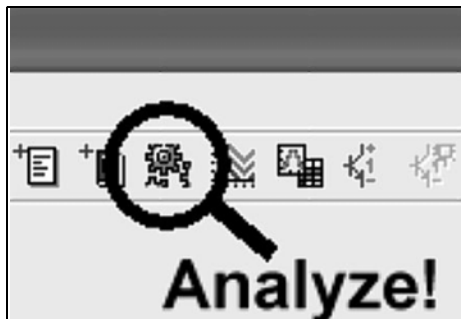


Fig 9: Once all changes has been made, press the Analyze button.



Fig 10: After the analysis, press the create report button.

you would like to see the results. To do this press “Create report” (see Fig 10) to see the options available. Take a look at the options for “Display type”, the default is a “Rectangular plot” but it can be changed if required to Smithchart for example.

Leave it as “Rectangular plot” this time and confirm by pressing OK. Then there is the agony of more choices; the program shows precisely what it can do (Fig

11) with a selection of more than a dozen variables that can be shown. Select reflection factor “S11” under S-parameters and add to the task list with “ADD trace”, repeat with S21. Pressing “Done” finally shows Fig 12 on the screen. With hindsight it is easy to see that the steps in the circuit are causing a problem. This can be seen most clearly in the S11 curve, optimisation is required which will be discussed in the next paragraph.

3.3. Physical circuit definition and optimisation

The finished filter is to be built into an aluminium housing, therefore it needs a 50Ω input on the left and a 50Ω output on the right. Unfortunately this requires some additional work to add these.

Step 1: Store the project where you will be able to find it again using a reasonable name. From my own experience I use a new folder called “own work” in the designer SV Folder. I store the project in that folder with a file name like “BPF1575_01” with a description of “band pass filter 1575MHZ”. Using the mouse in the “project manager” the complete menu structure can be expanded as shown in Fig 13. All the items that you may need to use later can then be seen.

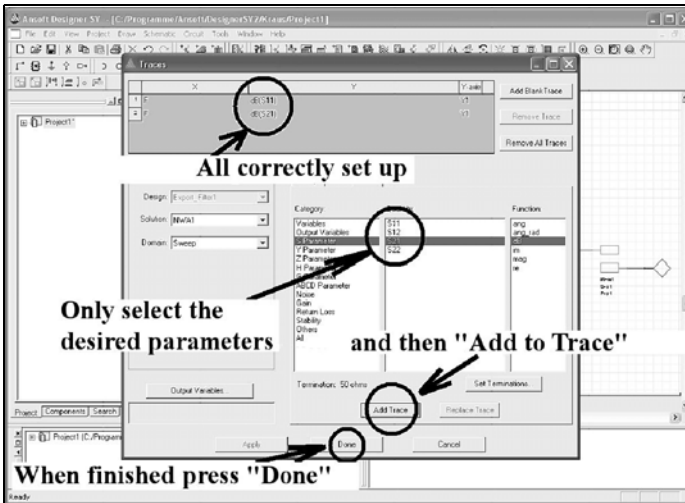


Fig 11: Select S11 and S22 from the list of parameters to display.

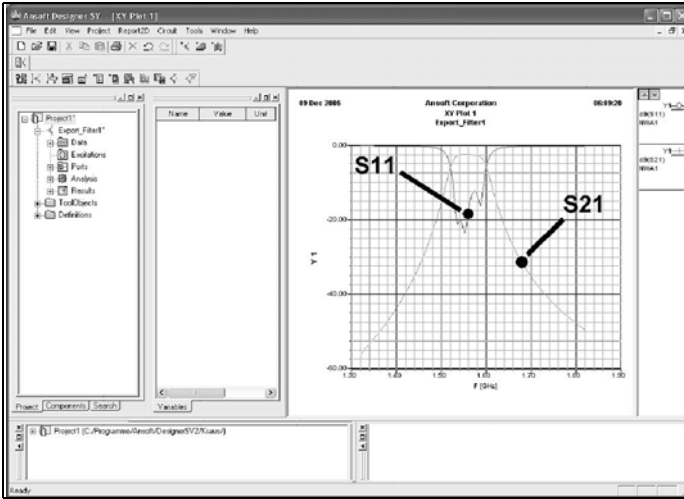


Fig 12: The display of the filter shows the problems with the first simulation.

Step 2: Double click the port 1 symbol on the circuit diagram, this shows the port definition (Fig 14) where it can be marked as a Microwave port, pressing CTRL+R rotates the port symbol to your desired position (Fig 15). Repeat this for port 2. To create space for the inputs and steps, delete the feeder lines by clicking on them to select them and use the delete key.

Step 3: In the “Project Manager” win-

dow, change to Components / Microstrip / Transmission line and then locate the "Transmission Line, physical length version" with the exact name:

MSTRL: Ms of Trans. LINE,Physical length

With drag and drop (press the left mouse button while over the option and move it by moving the mouse to the desired location and release the mouse button) move this physical length version of a transmission line onto the circuit diagram in two places. When it is dropped there

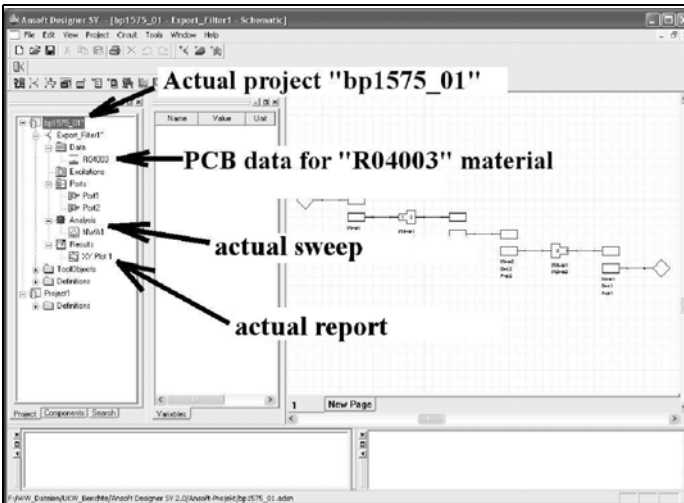


Fig 13: All the data about the project is available from this menu.

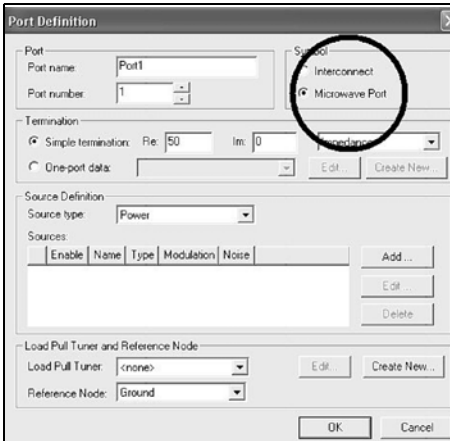


Fig 14: The port definition can be changed from "Interconnect" to "Microwave Port".

will be a question, "Merge Layers", this should be accepted. Now locate (under the same path) the Microstrip step option:

MSSTEP: Ms step

Again move two microwave steps onto the circuit, there are two possibilities:

1) Move the Microstrip step over the circuit diagram until it lines up with the

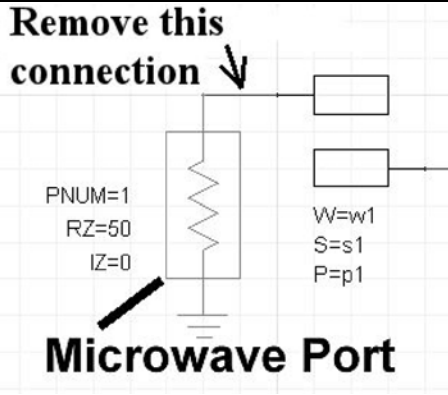


Fig 15: Remove the connection so that the steps can be added.

adjacent component. When releasing the mouse button the connection is automatically made and it will remain connected if you move the component.

2) The component can be placed and connected using the "Wire" tool. Click on the "Wire" tool to select it and then hold down the left mouse button while over one connection point, unrel the wire by moving the mouse to the second connection point then release the mouse button, see Fig 16. To stop the wire connection, press ESC.

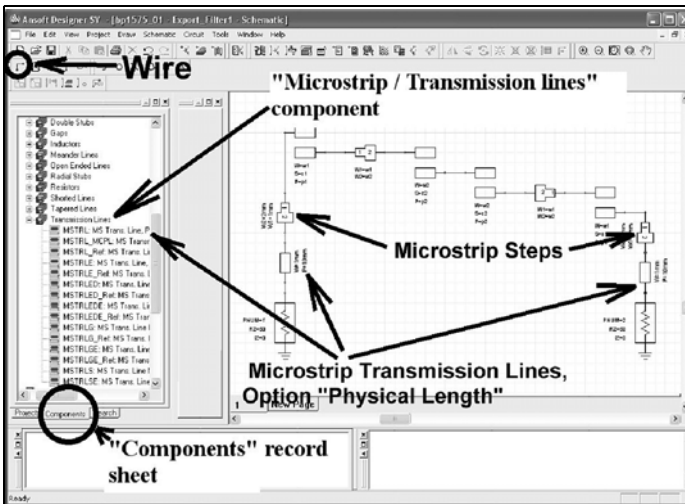


Fig 16: The microstrip steps and microstrip lines are added.

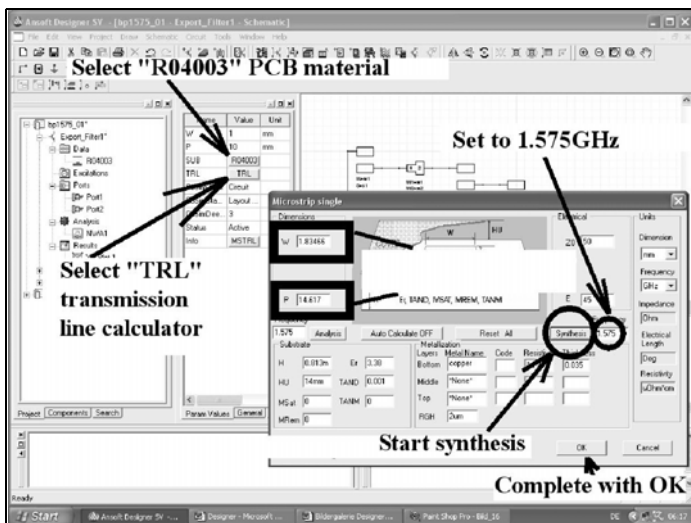


Fig 17: The transmission line calculator, TR85 is used to calculate the line parameters and add them to the line and it's properties.

Step 4: Click on port 1 to open the properties menu for the left hand 50Ω stripline, then click on the R04003 button (fourth line) to open the PCB attributes edit window, check that everything is correct and press OK to return to the properties menu. Press the CTRL key (fifth line of properties menu) to open the familiar transmission line calculator, TRL85. The PCB data is displayed as shown in Fig 17, it is only necessary to change the operating frequency to 1.575GHz. Check that the characteristic impedance is 50Ω. The electrical length is calculated for 45 degrees using the

“Synthesis” button, one press of OK calculates the result (width $W = 1.83466\text{mm}$, length $P = 14.617\text{mm}$) and moves them into both the circuit diagram and properties menu, see Fig 18. Repeat this procedure for port 2.

Step 5: Next the two Microstrip steps must be adjusted. Click on the left hand Microstrip step to open its properties menu, see Fig 19. For the line W1 enter “w1” which is the conductor for the first pair of coupled lines. On the next line, W2, enter the width of the 50Ω input as 1.83mm, there is no need to enter more

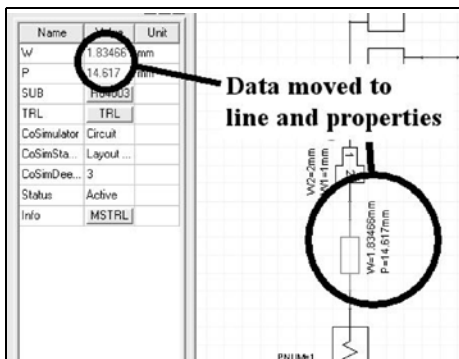


Fig 18: The line data is added to the circuit and properties.

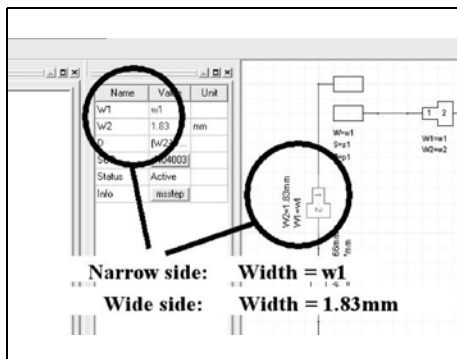


Fig 19: The data for the microrstrip steps are added.

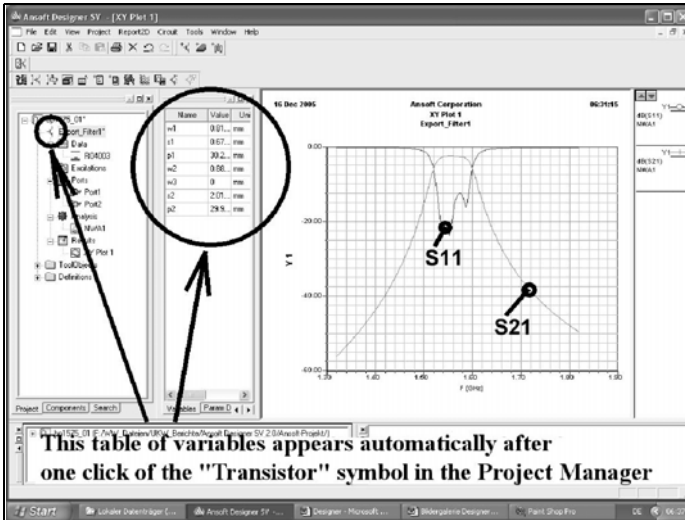


Fig 20: The new simulation show the affect of the steps. Optimisation is required.

than two places after the decimal point because no human can etch a stripline that accurately. Repeat the same procedure for the right hand microwave step.

Step 6: This step is the same as previously, press the “Analyze” button to let the program analyse the circuit and display the S11 and S22 curves, see Fig 20. The deviation from ideal has naturally become greater because of the additional steps but this can be eliminated by manual optimisation.

First all of the mechanical dimensions should be rounded to two decimal places.:

- Leader width w1 = 0.82mm
- Gap width s1 = 0.68mm
- Leader length P1 = 30.22mm
- Leader width w2 = 0.89mm
- Gap width s2 = 2.02mm
- Leader length p2 = 29.97mm

Step 7: Do not proceed with the optimisation blindly or like a bull in a China shop but use the following system:

- 1) The first goal is an approximation for S11, a value of -16.4dB in the pass

band with equal peaks is good. The centre frequency should not be changed so first change the conductor lengths of the coupled lines until the peaks are the same height. Then make careful changes to the gap widths until the ideal of -16.4dB is reached..

Normally this is sufficient; if that should not be the case, the leader widths can be changed:

- 2) It is only necessary to worry about the centre frequency of the filter, usually any errors cause this to be shifted towards the higher frequencies. All that is needed to adjust this is to use a pocket calculator to shorten the conductor lengths in the relationship of the simulated centre frequency to the desired centre frequency. Normally only minimum changes are required.

The function of the circuit will now be investigated. First we must look at the S11 and S22 curves (if these are not visible double click on “XY Plot 1” under “Results” in the “Project Manager”). To read the information more easily you can zoom in by clicking the right mouse button over the curve, the desired zoom level and position can be selected with the mouse. It is also possi-

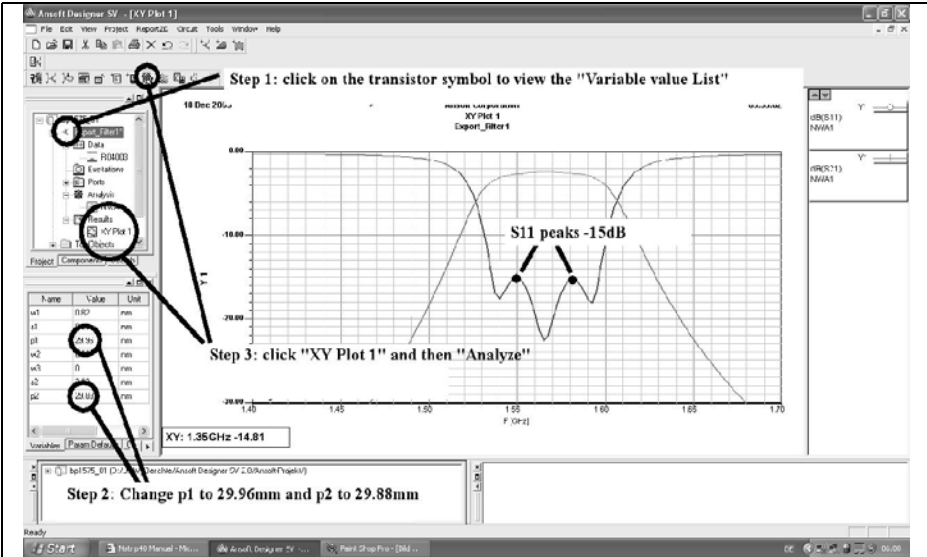


Fig 21: The changes made show that the S11 curve is now symmetrical.

ble to manually change the x or y axis divisions, e.g. double click on the y axis to display the “ Y axis Properties (Y1)” where the scaling can be found. Here you

can switch off “Autoscaling” and under “Manual Scaling” change not only the range and scaling but also the fine scaling. Please take time play with this and

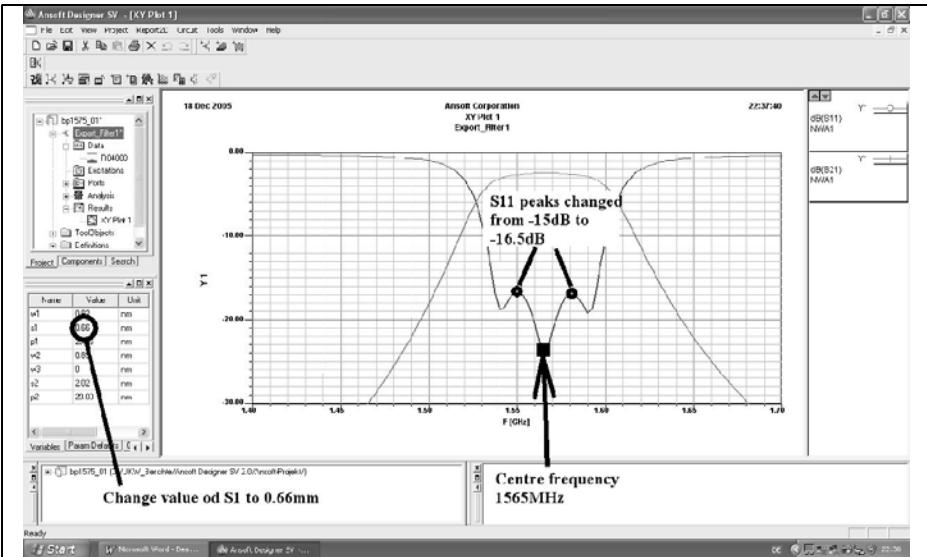


Fig 22: The changes made show that the S11 maxima are now acceptable.

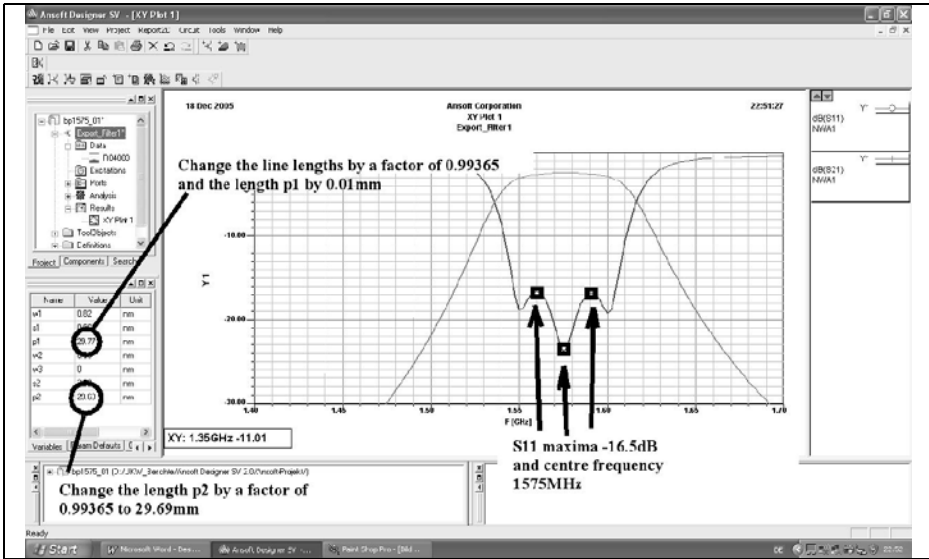


Fig 23: The final adjustment correct the centre frequency of the filter.

repeat for the x axis.

But now, see Fig 21:

- 1) The transistor symbol in the “Project manager” is clicked to see the list of the variables used.
- 2) In this list change the length P1 to 29.96mm as well as the length p2 to 29.88mm.
- 3) Then click on “XY Plot 1” to use the changed values correctly. Afterwards press the “Analyze” button to start a new simulation.

The new simulation results appear immediately and it can be seen that this is successful. The two S11 maxima are equal but have a value of -15dB which is still not good enough. Therefore the procedure is repeated:

- 1) Clicks the transistor symbol.
- 2) Change the value of the first gap width of s1 = 0.68mm in the variable list to s1 = 0.66mm.
- 3) Click on “XY Plot 1” and then on the

“Analyze“ button, now we see Fig 22 with the curve exactly as required but the centre frequency is 1565MHz instead of 1575MHz.

Therefore pick up the pocket calculator to shorten both conductor lengths by the factor:

$$\frac{1565MHz}{1575MHz} = 0.99365$$

And repeat the simulation as above. The result gives the desired centre frequency of 1575MHz, but again two somewhat unequally high S11 peaks. That is not a problem, because we already know that a small correction to the length of P1 by 0.01mm will eliminate the problem, see Fig 23.

Finished!

Thus a prototype PCB using the following line data can be made:

- Leader width w1 = 0.82mm
- Gap width s1 = 0.66mm



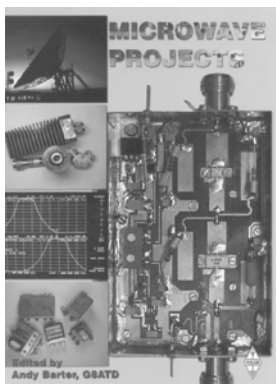
- Leader length $P1 = 29.77\text{mm}$
- Leader width $w2 = 0.89\text{mm}$
- Gap width $s2 = 2.02\text{mm}$
- Leader length $p2 = 29.69\text{mm}$

By the way: Did you notice, changes of only 0.01mm have a critical affect? This is a very high requirement for a layout or PCB manufacturer, but be prepared for a disappointment: despite all the effort put into the simulation, the first PCB manufactured will probably have clear deviations from the ideal results. The complete manufacturing process will expose inaccuracies, conductor under etching and

changes in the material data play a very large role. These additional errors can be corrected by further optimisation and after analysis of the Network analyser measurement results. Therefore the old developer wisdom is correct: only the third PCB version can be released for manufacturing of a small batch of product.

To be continued.

Microwave Projects Books edited by Andy Barter, G8ATD



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

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

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

- Signal Sources
- Transverters
- Power Amplifiers
- Test Equipment
- Design

Microwave projects is a must have book for all those who are already active on the microwave bands.



Following the huge success of Microwave Projects Andy Barter, G8ATD brings you more innovative projects from around the world. If you are interested in building equipment for the amateur radio microwave bands, the designs in this book are sure to please you. Projects have been selected from international authors and all of the projects use modern techniques and up to date components. Details of how to obtain ready-made boards are included with most projects. The book includes chapters covering:

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

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Johannes Schad, DG6NDS

Design of a Quad Yagi: Part 2

Continued from 1/2006

2.3. Unification of Long Yagi and Cubical Quad into the QUAGI

If the transverse electromagnetic wave system of a Long Yagi is combined with a Cubical Quad exciter, the result is a Quad Yagi or “QUAGI”.

The behaviour of the first director D1 when it is fed by a Quad loop is almost the same as when it is fed with a half wave dipole or folded dipole. This is explained because the Quad loop excites the antenna with an equivalent phase to the half wave dipole. The arrangement to feed a Long Yagi with a Quad loop is shown in Fig 18.

2.3.1. Gain of a Quagi

The gain improvement of a Quagi over an antenna of the same length and/or number of parasitic elements is essentially the same as the gain over a non-optimised Yagi. In various literature the statements about the excess gain of a Quagi in comparison to a conventional Long Yagi array of the same length contradict themselves.

In the Red Sheep antenna book [2], the chapter “Directional Antennas for VHF and UHF” on page 543 it has the following statement:

“For Yagi arrays with lengths over 1λ the

type of fed element has no influence on the gain, but does affect the frequency range and adjustment of the antenna. ... With short Yagis an increase in gain is achieved, if the fed element has an increased gain (e.g. Quagi). This increase in gain reduces with increasing antenna length “.

On page 580 of the Red Sheep antenna book it states that the Quagi has overall an excess gain of approximately 1dB compared to the Long Yagi.

This assumption is based on the following:

In the chapter on the Cubical Quad, the book states that it has a gain of 7.3dBi. From 2.1.1 we know that a dipole with reflector exhibits about 5.75dBi gain. Logically the conclusion is that the excess gain of a Cubical Quad at 1.55dB positively affects the gain of a Quagi. Since the gain of a Cubical Quad is greater by around 1.55dB, this means that the wave field running forward has higher energy thus a higher current flows in the directors. This current gives rise to higher resistive losses. This would mean that with an increased number of parasitic elements the Cubical Quad reduces the additional gain. Nevertheless I think an increased gain of 1dB is realistic for antennas up to 4λ long.

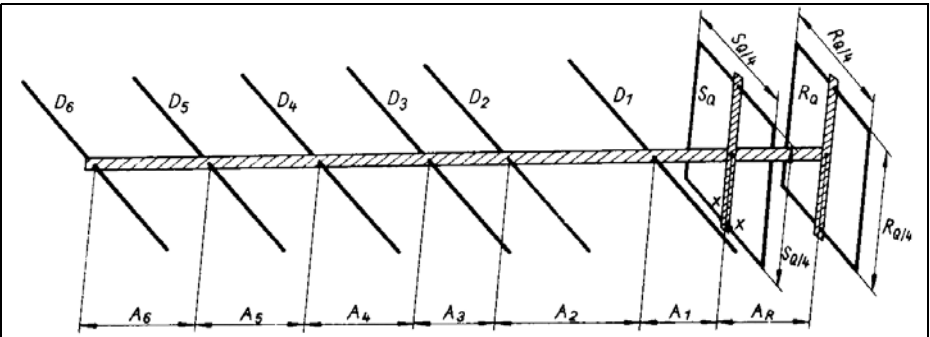


Fig 18: A Yagi antenna with a Cubical Quad feed - a Quagi.

Fig 19 shows that the increase in gain of a Yagi reduces as the number of elements increases. For this reason one hardly ever finds antennas with more than 12 to 15 directors for a given frequency range. Strictly speaking one would have to build an antenna twice as long for 3dB gain.

In practice a reflector or a combination of reflectors has only a very small influence on the gain and the directivity of an antenna. In most cases it is not worth the additional expenditure. If a highly reflective surface is used as a reflector an improvement can be achieved. However the reflected energy usually goes into the side-lobes therefore hardly any improvement in the front to back ratio is achieved.

2.3.2. Input impedance of the fed element

As already described, the distances of the parasitic elements affect the input impedance of the fed element. Because of their proximity to the driven element the reflector and first director have more influence than the other elements. The input impedance of a Cubical Quad is between 60Ω and 120Ω before this it is combined with the transverse electromagnetic wave system of a Yagi array, this can lower the characteristic impedance by adjusting the first director D1. Ideally the antenna would be fed directly using 50Ω coaxial cable.

The Quad loop is actually a balanced antenna, but it can also be feed unbalanced. A Quad loop will produce an offset field when fed like this. It is however not very pronounced because it is in a transverse electromagnetic wave system. Any surface waves produced could be eliminated with a surface wave trap.

On one hand the feed impedance can be adjusted to a useful value by moving the director. On the other hand the distances of the reflector and first director, from the fed element, set the gain of the antenna. Additionally the bandwidth of the fed element is dependant on the two parasitic elements. This is no problem in

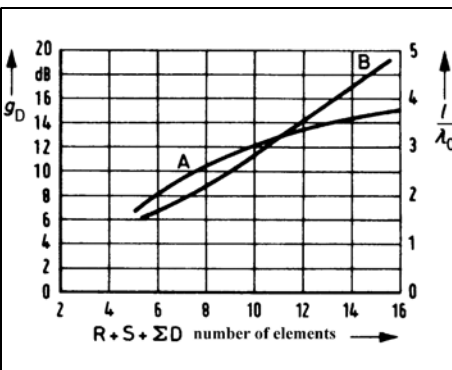


Fig 19: Antenna gain with increasing antenna length.

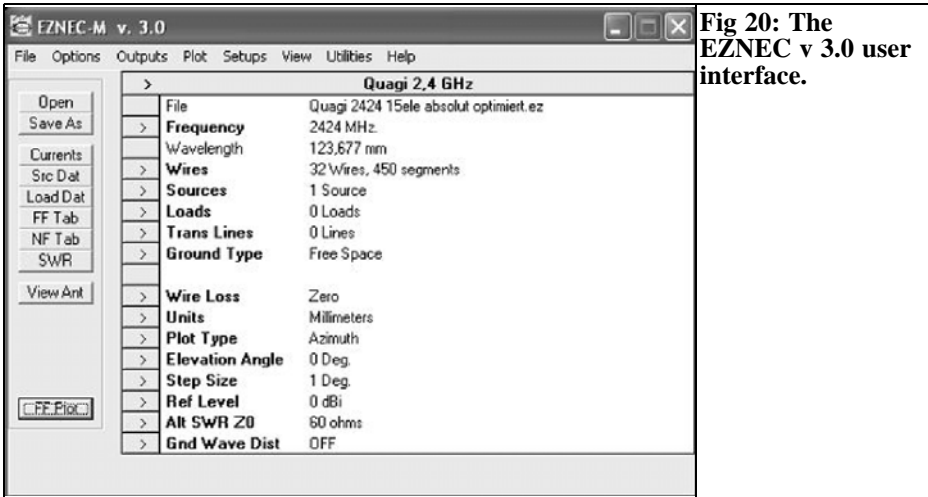


Fig 20: The EZNEC v 3.0 user interface.

practice since the Quad antenna is relatively wide band.

It is apparent that all of the dependencies make a mathematical solution nearly impossible. The dependencies only increase with an increasing number of elements. It is therefore impossible to manufacture an optimal Quagi, the desired characteristics must be specified and the necessary compromises that are required must be accepted.

The description of the Quagi antenna defined a few of mathematical formulas. During the construction and simulation of a Quagi many optimisation factors are required, this can only be achieved by computer assistance.

2.4. Simulation and optimisation

The program EZNEC PRO 3 by Roy W. Lewallen, W7EL was used for the simulation of the Quagi antenna. With this high performance program one can copy and simulate almost all antenna types. The antenna simulator routine makes fast computation of gain, impedance, SWR, current and power distribution as well as showing the radiation patterns. The operator interface of the program is shown in Fig 20.

The basis of the program is the so-called “Numerical Electromagnetics Code” or the “NEC engine”. It uses a moments method, calculus is used to generate a set of equations into matrix that can be solved numerically. Harrington who wrote the book “Field computation using the moments method” devised the moments method. An advancement of the NEC is the MININEC engine. It resulted from a program reduction of the large computer program NEC and forms the basis of today’s antenna simulation programs.

With the antenna simulation the antenna is developed from one or more wires, their diameter and material can be selected. The position of the wires is specify using a three dimensional coordinate system. The computation by EZNEC divides each of these wires into segments with a uniform current flow. The accuracy increases with the number of these sections, however this increases the computing time by a square factor. The author does not recommend going over 10 sections per single wire. The extensive adjustment facilities make this program one of the best antenna simulation tools on the market at present.

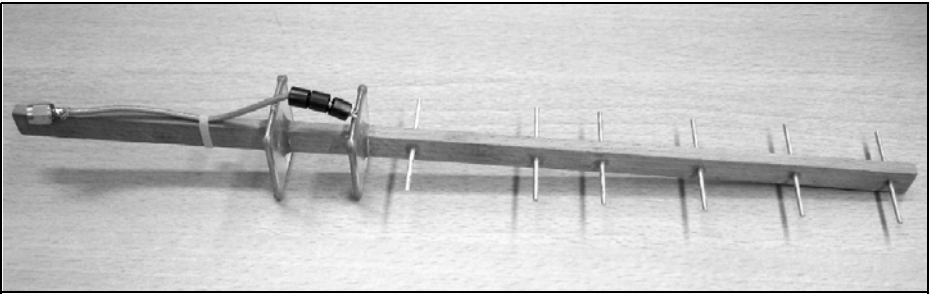


Fig 21: The 8 element Quagi for 2.4GHz with a Quad reflector, it is 25.1cm long.

2.4.1. The optimisation

I did not re-invent the wheel to start the simulation, the publications by Wayne Overbeck, N6NB [3] was used as a starting point. In his article “The Quagi Antenna turns 30” he describes 8 and 15 element Quagis. These are from older articles in QST, see introduction. Wayne Overbeck only computed and built Quagi antennas up to a frequency of 1.3GHz, now these had to be transferred to 2.4 and 5.8GHz. The 15 element long boom Quagis for 432MHz and 1296MHz formed the basis of the following simulation.

As a first step the data from Wayne Overbeck’s 432MHz Quagi was converted into metric units and entered into EZNEC. The results of the simulation corresponded, with small deviations, to the values in his publication.

Subsequently, this antenna was converted

by EZNEC to 1296MHz. Naturally the gain, VSWR and radiation pattern did not change. The dimensions of this antenna were compared and it turned out that the lengths and distances of the individual elements of the antennas became smaller by the factor 3. Naturally the element thicknesses were reduced by the scaling factor. This shows that Wayne Overbeck’s optimisation was good otherwise things may have been different. Because the computed values for the 1.3GHz Quagi almost agreed, the antennas for 2.4GHz and 5.8GHz were converted.

The 2.4GHz and 5.8GHz Quagis developed (Figs 21 to 23) were useful but only in the theory. Converting the distances and element thicknesses so that the same electrical characteristics as the original 432MHz antenna are maintained gives unrealistic values. A distance of 23.87393mm between two directors or the diameter of all elements of 0.78267mm would have to be changed.



Fig 22: The 15 element Quagi for 2.4GHz with a plate reflector, it is 59.5cm long.

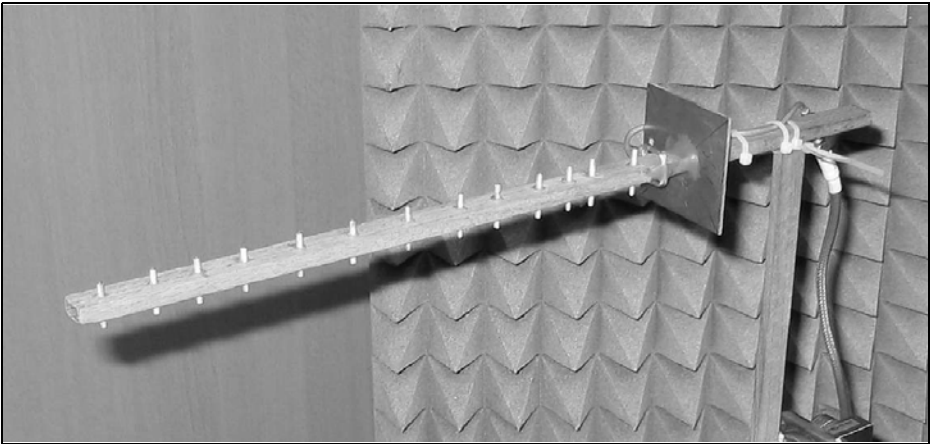


Fig 23: The 15 element Quagi for 5.8GHz with a reflector plate, it is 24.9cm long.

Unfortunately there is no simulator that will take a basic antenna design and produce an optimised design. So the optimisation must be done, in order prevent spending hours changing values in a senseless way the following procedure is advisable:

First, round all values to the maximum possible manufacturing accuracy. All elements should be the same diameter. For 2.4GHz and 5.8 GHz it is recommended that the element diameters should not be larger than 2mm.

Now try to get the feed element into resonance. As mentions above, the reflector and the first director affect its resonant frequency substantially. Therefore make small changes and watch the antenna radiation pattern and VSWR. Once the driven element is in resonance, try to adjust the length of the first and second directors for maximum gain. The resonance of the feed element must be examined and adjusted if necessary.

In principle, optimise the elements working through from the fed element forward. The reflector is less critical and can be optimised later. The effort and possi-

bilities for optimisation increases with the number of elements.

If you reach a point where optimisation is not progressing, i.e. a change causes a decrease; you can try to change the distance between the two elements. I advise against this however under normal conditions, since one will interact and cause additional optimisation problems.

When you reach a point where optimisation makes only changes of approximately one tenth of a dB, stop the optimisation. An additional few tenths of a dB further improvement in the antenna does not compare to the necessary additional effort that idealists would have you spend. Due to the manufacturing tolerances and measurement inaccuracies this possible increase in gain is probably not realistic.

After the simulation the antenna should be made with greatest possible manufacturing accuracy. With frequencies above 1GHz each small deviation affects the behaviour of the Quagi antenna. Aluminium was chosen to make the antenna, since it exhibits very good electrical characteristics and in addition is very

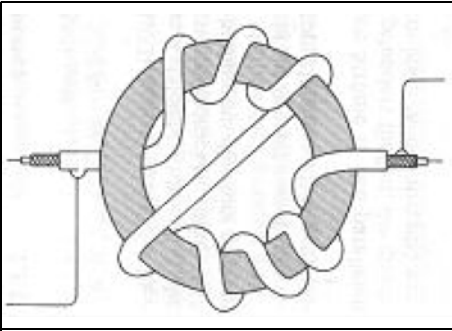


Fig 24: The ferrite ring trap used on the 15 element Quagi.

easy to work. The boom should be made from non conductive material. Instead of the reflector Quad a highly reflective surface made from printed circuit board material can be used.

The Quagi antennas were tested with different reflectors. Pine wood poles were used for the boom and the elements were made from 2mm diameter aluminium rod. Brass was used for the feed element because it is easily soldered. 2mm lacquered copper brazing rod could have been used instead of aluminium rod but it was not available. The best accuracy for individual components and the boom were obtained from FH Engineer-

ing with 100µm accuracy achieved.

Teflon coaxial cable type RG316 was used for the feeder. This has a very high attenuation but it can be used without problems in short lengths such as 15cm. Gold plated SMA plugs were used. Ferrite beads threaded over the coax of the 2.4GHz 8 element antenna were used as a trap to suppress any surface waves. A ferrite ring as described in [2, p.158] (Fig 24) was used as a trap for the 15 element antenna. The effectiveness of both traps could not be tested for the 5.8GHz antenna because the ferrite material is most probably no longer electrically suitable at this frequency.

2.5. Comparison of simulation and measured results

2.5.1. VSWR behaviour

If the Quad feed described by the formulae in 2.2.3 is made and tested it will have a resonant frequency slightly higher than required. If it is fitted to the Quagi system, the resonant frequency falls because of the proximity to the first director and/or reflector. If the computed spacing is used you can decide whether the element must be extended or shortened. Because the smallest changes to the elements have an affect it is difficult to find the best adjustment. If necessary the distances from director and reflector can be changed slightly. With the experimental set-up a voltage standing-wave ratio of 1:1.74 at the operating frequency should be sufficient, see Fig 25.

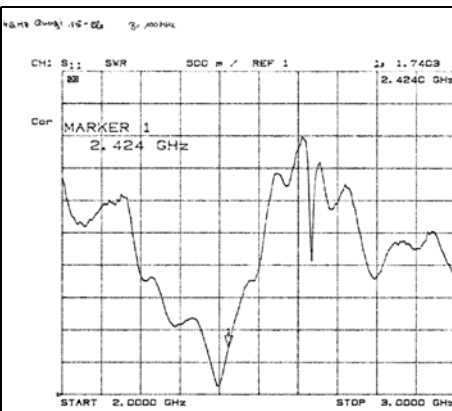
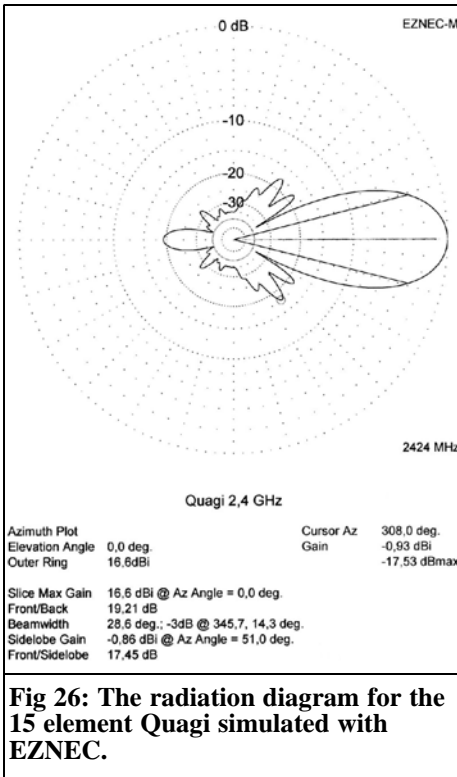


Fig 25: VSWR diagram for the 15 element Cubical Quad.

The bandwidth of the Quagi where the VSWR is better than 1:2 is approximately 80MHz. The Red Sheep handbook says [2, p.390] that a Quad array has a greater bandwidth than a half wave dipole. This behaviour can be confirmed because a Long Yagi with a half wave dipole feed element was also tested. The bandwidth of the Yagi was so small that an exact measurement was not possible. This was one of the reasons that it was



decided to make the following Quagi antenna.

2.5.2. Simulation of the radiation pattern of the 15 element Quagi for 2.4GHz

The result of the simulation using EZNEC is shown in Fig 26. The simulation was carried out in free space. In practice the antenna will be used without ground effects so this a valid simulation. According to EZNEC, the gain of the antenna is 16.6dBi. From Fig 19 a 15 element Yagi would have a theoretical gain of approximately 15dBd or 17.15dBi. Since the values from the diagram cannot be achieved in practice, both statements almost agree.

The gain can be increased in simulation to 19.21dB by adding numerous reflector

elements. This gain improvement is easy but it makes the forward radiation pattern worse, it is better to aim for 17.45dB to 18dB. It is up to the user to optimise the antenna to suit the particular application. Here great importance was attached to a good front to back ratio, because it is to be used in a radio relay system which is normal practice for short distance links. These links are limited by resonance absorption of gases in the atmosphere, particularly at higher frequencies. This can limit the range to a few kilometres, more information on this topic can be found in a thesis by Thomas Bergmann, DG8NTB [6].

The beamwidth of the simulated antenna is 28.6° . From the theory an increase in the number of elements reduces the beamwidth.

The front to side-lobe relationship or the side-lobe suppression of the simulated Quagi is 17.45dB and is uniform. This measurement only indicates the worst side-lobe suppression. To be accurate the relevant azimuth angle should also be specified. In practice a value of over 15dB is acceptable but the front to side-lobe suppression should be as high as possible, which often leads to optimisation problems. The advantage of a high side-lobe suppression is that stations to the side of the antenna are suppressed as much as possible. An ideal antenna would exclude everything except the forward direction, except for the horn antenna this is not practically possible.

The measured radiation pattern of the 2.4GHz 15 element Quagi is shown in Fig 27. The antenna was mounted centrally and the receive levels noted every 5° , the gain was measured into a short circuit.

Comparing the radiation patterns of the simulation and measured values it must be noted that EZNEC uses a logarithmic dB scale. With linear presentation the

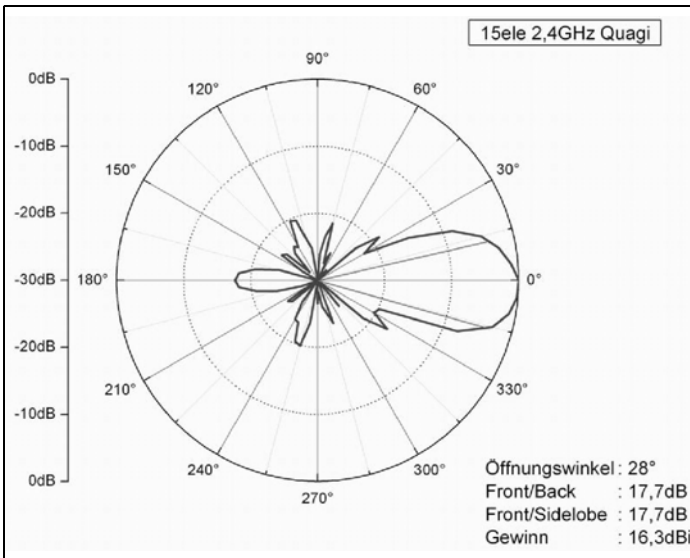


Fig 27: The measured radiation diagram for the 15 element Quagi.

side-lobe suppression is apparently worse, which is not the case.

The antenna plots were produced with ORIGIN 7.5, a data analysis and plot routine from OriginLab corporation.

Most side-lobes are approximately 20dB lower than the maximum level in the forward direction. The lobe at the rear of the antenna could be made better than 17.5dB. This would increase the energy in the forward direction but would also increase the two side lobes at, for example, 40° and 325°. Since these are around 18dB lower than the level in the forward direction, this is rather insignificant in practice.

The antenna has a 3dB radiation angle of 28°. This compares well with the simulation of the antenna. If the measured gain is compared with the simulation, they differ by about 0.3dB. On one hand this is connected with the fact that EZNEC simulates the antenna with a finite number at reflectors (15), but in practice a reflector of high conductivity copper laminate was used. On the other hand the antenna may not be manufactured accu-

rately to the simulated dimensions. For this reason the values for the front to back relationship and front to side-lobe relationship can be considered identical to the simulated values.

The horizontal radiation pattern of the 2.4GHz 8 element Quagi for 2.4 GHz is shown in Fig 28. It is noticeable that there is almost no commonality with the radiation pattern of the 15 element Quagi. That is because the 8 element Quagi was developed using it's own optimisation process, i.e. one cannot omit the first seven elements of an optimised 15 element Quagi and produce an optimised 8 element Quagi. The values are only a few millimetres different but that makes all the difference.

The gain is 13.8dB which is 2.5dB worse than the 15 element Quagi but the antenna is 25.1cm long, not even half as long as the larger Quagi (59.5cm). Chapter 2.3.1 showed that doubling the length of an antenna should, in the theory, give a gain increase of 3dB; in practice however only about 2.2dB is achieved. Additionally the reflector used was a Quad reflector not a metal plate. The radiation pattern clearly shows that this widened the rear lobe.

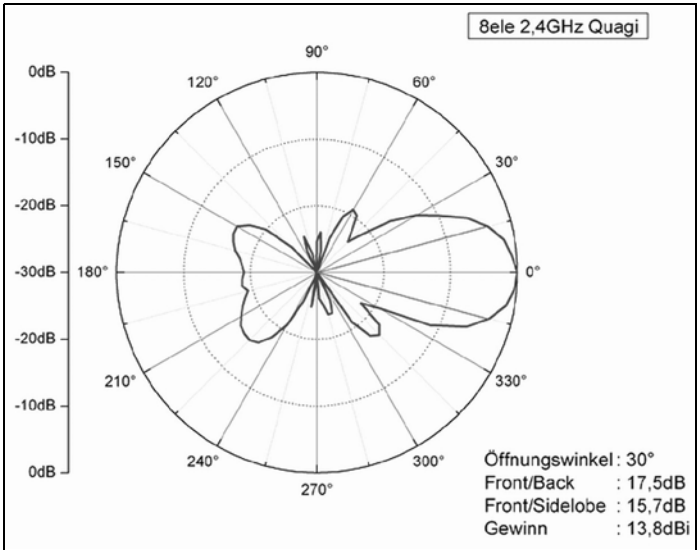


Fig 28: The measured radiation diagram for the 8 element Quagi.

The front to back relationship is 17.5dB which is amazingly good. On the other hand the strongest side-lobe is at an angle of 225°.

Also the statement that the beamwidth becomes smaller with increasing gain can be verified. The measured results showed only a very small difference of 2°, however this deviates from the simulation, results by 9.2°.

The radiation pattern of the 5.8GHz 15 element Quagi is shown in Fig 29. A clear main lobe is to be seen, however there are also large side and back lobes. The simulation resulted in similar values as for the 2.4GHz 15 element Quagi with a gain of 16.6dBi and side-lobe suppression of 17.3dB. There is a large deviation between the measured and simulated results. The following explanation describes why the results are so different:

The size of the antenna and / or the length of the individual elements plays a crucial role. The first is 1.96cm long and the last one is only 1.6cm long. The pine wood boom is 1.2cm wide and 0.6cm high which means that over a third of the

element length is in the wood. Thus the forward wave field is dampened, which naturally diminishes the gain. The side-lobes use more energy reducing the gain in the forward direction.

Although all the dimensions from the simulation were used exactly, small deviations occurred during assembly. The finished antenna is not just one piece, the individual elements had to be fitted by hand into the holes in the boom so that the distance on either side of the boom was the same. The problem becomes clear in Fig 30 and Fig 31. The last director on the 2.4GHz Quagi is approximately 4.3cm long and it can be seen that the wooden boom is not large in comparison. The last director of the 5.8GHz Quagi has over 30% of its length in the wooden boom.

The sides of the Quad feed element are only 6.7mm long and rest on the wooden boom. The Epoxy adhesive needed to fasten the elements covered more of the metal. Thus the electrical behaviour of the antenna is not as expected.

Making a Quagi antenna for frequencies

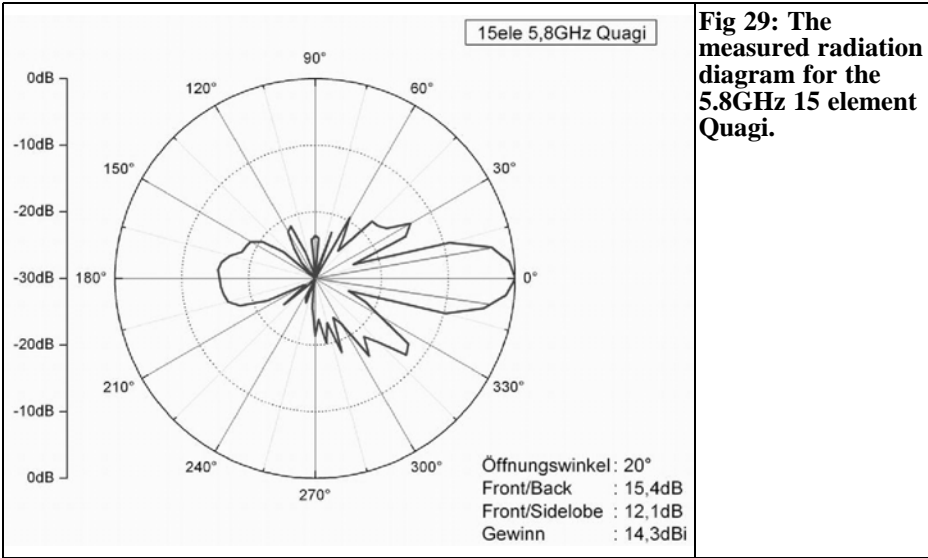


Fig 29: The measured radiation diagram for the 5.8GHz 15 element Quagi.

above 4GHz causes so many mechanical difficulties that can only be overcome by machine assembly. If a high gain antenna is needed for frequencies higher than 4GHz it is probably better to use a different antenna system e.g. Patch or reflector aerials.

Since ground effects could not be eliminated when measuring the antennas, no vertical radiation patterns of the Quagi antennas were taken.

2.5.3 Summary of the most important characteristics of Quagi antennas

The characteristics of this antenna are achieved by combining the advantages of a Yagi and a Cubical Quad. The Cubical Quad has a relatively large frequency range, combined with an uncomplicated excitation method as well as the high absorption of the reflector. The Yagi has

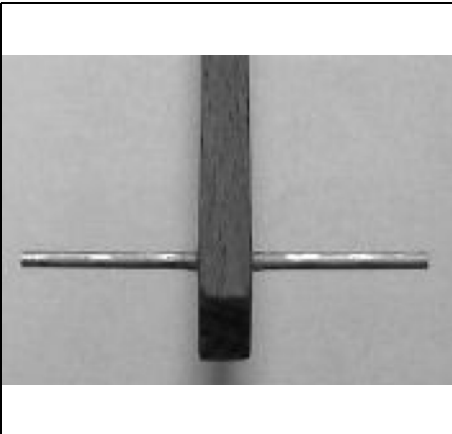


Fig 30: The last director in the boom of the 2.4GHz Quagi.

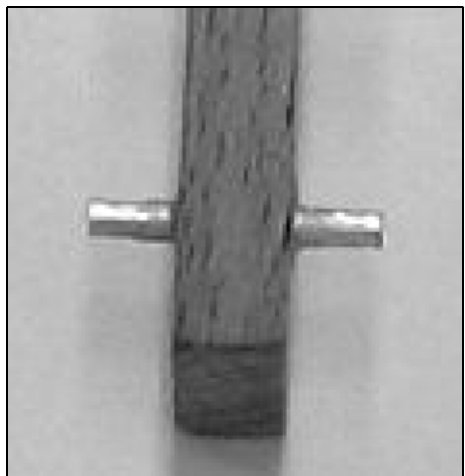


Fig 31: The last director in the boom of the 5.8GHz Quagi.



an economical transverse electromagnetic wave system with optimum gain; adjusting individual elements can carry out alignment. In addition the antenna can be fed using coaxial cable without the need for a balun or additional contacts or connections.

The Quagi has an additional 1dB gain compared to an equivalent Long Yagi. Because the Quad feed element is a closed loop, it is less sensitive to static loadings and disturbances making the antenna less sensitive to environmental influences and polarization fluctuations.

3.

Summary

Due to its electrical characteristics, the Quagi antenna is an outstanding directional antenna. High gain and a tidy radiation pattern make this antenna interesting. As shown the 2.4GHz 8 element Quagi has a gain of 13.8dBi and is only approximately 25cm long. Because of these dimensions this is a very inconspicuous antenna and is hardly noticeable compared to reflector aerials.

In the long run everyone must decide which antenna system suits their purpose. Depending upon place of use one must decide the pros and cons of an antenna system. One antenna does not suit every purpose and therefore further discussion is needed in the future.

4.

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[1] Edmund Stirner, "Antennas", volume 2, Hüthig publishing house, Heidelberg

[2] Rothammels Antennabuch (Red Sheep antenna book) 12th edition, DARC publishing house Baunatal

[3] Wayne Overbeck, N6NB "The Quagi Antenna turns 30"

[4] QST number 8/1981

[5] "Wireless Communication II", Professor Dr. – Ing. J. Jirmann

[6] Thesis to the FH-Coburg; Thomas Begmann, DG8NTB



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Prof. Dr. Ing. Gerd Janzen, DF6SJ

Reciprocal effects between antennas and surrounding metal objects

Part 2 continued from issue 2/2006

3.3. Dipole with a diagonal disturbing wire

The two cases of a resonant disturbing wire parallel and transverse to a dipole were examined in detail in the preceding chapters. Now the behaviour of a wire diagonally to the dipole like wire 8 in Fig 2 will be examined.

The data; End1: 0.17/0.23/0 and End2: 0.17/0.57/0 which is a wire 0.48λ long running 45° diagonally to the dipole was defined using the WIRES menu of EZNEC. The centre of the wire is $y=0.4\lambda$ away from the dipole as shown in Fig 10. The resulting feed impedance of the antenna is $Z_A = (81.61-j0.16)\Omega$ the current flow in the centre of the disturbing wire is $0.388A$ and the radiation pattern shown in Fig 11 is clearly asymmetrical.

Without examining other angles of disturbing wires, it can be concluded that only wires perpendicular to and in the main radiation direction of the dipole do not have any reciprocal effect on the dipole, and that wires parallel to the dipole have an increasing effect the closer they are to the dipole.

3.4. Dipole with a vertically arranged disturbing wire

How does a wire arranged vertically in front the dipole behave e.g. wire 4 in Fig

2. The data; End1: 0/0.1/-0.24 and End2: 0/0.1/0.24 defines a 0.48λ long wire in z axis. The centre of the vertical wire is $y=0.1\lambda$ away from the dipole as shown in Fig 12. The resulting feed impedance of the antenna is $Z_A = (71.39-j2.96)\Omega$ and is therefore completely unchanged. Also the current flow in the vertical wire is zero. The dipole and the wire are completely decoupled, there are no electrical fields that could cause current to flow in the vertical wire.

What happens if the vertical wire is not directly in front of the dipole, but off centre? Examples are wires 5, 7, 9 in Fig. 2. The data End1: 0.1/0.1/-0.24 and

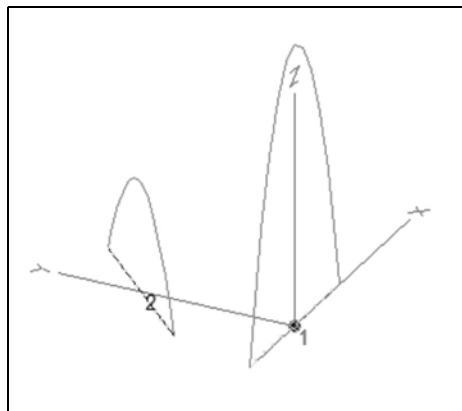


Fig 10: A disturbing wire at 45° to the dipole.

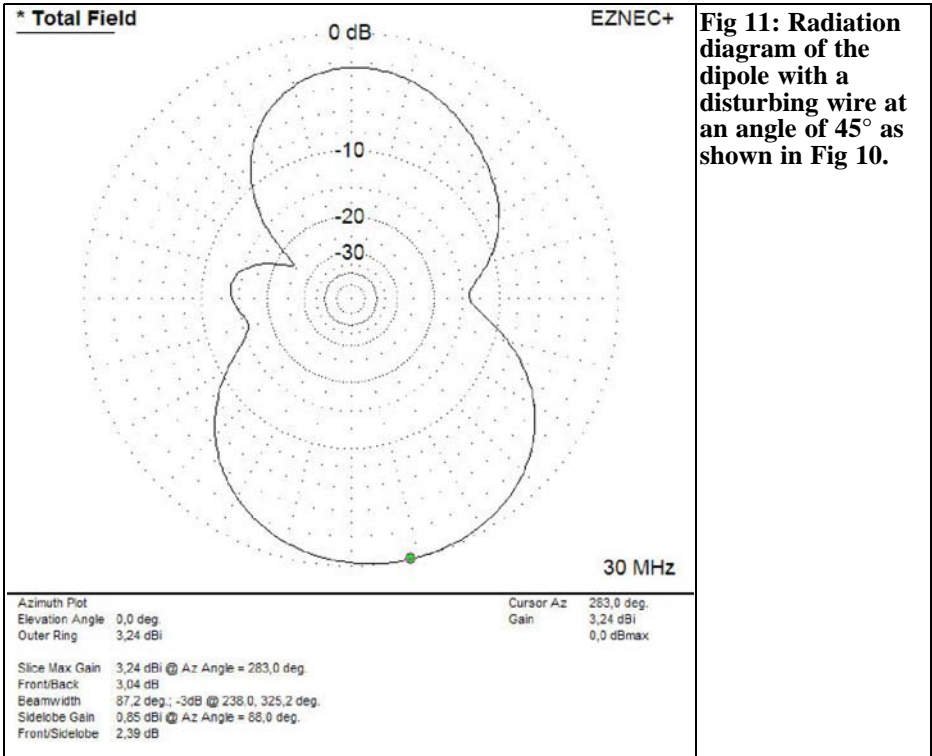


Fig 11: Radiation diagram of the dipole with a disturbing wire at an angle of 45° as shown in Fig 10.

End2: 0.1/0.1/0.24 defines a wire offset from the symmetrical axis. Most radio amateurs know that two wires at right angles to one another are not coupled, this has been confirmed by the previous example and Fig 12.

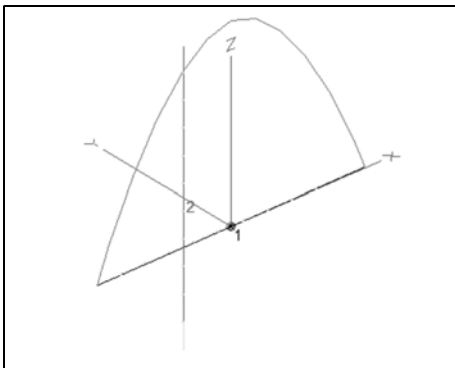


Fig 12: A vertical disturbing wire 0.1λ away from the dipole.

But something very strange happens as shown in Fig 13: There is a current flowing in the vertical wire, it is not the typical cosine waveform but has two peaks, even though the vertical wire with 0.48λ halves is 1λ long! What do we learn from this? The well known decoupling of wires/poles perpendicularly to one another only applies in the far field. In the near field of an antenna there are many more complicated relationships than is commonly known. If the vertical wire is moved towards the end of the dipole end ($x=0.24\lambda$); the current flow continues to rise with the double current waveform.

The conclusion from this surprising simulation result is that “actually” decoupled metal structures can interact in the near field of the antenna and that quarter wavelength structures can be resonant.

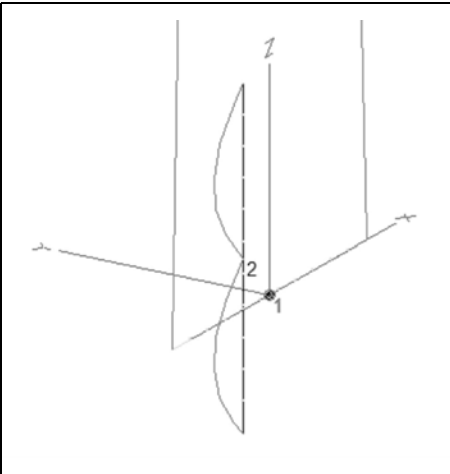


Fig 13: Surprising affect when when the vertical disturbing wire is offset.

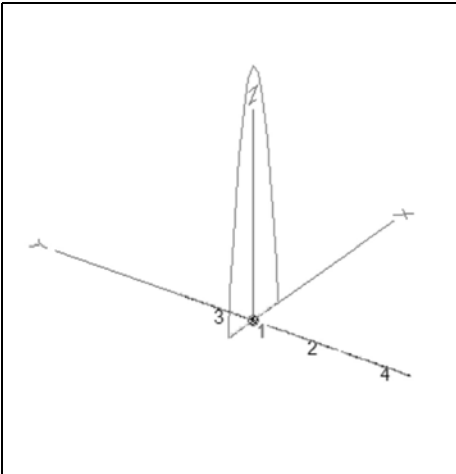


Fig 14: Dipole with 3 perpendicular disturbing wires,

3.5. Unexpected coupling of disturbing wires

From chapter 3.2. we know that wires arranged perpendicular to the fed dipole on the y axis do not have any affect on the dipole. This fact is used, as mentioned, for the boom of a directional antenna. An arrangement of three identical 0.48λ wires with the data: Wire 2: $0 / -0.1/0 - 0/-0.58/0$, wire 3: $0 / 0.05/0 - 0/0.53/0$ and wire 4: $0 / -0.6/0 - 0/-1.08/0$ gives a current diagram as shown in Fig 14. As expected no current flows in the passive wires 2, 3, 4 even with a dipole

current of 1A and when zooming in on he wires (with zoom Current in VIEW ANTENNA). If an additional wire 5 is added as shown in Fig 15 with the coordinates: $0.24/0.05/0 - 0.24/0.53/0$ which is off centre to the dipole, then the current situation changes. The off centre wire 5 couples energy from the dipole (current maximum 0.54A) and couples part of this to the parallel wires 2, 3, 4. Depending upon proximity of the off centre wire the current flow and thus the affects on impedance and radiation pattern can be substantial. The conclusion of this for practice: the boom of an adjacent antenna can absorb current and start to radiate, this could explain the radiation pattern shown in Fig 16.

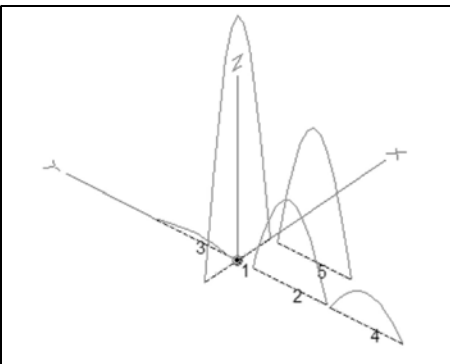
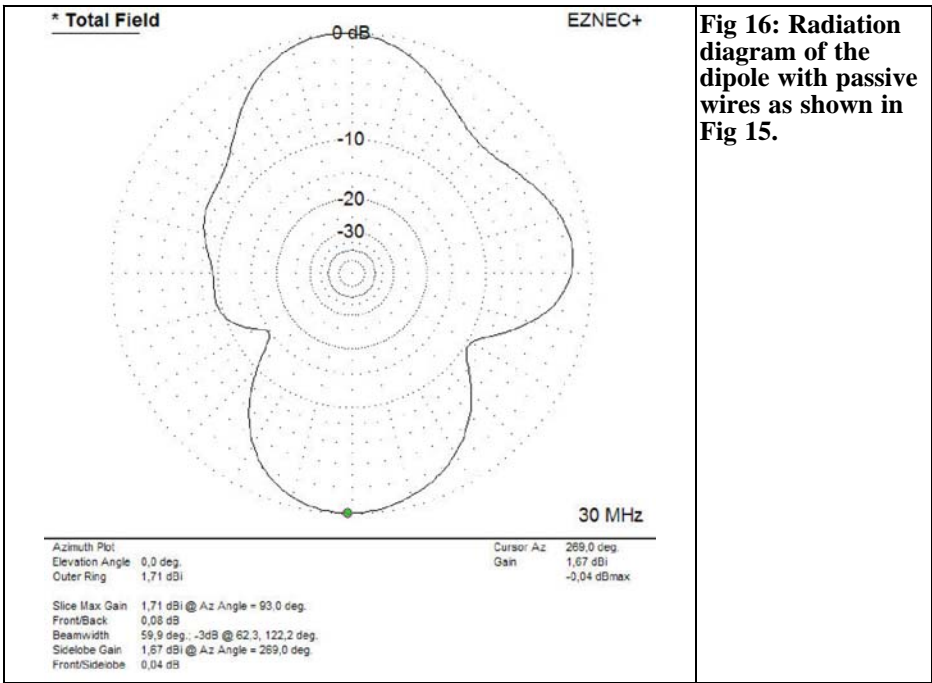


Fig 15: Affect of adding wire 5, it induces current in wires 2, 3 and 4.

4.

Influence of a parallel wire on a dipole in free space

In the preceding investigations it has been assumed that the disturbing wire



was a resonant length, $2 \times 0.24\lambda$. Now the effects of different lengths of wire will be examined. So that not too many parameters are examined at the same time, first the two distances $y=0.2\lambda$ and $y=1\lambda$ were specified.

All results of the simulations with different disturbing wire lengths at the distance $y=0.2\lambda$ from the antenna are shown in Table 4. The most interesting thing is the current flow in the disturbing wire. Ex-

Table 4: Results of simulation with disturbing wire 0.2λ from the dipole.

Length of disturbing wire in λ	Impedance Z_A in Ω	SWR	Maximum current in disturbing wire in A
0.1	71.33-j3	1.43	0.006
0.2	70.92-j3.3	1.43	0.023
0.3	69.49-j4.34	1.40	0.065
0.36	67-j5.88	1.36	0.124
0.44	50.86-j7.56	1.16	0.419
0.46	39.23+j1.73	1.28	0.626
0.48	42.13+j25.76	1.78	0.763
0.5	63.25+j30.87	1.80	0.628
0.52	72.87+j23.75	1.72	0.468
0.6	76.52+j10.65	1.58	0.229
0.8	75.2+j5.97	1.52	0.126
1.0	74.68+j5.17	1.51	0.098
1.3	74.28+j4.46	1.50	0.063
1.4	74.74+j3.76	1.48	0.046
1.5	71.92+j6.26	1.46	0.212
1.6	73.89+j5.9	1.50	0.144
1.7	74.07+j5.51	1.50	0.121
2.0	74.1+j5.14	1.50	0.100

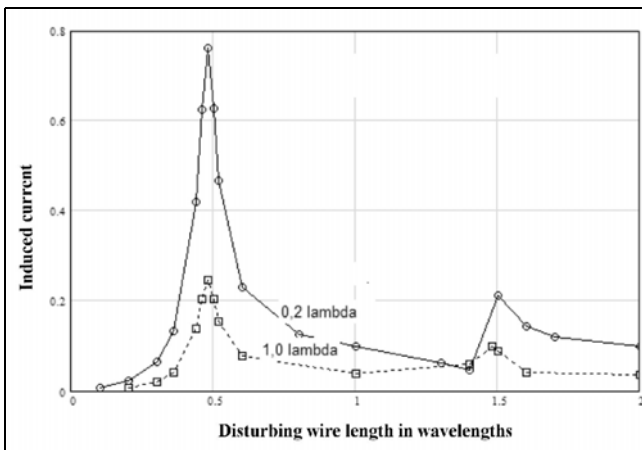
**Table 5: Results of simulation with disturbing wire 1λ from the dipole.**

Length of disturbing wire in λ	Impedance Z_A in Ω	SWR	Maximum current in disturbing wire in A
0.1	71.39-j2.96	1.43	0.002
0.2	71.41-j2.91	1.43	0.007
0.3	71.51-j2.76	1.43	0.020
0.36	71.7-j2.52	1.44	0.040
0.44	73.34-j1.85	1.47	0.137
0.46	74.85-j2.51	1.50	0.205
0.48	75.21-j5.13	1.52	0.247
0.5	73.16-j6.26	1.48	0.204
0.52	71.94-j5.82	1.46	0.153
0.6	71.11-j4.59	1.43	0.077
1.0	70.93-j4.03	1.43	0.039
1.4	70.59-j4.63	1.42	0.058
1.48	68.94-j3.97	1.39	0.100
1.5	69.28-j3.23	1.39	0.087
1.6	70.52-j3.23	1.42	0.042
2.0	70.8-j3.53	1.42	0.035

aming the table shows that the maximum current flows is at lengths of 0.48λ and 1.5λ . This fact is what led to the use of 0.48λ wires in the preceding chapters. Table 5 shows the results for the distance $y=1\lambda$, again the current maxima can be seen at lengths of 0.48λ and approximately 1.5λ .

Tables of figures are difficult to read so they have been turned into the graphs shown in Figs 17 and 18. The induced current as a function of the length of the disturbing wire are shown in Fig 17, the line with circles is for the smaller distance $y=0.2\lambda$ and the dotted line with

squares is for the larger distance $y=1\lambda$. The current maxima show up at lengths around $\lambda/2$ and a smaller maximum at $3\lambda/2$. The SWR of the dipole is shown in Fig 18. The greatest influence is found at the lengths of 0.5λ and 1.5λ . This means that resonant lengths react most violently with the electromagnetic field of the antenna. The width of the disturbance pattern is shown best in Fig 17, this means that metal structures with lengths within 20% of the resonant length can induce a current flow. For the 10m band that means that metal parts of the length 5m will interact strongly, in addition, lengths from 4m to 6m can have a clear

**Fig 17: Graphs of induced current in a disturbing wire at distances of 0.2λ and 1λ .**

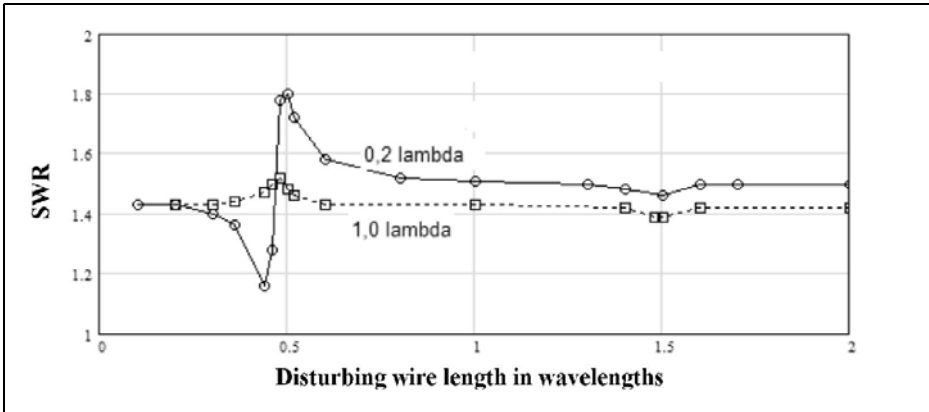


Fig 18: Graphs of SWR of the dipole with a disturbing wire at distances of 0.2λ and 1λ.

influence on an antenna. In the 160m band the “dangerous” length would be between 65m and 100m. Table 6 shows the effects of wires for different amateur radio bands.

Looking at the table shows that almost any metal structure in the near field of an amateur radio aerial for the 80m to 6m band will have an interaction. This can mean your own antennas as well as gutters, rainwater pipes, metal window frames, balcony balustrades or sun screens.

5.

Influence of a parallel disturbing of wire on a directional antenna

The influence of metallic structures on an antenna depends on how far the near field of an antenna extends. Approximate near field ranges of simple dipoles are indicated in Table 6. The effects of how metal structures in front and behind a directional antenna will now be examined by example.

Table 6: Near field and critical lengths of disturbing wires for various Amateur Radio Bands.

Amateur Radio Band	Approximate near field			Critical length	
	0	0.2λ	1λ	0.4λ	0.6λ
160m	0 to	30m to	160m	65m to	97m
80m	0 to	16m to	80m	33m to	49m
40m	0 to	8m to	40m	17m to	26m
30m	0 to	6m to	30m	12m to	18m
20m	0 to	4m to	20m	9m to	13m
17m	0 to	3.4m to	17m	7m to	10m
15m	0 to	3m to	15m	6m to	8.5m
12m	0 to	2.4m to	12m	5m to	7.2m
10m	0 to	2m to	10m	4.1m to	6.2m
6m	0 to	1.2m to	6m	2.4m to	3.5m
2m	0 to	0.4m to	2m	0.83m to	1.24m
70cm	0 to	0.14m to	0.7m	0.28m to	0.41m



Table 7: Parameters for non-optimised 3 element directional antenna.

```

EZNEC+ ver. 4.0
3elYagimitParallelstab          13.06.2005    18:34:40
----- ANTENNA DESCRIPTION -----
Frequency = 30 MHz
Wire Loss: Zero

----- WIRES -----
No.   Conn.  End 1  Coord. (wl)  Z   Conn.  End 2  Coord. (wl)  Z   Dia (wl)  Segs  Insulation
      X      Y      Z      X      Y      Z
1     2     -0,25, -0,06, 0   2     0,25, -0,06, 0   0,0005  11    1  0
2     2     -0,24,  0,  0   2     0,24,  0,  0   0,0005  11    1  0
3     2     -0,22,  0,2,  0   2     0,22,  0,2,  0   0,0005  11    1  0

Total Segments: 33

----- SOURCES -----
No.   Specified Pos.  Actual Pos.  Amplitude  Phase  Type
      wire # % From E1 % From E1 Seg (V/A) (deg.)
1     2     50,00     50,00     6         1         0         I

No loads specified
No transmission lines specified
Ground type is Free Space
    
```

A 3 element directional antenna was sketched using EZNEC, this is not a perfect antenna but serves the purpose of these experiments. The data for the antenna was: Reflector length $2 \times 0.25\lambda$ at $y=-0.06\lambda$, fed element length $2 \times 0.24\lambda$ at $y=0$ and director length $2 \times 0.22\lambda$ at $y=0.2\lambda$ resulting in an antenna with a gain $G=7.64\text{dBi}$ and a front to back ratio $V/R=22.42\text{dB}$. For 30MHz the elements of the Yagi antenna are approximately 5m, 4.8m and 4.4m long spaced from

0.6m and 2.2m from the fed element. The wire diameter used for the in simulation was $0.0005\lambda = 5\text{mm}$. The data for this antenna are summarised in Table 7.

The +y direction is in front of the 3 element antenna and the -y direction is behind the antenna. The antenna is in free space, a disturbing element parallel to the elements has a length of $2 \times 0.24\lambda$ and is moved in y-direction along the axis of the antenna from infinity to directly in front and then behind the antenna. The wire is a resonant length of 0.48λ , so that it has maximum influence.

The influence on the antenna feed impedance Z_A , standing wave ratio $s=\text{SWR}$, radiation pattern and, above all, the value of the current flow in the passive element as a function of its distance to the fed were examined. The arrangement of the antenna (wires 1, 2, 3) with the disturbing wire 4 is shown in Fig 19.

If the passive wire is infinitely far from the antenna, then an antenna impedance is $Z_A = (10.75 - j7.62)\Omega$ The antenna is resonant and fed with 50Ω giving an $\text{SWR } s = 4.77$. (Reference: in practice

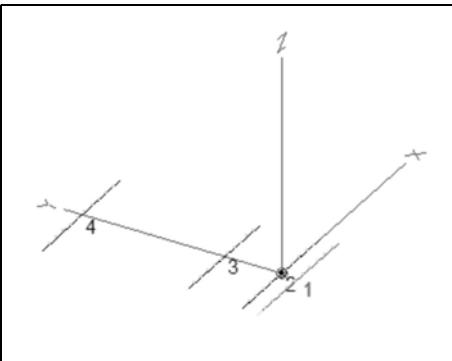


Fig 19: Diagram of the 3 element directional antenna showing the elements 1, 2 and 3 plus the disturbing wire 4.



Table 8: Simulation results of a disturbing wire being moved from the back to the front of a 3 element directional antenna.

Distance of disturbing	Impedance $Z_A \Omega$	SWR	Induced Current A	Gain dBi	V/R dB
infinity	10.75-j7.62	4.77	0	7.64	22.72
-10	10.75-j7.62	4.77	0.001	7.62	22.71
-5	10.75-j7.62	4.77	0.003	7.61	22.71
-2	10.75-j7.63	4.77	0.007	7.57	22.81
-1	10.74-j7.62	4.76	0.011	7.54	23.25
-0.75	10.76-j7.62	4.77	0.015	7.74	22.81
-0.5	10.74-j7.63	4.77	0.013	7.57	24.01
-0.3	10.75-j7.63	4.76	0.011	7.60	24.36
-0.25	10.76-j7.63	4.77	0.012	7.63	23.95
-0.15	10.75-j7.61	4.76	0.025	7.72	21.15
-0.1	10.79-j7.66	4.75	0.062	7.55	20.77
-0.06	Position of the reflector				
0	Position of the fed element				
0.2	Position of the director				
0.25	1.62-j3.68	31.04	0.302	8.4	3.3
0.3	2.85-j.29	17.53	0.344	7.98	3.12
0.4	13.74+j7.3	3.72	0.554	7.15	3.83
0.5	22.64-j12.77	2.39	0.542	6.64	5.91
0.7	7.41-j9.83	7.01	0.216	6.79	11.5
0.75	7.26-j8.04	7.02	0.202	6.94	11.19
0.8	7.72-j6.45	6.59	0.179	7.07	10.39
1	13.41-j6.55	3.8	0.212	7.26	10.07
2	11.27-j7.19	4.53	0.099	7.54	14.0
5	10.82-j7.54	4.73	0.039	7.63	18.08
10	10.76-j7.6	4.76	0.019	7.64	20.08
infinity	10.75-j7.62	4.77	0	7.64	22.42

this would not be an acceptable SWR, but this is a non-optimised antenna and would need a matching network (e.g.

with a series inductance of $+7.62\Omega$ and then with a suitable transformer for 50Ω). The radiation pattern of this direc-

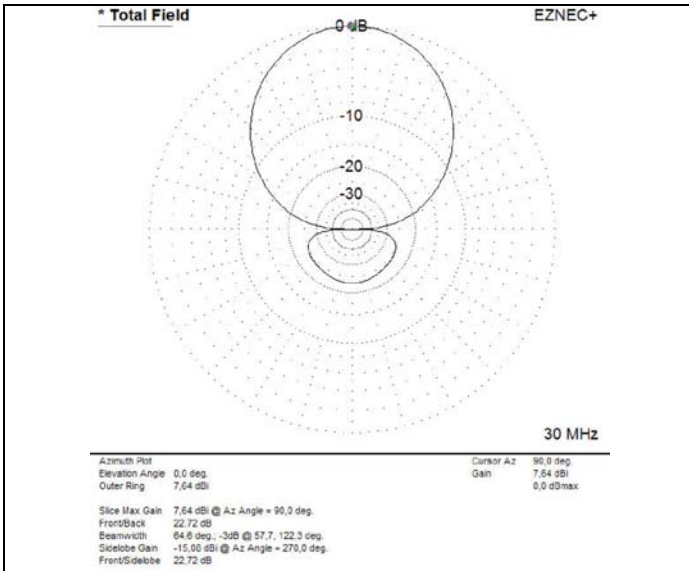


Fig 20: Radiation diagram of the non-optimised 3 element directional antenna.

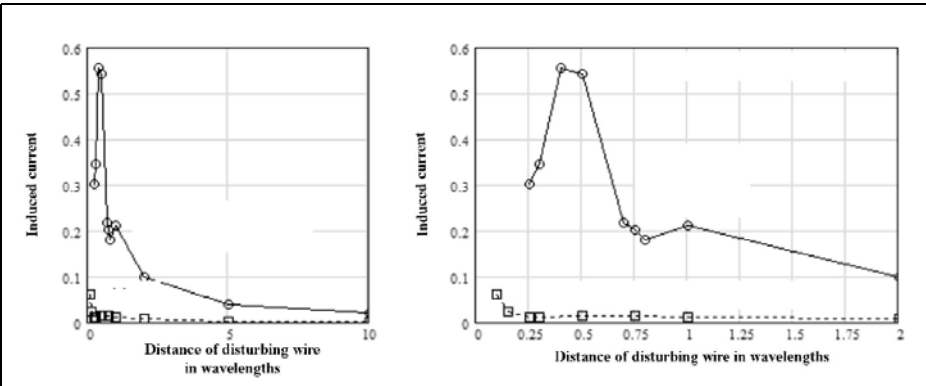


Fig 21: Graphs of the induced current in the disturbing wire when moved from the back to the front of the 3 element directional antenna.

tional antenna can be seen in Fig 20.

5.1. Measurements with wire behind the antenna

The simulation results as a function of the distance of the disturbing wire during measurements from the rear (negative y values) to the front (positive y values) are summarised in Table 8: Antenna feed impedance Z_A , SWR s in a 50Ω system, current maximum in the centre of the disturbing wire, antenna gain G in dB and front to back ratio V/R in dB.

There is hardly any affect as the direc-

tional antenna is approached from behind with the disturbing wire. Only when it is less than a fifth wavelength ($y=-0.2\lambda$) does the current induced in the disturbing wire rises to over 20mA (with an antenna energizing with 1A), and the front to back ratio changes slightly. The disturbing wire cannot be taken closer that $1/10$ of a wavelength because it will contact the reflector. The reaction of the antenna to the disturbing wire is easily understood because the front to back ratio of the antenna is over 20dB thus everything approaching from behind has little affect. The antenna is "deaf and blind" from the rear.

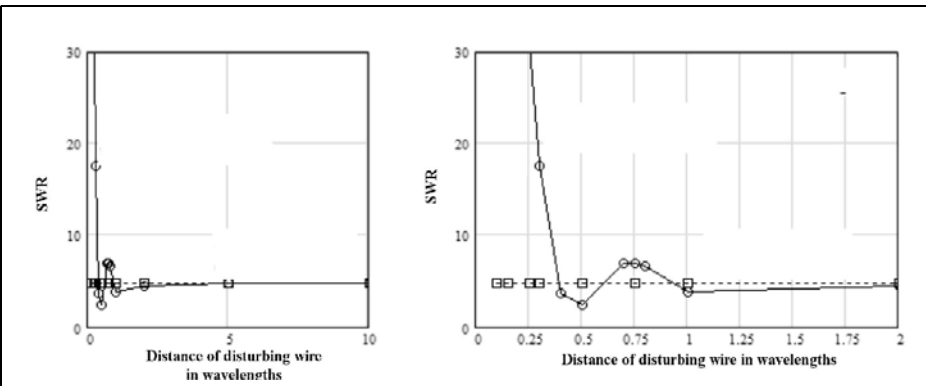


Fig 22: Graphs of the SWR of the 3 element directional antenna when the disturbing wire is moved from the back to the front.

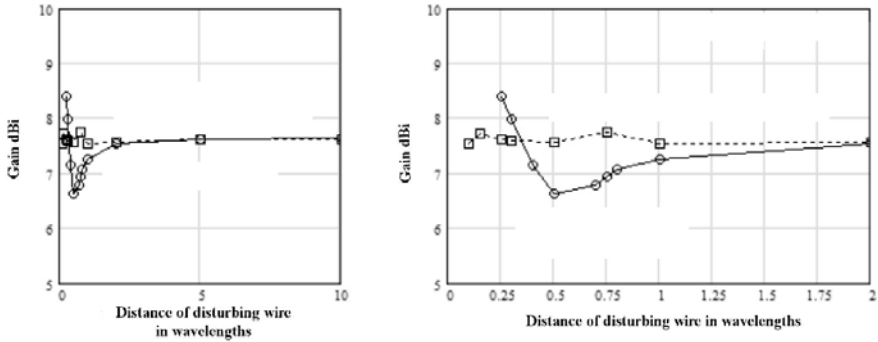


Fig 23: Graphs of the gain of the 3 element directional antenna when the disturbing wire is moved from the back to the front.

5.2. Measurements with the wire in front of the antenna

A clear influence on the antenna can be seen at a distance of 10 wavelengths from the front, the current in the disturbing wire is 19mA (it was only 1mA at the rear) Impedance Z_A , SWR s, gain G and front to back ratio V/R are only slightly affected at this distance.

As the disturbing wire is moved closer the affects can be clearly seen. If the induced current in the disturbing wire at a distance of one wavelength in front ($y=+1\lambda$) and behind ($y=-1\lambda$) the antenna

are compared it is nearly 20 times larger, in the forward direction at 0.212A compared to 0.011A.

5.3. Comparison of the disturbing wire in front of and behind the directional antenna

Diagrams are far better than tables to see the affects measured. The different influence becomes particularly clear in the diagrams, the maximum current flow at the centre of the disturbing wire in Fig 21, the SWR s in Fig 22, the antenna gain G in Fig 23, and finally the front to back ratio V/R in Fig 24 In the diagrams the

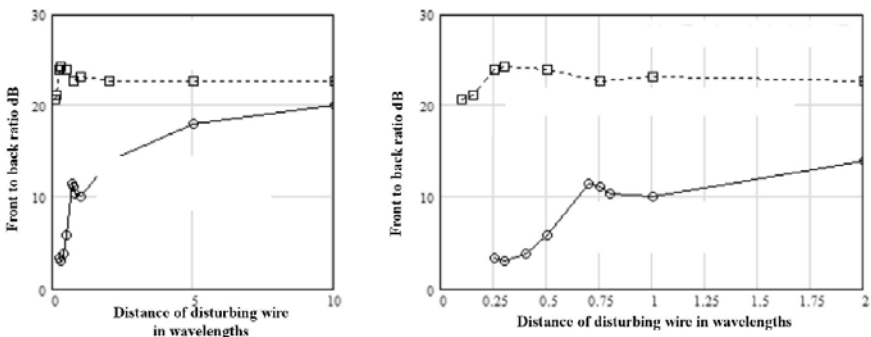


Fig 24: Graphs of the front to back ratio of the 3 element directional antenna when the disturbing wire is moved from the back to the front.

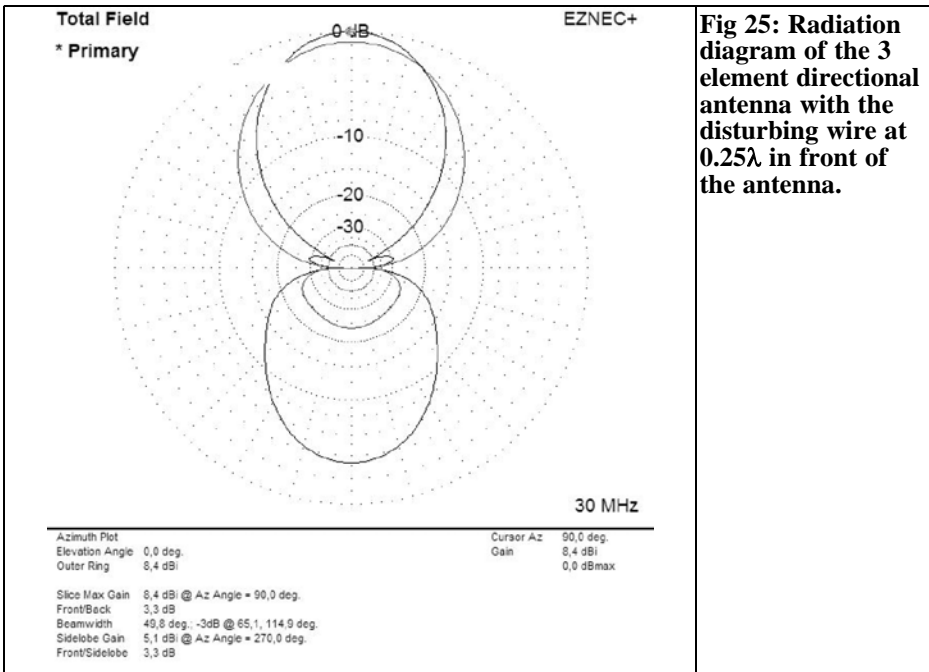


Fig 25: Radiation diagram of the 3 element directional antenna with the disturbing wire at 0.25λ in front of the antenna.

data for the disturbing wire in front of the antenna are shown by a curve with circles and the data for the disturbing wire behind the antenna are shown by a dotted curve with squares. Each measurement is shown with two diagrams. On the left the diagram represents the distance range from 0 to 10 wavelengths and on the right a close up showing the distance range from 0 to 2 wavelengths.

Fig 21 shows that the current induced in the disturbing wire falls slowly towards zero as the distance increases where the wire has no more effect on the antenna. In the right hand diagram the resonance affect can be seen at about half a wavelength ($y=0.5\lambda$).

5.3.1. Influence on the SWR

The SWR is the reaction of a mismatch in the system. The SWR shown in Fig 22 is more or less unaffected by the disturbing wire at the rear of the antenna but strong influences can be seen when the

disturbing wire is at the front of the antenna particularly at a distance of three-quarters to half a wavelength.

5.3.2. Influence on the antenna gain

Fig. 23 shows the influence of the disturbing wire on the antenna gain G. When the disturbing wire is to the rear of the antenna it has no affect but when it is in front of the antenna at distances less than approximately 1 wavelength the antenna gain falls by around 1dB and rises then continuously to approximately 2dB which shows the gain improving effect of the disturbing wire, which works as the second director element in the directional antenna.

5.3.3. Influence on the front to back ratio of the antenna

The front to back ratio is affected most by the disturbing wire, this is shown in Fig 24. For distances of one quarter to three quarters of a wavelength from the

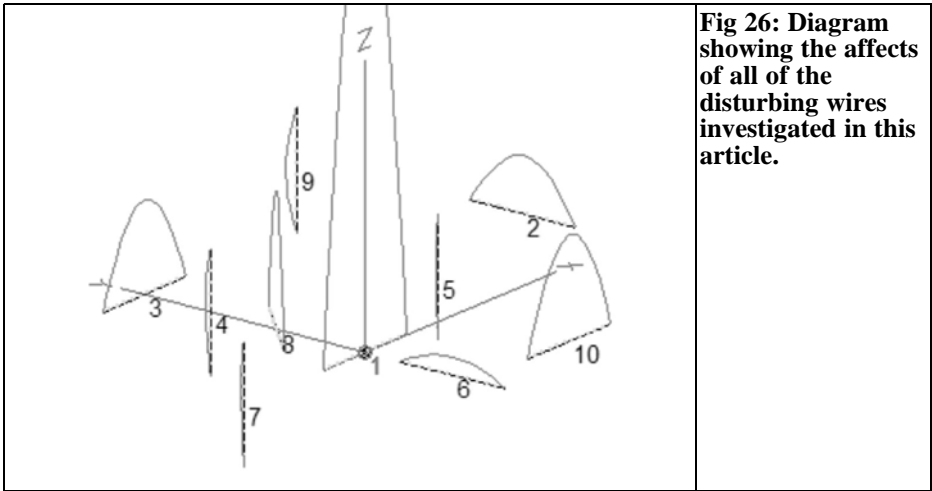


Fig 26: Diagram showing the affects of all of the disturbing wires investigated in this article.

rear of the antenna the ratio increases. When the disturbing wire is at the front of the antenna it starts to affect the ratio at 10 wavelengths. As the disturbing wire approaches three quarters of a wavelength the front to back ratio changes dramatically and it destroys the directional characteristics of the antenna. The radiation pattern with the disturbing wire 0.25λ in front of the antenna is shown in Fig 25.

6.

Summary

In order to gain an impression of the influence of disturbing metal parts in the environment of an antenna, the antenna simulator program EZNEC was used with a simple dipole and a non-optimised 3 element directional antenna and disturbing wires of different lengths and orientations. The reaction of the antenna was examined for it's input impedance, standing wave ratio, gain and front to back ratio of the directional antenna. As an important indicator of the reciprocal effect between the antenna and the dis-

turbing wire the current flow induced by the antenna in the disturbing wire was measured.

In a first simulations disturbing wires of different orientation (parallel, perpendicularly, crosswise, lengthwise, vertical) were examined in relation to a horizontal dipole wire hanging in free space. As expected the parallel wires showed a strong reciprocal effect with the dipole. Less expected was the influence of a "decoupled" wire on perpendicular wires. The effect of a second wire in the near field on symmetrical wires was not expected, it had a strong reciprocal effect between all wires. This phenomenon can have clear practical affects on the radiation of a directional antenna due to the effect of the boom of another antenna. A summary of the affects of the disturbing wires described in Fig 2 are shown in Fig 26. The dipole (wire No. 1 in the centre of the coordinate system) is the only fed wire. All current flows shown in the disturbing wires are full scale with reference to 1A fed into the dipole.

A second simulation showed the effects of disturbing wires of different lengths parallel to a dipole antenna. The reaction of the dipole to the disturbing wire was largest if this was a half wave and thus



resonant. Since the affects were quite large even when the wire was not a half wavelength it is easy to see how the environment around an amateur radio antenna will always affect their antennas particularly at higher frequencies. This affect on the antenna is not only to it's input impedance but particularly the radiation pattern.

The influence on the radiation pattern becomes still clearer with a disturbing wire and a directional antenna. The affect is clearly different when the disturbing wire is in front of rather than behind the directional antenna. Depending on the gain of the directional antenna the affects can be seen at a number of wavelengths. The front to back ratio of a directional antenna is particularly sensitive to disturbances as far away as 10 wavelengths and more. Disturbances behind the directional antenna are not practically noticeable.

Note: all disturbing metal structures in this article were highly conductive. Lossy foreign elements will affect antennas, but requires further investigations.

7.

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John Fielding, ZS5JF

Diode Multipliers

1.

Introduction

Semiconductor diodes of various types can be used for multiplier applications. Many previously published articles have omitted details of how to extract the optimum performance from simple diode multipliers.

Some of the designs previously published give details of expensive diodes but often a cheaper diode can be pressed into service. Exotic diodes such as Step Recovery Diodes or Varactor diodes can often be changed for more mundane types with little loss in performance.

For amateurs involved with VHF and UHF construction the need to multiply up from a base oscillator frequency to reach a higher frequency is a common task.

Whereas diode multipliers may not be the best approach they are still the basis of many designs. With an understanding of the limitations a better insight in the best method is possible.

2.

Simple diode multiplier

A typical diode multiplier circuit is shown in Figure 1. This circuit is often used where a medium step in multiplication is required; say 3 times the input frequency.

The diode D1 is either a hot carrier or Schottky diode. The input drive comes from a low impedance source such as the emitter follower TR1 to drive a large current into the diode. The diode is

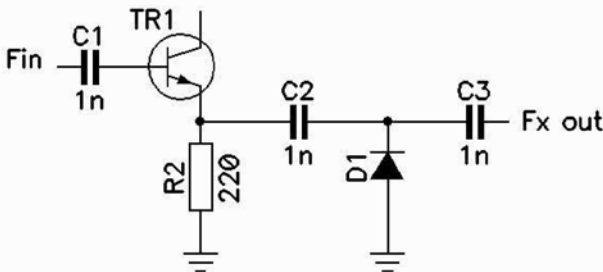


Fig 1: Basic diode frequency multiplier.

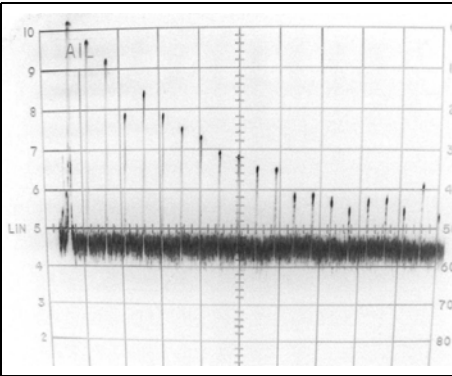


Fig 2 Spectrum generated by simple shunt diode multiplier.

connected in shunt across the input and output and the high circulating current in the diode generates harmonics of the input signal. The output of the multiplier is then connected to a selective band pass filter via C3. If the spectrum of Figure 1 circuit is examined the response is typically that as shown in Figure 2. This was taken from a 50MHz crystal oscillator driving a very low impedance buffer amplifier feeding a hot carrier diode multiplier.

The extreme left-hand signal is the zero Hz marker on the spectrum analyser, the first spike after this is the 50MHz oscillator. The next signal is the 100MHz (2nd harmonic) product and the next is the 150MHz (3rd harmonic) product. The drive level into the diode is +17dBm (50mW) at 50MHz; the spectrum analyser is set to +20dBm full scale. The third harmonic signal is lower in power level than would be extrapolated by the other harmonics and is approximately 17dB below the 50MHz fundamental. The vertical scale is 10dB per division and the horizontal scale covers from 0Hz to approximately 1GHz. So we can see that for 50mW input power the 3rd harmonic is only 0dBm (1mW), or 1/50th of the input power. Hence, the efficiency as a tripler is very poor at approximately 2%.

This multiplier requires a severe amount

of filtering to reject the fundamental, 2nd and 4th harmonic. The order of filtering required is approximately 3 narrow band tuned circuits to reject the unwanted products to less than -60dBc. Even with a very good filter the output power of the wanted 3rd harmonic is going to be quite low and will require considerable gain at 150MHz to bring it back to a usable power level for the next multiplier stage. If the specification demands all unwanted products, apart from the 3rd harmonic, to be at least 60dB down, then the attenuation required for the fundamental signal is at least 77dB and the attenuation for the 2nd harmonic is at least 73dB. This order of filter will have considerable insertion loss to the wanted 3rd harmonic, perhaps as much as 6dB in a practical filter, so the output power will be of the order of -6dBm after the filter.

The basic problem is that a small signal hot carrier or Schottky diode is not inherently a multiplier, rather it is a divider. If a simple half wave rectifier is used the diode produces one output pulse for every two pulses applied; hence it makes an efficient divide by two but a poor multiplier. To enhance the multiplication we need to do a few modifications to the simple circuit.

From Fourier transforms we know that we can construct a square wave signal by adding together the various sine wave harmonics in the correct amplitude. The value of the 3rd harmonic content is hence quite small. Even though we are driving the diode very hard to generate a quasi-square wave current in the diode the amplitude of the higher harmonics is expected to tail off in a more or less linear manner. This can be clearly seen in the spectrum plot in Figure 2. The apparent lower amplitude at the 3rd harmonic is probably due to some other resonance occurring in the circuit parasitic inductance and stray capacitance. As the input drive level is varied these show as peaks and troughs at different parts of the spectrum. The spectrum plot shown is that of a 50MHz calibrator for

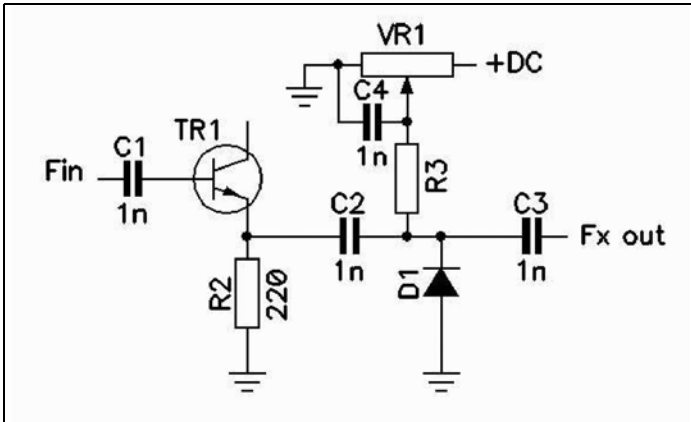


Fig 3: Reverse biased diode multiplier.

the spectrum analyser to allow alignment. It is known as a “Comb Generator” and ideally all the output products should be of similar amplitude. This can be achieved with an exotic diode such as a Step Recovery diode, but these are expensive.

The harmonic generation is dependent on the conduction angle of the diode. To improve the harmonic generation we need to reduce the conduction angle. A simple way of achieving this is to reverse bias the diode with a standing DC bias voltage. This is shown in Figure 3. The variable resistor VR1 and a high value resistor R3 acting as a RF choke applies the reverse bias to D1. No static current flows in D1 apart from the reverse bias leakage current, which is negligible in most cases.

Another version of the single ended multiplier is shown in Figure 4. This is

somewhat better but still not optimum when high multiplication values are required. This circuit is often used for the final multiplication of a microwave local oscillator chain, for example a 384MHz to 1152MHz chain for a 23cm converter using a 144MHz IF. The filter is often two $\lambda/4$ lines with the multiplier diode tapped well down on the input tuned line.

Diode D1 is pumped by a high level signal at the fundamental frequency. The diode is tapped well down on the input inductor L2. Inductor L1 is the DC return for the circulating DC diode current and needs to represent a high impedance to the input frequency but a low DC resistance for the diode current. The fundamental energy is dumped into the low inductance tap point on L2 so suppressing the power contained. The top coupled inductors L2 and L3 which form the band pass filter act as a high pass

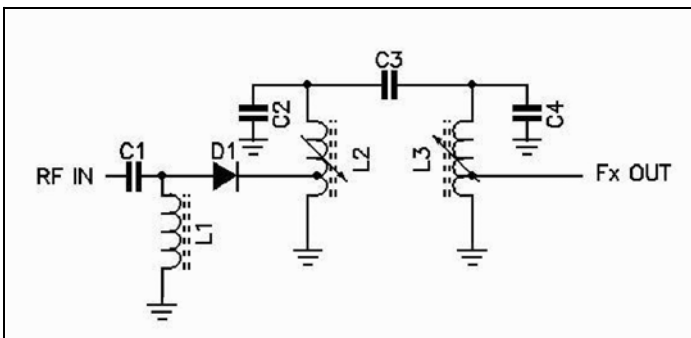


Fig 4: Improved diode multiplier.

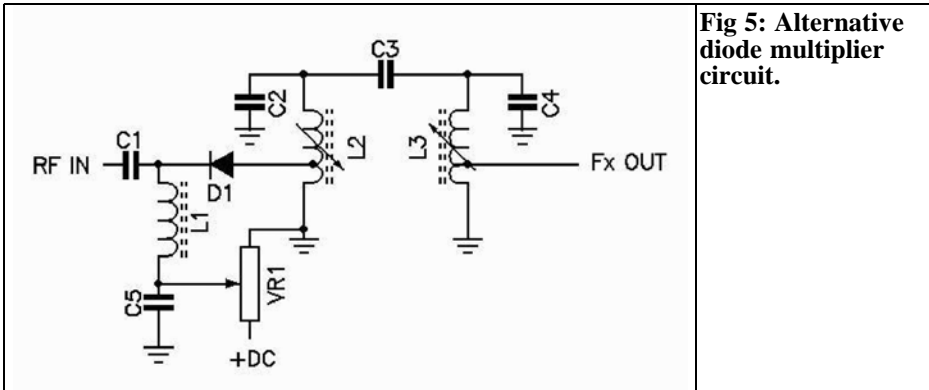


Fig 5: Alternative diode multiplier circuit.

filter to the higher harmonics and so the spectrum roll off on the high frequency side is not as good as it could be.

An improved circuit is shown in Figure 5 where a pre-bias is applied to the diode to enhance the multiplier efficiency. By varying the bias voltage the circuit can be made to peak at a particular input drive level. This in many cases can deliver up to 10dB more output power than an unbiased diode. The exact reverse bias needs to be found empirically and a good starting point is about 1 to 2V reverse bias. This of course assumes adequate RF input power is available from the preceding multiplier stages. A good rule of thumb is that the peak RF voltage swing is at least the same as the DC reverse bias, which in many cases means we need more than +10dBm of drive power.

3.

Limitation of simple diode multiplier

All practical simple diode multipliers have conversion loss. In many cases this can be overcome by simply increasing the input drive level, up to the power dissipation of the diode being used. If the input drive level exceeds the safe

diode dissipation the diode will heat up internally and may fail.

The basic problem with all these circuits is the type of diode used. A Schottky or hot carrier diode is not the best choice for a multiplier because of the small junction capacity, although they make good receiving mixers with a reasonably low noise figure. We know from a typical varactor diode power tripler that an input of 10W at 432MHz can yield approximately 6W at 1296MHz, an efficiency of 60%. What distinguishes the varactor diode from other types is the large junction capacity.

Energy pumped into the diode on one half cycle of the RF waveform is stored as charge and is released on the next half cycle with virtually no losses. Step Recovery diodes are similar but able to multiply to greater frequencies because the stored charge is released very much quicker than a normal varactor diode, so producing harmonics to a far higher order. This "*parametric conversion*" is the clue to making high efficiency multipliers. If the junction capacity of a Schottky diode is examined against applied reverse bias voltage it appears like the graph in Figure 6.

The amount of energy that can be transferred is directly proportional to the junction capacity. The curve in Figure 6 is that of the BAT-85 Schottky diode. If a typical small signal varicap diode char-

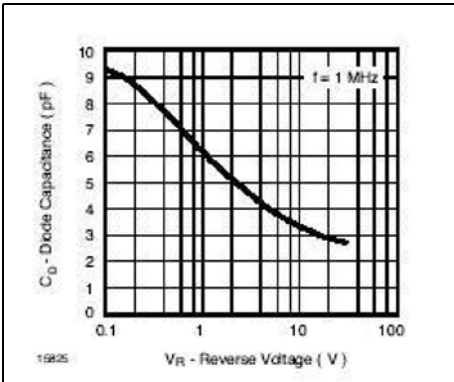


Fig 6: Typical small signal Schottky diode characteristics.

acteristics is examined we see a very different picture.

Figure 7 shows the junction capacity versus reverse bias voltage of the BB-809 varicap diode. With zero applied bias voltage the junction capacity is more than 50pF and with 10V bias it is approximately 13pF. By swinging the input RF voltage between approximately 1V and 10V we have a total capacitance variation of 30pF. Compare this to the Schottky diode curve in Figure 6 and we can see that it only swings approximately 5pF for the same DC and RF conditions. Hence, we can intuitively see that the BB-809 diode has about 6 times the capacitance variation and hence will perform better as a multiplier than the BAT-85. This can be borne out in experiments where a

greater harmonic power is generated when the varicap diode is used. The exact junction capacitance required is a function of the frequency being used, for microwave multiplier stages the junction capacitance will be proportionally less than for a low frequency multiplier. There are many low cost varicap diodes available from several manufacturers that can be used. For low frequency applications the BB-809 or the BB-132 in SMD are a good choice and for higher frequencies a good choice is the BBY-31 which is a SMD device.

4. Special Multipliers

Often we require to multiply by a fixed factor, for example times 2. For this application we can utilise the dividing property of the typical diode. If two diodes are arranged as in Figure 8 the two half cycles of the input sine wave are used and the predominant current pulses at the output are even harmonics. This configuration is known as a “push-pull” doubler. Because only the even harmonics are used the circuit offers a greater rejection to odd harmonic multiplication factors. It works well at 2x and 4x but very poorly at 3x and 5x because of the balanced configuration. If the input transformer has sufficient bandwidth the

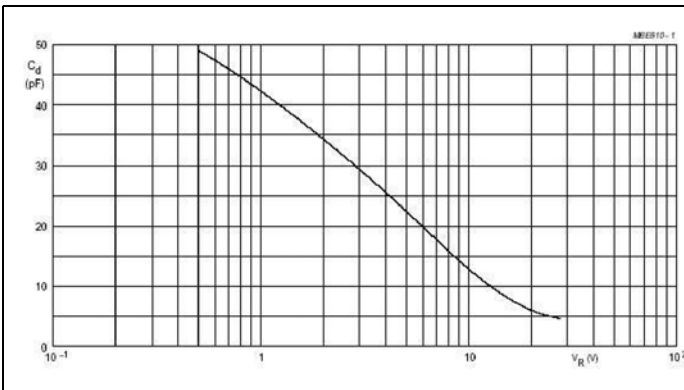


Fig 7: Typical small signal varicap diode characteristics.

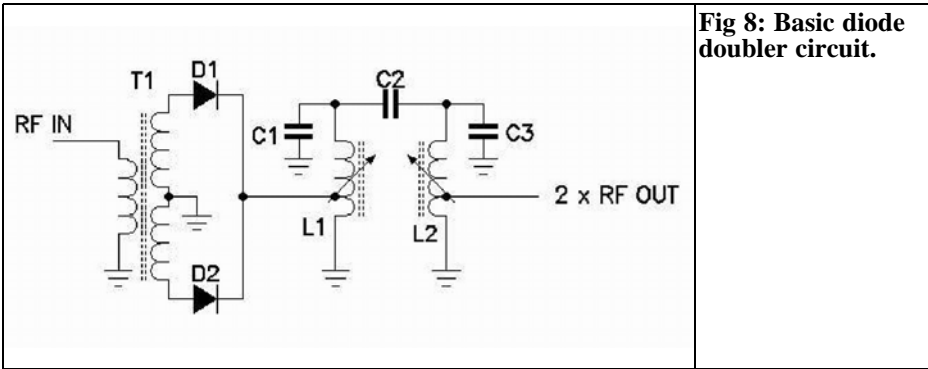


Fig 8: Basic diode doubler circuit.

circuit can work over an octave or more of bandwidth. Normally the transformer is wound on a small ferrite bead or core with good high frequency characteristics. The typical transformer is wound with tri-filar wire on a small two-hole ferrite core.

The input power is applied to the transformer T1, which develops a push-pull drive to two diodes. The summation of the two diodes at the tap point on L1 is rich in even harmonic power and the band pass filter selects the required harmonic, in this case the second harmonic. By selecting suitable values for L1, L2 and the tuning capacitors the circuit can be used as a 4x, 8x multiplier etc. However, the output power as a 4x multiplier is not as much as a 2x multiplier, typically it will be 6 to 12dB less than the 2x multiplier.

By substituting varicap diodes for D1 and

D2 with some pre-bias the output power can be greatly increased over the simple Schottky diodes types shown. This circuit is shown in Figure 9.

Diodes D1 and D2 are now small varicap diodes, such as the BB-809, and the reverse bias is applied via the preset VR2. C4 is a low reactance capacitor to RF ground the bottom of L1.

The basic problem with both of these circuits is making a suitable push-pull transformer to work at the required frequency. For low frequencies this is not too difficult but as the input frequency increases the winding of the transformer to obtain good balance becomes more difficult. Hence, the circuit is really only useful up to about 400MHz. At higher input frequencies a printed 180° 3dB coupler is a better choice, although it will occupy a fair amount of board area even at 1GHz.

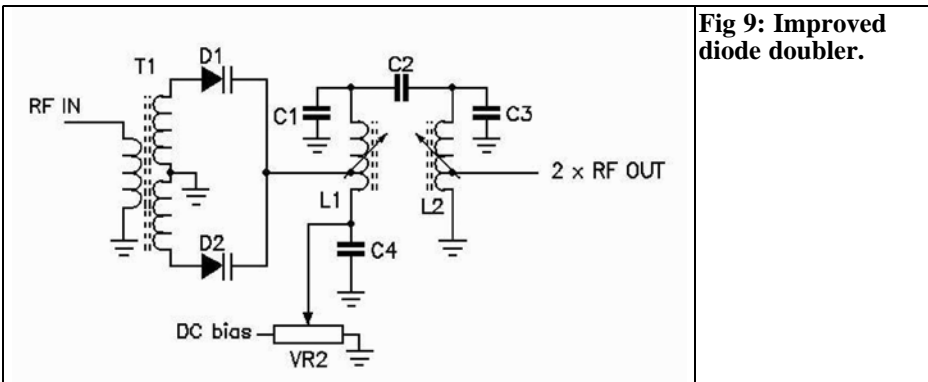


Fig 9: Improved diode doubler.

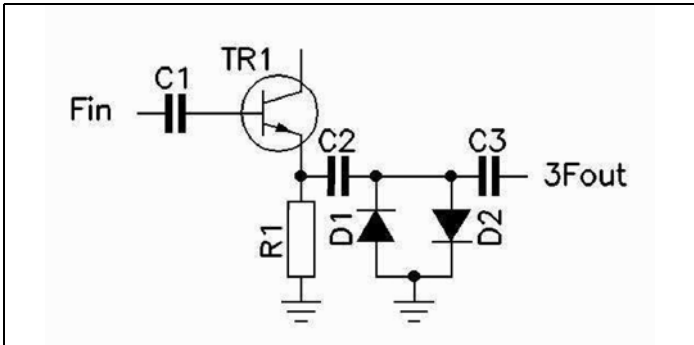


Fig 10: Push-Pull tripler.

5. Odd Harmonic Multipliers

When we need to enhance the odd order harmonics and suppress the even order ones then the “push-push” circuit is a good candidate. The basic circuit is shown in Figure 10. Here two diodes are connected in anti-parallel so both half cycles of the input waveform are acted on. In the experimental circuit two BAT-85 Schottky diodes were used because they were to hand.

The input spectrum to this multiplier is shown in Figure 11. This is the same 50MHz driver as used earlier but the spectrum analyser sweep has been re-

duced to show 0 – 300MHz. The extreme left-hand spike is zero Hz and 150MHz is in the centre of the display.

The output spectrum of the push-push circuit is shown in Figure 12, where it can be seen that the even harmonics have been suppressed by approximately 35dB, so making the filtering of the unwanted products much easier. The conversion loss of the circuit is approximately 6dB at the 3rd harmonic, which yields an efficiency of 25%, a considerable improvement over the simpler circuit.

Figure 13 shows the same signal after passing through a simple 2-section band pass filter. The products at 50MHz and 250MHz are now suppressed by approximately 40dB and all others by over 50dB. The following amplifier stage, with further filtering, will provide adequate drive

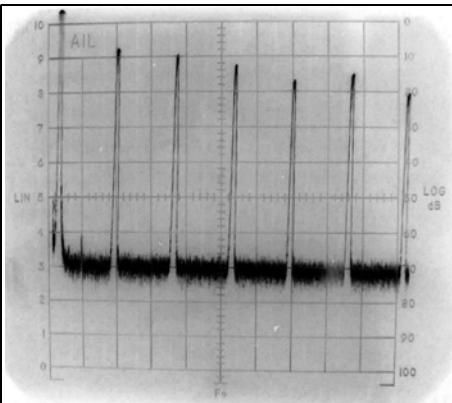


Fig 11: Input spectrum to push-pull tripler.

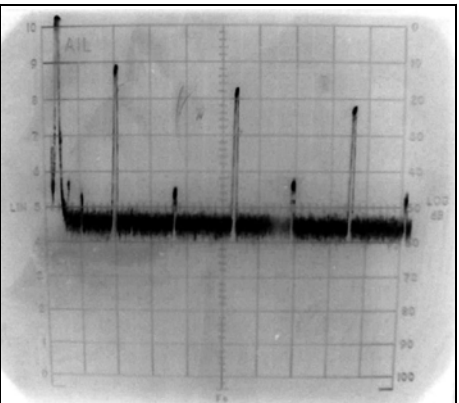


Fig 12: Output spectrum of the push-pull tripler before filtering.

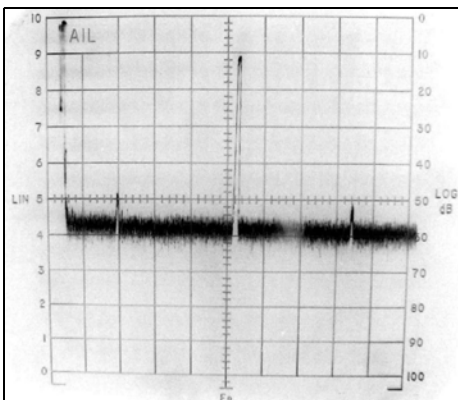


Fig 13: Output of diode tripler after filtering.

power to the next multiplier with additional suppression of the unwanted products.

6.

Enhancing Multiplier Efficiency

If the power varactor diode tripler multiplier is examined we see that there are not only tuned networks at the input and output but also an extra tuned network resonant at the 2nd harmonic. This is known as an “idler circuit” and it increases the 2nd harmonic current flowing in the diode. You can think of the parametric varactor tripler as a type of

mixer. The high 2nd harmonic current ($2f$) is added to the fundamental current ($1f$) and out pops the sum and difference products. The difference product ($2f - 1f$) is the fundamental and this reinforces the input current. However, the sum product is $1f + 2f$, which is the wanted output frequency. So the clue to efficient multipliers is to add idler networks to the input and output to reinforce the wanted harmonic. This is illustrated in Figure 14 where two idler networks have been added to the basic diode multiplier.

The input network $C1$ & $L1$ is series resonant at the fundamental frequency and so appear to be highly reactive to the second, third and higher harmonics, reflecting power back to the diode. Network $C2$ & $L2$ is series resonant at the 2nd harmonic generating a high circulating current through the varicap diode at that frequency. Pre-bias is applied to the varicap by the variable resistor and $R3$. The predominant energy is therefore contained in the 3rd harmonic and is filtered in a band pass filter to remove unwanted products.

When a circuit such as this is aligned using a spectrum analyser it will be noted that as $C2$ is adjusted the power at the 2nd harmonic will dip towards zero and the power at the 3rd harmonic will correspondingly peak. This is a much better signal spectrum to try and filter as the main unwanted signal is now at the fundamental and hence further away on the band pass filter slope so easing the

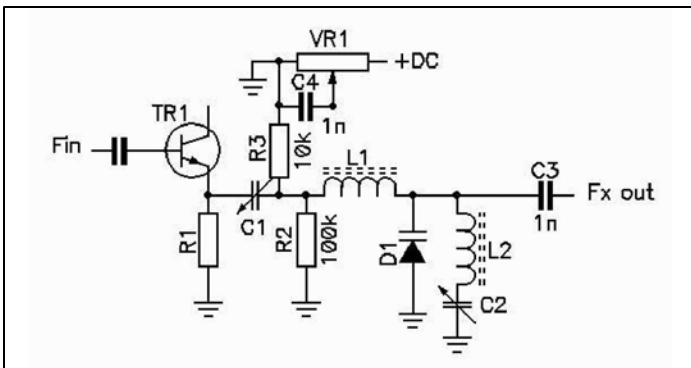


Fig 14: Multiplier with idler networks.

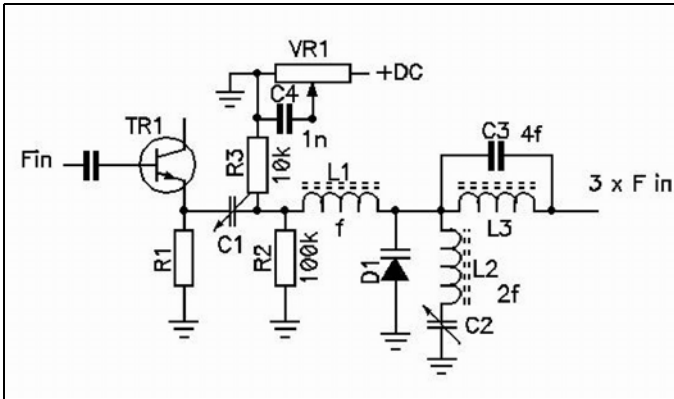


Fig 15: 4th harmonic idler added.

design of the filter.

Additional idlers can be added for the 4th and higher harmonics should they be necessary. Typically this network would be a parallel tuned “trap” inserted in series with the output and adjusted to null out the harmonic power. With sufficient Q in the idler networks more than 30dB of additional rejection can be obtained. The parallel tuned trap reflects power at the harmonic back into the multiplier diode and this is added to the wanted 3rd harmonic signal. This is demonstrated in Figure 15 where a trap for the 4th harmonic has been added.

7.

Other Diode Types

Whereas most designs published to date have concentrated on hot carrier, Schottky or varicap diodes there are other less expensive types that can work. As an experiment the writer substituted the BAT-85 diodes in the push-push tripler circuit shown in Figure 10 with a common switching diode, the 1N4148, to see what difference would result. It was rather a surprise to find that the 1N4148 actually performed far better than the BAT-85. The output power at 150MHz was almost 9dB greater and after changing the band pass filter component values

to provide more rejection to the close in products an overall improvement of nearly 8dB was achieved with better rejection of the unwanted products. It occurred to the writer that this is probably because the forward voltage of the 1N4148 is 0.7V whereas the Schottky BAT-85 is typically 0.35V. The junction capacity of the 1N4148 is also much greater than the BAT-85 or similar Schottky diodes so enhancing the parametric characteristics. Hence this bears out the previous measurements made with reverse biased diodes, where the greater the reverse bias the better the diode acts like a non-linear element. As the cost of the 1N4148 is approximately 10% of the BAT-85 this is a worthwhile modification for low frequency applications.

Figure 16 is a picture of the spectrum plot after the diodes were changed to 1N4148s and the band pass filter readjusted. The suppression of the products at 50MHz and 250MHz is improved to about 50dB. The 100MHz and 200MHz products are greater than 60dB below the 150MHz output. A possible choice for the dual-diode is the BAV-99, which contains two closely matched 1N4148-type diodes connected internally in a SOT-23 package. This is shown in Figure 17.

The configuration is ideal for the push-push tripler and pins 1 and 2 are grounded and pin 3 is the common point.

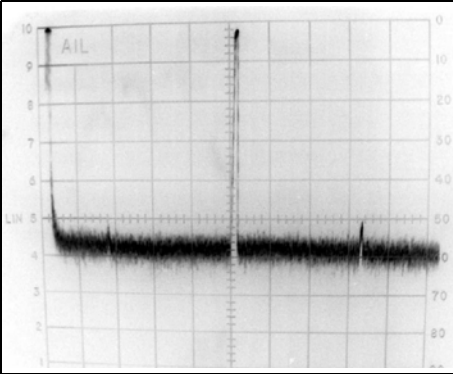


Fig 16: Push-pull tripler fitted with 1N4148 diodes.

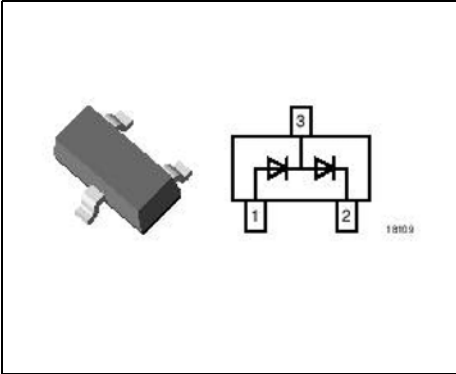


Fig 17: BAV-99 dual diode.

8.

Comb Generator

For circuit where the push-pull doubler extra components may not be desired a simpler modification is to make a slight change to the circuit shown in Figure 1. If a high value resistor is connected across the multiplier diode the circuit effectively becomes self-biasing. A typical resistor value is about 33kΩ. Figure 18 shows the output spectrum of a simple harmonic multiplier driven by a 100MHz low impedance source. This circuit was

built as a Comb Generator to calibrate a spectrum analyser. There are some 50MHz products apparent in the higher portion of the spectrum due to insufficient filtering in the 50MHz to 100MHz doubler stage preceding the final comb generator. This comb generator uses a 1N4148 switching diode to generate the 100MHz comb output, which extends up to at least 4GHz.

The output power from the Comb Generator is adequate for the purpose intended, to calibrate the frequency range of the spectrum analyser. Figure 19 shows the same comb generator covering 1.8 to 2.8GHz.

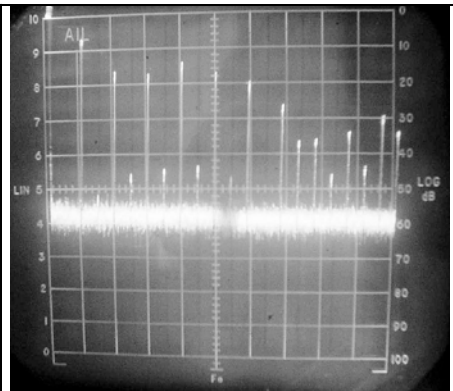


Fig 18: 100MHz comb generator output spectrum. The spectrum analyser covers 0 to 1GHz.

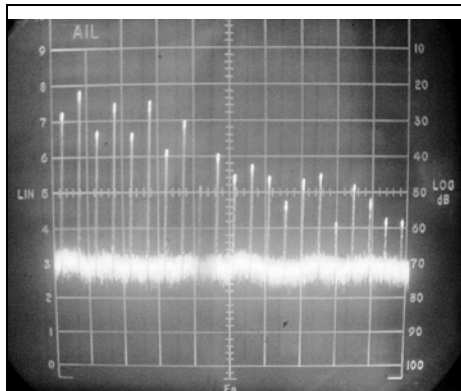


Fig 19: Comb generator output on 1.8 to 2.8GHz range.



Wolfgang Schneider, DJ8ES

Power detector covering up to 2.7 GHz

Signal detectors can be made with very few components using modern integrated circuits. This article describes a measuring head using an AD8362 True Power Detector from Analog Devices.

1.

Introduction

From the manufacturers data, the AD8362 detector has very good linear behaviour over a wide frequency range up to 2.7GHz. My own measurements confirm the good characteristics of this integrated circuit. This was described in [1].

According to the data sheet optimum results are achieved with a 1:4 Balun on

the input. However good performance is achieved with the recommended transformer from Mini Circuits. The circuit is useful for frequencies from 10MHz and drops by about 1dB at 1.9GHz. A suitable indicator circuit can compensate for this.

2.

Circuit description

The heart of the measuring head (Fig 2) is a True Power Detector, an AD8362 (IC1) integrated circuit from Analog Devices [3]. The external components required are limited to some decoupling capacitors (C3 to C6).

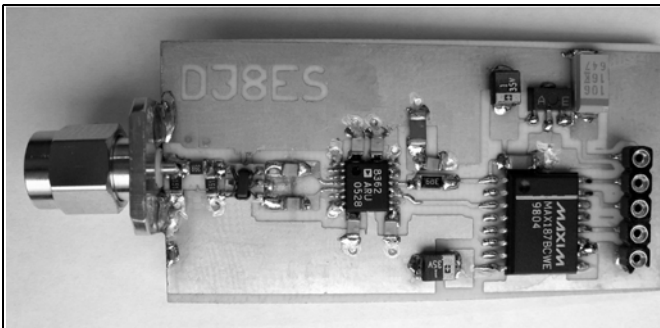


Fig 1: The completed 10 to 2700MHz power detector.

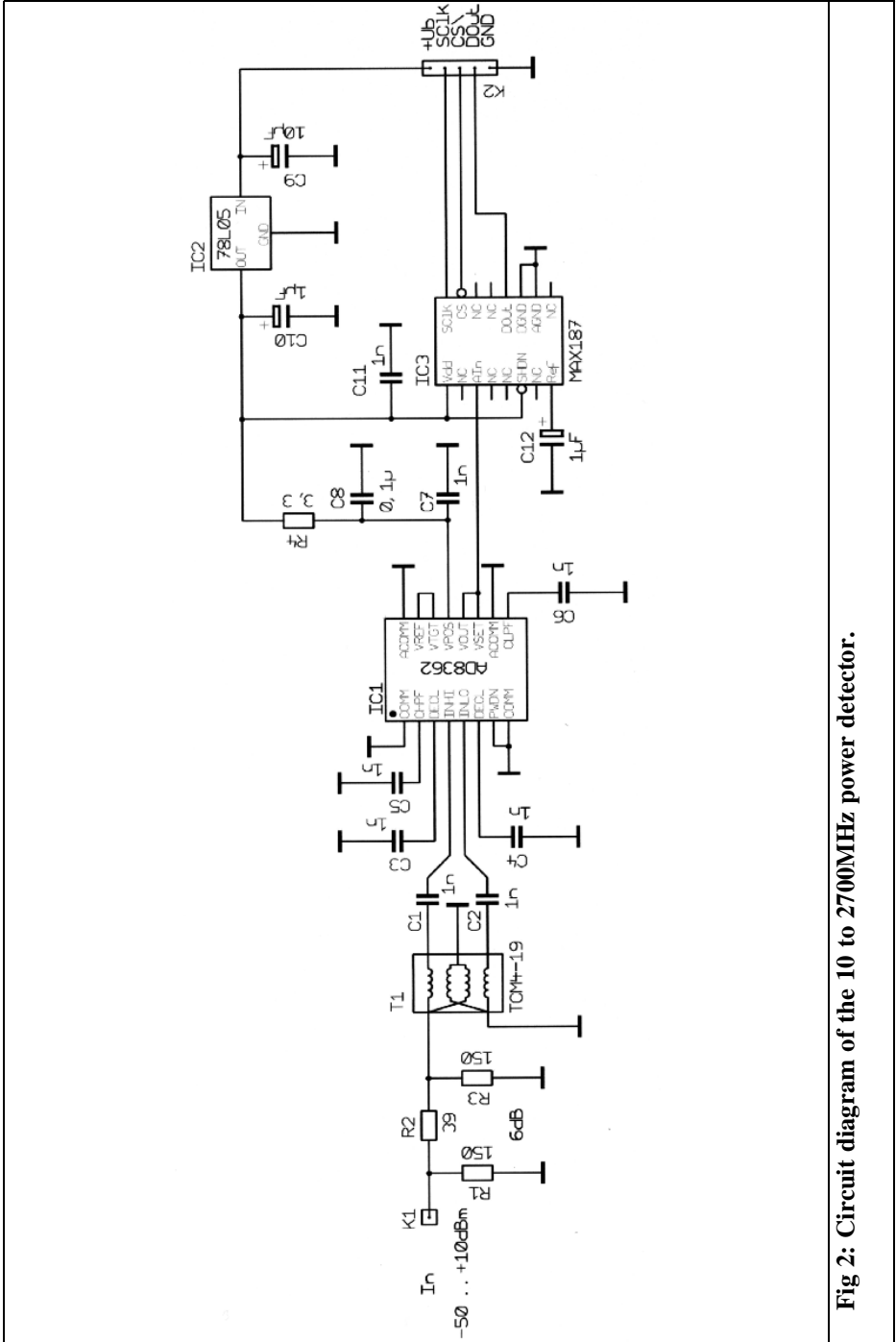


Fig 2: Circuit diagram of the 10 to 2700MHz power detector.

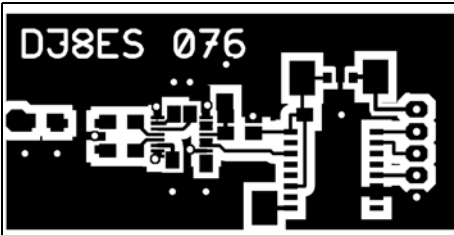


Fig 3: PCB layout for 10 to 2700MHz power detector.

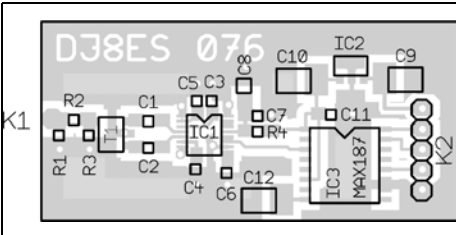


Fig 4: Component layout for 10 to 2700MHz power detector.

The input of the detector is fed via a 1:4 balun to match the 50Ω feed to that of the detector. This ensures that the maximum frequency range is achieved; Fig 6 shows the frequency response of the total circuit.

The 12 Bit A/D converter, a MAX187, (IC3) uses an internal reference voltage of 4.096V giving a resolution of 1mV per bit (12 bits = 4096 steps), or in this circuit approximately 0.02dB per single step. The interface to the A/D converter is a simple 3 wire serial interface (Clock, Serial data and chip select). This can be used for further processing with a micro controller or a PC.

The +5V supply for the two ICs is derived from a three terminal regulator type 78L05 (IC2). This makes the supply voltage range for the detector between +9 and +15V.

3. Construction

The printed circuit board for the detector (DJ8ES 076) is made from double sided epoxy material, it is 25 x 53mm and the layout is shown in Fig 3.

All components are SMD types because they give better RF performance. The components are mounted on the top side of the printed circuit board. The lower surface is an earth plane with the relevant RF connections to the top side being made with plated through holes for low inductance connections.

The first components to be fitted should be the resistors; capacitors, the voltage regulator and the 1:4 transformer in accordance to positions on the compo-

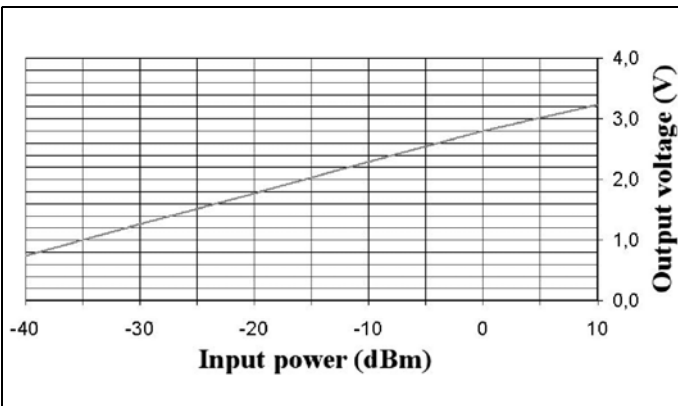


Fig 5: Transfer characteristics of the AD8262 true power detector.

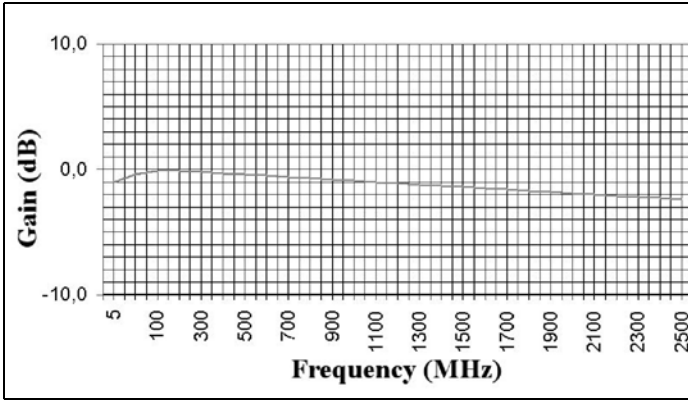


Fig 6: The frequency response only falls off in the GHz range.

ment layout (Fig 4). Before the comparatively expensive AD8362 and the MAX187 are fitted, their supply voltage should be checked.

It is important that all of the connections for SMD components are clean before they are fitted and a soldering iron with a fine 0.5mm tip is used.

The input is designed to use a flanged SMA connector. The +9 to +15V supply and the signal lines for the A/D converter are connected using a multi way cable so that the measuring head can be used remotely.

The measuring head can be built into a suitable housing, the PCB should be cut to fit the housing before assembly is started.

4. Parts list

- IC1 AD8362 True Power Detector IC, Analog Devices
- IC2 78L05 Voltage Regulator
- IC3 MAX187 12-Bit A/D Converter
- T1 TCM4-19 Transformer, Mini Circuits
- C8 0.1µF SMD Capacitor, 1206
- C9 10µF 16V Tantalum Capacitor
- C10,12 1µF 35V Tantalum Capacitor
- C1-7,11 1nF SMD Capacitor, 0805
- SMD resistors 0805:**
- G 1, R3 150Ω
- R2 39Ω
- R4 3.3Ω
- 1 x SMA flanged socket

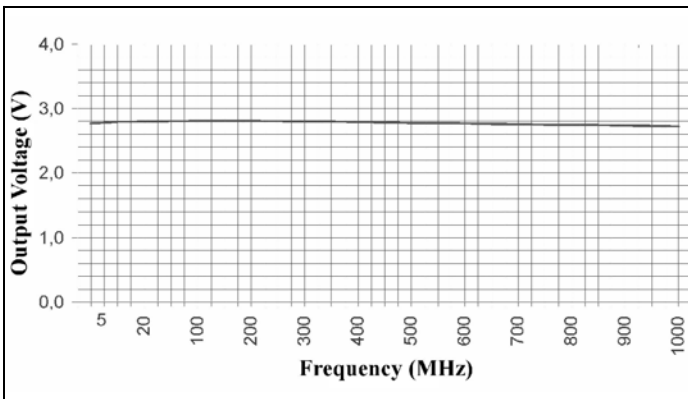


Fig 7: The frequency response is wonderfully flat up to 1GHz.



Dragoslav Dobricic, YU1AW

Ultra linear low noise preamplifier for 6m

Designs for 6m to 23cm in same family [5]

A new type of low noise preamplifier is described here, which is recommended for its exceptional noise and inter-modulation characteristics not only for normal DX operation, but also for operation under difficult conditions when there are a significant number of powerful local stations, for example during competitions. The amplifiers are designed to have low noise, unconditional stability and exceptional linearity, thanks to the use of special ultra-linear, bipolar, low noise transistors designed for TV signal amplifiers. Since they are widely used, they are readily available and low price. Construction is extremely simple with a small number of components, very simple adjustments and a high repeatability. This has been achieved using extensive computer non-linear and statistical optimisation. Designs are available for all amateur bands from 6m to 23cm with full English text [5]. The 6m preamplifier is featured in this article

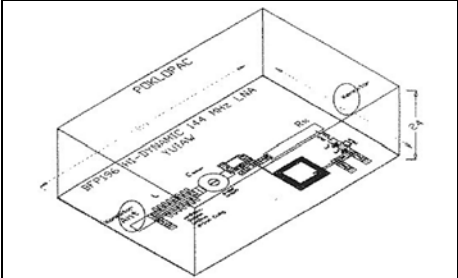


Fig 1: Mechanical layout of the preamplifier.

conditional stability on VHF and UHF frequencies [1 – 4]. However, with an increasing number of stations using greater output powers, especially during competitions, the majority of these low noise preamplifiers that are successfully used for DX, MS or EME activity, become overloaded. This is manifested by a large number of inter-modulation products that contaminate the band, this is attributed to other stations and especially those that use powerful amplifiers.

The problem, of course, could be in non-linear power amplifiers due to excess input power causing saturation. This generates a high level of inter-modulation products. However, in practice it is more frequently due to the receiver's excessively high amplification and insufficient linearity, i.e. its input stage is overloaded causing it to generate products that look as though they really exist on the band.

In order to understand how to cure this problem, it is necessary to know how,

1.0

Introduction

For three decades, MOS-FET or GaAs-FET transistors have been used almost exclusively in preamplifier designs. The reason for this is their superior noise performance and amplification. What we inevitably encounter when using GaAs-FETs, is a stability problem due to their



where and under what conditions it occurs. It turns out that the source of this problem is very high amplification, a feature that the majority of amateurs praise the most and should be praised the least or even avoided. Technically it is much more difficult to achieve the two other important properties of an amplifier: low noise factor and strong signal performance, i.e. linearity. These are the most important properties for an amplifier to those for whom decibels are not just numbers that cover ignorance.

How do we determine which amplifier is of good quality? In order to resolve this dilemma a measure of an amplifier's quality has been introduced which encompasses all of an amplifier's three characteristics: noise factor, amplification and the output level of a signal for a determined level of non-linear distortions. This measure of quality is called the *dynamic range* of an amplifier and represents a range in which the level of a signal on an amplifier's input can be changed, while the output signal degradation stays within defined limits. The lower limit of this range is determined by the minimum allowable signal/noise ratio of the output signal and it is directly determined by the amplifier's noise factor, and the upper limit is the allowable level of non-linear distortion.

The lower limit of a dynamic range is the level of the input signal that gives a previously determined minimal signal/noise ratio (S/N) at the output. If the lower limit value is a $S/N = 0$ (incoming signal and following noise are equal) and if the upper limit of this range is limited by the maximum output signal voltage, at which the amplifier, due to non-linear distortion, generates products equal to the level of noise on the output of the amplifier. Then this is the so-called SFDR (Spurious Free Dynamic Range) or a dynamic range free from distortion, i.e. products of inter-modular distortions or *IMD*.

Since the third order inter-modulation

distortion (*IMD3*) is dependant on the cube of the input signal, i.e. with each increase or decrease of the input signal by 1dB, the third order inter-modulation products increase or decrease by 3dB. It is therefore possible to calculate the maximum output level for different values of relations between products and the signal that is being used, or the value of *IMD3* products, at different output signal levels. Using an attenuator enables us to also check whether an amplifier is overloaded, i.e. recognize whether an audible signal on our receiver really exists on the band or whether it is simply the "imagination" of our overloaded receiver. This enables us to dispose of overload and *IMD3*.

Since the level of products rise faster than the basic signal, by increasing the input signal we reach the point at which third order inter-modulation products, *IMD3*, reach the level of a useful signal at the output and that point is known as *IP3* (Intercept Point). When the *IP3* value is quoted it is necessary to state if it is referenced to the input or output of the amplifier. These values naturally differ by the value of the amplifier's amplification. Occasionally, it is stated as the *TOI* (Third Order Intercept). This point is often taken as a measure of an amplifier's linearity and is highly convenient when comparing different amplifiers. Knowing the value of an amplifier's *IP3* enables us to precisely calculate the value of *IMD3* products at some arbitrarily chosen output or input signal level.

If excessive amplification is used e.g. a multi stage amplifier, a danger exists where the aerial noise and the noise of the first amplifier are amplified to such an extent that they exceed the limit of linear operation of the last transistor, at which point the amplifier is saturated with the noise itself without any signal.

The conclusion is clear: An amplifier is worth as much as its dynamic range value, rather than how great its amplification is!

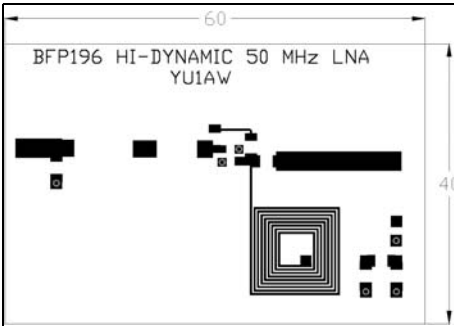


Fig 2: PCB layout for the 6m preamplifier.

2.0

Construction

If we want to construct an amplifier with the maximum amount of SFDR we have to fulfil the following conditions:

- make sure that the noise factor is as low as possible
- the IP3 is as high as possible
- has acceptable amplification.

On the one hand, amplification should be as large as possible, to prevent second degree influence on noise factor, and on the other hand it should be as small as possible so that the IP3 input is as high as possible, i.e. so that the amplifier should withstand the highest possible input signals without distortion. Compromise is essential and it usually ranges between 13–20dB amplification, depending on which parameter is more important for us.

If we want a low noise amplifier with a high dynamic range, then the choice of a corresponding transistor is extremely important. It is necessary to choose the type of transistor that besides low noise and sufficient amplification on the given frequency fulfils the condition of good linearity, i.e. high IP3 along with unconditional stability. Hitherto, MOS-FET and GaAs-FET transistors did not fulfil

this condition in a satisfactory manner. Specially built transistors for ultra linear working, primarily for CATV fulfil these criteria. For that reason, Siemens BFP196 bipolar transistors in SMD packaging were chosen. The Philips transistor BFG540/X corresponds closely to the Siemens device, it only requires slightly different base bias resistors. I should stress that BFG540 without /X could be used, but the layout of pins is different, i.e. it is not pin-to-pin compatible with the BFP196, therefore the printed circuit board has to be changed, which is not recommended.

Since we are talking about a broadband transistor whose Z_{in} and S_{11} values are relatively close to 50Ω , the input circuit has been chosen to optimally match the transistor with regards to noise, while at the same time it provides some selectivity at the input. By varying the circuit values a compromise is found which provides the highest selectivity with minimal degradation of the noise factor. On lower bands where the noise factor is not as important, the compromise was more in favour of selectivity which is more important than noise on these bands. The operating point of the transistor was also chosen as a compromise between minimal noise and maximum IP3. The output circuit is relatively broadband and it is implemented using a printed inductor to reduce coupling with the input and to provide high repeatability. In order to maintain optimal output matching that gives minimal IMD, any matching by trimmer capacitors or by variable inductances is forbidden on the output. In order to achieve unconditional stability, minimal IMD, optimal amplification and minimal noise, negative feedback is applied which cannot be changed arbitrarily.

The printed circuit board is made with the dimensions shown in Fig 2. Double sided board, type G10 or FR4 is suitable. The bottom copper surface is an unetched ground plane. SMD components are the 1206 type and the ground connec-

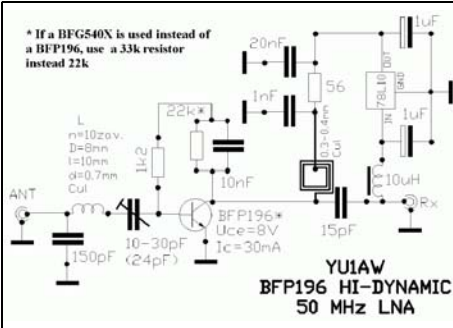


Fig 3: Circuit diagram for the 6m preamplifier.

tions are made using through plated holes or with wire links through the holes soldered on both sides. The parallel resistor and capacitor in the base bias circuit are soldered on top of each other and not next to each other. The transistor collector is connected to the wider track.

The trimmer used is either of the air or PTFE foil type, although a ceramic one can also be used if it has a suitable capacity range. It is especially important for the higher band amplifiers that the trimmer capacitor has a low enough minimum capacity.

The coil is wound, as shown in the Fig 3 with silver plated copper wire, thickness “d” and “n” turns with a body diameter “D”. The coil is to be expanded to length “l”. When the coil is fitted it has to be positioned so that the bottom is approximately 3mm above the printed circuit board.

The box for the amplifier is made so that the printed board is the bottom side of the box, as can be seen in Fig 1. The easiest way to do this is to solder 25-30mm wide copper or brass strip, 0.3mm thick, around the edges of the printed board. Connections are mounted onto the box created, and a lid is made out of the same kind of sheet metal.

Once everything is carefully soldered, check for any possible mistakes such as short circuits. Then connect the DC supply voltage and measure the collector

current and voltage. If everything is correct and properly connected and the transistor is functioning properly, the values should be close to the ones given in the circuit diagram. If the differences are within 10%, everything is OK. If the differences are greater, check the supply voltage and then reduce the value of the base bias resistor, which is in parallel with the capacitor. Make it lower to raise the collector current and vice-versa. Do not change the value of the other resistor in the base bias circuit. If the collector voltage is not correct at the correct value of collector current, adjust the value of the resistor in the supply line. Such corrections are extremely rare and are necessary only if the particular transistor used has different characteristics from the common characteristics for that transistor type.

When both collector voltage and current are within the expected range, connect an aerial to the amplifier input and the receiver to the output and adjust the trimmer capacitor for maximum received signal using a weaker station. This completes the final adjustment; the performance will be very close to the predicted values. The amplifier should be set to maximum amplification and then adjust it to a slightly lower frequency, i.e. slightly raise the trimmer capacity, until the amplification falls by 1-2dB.

The amplifier should always be mounted on the aerial pole and connected to the aerial with the shortest possible cable, using coaxial relays to switch the aerial from receive to transmit. Its power supply should be fed through the coaxial cable that connects it to the receiver, using the adapter shown in Fig 4.

As expected, very good noise characteristics have been proved in practical use, which mainly satisfy every requirement for serious DX work. This amplifier has shown exceptional linearity with IP3 values far exceeding 30dBm.

As a comparison, Figs 5 and 6 show the performance of this amplifier and a com-

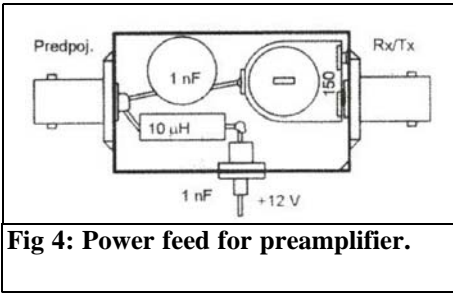


Fig 4: Power feed for preamplifier.

mon amplifier which uses an MGF1302 GaAs-FET. Both amplifier inputs have three signals of 7.1mV (-30dBm) to simulate “three strong stations on the band”. The graphs show what would be heard with an ideal receiver without its own IMD. With a real receiver, because of its possible IMD, things would look even worse! Before you accuse someone of “band wasting” check with an input attenuator whether your receiver might actually be creating IMD due to a strong input signal! The amplifier using the BFP196 is superior to the one using the MGF1302. The difference in the IMD products appearing on the output of an ideal receiver was over 30dB! Of course, in both cases the amplifiers had approximately the same amplification.

The component layout for the amplifier is shown in Fig 7. The predicted performance of the amplifier is shown in Figs 8 - 10. The values shown have been simulated on a computer, and in real life proven on a sufficient number of built

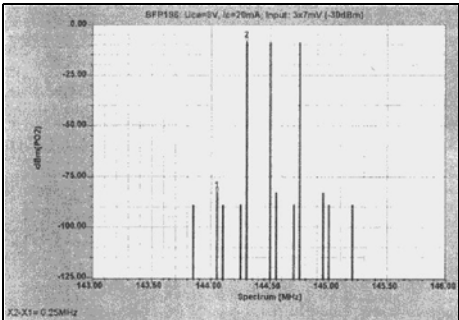


Fig 5: IMD products from a BFP196 preamplifier.

and measured amplifiers that they do not differ more than usual for this type of construction. Strict adherence to the guidelines given here will produce amplifiers with performance very close to those shown.

The final results achieved with these amplifiers in real life conditions largely depend on the IMD characteristics of the receiver used. If it has weaker characteristics, then the results may even be worse in respect to IMD because when signals, amplified in the preamplifier, reach the input of a bad receiver they cause overload and the IMD in it and the results are poor. That is why the minimum necessary amplification is recommended between this amplifier and the first mixer in the receiver or the transverter in order to preserve as much dynamics of the whole receiving system as possible.

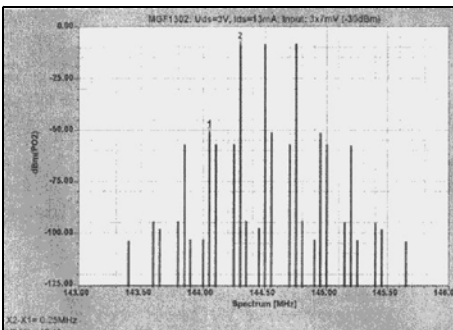


Fig 6: IMD products from MGF1302 preamplifier.

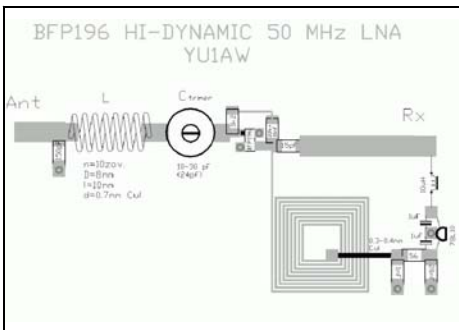


Fig 7: Component layout for 6m preamplifier.

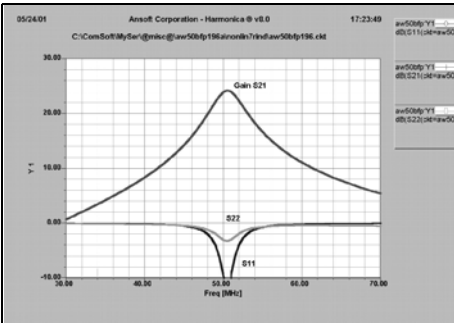


Fig 8: Amplification, input and output adjustment of the 6m preamp.

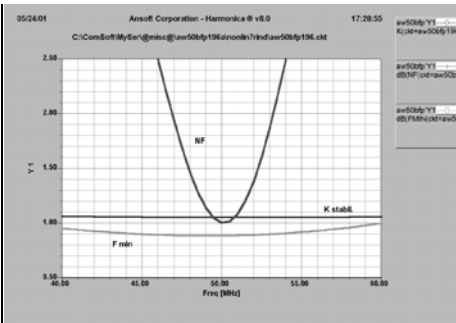


Fig 9: Noise figure, minimum noise and stability of 6m preamplifier.

If IMD is apparent in the receiver it is recommended to put a variable attenuator between the amplifier and the receiver, define the lowest attenuation at which it disappears, replace it with a fix attenuation of the same value and work in that manner. This method is highly efficient because the IMD products are attenuated three times faster than the wanted signal, so that it is possible to weaken the products to the level where they are not heard whilst preserving the useful signal with very little attenuation! Don't be afraid that you will not hear the desired signal, there is too much amplification as soon as IMD appears – feel free to lower it!

A miniature 100-500Ω trimmer potentiometer connected to the receiver input can be used in place of a variable attenuator. This can be built into the amplifier supply adaptor box as shown on Fig 4. This represents a very practical and rather elegant solution at least on lower bands. You can also use a variable 20dB attenuator used in CATV.

3.0

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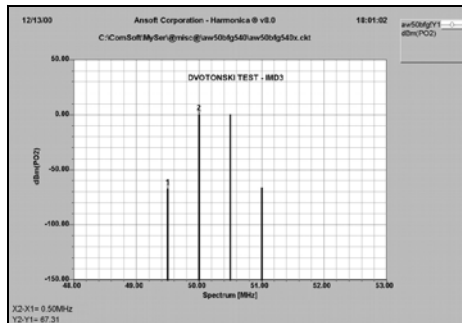


Fig 10: Two tone test and IMD products for the 6m preamplifier.



Gunthard Kraus, DG8GB

Internet Treasure Trove

ASI (Advanced Semiconductor)

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KM PUBLICATIONS,
63 Ringwood Road, Luton,
LU2 7BG, United Kingdom
Tel: +44 (0) 1582 581051
Fax: +44 (0) 1582 581051

Email:
andy@vhfcomm.co.uk

Editor

Andy Barter G8ATD

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AUSTRALIA - Mark Spooner c/o, W.I.A. SA/NT Division, GPO Box 1234, Adelaide, SA 5001, Australia Tel/Fax 08 8261 1998
BELGIUM - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133-77980. Fax: 09133-779833
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DENMARK - KM PUBLICATIONS, 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
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FRANCE - Christiane Michel F5SM, Les Pillets, 89240 PARLY, France
Fax: (33) 03 86 44 08 82 Tel: (33) 03 86 44 06 91

FINLAND - KM PUBLICATIONS, 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
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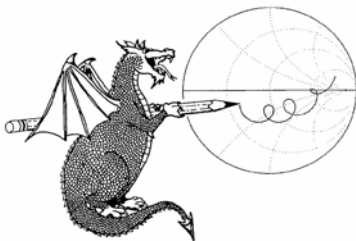
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