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VHF COMMUNICATIONS

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Editor button

Frequency setting

Microstrip

Segment dimensions in mm

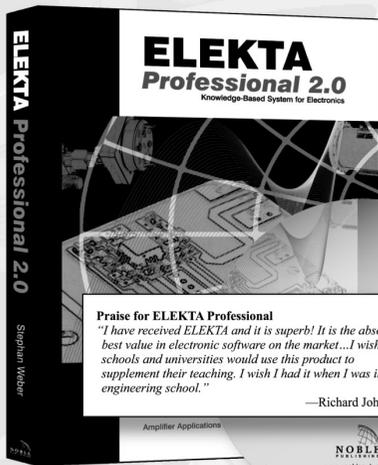
... and in wavelengths

PCB data

Layer	epsilon	loss[-3]	h[mm]
Layer 5	1.0	-1.0	0.0
Layer 4	1.0	-1.0	0.0
Layer 3	1.0	-1.0	0.0
Layer 2	1.0	-1.0	0.0
Layer 1	3.38	-1.0	1.52

ready, file in use: C:\Mstrip40\data\Kraus_01.STR

An Interesting Program MSTRIP40
Simulation and Analysis of Microstrip
Gunthard Kraus, DG8GB



ANNOUNCING

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I have had a number of request to translate the article on rotator control by DB6NT, so I have included it in this magazine. There is also an article from an author new to VHF Communications, Geoff Pike, who has also promised me an article on a 23/24cm amplifier for a future issue. I am always please to receive articles from new authors since it widens the scope of the magazine

73s - Andy



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Michael Kuhne, DB6NT

Simple Speed Control For Rotators

Standard commercial antenna rotators are too rapid and too inexact for the precise positioning of parabolic dishes e.g. for 10GHz or 24GHz. With relatively little design effort, and with the help of a liquid crystal display, rotators can be improved in this respect.

1.

Introduction

The rotating speed, and thus the accuracy of adjustment, of rotators is too imprecise for positioning of 10/24 GHz parabolic dishes with beamwidth of less than 3°. The display resolution can be improved through an additional LCD panel meter, and the accuracy of adjustment can be improved through speed control.

Since a frequency converter requires too much design effort, a pulsed motor control with a semiconductor relay was chosen. This means that no modifications are needed to the rotator.

The built in potentiometer for position sensing is relatively precise and the mechanical play of a new rotator is small.

The moving coil meter used for direction indication is replaced by a simple LCD panel meter.

The printed circuit board for the addi-

tional circuit with the mains transformer is built into the control unit. The rotating speed can be adjusted and using the 10 turn zero point and 360° calibration trimming potentiometers, direction calibration can be carried out using a beacon signal received from a known direction.

2.

Circuit description

The speed control was designed as a pulsed motor control with a semiconductor relay. The alternative, a frequency converter, needs too much design effort, and the idea was not taken any further.

Using 3 gates of a 74HC00, a pulse generator was assembled, the frequency and pulse length are adjustable. The fourth gate of the IC is used as an output driver and directly controls the electronic relay. A varistor is connected across the contacts of the relay.

The power supply line is cut in the control unit after the fuse, and is fed through the contacts and a 4.7Ω load resistor. The three motor conductors remain unchanged between the control unit and the rotator.

The supply voltage for the position indicating potentiometer in the rotator uses a

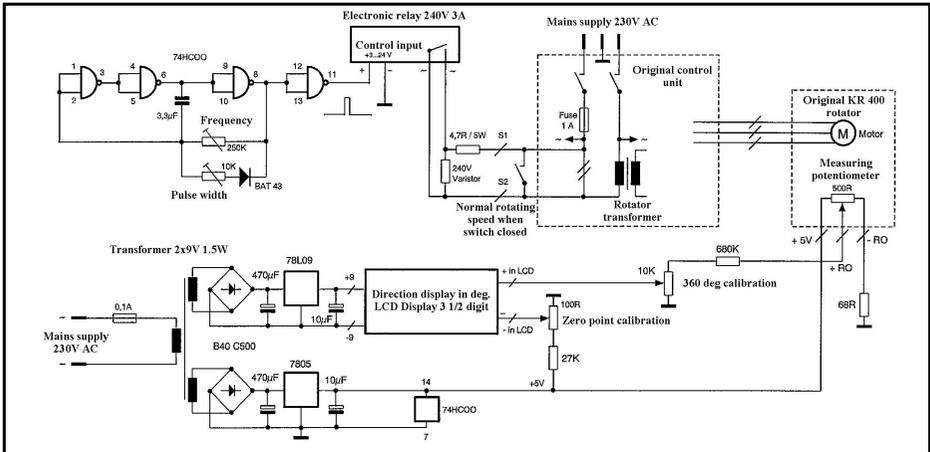


Fig 1 : Circuit diagram for DB6NT speed control.

separate transformer with rectification and stabilisation. A second winding of the transformer supplies the LCD panel meter with an isolated stabilised voltage of +9V.

The indicating voltage from the potentiometer in the rotator is connected to a voltage divider with the 360° calibration

trimming potentiometer connected to the positive input of the LCD meter. The negative input of the LCD meter is used for setting the zero point.

Parts list:

- 1 printed circuit board DB6NT/DG8NE19.98

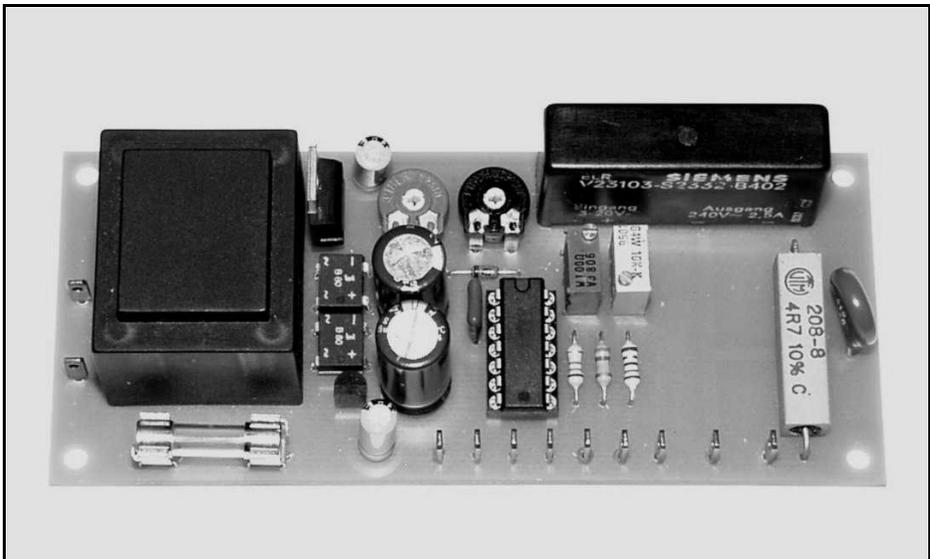


Fig 2 : Photograph of completed DB6NT/DG8NE1 PCB.

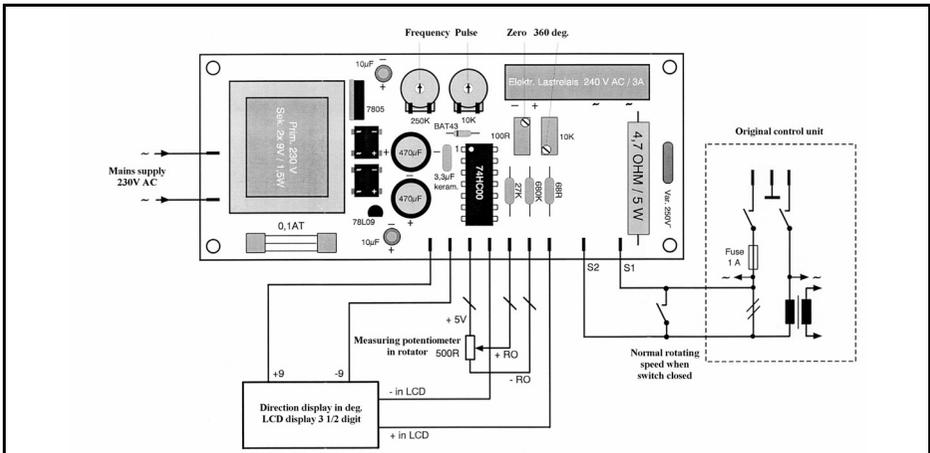


Fig 3 : PCB Component layout and external wiring.

- 1 mains transformer 2 x 9V / 85 mA 1.5W ET015
- 2 bridge rectifiers B80C800 DIL Type 88
- 1 voltage regulator 7805 TO220
- 1 voltage regulator 78L09 TO 921
- 1 IC 74HC00 DIL 14
- 1 diode BAT 43, DO 35
- 2 electrolytic capacitors 470 μ F 25V RM 5 mm.
- 2 electrolytic capacitors 10 μ F 16V RM 2.54 mm.
- 1 potentiometer 10k Ω PTC10Lh
- 1 potentiometer 250k Ω PTC10Lh
- 1 potentiometer 100 Ω Helipot type 3299 W
- 1 potentiometer 10k Ω Helipot type 3299 W
- 1 resistor 27k Ω 0.25 W
- 1 resistor 680k Ω 0.25 W
- 1 resistor 68 Ω 0.25 W
- 1 resistor 4.7 Ω 5 W
- 1 capacitor 3.3 μ F ceramic RM5 mm.
- 1 varistor S10 250 V RM 7.5mm.
- 1 electronic load relay 240V 2.5A Siemens
- 1 PCB fuse holder
- 1 LCD panel meter, 3 1/2 digit

3.

Assembly and calibration

The circuit, together with the power supply, is assembled on a 117mm. x 50mm single sided printed circuit board.

The fully assembled and tested printed circuit board is fitted into the original control unit and wired up in accordance with the connection drawing (Fig. 3).

To set up the unit, all you need to do is:

- Set the frequency trimmer and the pulse time to suit the motor in the rotator
- Adjust the zero point and 360 calibration potentiometers by turning the rotator from one end stop to the other. This will require several adjustment cycles to ensure that the calibration is correct.
- The simplest method of setting the directional calibration is to use beacon signal from a defined direction.



Gunthard Kraus, DG8GB

An Interesting Program

MSTRIP40

1.

Brief description

Name: MSTRIP40.ZIP

Area of application:

EM simulator using the moment method for the precise analysis of microstrip structures (main applications: calculation of patch antennas, slot antennas, couplers, etc.).

Its the tops:

The simulation automatically captures all physical effects and determines all important characteristics (e. g. in antennas the gain, the 3D radiated field pattern, any 2D diagram desired, the S Parameters, the cross polarisation etc.). Even slot antennas and structures with slot couplings can be correctly calculated, thanks to a refined trick.

But theres a problem:

In the present demo version, the simulation is restricted to double sided coated printed circuit boards (structures with an insulator plane and 2 metal planes). The full version creates multilayers with three insulator planes, but is no longer available free.

File type and file size:

Bundled program with download size of 2.6MB. Following installation, this gives a Windows program of about 8MB.

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

Bugs or serious computing errors:

In principle: No, but as soon as an entry was forgotten or the wrong entry was made anywhere, although there was generally no message concerning it, the program simply crashed with a runtime error message.

Suggested improvement:

Currently, printing is very poor DOS quality and is in sharp contrast to the lovely coloured graphics on the Windows screen. (Tip: you can use, for example, Paintshop Pro or some similar graphics program to help and to print the screen display directly).

User friendliness:

Easy to understand and familiarise yourself with, though some knowledge of basic principles is assumed with regard to the characteristics and limitations of modern EM simulators, and you need a lot of experience with such projects if the results are to be correct.



Aids:

Online help in the form of the English language manual (40 pages), which can also be printed out separately. Moreover, enough specimen layouts are supplied (Demo1 to Demo21), all of which you must print out, debug and analyse precisely before starting your own designs. Also the programmer (Prof. Splitt of Kiel College of Higher Education) has placed all his scientific articles and presentations on associated subjects on the homepage ready to be downloaded. Demanding, but highly recommended!

Procurement options:

Download from

<http://intek.e-technik.fhkiel.de/Splitt.htm>

2.

Brief overall view

All important characteristics and/or advantages can be found listed immediately in the first few lines of the manual. The extract reads as follows:

”.....the program is intended for the analysis and design of any type of multi-layer microstrip structures or microstrip antennas. This involves rigorous work on the solving of the necessary integral equations using the Galerkin method (moments method), and the correct Greens functions are used. All essential effects, such as radiation or surface wave generation in the substrate, together with dielectric losses, are therefore taken into account in the simulation....”

The user first has to divide the structure to be investigated into a lot of small but equal “cells” using known methods (the cells are called “segments” or “basic functions” here). Then a port is connected up and thus the circuit is activated, for example using a pulse func-

tion. The exact current density is determined in each element and then the reciprocal magnetic and electrical influences of the cells are calculated (here the assumed maximum “range” of the influence is 20 cells!). Finally the readings are totalled up to obtain the precise current distribution in the system, which leads to the determining of the electrical data such as radiation characteristics, input resistance etc.

But lets just leave aside the Galerkin method and the Greens functions used here. They can no longer be dispensed with in modern circuit design and nowadays they are used by many simulators and authors. The most important element in the foreword and in the method is actually that you can use it, for example, to obtain automatically all the radiation characteristics and electrical data for an antenna. This is also the case with the professional SONNET program, which we have already reviewed several times, but there these options were unfortunately blocked in the free “Lite” version. Thus the area of application of MSTRIP40 slowly becomes clear:

Everywhere where all radiation characteristics of an antenna are of interest, or where no substitution model for simulations is available in PUFF (e.g. slot antennas or patch antennas with slot coupling) or where we come up against the limitations of the SONNET Lite version, we reach for MSTRIP40. From my experience, however, the following things must be taken into account:

- The programmer recommends cell dimensions of between 5% and 20% of the wavelength. In practise they should not exceed 10%, as above that the errors increase. Most of the demo examples supplied have been simulated using cell dimensions of between 2% and 4%, and this clearly gives the best approximation to reality. You should not go down below 1%, as the calculation will take a very long time and the accuracy tends to

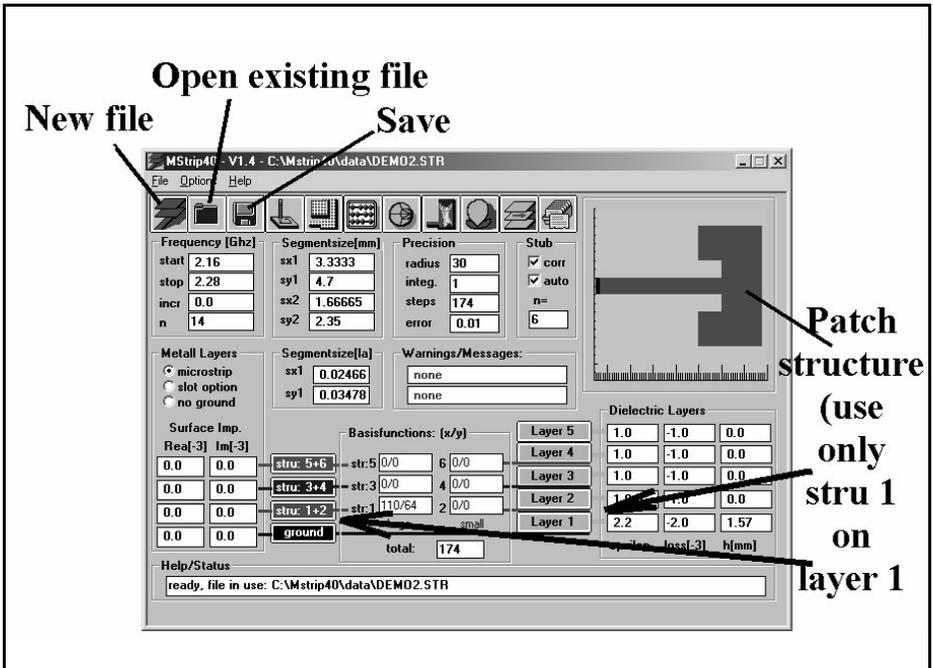


Fig 1 : Initial screen of MSTRIP40 for demonstration file DEMO2.

become progressively worse, even the program warns you about this. Moreover, if at all possible, a computer with a clock speed exceeding 300MHz should be utilised waiting around for results is just no fun at all. Another tip from experience: the height/width ratio of the two cell edges must lie roughly between 1:1 and 1:2 to retain the optimal precision of the simulation.

- Please do not make the sweep range too wide for the simulation. In my experience, the results can be trusted if the bandwidth is less than 30% of the mean frequency. This is connected with the fact that the “coupling of the activator signal into the structure” essentially generates a source of interference, which is compensated for by an automatically stub tuned to the mean frequency. Now we all know that matching using stubs is unfortunately a narrow band thing!

- For the same reason (interference due to coupling), you should never connect the input port directly to a structure, but connect it through a relatively long transmission line (e.g. with $Z = 50 \Omega$).
- A maximum of two different cell sizes can be used. “Structure 1” simulates examples of S_x and S_y on Layer 1 accurately, using the segment dimensions entered. This involves a great deal of hit and miss work with your pocket calculator until you have established values which are appropriate for both the microstrip structure and for the feed at the same time, without any serious discrepancies. And naturally you need a good stripline calculator (my tip: TRL85 from ANSOFT), since Mstrip40 can only analyse and operate with standards set by the developer. “Structure 2” also operates on Layer 1, but displays exactly half the dimensions of structure 1 and



must have it's own drawing in the circuit editor. It is then made congruent by entering a substitute with structure 1. (See corresponding demo examples).

3. Procedure using program MSTRIP40

3.1. Installation

Following download the programs and data must be unzipped on the hard disc in the usual way. It should be unzipped to the directory C:\Mstrip40. In this new directory, we find all the necessary exe files, together with two important folders:

- “mstrip40\data” contains all the simulation examples supplied.
- “mstrip40\ manual” has the operating instructions and the online help.

Before we start the program, we add an attractive icon to the Windows screen (You already know how: click on the right hand mouse button in an empty spot

on the screen, then “New link”, “Scan”, select “Mstrip40.exe”, open.....etc.). But we still can't start the program. There is something very important to do first.

The newly unzipped “mstrip40.ini” file must be copied into the C:\WINDOWS directory straight away. As a precaution, you should then open it immediately with a text editor and check and/or correct the settings. Table 1 shows the correct and comprehensive path settings for storage on hard disk C:\ in directory “mstrip40” with an automatic start up of example “DEMO2”. The path for the personal HTML editor in the fifth line may need altering. Please check everything!

3.2. Demonstration of characteristics and operation using example DEMO2

Now we start up the program and we should see Fig. 1. Unfortunately the black and white text does not reflect the blaze of colour which welcomes you. In Fig. 1 it is immediately clear that we can not only start up a new project or store a change. Using the centre button, we can actually load the various examples directly, a click brings up a long list of structure files with the suffix “str”. We make the selection by clicking on the

```
[Mstrip40]
PathName=C:\Mstrip40\data\DEMO2,STR
FileName=DEMO2.STR
EditorName=notepad.exe
HtmlName=C:\Programme\Internet Explorer\IEXPLORE.EXE C:\mstrip40>manual\manual.htm
ProgramPath=C:\MSTRIP40\

[Ms_smith]
PathNameC3d=C:\Mstrip40\data\DEMO2.slv
FileNameC3d=Mstrip40\data\DEMO2.slv

[Current3D]
PathNameC3d=C:\Mstrip40\data\DEMO2.slv
FileNameC3d=Mstrip40\data\DEMO2.slv

[Pattern3D]
PathNameC3d=C:\Mstrip40\data\DEMO2.slv
FileNameC3d=Mstrip40\data\DEMO2.slv
```

Table 1 : Contents of the Mstrip40 control file Mstrip40.ini.



corresponding file. The example is automatically loaded and the one previously in use is overwritten.

Please just call up various examples one after another and marvel at their structure, but return to “DEMO2.stu” at the end!

Lets start by examining the construction of DEMO2, which has to be analysed by the program. We are dealing with a patch antenna for 2.2GHz, in which the feeder goes directly to the optimal matching point and so the line structure of the antenna is used for transformation. This method is known from [1]. The main difference from the design at that time is the notch in the patch to connect the microstrip directly with the feed point. This is a simulation problem that PUFF cannot handle because it has no suitable model.

If you look more closely at Fig. 1, you can quickly make out the details of the structure: a printed circuit board material is used with a dielectric constant of 2.2, a loss factor of 0.002 and a thickness of 1.57 mm. The board underside is provided with a complete earthing surface and only large “basic functions” (elements or segments or cells) are used.

In Fig. 2 we see the settings of the frequency range to be swept. The user can pre-set, not only the start and stop frequencies, but also the number of simulation points. The frequency increment between two points is then automatically adjusted (provided the increment selected is not too fine, then only “0.0” is displayed). In addition, the evaluation option can be chosen for the special features of the structure to be examined (Microstrip / Slot / No ground).

Fig. 3 shows the entry fields for the segment sizes. Its worth remembering the following details:

- The edge lengths S_x and S_y may not exceed 20% of the wavelength, otherwise there are considerable errors. Dimensions below 5% of the wave-

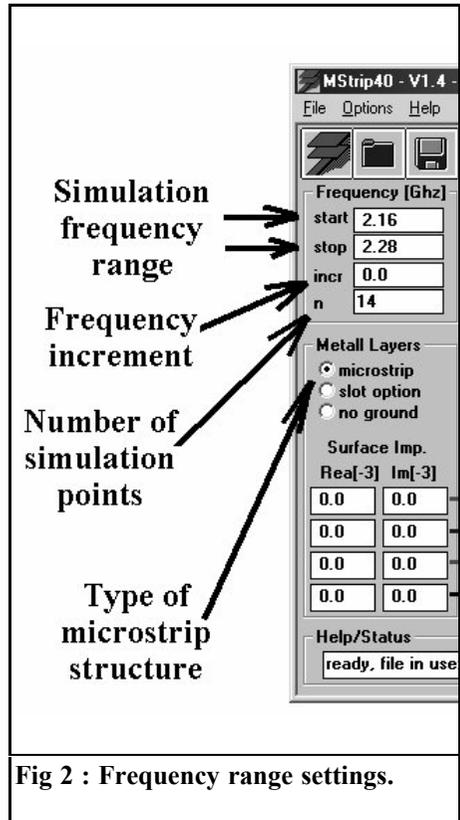


Fig 2 : Frequency range settings.

length increase the accuracy, but increase the computing time and should therefore be used with care (e. g. when the computer used has a clock speed exceeding 1GHz). Optimal dimensions are those which are approximately 2 to 3% of the wavelength.

- For better monitoring, therefore, the values of S_{x1} and S_{y1} (the “coarse” cells) are also based on the wavelength and are displayed separated in two windows.
- The S_{x2} and S_{y2} dimensions of the “fine” segments are always exactly 50% of S_{x1} and S_{y1} . This ratio can not be altered.

In Fig. 4 the settings for the simulation accuracy are shown on the left of the circuit. The definitions are as follows:

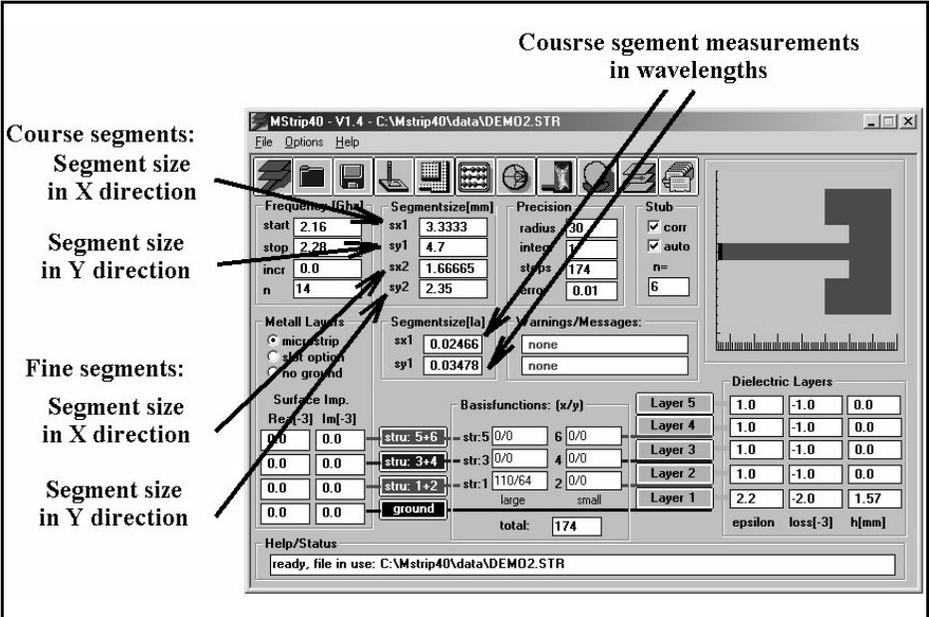


Fig 3 : Entering segment size information.

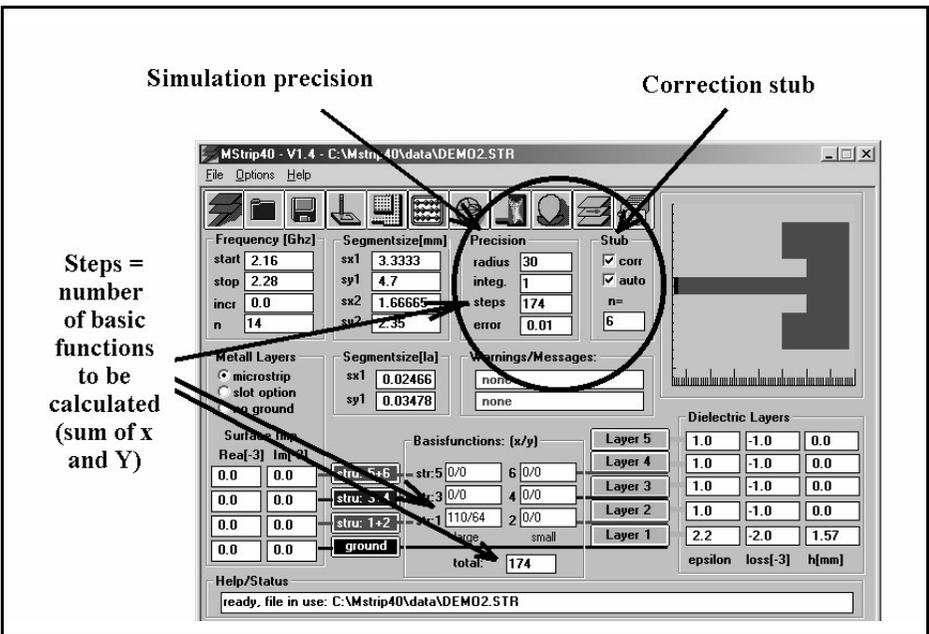


Fig 4 : Simulation precision settings and correction stub.

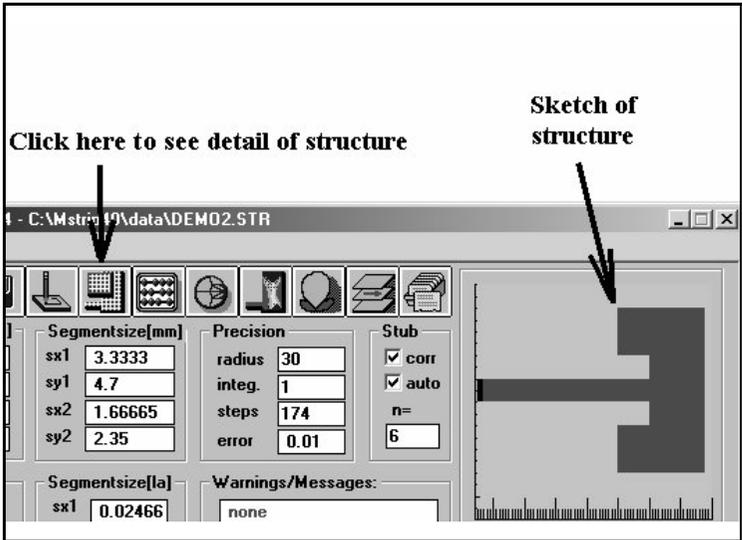


Fig 5 :
Structure
examined by
DEMO2.

- The “Precision Radius” determines (in the form of the number of segments) the “range of reciprocal influencing and coupling” in a cell. On this point, the manual supplies the information that sufficiently precise results can be expected if this radius covers two-thirds of the structure to be investigated.
- The “Integration Precision” factor is nominally 1 and lays down the numerical computing accuracy in solving the reaction integrals. Only for extremely small or extremely large basic functions or for very thick layers can it be increased up to 5 (once again at the expense of computing time).

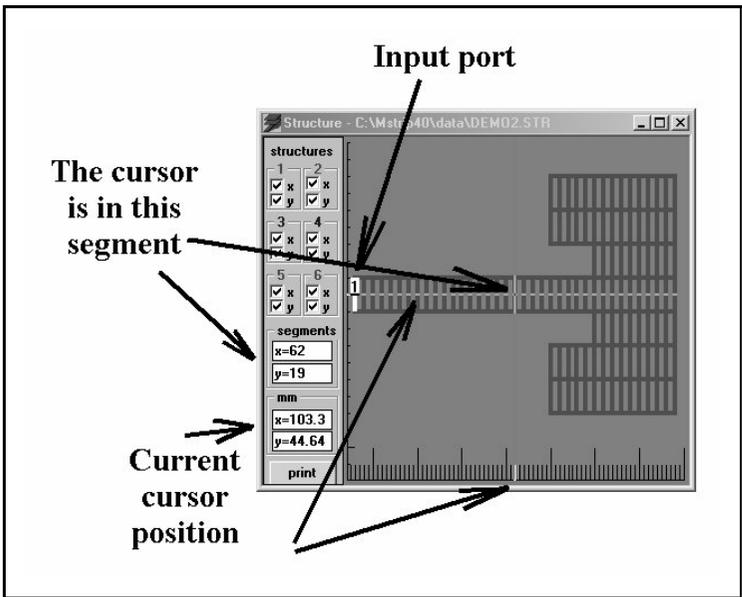


Fig 6 :
Detail
view of the
structure.

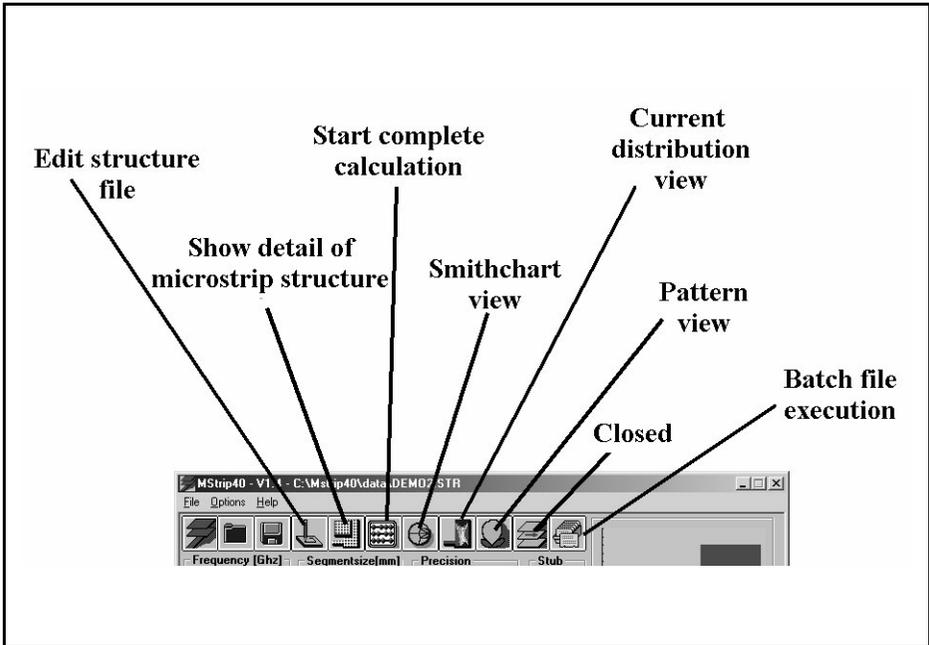


Fig 8 : The function of remaining buttons.

Very important:

This display is only there to provide precise information to the user and can not be altered. All the details, together with the type of structure or data entry, can be seen only if we press the button with the pencil (it is directly to the left of the button just used). In Fig. 7 we can see that, amazingly, everything is done with only a simple ASCII file as used by any text processing program! Its easy to see the values transferred from the screen input here for the frequency range being examined, the printed circuit board data, the segment dimensions and the structure 1 selected for this project (largest segments on layer 1). The selected entry form for the microstrip structure is especially amusing, it is embedded in a field of dots (these form, so to speak, the printed circuit board with the lower continuous earthing surface, i.e. the “ground” and can be imagined as going on for ever). Each hash sign entered

represents a segment and the “1” at the beginning of the feeder is naturally port 1 for activating the structure.

A tip:

Those wishing to experiment further should make life easier for themselves and store the demo example under another name. Then open this structure file and experiment by altering the values entered or the “drawing”. In this way, you can save yourself the trouble of starting a new file and keying in all the variables required that would take up a lot of time!

If we now look at Fig. 8, we find the functions of all the remaining buttons explained. Anyone with a lot of time to spare can re-start the entire calculation using the next button in the row (symbol: little computer) on relatively old computers, this is an opportunity for an extended coffee break! It is interesting to print out the Smith chart of the results (Fig. 9). As

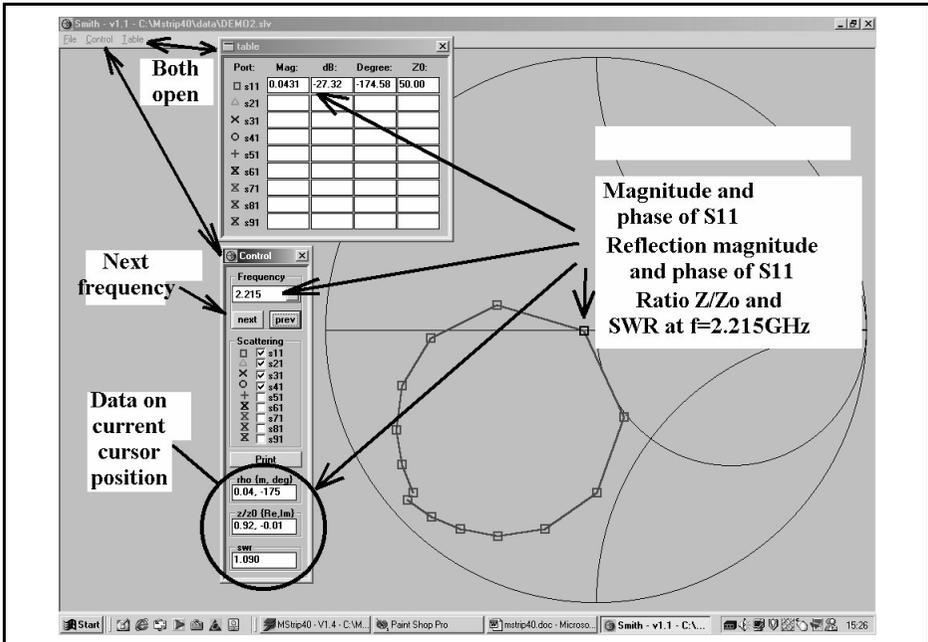


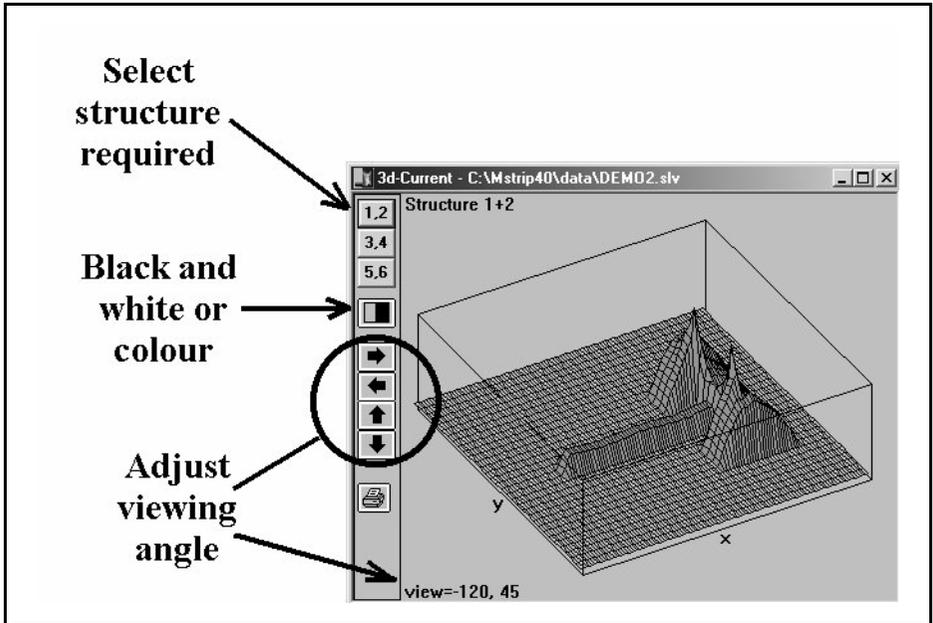
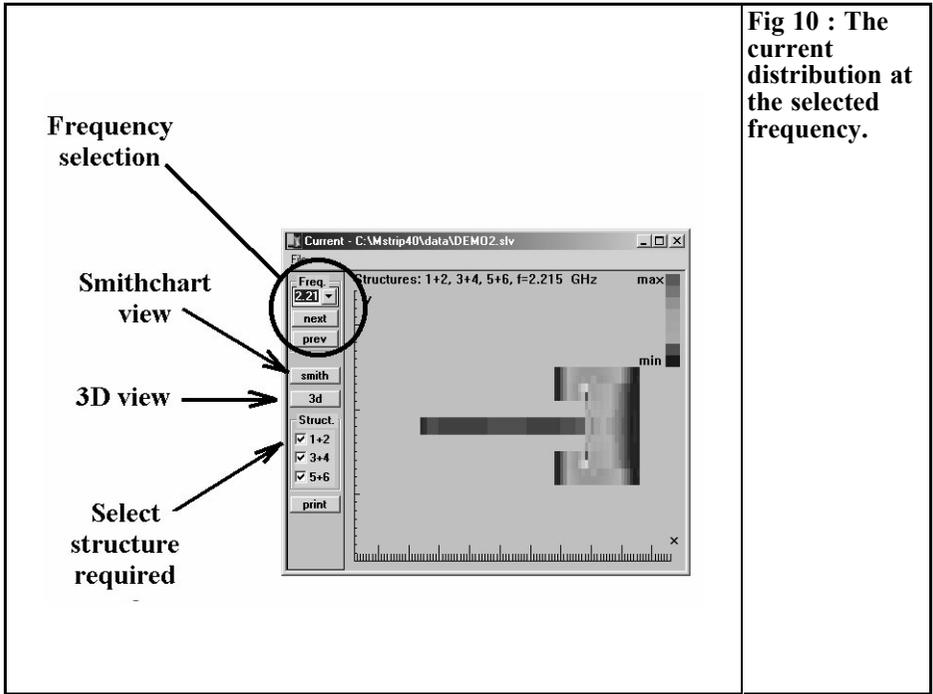
Fig 9 : The Smithchart view, once displayed open the "Control" and "Table" menus.

soon as you zoom the graphics to fill the screen, you have to open the two pull-down menus "Control" and "Table". "Table" shows the simulated S-parameters in table form for the frequency just set, whilst additional useful information is supplied in "Control". The buttons "next" and "previous" can be used, for example, to alter the frequency, with the active measurement point being brought out in black on the Smith chart.

There is something more to say about the cursor display, marked with a circle (reflection factor with level and phase, SWR and Z/Zo split into real and imaginary sections). The data displayed within the circle is only valid for the frequency actually selected if the cursor has not left the "Control" menu area. As soon as you roll over the edge into the Smith chart, the values associated with the current cursor position on the screen (based on the centre point of the Smith diagram!) are displayed. Definitely something you have to get used to, because rolling back

to "Control" does not change the display back again. You actually have to click on "prev" or "next" to switch the display back to a value from a table!

Now click on the button for current distribution. Fig. 10 cannot not show the beautiful colours on the screen, but the basic graphics principle is easy to see, as well as the additional options. Not only can you call up the Smith chart graphics onto the screen as an aid to selecting the desired frequency, but you can also call up a 3D radiated field pattern (Fig. 11). First, you select the structure to be examined (here structures 1+2, i.e. large and small segments on layer 1). You can choose between black and white or colour representation, and rotate the image using the four arrow keys. Looking at the display you will immediately see the two enormous peaks in the current density. The current from the feed connection flows in the direction of the radiating patch edges and round the corners of the notch. The display enables you to see



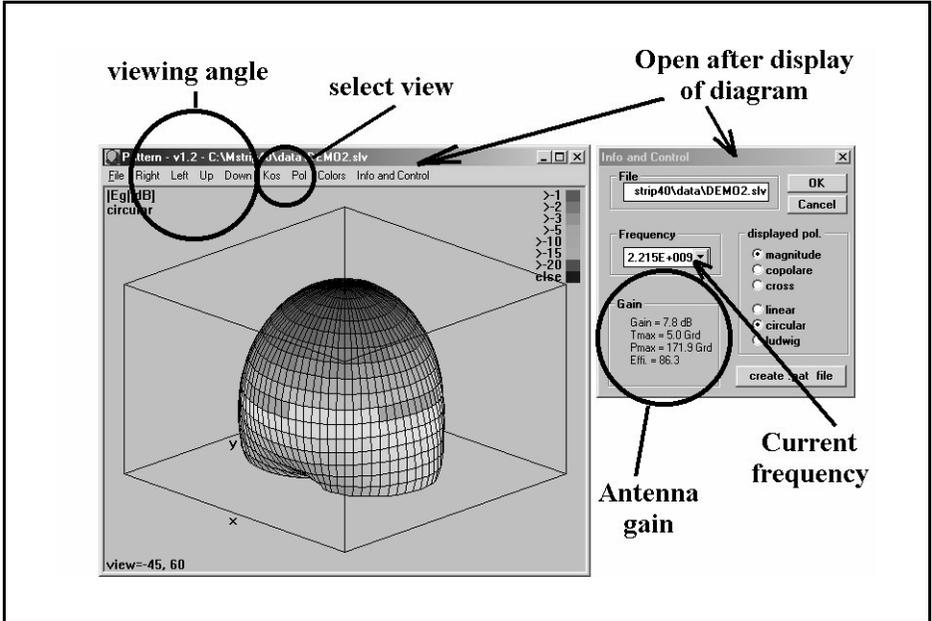


Fig 12 : Radiation pattern view with information on antenna gain.

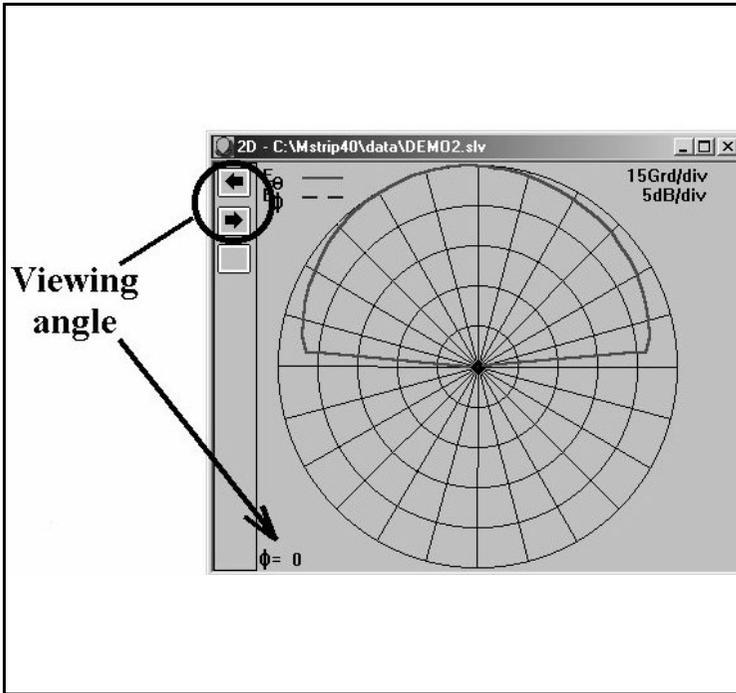


Fig 13 : 2D view of radiation pattern.

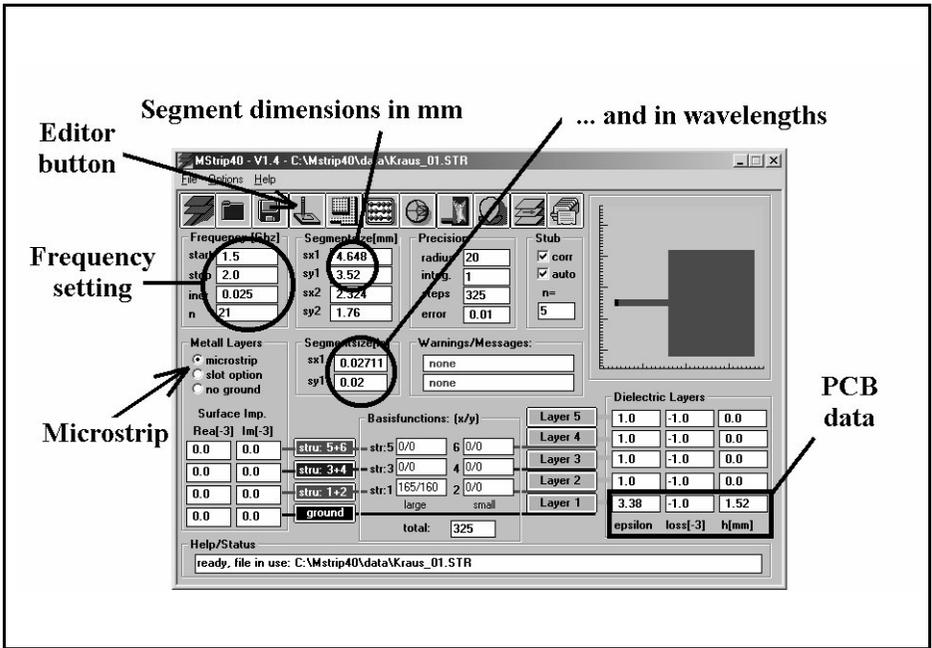


Fig 14 : Details of example entered into main input screen.

exactly what happens.

The next button brings us to the radiation pattern diagrams. Once the menu is opened, call up “Infos and Controls” and immediately read off the prediction for the antenna gain (Fig. 12) together with the details of the 3D radiated field pattern. The four arrow keys can again be used for rotation i.e. the angle of observation can be changed. But you should just play about with the various simulation options (right up to cross polarisation). To be honest, for individual menu items you really need to look through the manual or look in the appropriate literature in order to grasp the deeper meaning.

Incidentally:

If you miss your beloved 2D directional diagram, click the option “2D Pattern” under “Kos”. Fig. 13 will then make you very happy, since here too the angle of observation can be varied using the ar-

row keys.

3.3. My own example:

Determining the data for a patch antenna for METEOSAT

As a stimulation for your own work, the specimen antenna from the article in VHF Communications [1] is to be simulated here using Mstrip40, to check how far we are from the actual measurement taken at the time. The material data are as follows:

- Printed circuit board material: Rogers R04003
- Relative permittivity: 3.38
- Board thickness: 1.52mm
- Dielectric loss factor $\tan\delta$: 0.001 at 1,600MHz

The patch created and investigated had a width of 58.67mm and a length of 46.48mm, this length giving precisely the frequency of resonance of 1,691MHz (see simulation in Fig. 31 on Page 76 in

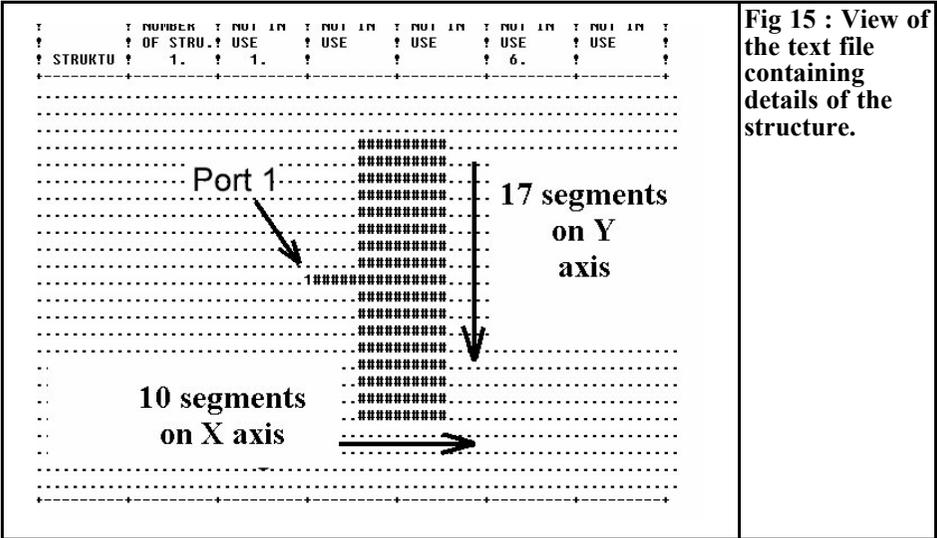


Fig 15 : View of the text file containing details of the structure.

Part 2 of the article referred to). For a feeder with $Z=50 \Omega$, we need a microstrip line with a width of 3.52 mm.

1st step:

Open the specimen “DEMO1.str” and save it using “SAVE AS”, giving it a new personal name.

2nd step:

Enter the board data correctly, split the frequency range from 1.5 to 2GHz into 21 segments and adopt the precise width of the 50Ω feeder (i.e. 3.52 mm) for the segment dimension, S_y . Divide the radiator length of 46.48 mm into 10 equal sections, thus giving a value of $S_x = 4.648$ mm.

Very important: you must save everything immediately afterwards by “pressing the key with the diskette”.

Fig. 14 shows the screen with these entries. Both the step width for the frequency steps and the values of S_x and S_y based on the wavelength are entered automatically. It can also be seen that the requirements listed at the beginning of this article (dimensions between 2% and 4% of the wavelength, and no side ratio exceeding 1:2) are correctly fulfilled.

3rd step:

Now open the circuit editor with the specified button and look at the top half of all the simulation objectives.

The antenna is in the bottom half of the page change the existing structure there with the keyboard. In the horizontal direction, enter 10 hash signs in each line (10 x 4.648 mm = 46.48 mm). There is an inaccuracy in the vertical direction, $S_y = 3.52$ mm and for a patch width of 58.67 mm we would actually need 16.66 segments. So we use 17 by filling 17 lines with a hash signs. Finally, please don't forget the feeder (in line 9) and the “1” for the input port. Fig. 15 shows what the result should look like.

4th step:

First carefully save everything and then close the editor. Press the “key with the computer” to start the simulation and, following an appropriate compulsory coffee break, call up the Smith chart graphics for the S-parameters. After calling up “Control” and “Table” as well (as previously described), you should have a graphic similar to Fig. 16. The input reflection loop, which is typical of patch antennas, is very easy to recognise, and

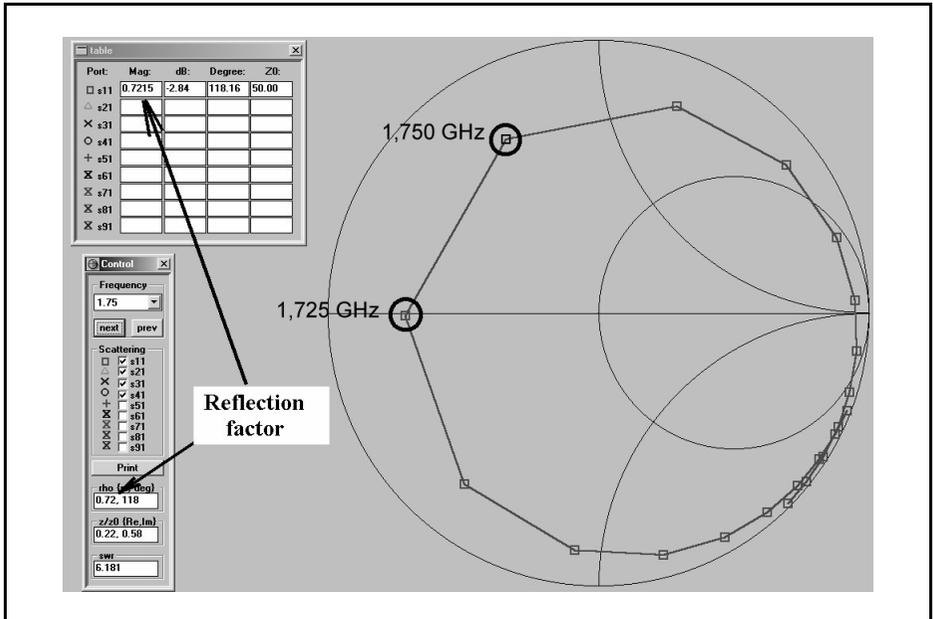


Fig 16 : The first simulation shows the resonant frequency of the antenna between 1725 and 1750MHz.

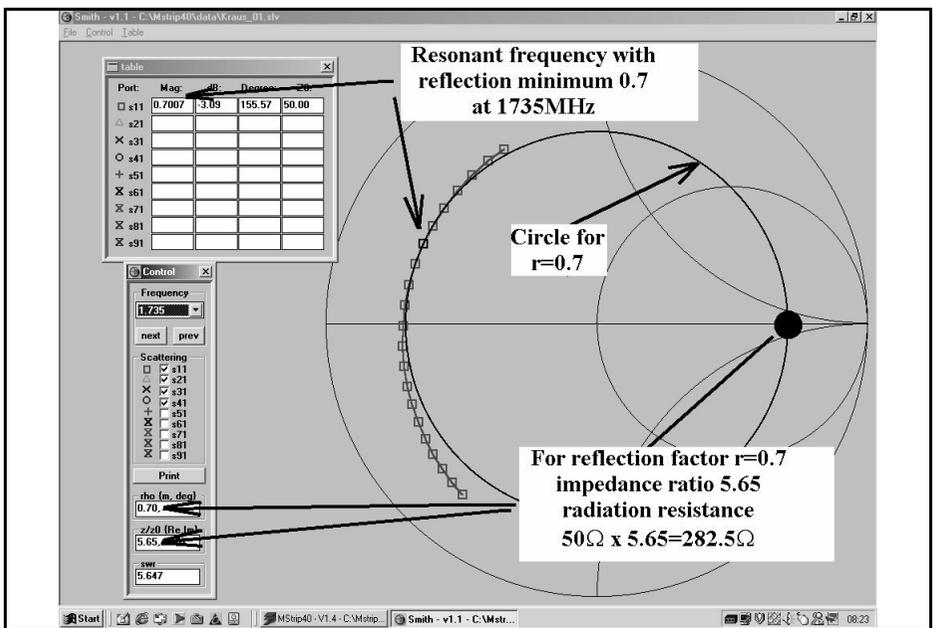


Fig 17 : The narrowband simulation show the resonant frequency of the antenna as 1735MHz and the radiation resistance as 282.5Ω.



you have only to look for the minimum of $|S_{11}|$ to find the natural resonance. According to the simulation, it lies somewhere between 1.725 and 1.75 GHz and unfortunately the selected scale division is too coarse there. Consequently, we use a new simulation with a narrower frequency range of 1.700 to 1.750 GHz. Please don't forget to save the new entries again before starting the simulation on the basis of my experience, it is perfectly reasonable to open the editor files again and check whether everything has been correctly transferred. The program frequently develops an unexpected life of its own and simply rejects such commands, but does not report this on the screen. Now and then, though, the message flashes up that, if there are problems or crashes, the two files “*.rea” and “*.slv” should be deleted from the specimen directory with Explorer to make everything work correctly again.

5th step:

In Fig. 17 we can see not only the result of this narrowband simulation, but also the next action we have to take. We now call up the individual frequencies one after another and look for the smallest value of $|S_{11}|$. This can be found at the frequency $f = 1,735\text{MHz}$, where there is a minimum value of $r=0.7$. This can be used immediately, just roll the cursor along the horizontal axis of the Smith chart from its centre point, moving right until you reach the display “rho = 0.07”. The ratio of the real component to the system resistance is then automatically displayed in the “Control” field as 5.65. The patch antenna consequently displays a real input resistance of $50\Omega \times 5.65 = 282.5\Omega$ at the feed point, since by this action we have simply “turned back” the 50Ω feed line right up to the radiating patch edge. If you take another look at the screen photo, Fig 28 on page 73 of VHF Communications 2/2002, it shows the network with its reflection factor of precisely 10% at the resonance point. Then do some calculations, you will

come up with an interesting result. The actual radiation resistance at the feed point of the antenna is actually 20 % higher here than it was assumed to be. If the start value selected for the simulation is 200Ω , in reality this would be $1.2 \times 200\Omega = 240\Omega$ - so were not all that far away with the Mstrip40 simulation of approximately 280Ω . Thus the overall result is entirely usable, even if Mstrip40 also yields a self resonant frequency of 1,735MHz instead of 1,691MHz (which is just 3 % too high).

One more interesting thing: If we now repeat the simulation with a shorter or longer feeder, or with shorter or longer time measurements, then we are nevertheless still faced with a result which gives us a frequency of resonance of between 1,720 and 1,750MHz and an input resistance of between 250 and 300Ω .

6th step:

What would all this trouble be worth without the antenna gain? Correctly calculating the radiation characteristics is certainly one of the most important properties of Mstrip40. If you press the “Pattern key” in the main menu, and then open “Info and Control” and enter a frequency of 1,735MHz, you obtain the information “Gain = 6.6 dB”.

You can then also analyse the 2D and 3D directional diagrams in any way you wish, or check the current distribution on the patch at the precise resonance.

3.4. Summary

Mstrip40 only supplies really good analyses of the extremely diverse microstrip and antenna structures if the procedures discussed are adhered to. Solid knowledge regarding the subject matter is indispensable here, not just with regard to the objects to be simulated, their main behavioural characteristics and the results to be expected, but likewise with respect to the characteristics and limitations of the EM simulators. Otherwise,



the simulation result easily becomes a lottery! Before beginning any work of your own, it is indispensable to carry out a very thorough analysis of all the specimen programs and of the manual. This is the only way to get to the bottom of the mysteries of the simulation of normal and abnormal patch antennas, slit antennas, arrays, or any type of aperture-coupled structures, etc. Who would have previously suspected that, for example, the option “no ground” could have been used to simulate the main outlines of the behaviour of microwave dipole antennas in the open air? All in all, this requires a great deal of preparatory work, but it is rewarded with a considerably deeper understanding of the relationships involved and with usable preliminary plans for a practical project.

4. **Literature:**

- [1] Gunthard Kraus: “Modern Patch Antenna Designs, Part 1 and Part 2”. VHF Communications 1/2001 and 2/2001
- [2] Mstrip40 manual
- [3] SONNET-LITE manual
- [4] APLAC manual



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Wolfgang Schneider, DJ8ES

Speech Store with Integrated Sequencer

Even on its own, a digital speech store is a practical aid, the project becomes even more interesting if a sequencer and a “Roger peep” are integrated. A micro controller is responsible for controlling the system, as in the original version [1].

1. Circuit description

The circuit diagram of the speech store with the sequencer is shown in Fig. 1. The main elements of the speech store have already been described in [1], so they are merely summarised here.

The IC ISD 2560 speech memory stores a maximum of 1 minutes speech and can play this back as often as desired. The contents of the memory are still retained even if the power fails. When the change over switch is operated, the system changes from receive to playback, or vice versa. The external analogue wiring of the IC is for filter functions, and for matching the level to the microphone input of the transceiver.

The complete control system, both for the speech store module and to the exterior, through the sequencer is operated by the AT90S2313-10PC micro controller.

The integrated sequence control serves to guarantee when the transceiver, or any additional equipment, such as a power amplifier or a masthead preamplifier, is switched on or off.

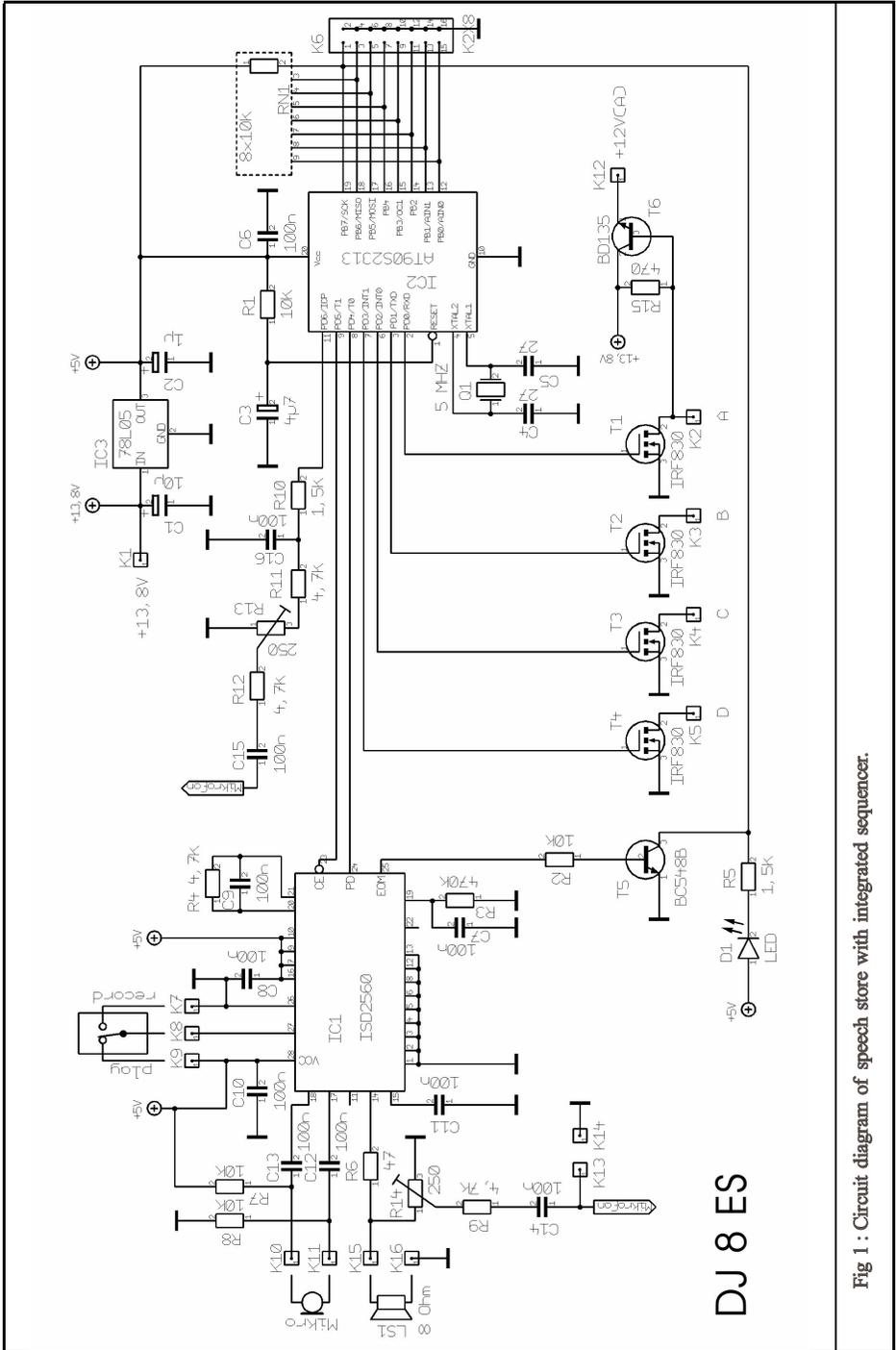
When the PTT key is activated, circuit outputs A to D are activated in a precisely co-ordinated timing sequence, and are de-activated again once the key is released. The power MOSFETs T1 to T4 are used to provide suitable amplification for the switching signals which control the PTT connection of the transceiver, the power amplifier, the antenna relay, etc.

The time cycle of 50ms, for the switching sequence of the sequencer, is derived from the crystal frequency (Q=5 MHz) through the software. The diagram in Fig. 3 makes the time sequence of the switching procedures clear.

The microphone PTT key is connected to port PB7 of the micro controller (IC2, Pin 19).

If desired, the function of a “Roger peep” can be selected through connections PB3 and PB4, as follows:

Roger Pip	PB4	PB3
no	Jumper	Jumper
pip	Jumper	-
cancelled	-	Jumper
"k"	-	-



DJ 8 ES

Fig 1 : Circuit diagram of speech store with integrated sequencer.



Fig 2 : Completed speech store in an aluminium case as produced by Frank Peter Richter, DL5HAT.

The circuit also offers the option of remotely powering an antenna amplifier though the download or through a separate cable. An operating voltage of 12V DC is available for the masthead pre-amplifier to do this, by means of the T6 (BD 135) transistor.

The “all stations call” key at port PB6 (IC2, Pin 18) of the micro controller initiates the transmission of an individual sequence, both for the initial reception of the “all stations call”, and for the play-back.

The “all stations loop” key PB5 (IC2, Pin 17) initiates an endless loop. By means of a jumper, the pause length of 5, 10, 15 or 20 seconds can be set between two calls. In the intervals, the PTT connection is always inactive and the transceiver switches to reception.

The “Stop” key PB4 (IC2, Pin 16) interrupts the endless loop or any other sequence. Thus an “all stations call” in progress can be interrupted at any time.

The “Stop” key also activates the ending of speech input.

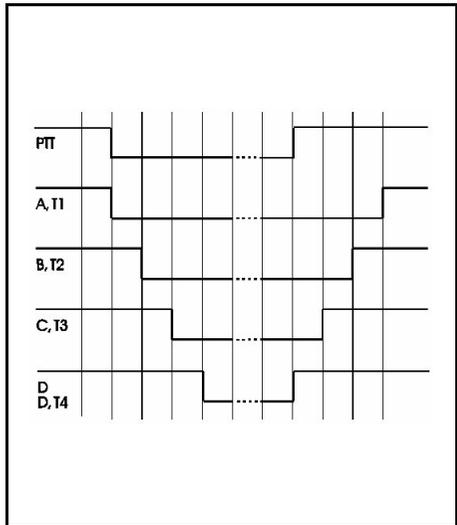


Fig 3 : Timing diagram showing the displacement of each signal.

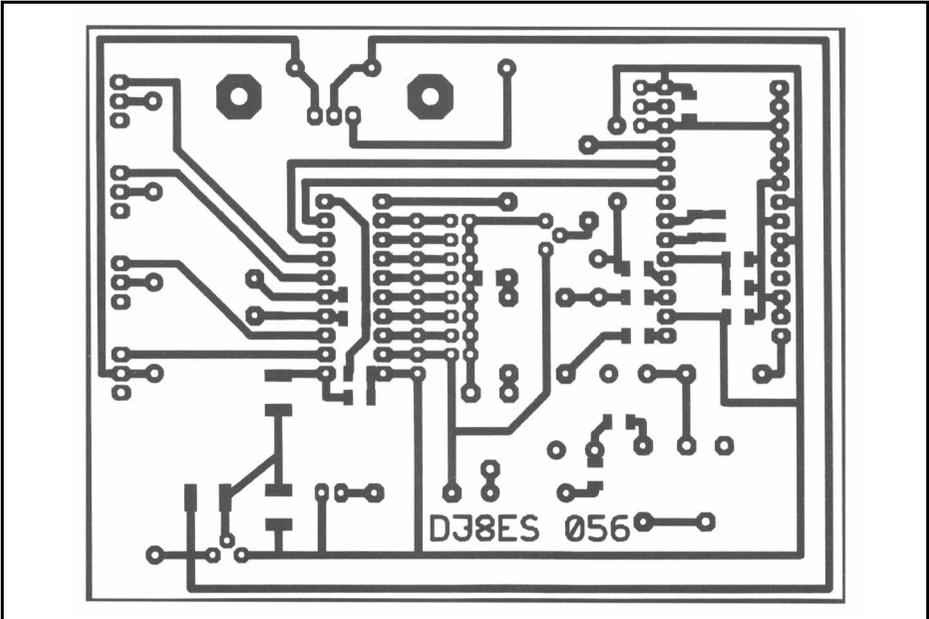


Fig 4 : PCB layout for the speech store.

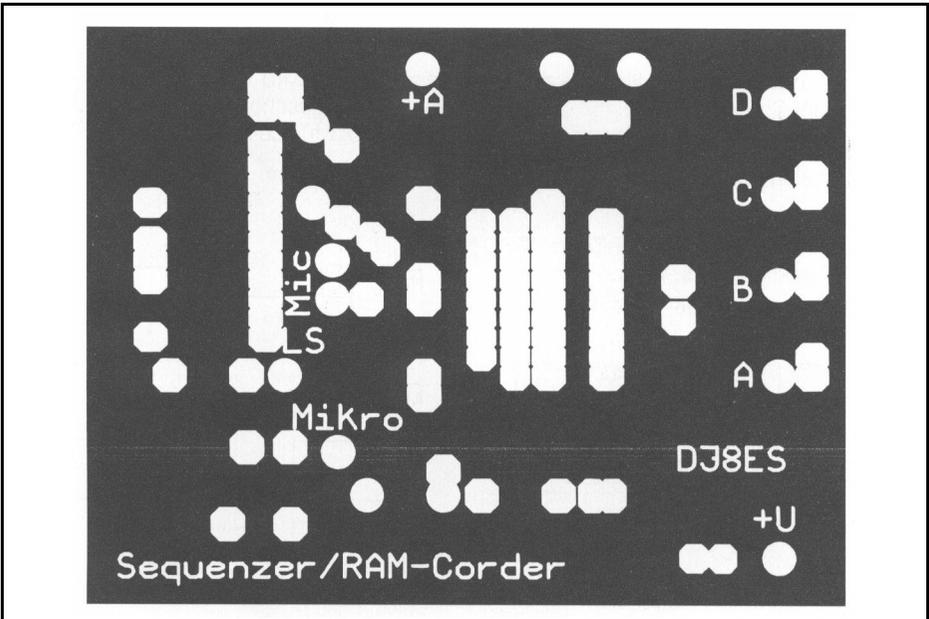


Fig 5 : Ground plane side of PCB.

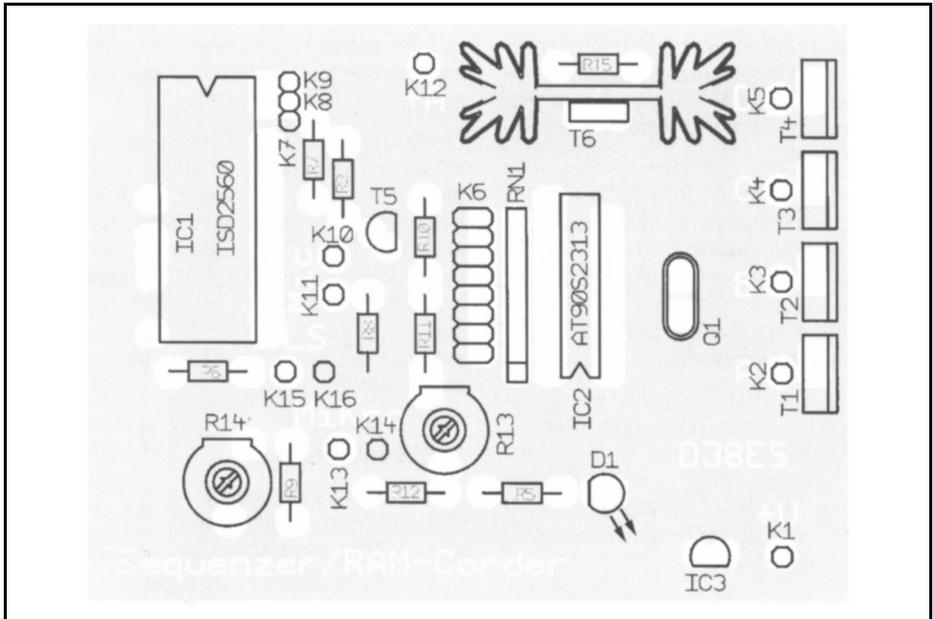


Fig 6 : Component layout for the speech store.

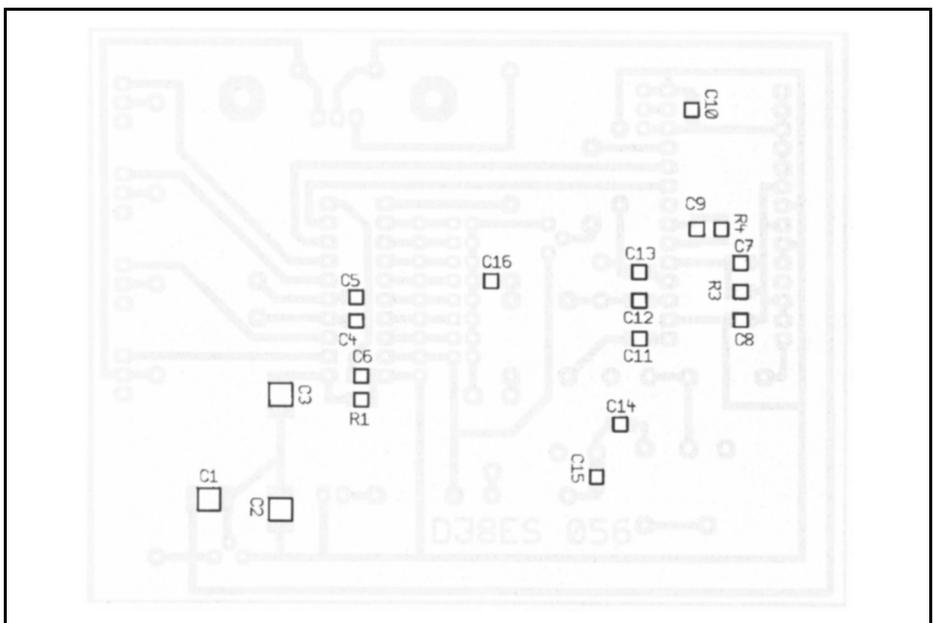


Fig 7 : Layout of SMD components.

**2.****Assembly instructions**

The speech store with sequencer is assembled on a double sided copper coated epoxy printed circuit board (Figs. 4 and 5) with dimensions of 100 mm x 75 mm. The fully coated side is an earthing surface. The earth connections are soldered in accordance with the components drawing (Fig. 6).

The two IC sockets and all the other components, with the exception of the speech store (IC1) and the micro-controller (IC2), are inserted and soldered in accordance with the components drawing, Fig. 6.

Mounting of the SMD components on the foil side is shown in Fig. 7. The inputs and outputs are fed using 1mm solder studs. The selection of the delay (waiting time) and "all stations loop" can be made by means of jumpers.

The T6 (BD135) transistor should be insulated, using a mica washer (and heat conducting paste).

The micro controller and the speech memory IC are not inserted until later on. First the function of the 5V voltage regulator is tested with a digital multimeter. To do this, the supply voltage of +12V DC, or better 13.8V DC, is applied and the stabilised voltage of +5V is measured at IC 1 (Pin 20) and IC 2 (Pin 28). Only now can the two ICs be inserted into their bases.

2.1. Component list

- 1 x Printed circuit board DJ8ES056
- IC1 ISD 2560, speech memory IC
- IC2 AT90S2313-10PC,
micro controller
- IC3 78L05, voltage regulator
- T1-T4 IRF830, power MOSFET
- T5BC548B, transistor
- T6BD135, transistor
- Q1 Crystal 5.0000 MHz
- R13,14 Trimmer 250Ω, RM 5/10mm,

- horizontal
- R15 470Ω/0.6W, RM 10mm
- RN1 Resistance array 8x10kΩ
- 1 x Loudspeaker 0.5W/8Ω
- 1 x Electret microphone
- 1 x LED red, low current
- 1 x IC holder, 20pin
- 1 x IC holder, 28pin
- 1 x Post strip, 2x8pin
- 1 x Heat sink PR-32/25,4
- 1 x Change over switch, 1pin,
On/On
- 4 x Hand actuated auxiliary switch
- 15x Soldering stud 1mm
- Resistors 1/4W, RM 10mm:
- 1 x 47Ω
- 3 x 4.7kΩ
- 2 x 1.5kΩ
- 3 x 10kΩ

SMD resistors, 1206:

- R1 10kΩ
- R3 470kΩ
- R4 4.7kΩ

SMD capacitors:

- C1 10μ/25V; SMD tantalum
- C2 1μ/16V; SMD tantalum
- C3 4.7μF/16V; SMD tantalum
- C4,5 27pF, ceram. SMD 1206
- 11x 100nF, ceram. SMD 1206

3.**Putting into operation**

Once the circuit has been assembled, the functioning of the voltage regulator checked and the ICs inserted, the equipment can be put into operation.

The waiting time for the "all stations loop" and the audio level for the unit are set at trimmer R14 or R13 for the "Roger peep". The audio output should drive the transceiver to full output, but should just not over modulate it (key word - bandwidth)! This audio output is applied parallel to the microphone at the microphone jack of the transceiver.

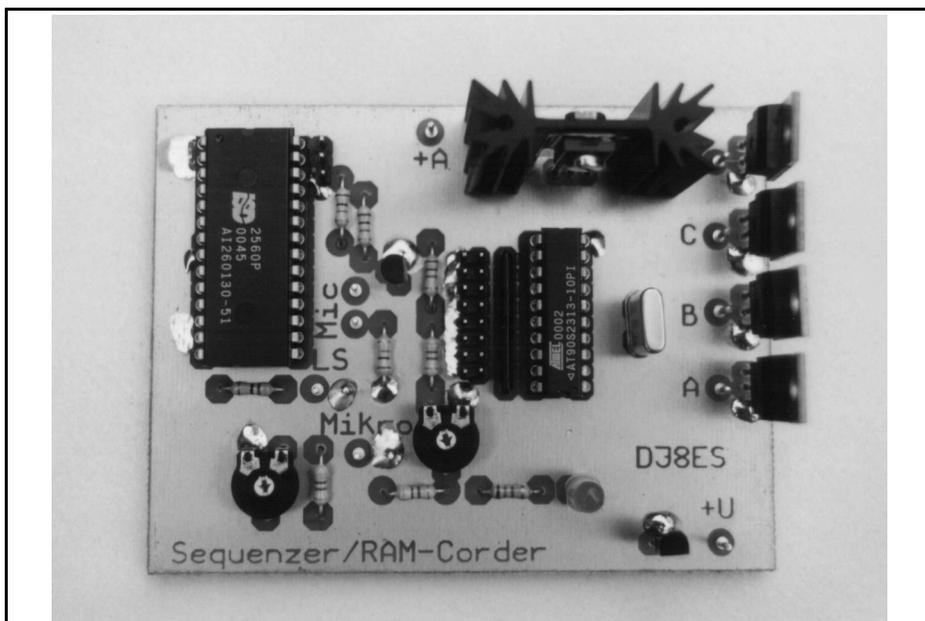


Fig 8 : Photograph of completed printed circuit board for the speech store.

The current consumption of the unit with +12 V (13.8 V) supply, in the various operating conditions is as follows: zero signal current 10mA, when PTT key is pressed 40mA and during an “all stations call” and /or when “all stations loop” is activated approximately 70mA.

The sequencer is connected into the microphone line. To do this, the PTT line should be modified. The PTT contact on the microphone controls the circuit and thus ensures that the station components are switched on and off at the correct times. The control outputs can be wired up, for example, as follows:

- A Antenna relay and/or preamplifier
- B Power amplifier
- C Transverter
- D Transceiver

If you don't want to test the functioning of the sequencer in practical operation straight away, you can initially wire up the outputs A to D with light emitting diodes (1k Ω to +12 V). The time se-

quence and thus the functioning of the switching outputs is visible.

In order to make and / or save an “all stations call”, proceed as follows:

- Switch in “Record” position
- Start recording using “all stations call” key
- Speak text (all stations call) into microphone
- End recording using “Stop” key

The LED lights during the recording. The maximum recording time is 1 Minute.

Note: In recording mode, no radio equipment should be connected, otherwise the transceiver automatically switches to transmission through the PTT connection when the “all stations call” key is pressed.

In playback mode (switch in “Play” position) pressing the key can be selected



Fig 9 : Rear view of the completed unit.

to bring about a single “all stations call” or an “all stations loop” (endless loop). The desired pause length between two calls can be set individually:

Pause	PB1	PB0
5sec	Jumper	-
10sec	Jumper	-
15sec	-	Jumper
20sec	-	-

During the pause, the PTT connection goes to “high” and the transceiver switches to reception.

The “Stop” key interrupts the endless loop or other sequences.

Finally, once again my sincere thanks to Frank Peter Richter, DL5HAT, for developing and testing the software for the speech memory with the integrated sequencer and the “Roger peep”.

4.

Literature references

[1] Wolfgang Schneider, DJ8ES: Digital Speech Memories for Radio Applications, VHF Communications, issue 2/2001 pp 87 - 91

[2] Wolfgang Schneider, DJ8ES: Micro controlled: Sequence Control (Sequencer) with Roger Peep, *Funkamateure* 49 (2000)



Prof. Gisbert Glasmachers

The Sensitivity of Radio Equipment

The technical options for modern radio equipment are mainly dependent on the circuit technology, antennas and the type of modulation. Other essential factors are the signal-to-noise ratio and the selection of the appropriate transmission method.

1. Overview

The following article describes the technical options for modern radio equipment to give the greatest possible sensitivity. It takes in the influence of the antenna, the circuit technology of the receiver, the type of modulation and the decoding. Tips are given on the optimum design of reception systems. The central focus is on the signal-to-noise ratio required for the individual application. In addition, some relationships are pointed out involving the noise, the noise figure and the signal-to-noise interval. Improvements are suggested that can be introduced using digital transmission methods.

One basic requirement for radio equipment is that the range should be as great as possible, provided a specific transmitting power is maintained. Thus only the antenna and the receiver are available for optimisation.

Antennas have a maximum power level that can be transmitted to them, determined by their positions and sizes and by any directional characteristic. The limits are defined by costs and handling.

For receivers, the sensitivity limits are predominantly defined by the noise. This limit varies a good deal, depending on whether analogue or digital modulation is used for data transmission. Using high quality components with good design of amplifiers, frequency converters and filters, and finally with good modulation methods and error correction, sensitivity can now be taken far beyond the previous margins. This article is intended to offer a general survey of the current state of the art.

2. Antenna equipment

Depending on the operating range, radio antennas can take widely varying forms. Receiving equipment for signals from space ranges from the 60cm dish for satellite reception all the way to the radio telescope with a diameter of 100m. Roof antennas for VHF and television reception are also widespread.

Then there are the tiny rod or flat top antennas in mobile telephones, as well as the considerably larger antennas on the



radio masts of base stations. Here too, the general rule is that the reception field strength available must be converted into as high an antenna power level as possible. There are basically several options available for bringing this about [1],[2]:

The antenna should be as big as possible

A large antenna can pick up more power from the electromagnetic field than a small one. Consequently, a large antenna increases the sensitivity of the reception equipment.

One absolutely classic example is the combination of the mobile telephone and the base station. Both pieces of apparatus are designed in such a way that they have approximately the same range during transmission and reception, and thus also have the same limiting sensitivity. The ratio of the transmission power levels here varies very widely, the base station has a transmission power level about 100 times that of the mobile phone. Moreover its antenna is much larger than that of the mobile. The base station needs a high transmission power level because the mobile telephone, with its tiny antenna, can not receive much power. The base station also needs a large antenna in order to receive the weak signals from the mobile phone. In this case, the limits are determined by peoples comfort. The mobile must be small, because its user would not accept a large antenna.

The antenna should have a directional characteristic

Paraboloidal reflector antennas have a strong directional characteristic. This effect can easily be observed when a satellite dish is being aligned. The signals from the satellite itself cover an area totalling millions of square kilometers in all. However, its signals are received only if the dish is accurately aligned with the satellite.

If transmitters and receivers never change their location, this means the sensitivity can be increased many times over, as opposed to a simple "all round

reception". It is also easier to tune out any sources of interference.

Even if such ideal pre-conditions are not present, it is sensible to use a directional characteristic. We can thus make sure, even for mobile telephones, that as little power as possible is emitted directly into the head of the telephone user. The major lobe points in the opposite direction, so that any possible biological damage is kept as slight as possible.

The antenna must be optimally matched to the receiver

The impedance at the connection point of the antenna must be resistive, in order to guarantee an optimal connection to the receiver's feeder. Any reflection reduces the available power at the receiver input.

Since antennas often have a marked frequency response, the phase angle of the impedance is only relatively small in the vicinity of resonance. In other words, the antenna should be carefully tuned to the transmission frequency. This expense is worthwhile for systems with a narrow frequency range, in particular, such as, for example, on the bands licensed for general purposes at 433 MHz and 869 MHz.

The difference in range between a section of "thrown out wire" and a carefully tuned and directed antenna is considerable.

3.

The signal-to-noise ratio

The heterodyne principle is nowadays in use practically everywhere. Whether single or double conversion is involved has little to do with the sensitivity. The deciding pre-conditions for high sensitivity are laid down in the input stage before the first mixer.

The variable that limits everything is the signal-to-noise ratio (S/N). This ratio can



become progressively lower, and thus worse, during the passage through the various stages of a receiver. Pre-amplifiers (LNA), filters, mixing stages and local oscillators are decisive for a sufficiently high receiver S/N. The S/N should be pre-set initially [3]:

The definition of the S/N is based on a comparison between the signal power and the noise power:

$$\frac{S}{N} = \frac{P_{SIG}}{P_{NOISE}} \quad (1)$$

In general, both the signal power and the noise level are functions of the frequency.

A generally recognised example of a signal is human speech. It encompasses a frequency range between approximately 50Hz and 10kHz. The intensity within this spectrum is variable. In addition, there are variations between men and women, children and adults, soft speech and loud shouting. Because in these circumstances we can not arrive at a uniform definition, we usually prefer instead to take one or more pure sinusoidal oscillations with a precisely defined amplitude as a signal, especially for the purposes of measurement.

Noise is also a variable distributed over the frequency range. An output uni-

formly distributed over the entire frequency range is described as "white noise". Fig. 1 shows, in simplified form, the spectrum distribution of a speech signal and the noise on the frequency axis. The area below the curve in question provides a measure of the output.

The interference to a speech or music transmission is considered to be mainly dependant on the signal-to-noise ratio. For a telephone connection on the margin of comprehensibility, a level of about 10dB must be attained. That means that the signal level must be 10 times higher than the noise level. For an acceptable television image, we need 45dB. For radio transmissions in stereo with hi-fi quality, we need 60dB. CD players give values of over 80dB. The parameters for acoustic and visual reproduction systems are thus defined.

4. The noise figure

Electronic equipment such as amplifiers, mixers and demodulators essentially generate a noise power at their outputs. This behaviour is described as the noise figure (NF). For definition, the reference value of the noise power is a passive dipole, e.g. an ohmic resistor. At room temperature, it generates an available noise power density of 174dBm/Hz. This value is determined on the basis of physical principles. We speak of "thermal noise" or "white noise", because the origins lie in the kinetic energy of the atoms, which increases in direct proportion to the temperature. Essentially, a noise source of this type exists in every electronic component.

The model of an amplifier includes a virtual noise source at the input. The amplified input noise then appears at the output - see Fig. 2.

With logarithmic scaling, e.g.. in dB, the

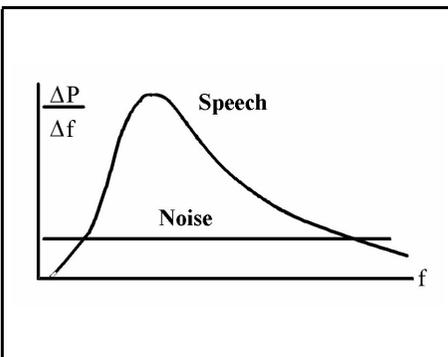


Fig 1 : Spectral distribution of speech and the white noise.

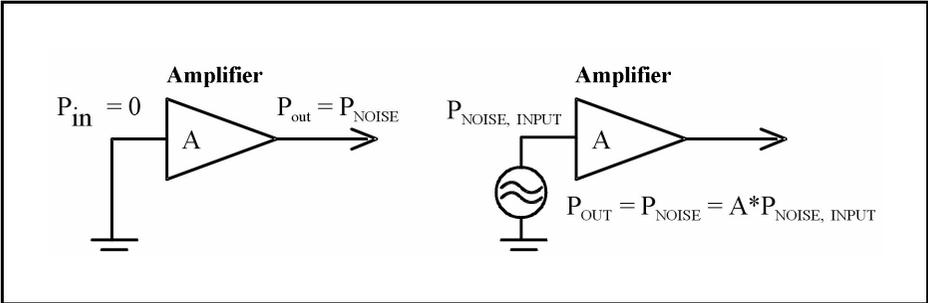


Fig 2 : Schematic of an amplifier showing noise amplification.

multiplication becomes an addition:

$$\frac{P_{NOISE}}{Hz} = \left(-174 \frac{dBm}{Hz} \right) \cdot A$$

Hence:

$$P_{NOISE} \left(\frac{dBm}{Hz} \right) = -174 + A[dB] \quad (2)$$

The power can be obtained by multiplying with the bandwidth, B:

$$P_{NOISE} [dB] = -174 + A[dB] + 10 \cdot \lg(B[Hz]) \quad (3)$$

In addition to the inevitable noise described above, real amplifiers also produce an internal noise. This is described using a multiplier NF, which is bound to exceed one:

$$P_{OUT} = P_{NOISE} \cdot NF$$

Hence formula (4):

$$P_{OUT} [dB] = -174 + A[dB] + 10 \cdot \lg(B[Hz]) + NF(dB)$$

A noise figure of NF= 0dB tells us that the amplifier has no additional internal noise. Three dB means that the noise power at the output is twice as high as we would expect on the basis of the thermal noise alone.

Information on the noise figure is provided for all essential components of a radio receiver. The first amplifiers in the input stage are always given a particularly low noise figure and are known as “low noise amplifiers” (LNA). The best

industrial products attain values approaching 1dB here.

4.1. S/N for circuits with filters and amplifiers

The concept of a voltage divider is usually associated with two ohmic resistors. Here it should be taken as referring to all passive circuit elements that have a lower voltage at their outputs than at their inputs. So these even include lines and filters. These components essentially worsen the signal-to-noise ratio. For the sake of simplification, all the equations below are based on the pre-condition that line matching has been carried out. First, we shall investigate the behaviour of a voltage divider with the attenuation factor D (Fig 3).

At the input a noise-free signal voltage is applied. At the output, thermal noise is obtained (174dBm/Hz). This is com-

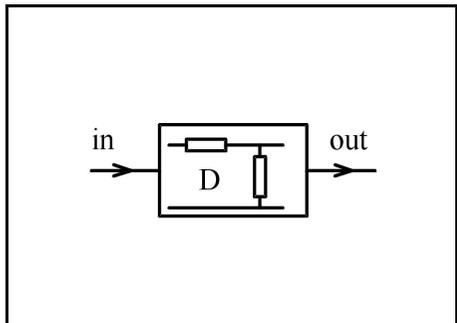


Fig 3 : Schematic of voltage divider.



pletely independent of the attenuation factor D. So if there is a signal-to-noise ratio of S/N1[dB] at the input of the voltage divider, then there is a worse value of:

$$\left(\frac{S}{N}\right)_2 [dB] = \left(\frac{S}{N}\right)_1 [dB] + D [dB] \quad (5)$$

at the output. The reduction in the signal-to-noise ratio becomes particularly relevant in the input stage of radio receivers, where the level is often not much higher than the thermal noise. Two circuit variants can be found in practise - see Fig. 4. On the left is the band filter before the LNA, and on the right it is after it.

Let us assume that a noise-free signal is applied at the input. Because some of the following equations include addition and multiplication or division, they can not be written in logarithmic form (in dB). The amplification and the attenuation must then generally be defined for output. The signal-to-noise ratio S/N for the left-hand circuits can then be determined as:

$$\left(\frac{S}{N}\right)_{LEFT} = \frac{P_{SIG} \cdot D}{\left(-174 \frac{dBm}{Hz} \cdot B\right) \cdot NF} \quad (6)$$

As can be seen, the amplification factor has no influence on the S/N in the left-hand circuit. On the other hand, the attenuation, D, of the filter and the noise

figure, NF, do have a negative effect. So we must make sure that filters with as little insertion attenuation as possible are used. For SAW filters, this is usually in a range between -2 and -5dB. In the right-hand circuit, we obtain:

$$\left(\frac{S}{N}\right)_{RIGHT} = \frac{P_{SIG} \cdot A \cdot D}{\left(-174 \frac{dBm}{Hz} \cdot B\right) \cdot (1 + NF \cdot A \cdot D)} \quad (7)$$

For a high amplification factor A, a practical approximation can be obtained:

$$\left(\frac{S}{N}\right)_{RIGHT} = \frac{P_{SIG}}{\left(-174 \frac{dBm}{Hz} \cdot B\right) \cdot NF} \quad (8)$$

In this case, the negative influence of the attenuator is absent. The right-hand circuit is therefore better than the left-hand one. A more precise analysis can be obtained if the two equations (6) and (7) are cross-divided:

$$\frac{\left(\frac{S}{N}\right)_{LEFT}}{\left(\frac{S}{N}\right)_{RIGHT}} = \frac{1}{NF \cdot A} + D \quad (9)$$

From this result, the factor can be determined by which the two S/N values differ. A plausible numerical example should make the difference clear:

$$D = -3dB, A = 18dB, NF = 2.5dB$$

The S/N for the left-hand circuit is then

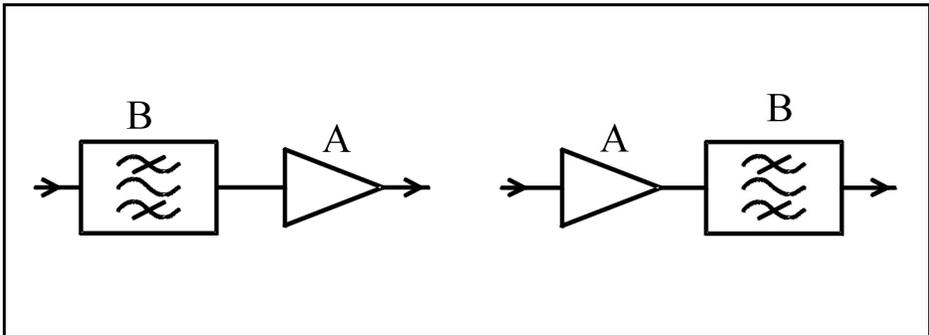


Fig 4 : Two methods of using a bandpass filter.

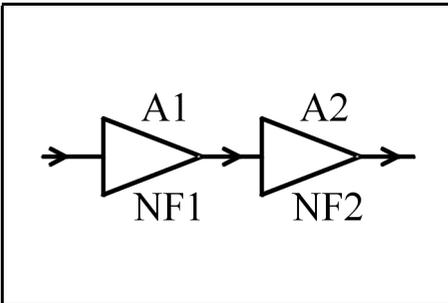


Fig 5 : Schematic showing the use of two amplifiers.

only about half as large as that for the right-hand circuit, i.e. 3 dB worse. Naturally, other arguments can be put forward in favour of the left-hand circuit. It is better protected against powerful external transmitters outside its own frequency band.

4.2. The mixing stage

To change the frequency into the intermediate frequency, a mixer and a local oscillator are used. The most important factor is that the mixer has a considerable influence on the signal-to-noise ratio. In principle, switched multipliers have considerably worse noise figures than a linear amplifier. Normal values are between 10 and 18dB. In simplified terms we can use network of two elementary amplifiers - see Fig. 5.

The first amplifier represents the LNA and the second the mixer stage. The two have different amplification levels (A_1 , A_2) and noise figures (NF_1 , NF_2). The attenuation of the filters is likewise left out of consideration here, as is the conversion to another frequency. For this configuration, we have the following signal-to-noise ratio at the output:

$$\frac{S}{N} = \frac{P_{SIG} \cdot A_1}{\left(-174 \frac{dBm}{Hz} \cdot B\right) \cdot NF_2 \cdot [1 + A_1 \cdot NF_1]} \quad (10)$$

One general requirement for a radio receiver is that the stages after the pre-amplifier should not make the signal-to-noise ratio any worse. If this require-

ment could be completely met, we would obtain:

$$\frac{S}{N} = \frac{P_{SIG}}{\left(-174 \frac{dBm}{Hz} \cdot B\right) \cdot NF_1} \quad (11)$$

To make the result of equation (10) as close as possible to that from equation (11), the following approximation must be fulfilled:

$$A_1 \gg NF_2 \quad (12)$$

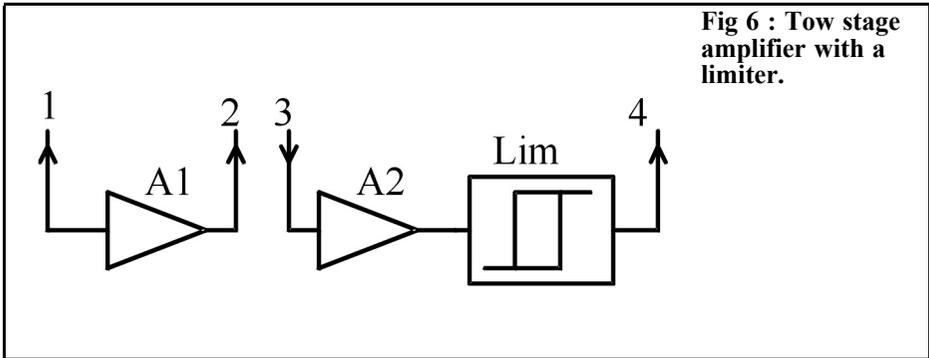
To put it in words, the amplification of the LNA must be considerably higher than the noise figure of the mixer. This is a problem with many standard receiver designs where the two figures are of the same order. If the insertion attenuation of one or more filters is then added in, the mixer stage, with its poor noise figure, frequently determines the actual signal-to-noise ratio of a receiver.

To remedy this, the LNA should be two stages, so that an effective amplification of at least 25dB is present before the mixer stage.

The local oscillator introduces an additional requirement. The mixer amplification is linearly dependent on the level of this oscillator. This should therefore have the highest permissible limit. In the standard commercial mixers, this is at least 10dBm. Integrated oscillators usually attain a level of only 0 to approximately 5dBm. In these cases, an amplifier between the oscillator and the mixer is very definitely recommended. Otherwise the mixer amplification could easily be less than one. Then, in certain circumstances, the intermediate-frequency amplifier would also have a considerable influence on the signal-to-noise ratio.

4.3. The intermediate-frequency amplifier

The considerations applying to the mixer stage are also valid for intermediate-frequency amplifiers. Their noise figures must therefore be lower than the product of the pre-amplification and the mixer



amplification. This condition can easily be fulfilled using a respectable receiver design. Many integrated intermediate-frequency amplifiers have an AGC function and/or a limiter, which an automatic amplifier regulator (AGC automatic gain control) and amplitude limitation.

Today the medium wave no longer plays much of a role on radio. Because its amplitude modulation is subject to so much interference, those using VHF have converted to frequency modulation. But analogue television also still has amplitude modulation. More modern systems rely entirely on frequency or phase modulation.

But there are also quite a few standards for digital modulation, e.g. QPSK, which alter both the frequency and the amplitude. These variations should be taken into account when selecting the intermediate-frequency amplifier. A limiter is incompatible with any type of amplitude modulation. The time constant for the AGC must be set in such a way that the regulation system adjusts fast enough to a different signal level, without modifying the modulation to a significant extent. All types of modulation with constant amplitude, for example FM, MSK and GMSK, are free from all these problems. So at low signal-to-noise ratios these still lead to acceptable results. The limiter should ensure that the amplitude is restricted to a maximum value at all levels. Amplifiers of this type will now be considered in greater detail.

Many integrated intermediate-frequency amplifiers are manufactured as two-stage units - see Fig. 6. A band filter can be wired up between connections 2 and 3. The total $A1 \cdot A2$ amplification attains about 90dB. The output voltage at terminal 4 is typically 200mV_{SS}. For the calculations below, the following data have been selected:

- Pre-amplification with mixer amplification 30dB
- Intermediate-frequency amplification 90dB
- Intermediate-frequency band width 100kHz
- Effective noise figure 6dB

The effective noise figure takes into account the influence of the mixer and the intermediate-frequency amplifier on the total noise. Thus we obtain a level of +2dBm before the mixer. This is practically sufficient to control the limiter, on the basis of the noise output alone. The receiver apparently delivers a signal, although no antenna is connected. Since an effective signal even if it should be a thousand times more powerful than the noise - also appears to lead to the same voltage curve at the limiter output, the signal-to-noise ratio can no longer be detected at the limiter output. Meaningful measurements can not be carried out again until demodulation has taken place.

Based on these considerations, we can



also find the limits at which we can still obtain a meaningful value for the intermediate-frequency amplification. In the ideal case, the amplification should be so slight that the thermal noise can not drive the limiter all the way to its limits. In addition, the signal amplitude should be so high that the limiter always restricts the level. To make sure that if the signal is too weak, or is not present, the noise is not received instead of the signal, many intermediate-frequency amplifiers have an output that gives a voltage level proportional to the level of the input signal. A threshold value can be set for an input power that is still acceptable. Below this threshold, the receiver will not operate. This function is carried out on all mobile telephones.

If the signal-to-noise ratio at the demodulator output is analysed, it is considered in a different way. By “signal” in the pre-amplifier or the intermediate-frequency amplifier, we understand the power of the entire spectrum, i.e. the carrier and the sidebands. In the demodulator, only the energy of one or two sidebands is evaluated. The signal power is thus considerably reduced. In addition, conversion losses are also involved. The degree of modulation likewise has an influence. On the whole it can be shown that a demodulator considerably reduces the signal-to-noise ratio.

5. Digital data transmission by means of radio systems

In the above analysis, the signal-to-noise ratio at the output is decisive for the efficiency of the receiver. We have already referred earlier to how high it should be for analogue radio and television applications.

Modern systems predominantly rely on digital transmission, even for speech and

images. The “poor sound quality” on radio when the signal-to-noise ratio is low is reflected in an increased bit error rate (BER) in digital systems. On this subject, here is a somewhat simplified example from MSK: a signal-to-noise ratio of 0dB is assumed, with the signal output being exactly the same as the noise output. On average, there will then be an error in every tenth bit. Even if the signal power is only doubled ($S/N = 3\text{dB}$) the ratio improves to give an error in every fifty-first bit.

A further advantage of digital systems lies in the fact that bit errors can be corrected to a certain degree. There are error-correcting codes e.g. the Hamming code - which, under certain conditions, can eliminate all bit errors[4].

6. Decoding

With analogue systems, we receive the output signal after demodulation. True, it then has to be re-amplified in order, for example, to be easily audible through a loudspeaker, but this amplification no longer reduces the signal-to-noise ratio.

The situation is different with digital systems. Here demodulation is followed by decoding, which turns the output signal back into a genuine digital signal. The lower the signal-to-noise ratio is at the decoder input, the more bit errors it will cause. However, decoders with a high level of efficiency can be assembled by means of ingenious techniques. These are in a position to decode digital signals, largely free from errors, even with a low signal-to-noise ratio. We might mention the matched filter technique and cross-correlation. There follows a brief description of these two methods [5]:

6.1. Matched filter technique

To restrict the bandwidth, digital signals are filtered before transmission along a



radio path. Instead of square wave pulses, this gives rise to other forms with gentle flanks. These are digitised in the receiver by means of scanning. In the matched filter method, the form of the transmission pulse and the filter function in the receiver are combined in such a way that the correct value can be detected, with a high degree of probability, with a single scan of each bit (the choice is only zero or one!). Relatively little technical effort is required.

6.2. Cross-correlation

With cross-correlation, the receiver forms the data itself, as if it were the transmitter. But because it does not know and can not know in advance whether a zero or a one will be transmitted at any given moment, it forms both values simultaneously and keeps them ready in two register banks. When the transmission signals arrive hit, the receiver checks, using a special algorithm, cross-correlation, whether the bit transmitted is more like a one or more like a zero. This then forms the basis for the digital decision. The method makes it possible to extract signals, even when buried deep under the noise. Even with an S/N considerably below zero dB, good results can still be obtained. This is an enormous step forward, when compared with all analogue systems.

The disadvantage lies in the high-level circuitry required. Cross-correlation can be implemented in real time only with very efficient field-programmable gate arrays (FPGA).

7.

Data compression

A basic disadvantage of digital systems is their requirement for increased bandwidth. It has already been pointed out that human speech has a bandwidth of approximately 10kHz. But digitised

speech signals display a much higher bandwidth, which for average quality lies at about 100kBit/sec. The increased bandwidth puts an unnecessary strain on the expensive frequency bands in radio systems, and thus increases the noise power of amplifiers. This leads to a lot of expenditure to reduce bandwidths. With mobile telephones, the effective bandwidth is restricted to about 23kBit/sec. The key phrase here is data compression. Thus redundant (i.e. superfluous) information is removed from the signal and only the absolutely essential data are transmitted. But on the receiver side the signal must be converted back into its original condition (as far as possible). Data compression does not function under strict real-time conditions. The compressed signal is created only some time after the original signal. For transmissions between a transmitter and a receiver without talkback, this delay causes no problems. It becomes more difficult especially with a high degree of compression if two people are carrying on a live conversation. Further progress should be expected in the future in the field of data compression [6].

8.

Outlook

The world of wireless telecommunications has experienced an enormous upswing, thanks to modern techniques in combination with digital data transmission. Even radio amateurs can profit from this. Many circuits can be obtained at favourable prices, and many can even be acquired free of charge as samples. It is also worthwhile for the committed radio amateur to gain a wider understanding of the world of modern radio technology through his/her own experience.



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Carl G. Lodström, KQ6AX & SM6MOM

A Simple Noise Figure Meter

1.

Introduction

Since I have built a detector using the new and interesting Logarithmic Detector ICs from Analog Devices (AD8307 and AD8313 [1]) I realised that the same detector could be used directly in a noise figure meter. This is because subtracting the logarithms of signals is the same as dividing them by each other.

I believe that there are at least two shortcomings of this instrument, as described, but it works so well that I do not think it matters too much.

The detector does not treat the noise signals as True RMS, but as Averages, introducing a small error.

The detector does not have a zero output for a zero input signal. A small offset error is introduced, but it can be trimmed.

2.

Construction

The Logarithmic Detector IC (AD8307) and a dual OP amplifier (LT1630) for offset and scaling of the signal and a 5V regulator are assembled as per the sche-

matic diagram in Fig 1. These detectors run on 5V (LM78L05) but the rest of the circuit is a ± 12 or ± 15 V design. It is probably not necessary to point out to the VHF Communications readers that the two 1 nF capacitors and the 10Ω resistors are to be close to their pins on the package of the AD8307 (or AD8313). The LT1630 is a very good dual OP-amp, bipolar, “rail to rail” on both inputs and output. I prefer the LM78L05 regulator because it, as well as the OP, can handle 32 volt, so make the capacitor on the input a 35V one if you want to use this detector as the “General Purpose”, free standing and very useful instrument it is. It is my design, and it is trademarked, registered with the U.S. Patent Office, as LogProbe so should you feel like entering a commercial arrangement using this, please let me know! I am agreeable!

With the values shown (24k and 56k) the output will be very close to 10 dB/Volt. The 10k/1k1 sets the output DC level, and the “bottom” of this ladder may be a 2 - 5k trimpot. The 47Ω on the output is to avoid capacitive loading of the OP. It is a pretty fast one, the LT1632 is even faster, and it can handle some capacitance, but this way there will be no problems. Besides, it may survive long short circuits of the output.

A noise source is constructed. This is done best using a noise diode from the

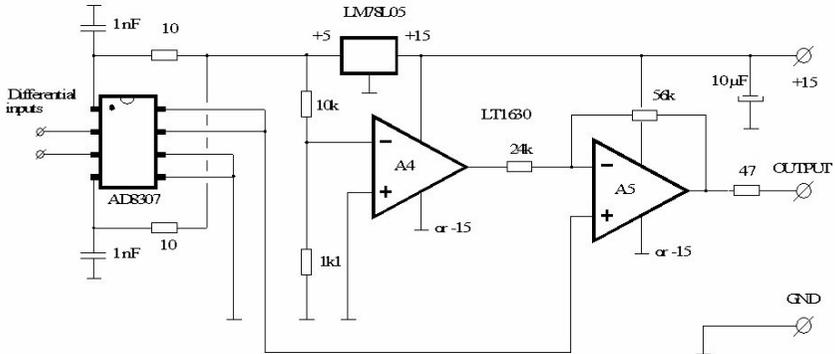


Fig 1 : Schematic diagram of logarithmic detector.

people who know how to make them [2] but I seem to have had good luck with a UHF transistor. Using the B-E junction in reverse forms a Zener diode at about 6v. Just as it turns on; it has a very nice noise! Measure up “your” diode and be aware of that the noise level can change dramatically with just a few tenths of a volt supply variation! Maybe a constant current (in the 10µA region) is the way to drive them, I have not ventured into this area. I tune the trimpot for max smoke. Although regular Zener diodes (above some 5-6V) seem to give plenty of noise of varying intensities, but they fall flat above HF bands, some 30MHz, probably due to high internal capacitances. After all, they are designed to keep a voltage stable and to handle as much current as possible.

The schematic diagram of the noise source is shown in Fig 2. It can, or should, be enhanced regarding A1 output rise and fall times as well as limiting the voltage swing on the noise diode to a minimum. One may also contemplate stabilizing of the “On-voltage” since it can be pretty sensitive. The switches on the meters switch in different ranges of

sensitivity. The signals from a radio receiver may be much stronger than from a 10dB LNA.

Using the “Alt. In” (MAX4527 pin 7 instead of pin 6) will switch which capacitor stores the S and which stores the (S+N), reversing the M2 reading. If M2 is offset (for example by means of making the Reference input of the A3 negative) to zero, the NF reading would go “the right way”, showing 0 for 0 dB and more for larger NF values. The non-inverting input of A2 would then have to be connected to the other of the two electrolytic capacitors.

The output of the noise source is passed through a 20dB attenuator [3], of the surface mount “wind mill” type, this ensures a very good S11 on the source. Theoretically < 40dB return loss, but in reality depending on how accurate to 50Ω the attenuator is, probably better than a RL of 36 dB.

In the design of the chopped noise source, it is a good idea (that I have not yet achieved) to see to that the DC on the output does not vary more than necessary. In other words: do not put the

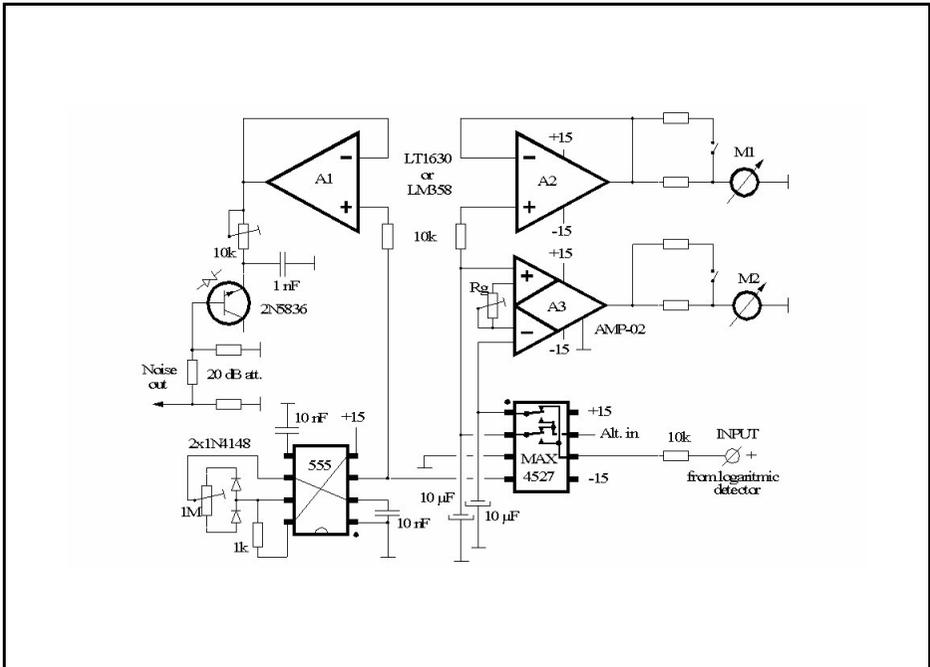


Fig 2 : Schematic diagram of the noise source.

whole 0-6v on/off waveform on the attenuator! The noise may be totally off at 5.7V and on at 5.9V! Measure this and do not apply (or turn off) more DC than you need for to turn the source on and off. It may also be a good idea to use the A1 amplifier for this, even if the pin 3 on the 555 has plenty of signal. The extra amplifier is there anyway. A voltage divider with a permanent voltage, to which is added or subtracted a little by the square wave, should be easy to make using the A1. One can add R and C to slew the output ramps a little, whereas the 555 has a very fast transient turn-on and turn-off. When you measure a receiver you do not want any spikes like these on the antenna input. It may well be good to make a 10µs rise and fall time on the DC that is pulsing the noise source. That is a very small part of a period of 16Hz anyway. In combination with the DC variation being kept to a minimum, this works very well on receivers. On an

LNA for UHF it would not matter either way.

A simple oscillator that turns the noise source on and off (slowly, <100Hz) is constructed. Although I am not a big fan of the “555”, a circuit that may create as many problems as it solves, here is a good place for it. The newer “CMOS-555” ought to work as well. As will any other square wave generator. The version I built here has the duty cycle (D.C.) adjustable widely around 50%, but this is probably not necessary. Anything from 20 - 80% is likely to work.

3.

Function

A logarithmic detector converts AC (or DC) input signals to a DC signal that is



proportional to the logarithm of the input signal level. For AD8307 this is true from about $50\mu\text{V}$ to about 2.2V (differential input) over a frequency range to nearly 1GHz. A 92dB range with very good linearity and temperature compensation, quite good! The AD8313 has a lesser range, some 70dB, but goes to 2.5GHz.

With a dual OP amplifier the DC detector output, corresponding to the logarithm of the signal amplitude, can be shifted and scaled as one desires.

The noise source is turned on and off at a relatively slow speed (it happened to be 16Hz in my version). It is fed to the receiver or Low Noise Amplifier under test. The Logarithmic Detector is placed on the output of the amplifier (or receiver). The signal from the detector is thus a DC voltage that alternates in strength, corresponding to $\log(S)$ and $\log(S+N)$.

The $S/(S+N)$ is a measure of the sensitivity of the device. Since the signals are present in their logarithmic form, one can perform the division by subtracting them!

Using an analogue switch, synchronous to the 16Hz noise switching, the signals are stored in either of two large capacitors. A resistor in series with the common line makes sure the display is nice and smooth, letting the capacitor voltage average out over a few cycles of switching! The signals are subtracted in an Instrumentation Amplifier with a gain of 1. The signal from this is the "Noise Figure Signal" and it is rather "1 NF" so to speak. A reading on this M2 meter will be identical to the other meter (the M1 Gain meter) if a device is perfect, adding no noise of its own. Connecting the detector directly to the noise source for example. The lower the display of the M2 NF meter (compared to the value of the M1 Gain meter), the larger (worse) the NF of the device.

The output to the M1, the Gain reading, is simply the voltage of the applied noise,

as amplified by the LNA (or receiver) and as detected by the logarithmic detector.

4.

Applications

The pulsating Noise Source is connected to the input of a LNA (or receiver or any other device under test) and the Detector to its output. The LNA is (usually) tweaked for best NF and as good a gain as possible. To get an idea of "where it is", connect the detector directly to the noise source and note the readings. If they are not equal, it is most likely because the detector has a little output even for no input signal. Adjust the 1k1 in the detector schematic diagram! Connect again to the LNA!

To see how much difference a change NF reading makes, connect an attenuator of 1 or 2dB and see "how much more" M2 drops when the NF is worsened by the amount of the attenuator. This shows the slope of the meter.

For a sanity test, connect an attenuator to the output to see the gain worsened by the same amount as the attenuator value. If the attenuator is <10dB (or so) the NF reading should barely be affected. If the attenuator is connected to the input, both M1 and M2 should drop by the same amount. This is a serious test of the instrument!

When connected to a radio one can actually "listen" to the audio with the logarithmic detector! Be careful if connected to the loudspeaker output not to turn up the volume too much, and be careful about pops and clicks that may kill the detector. Personally I use a small 1:1 audio transformer, mounted on a $\frac{1}{4}$ inch audio plug, with a centre tap on the secondary. I connect this tap to ground with an electrolytic capacitor because the inputs of the Logarithmic Detector like to

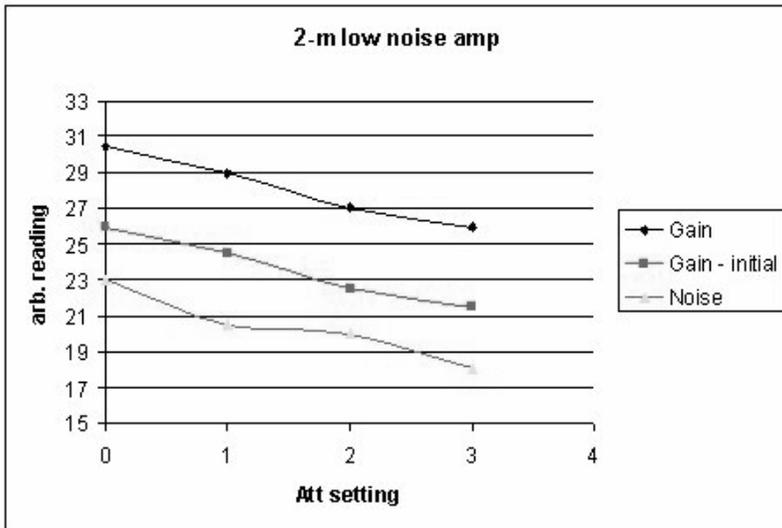


Fig 3 : Example reading testing a 2m LNA using a BF998 dual gate MOSFET.

float around at a few volts. The IC works from 2 to 5V common mode, but in this case there is no reason to steer them to any particular DC voltage. A couple of crossed diodes (1N4004 or similar) across the transformer primary may be a very good idea! They can pay for themselves sooner than you think! A tape recorder output is safer to use, and furthermore independent of the volume control setting.

The receiver has almost certainly AGC, which should be disabled for true measurements, but it may very well work anyway, especially on low input levels. One can also speed up the chopping rate a bit (make it variable!) to 50Hz or so, where the AGC is not fast enough to make much difference. It is most interesting to take a look at the Gain and NF of a "General Coverage Receiver" to see how both vary over any one band and from band to band! That wonderful receiver is suddenly not so wonderful anymore, even though it still works fine!

5.

FM receivers

FM receivers are limited in the IF so no matter what strength of the input noise signal, the audio (noise) will be the same! So, the Logarithmic Detector has to be used to pick up either the RF signal after the front LNA or the IF after the mixer, before the signal reaches the limiter. Put a small coil between the inputs of the IC and "sniff" around for readings on the Gain meter! The detector is very sensitive and you can most likely find enough signals. Then: just trim the LNA for the best Gain/NF combination as usual!

6.

Example readings

Fig 3 shows measurement on a 2m LNA.

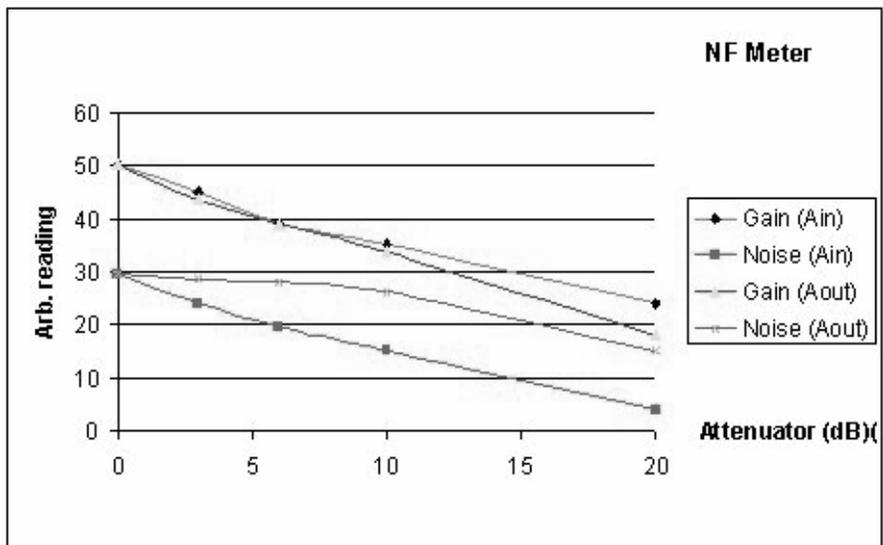


Fig 4 : Extrapolation of real NF.

The readings on the Y axis are arbitrary readings of the meters (identical) M1 and M2 numbers but the X axis shows the actual settings in dB of the attenuator. As expected, when the gain reading is independent if the attenuator is placed on the input or the output (the upper two graphs). For the “1-NF” reading, the lower two graphs show the reading dropping at once with the attenuator on the input and at about 10dB when placed on the output. This is the ultimate test of an instrument like this, and a proof of its function. It is likely that the LNA has a NF of a dB or two. If so, the noise output from the source is ~12dB ENR (Excessive Noise Ratio). This means 12dB more than the thermal noise from a 50Ω resistor.

In the graph shown in Fig 4, I have tried to extrapolate what the real NF may be by adding in an attenuator with 1dB steps in the input.

7.

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Geoff Pike, GI0GDP

S12

1.

Introduction

Of the four main s-parameters, the other three being S11, S21, and S22, S12 is often ignored because of its modifying effect on S11 and S22.

Although it will result in a quick match for an amplifier the result is often wrong and the amplifier only partly works this being due to the adjustment available from trimmer capacitors. It can be shown using Puff and with some Excel calculations how easy it can be to produce an amplifier with a better match (and often a simultaneous conjugate match if $K > 1$) first time round without resorting to “tweaking and lets see” tactics.

As the BFG135 exhibits quite an extreme change in S22 and S11 when a conjugate match is attempted it will be used as an example in making an amplifier for 1296MHz.

This will be done in three different ways:

- Using the Smith chart or Puff the conjugate of S11 and S22 will be determined and using 50 lines then put together to produce an amplifier that totally ignores S12.
- The same can be done using $\lambda/4$ lines and capacitance at the line ends or

start as needed using a basic program that ignores S12.

- Use the new S11 and S22 values under the influence of S12 to produce the amplifiers. The effect of this is quite dramatic as with this device it will change the imaginary part of the output impedance from $+j$ to $-j$ with an associated change in the real part which to some extent makes matching easier (or more convenient) to achieve.

2.

Matching methods

The s-parameters for the BFG135 were obtained from Philips semiconductors web site these were taken at 8v and 100mA.

The values for 1300MHz being interpolated from the 1.2 and 1.4GHz values:

f	1.3GHz
S11	.569 \angle 147.8°
S21	2.276 \angle 50.9°
S12	.190 \angle 58.1°
S22	.281 \angle 165.5°

These reflection coefficients can be plotted on a Smith chart or computed to yield 0.3+j0.26 or 14.8+j13.1 Ω input imped-

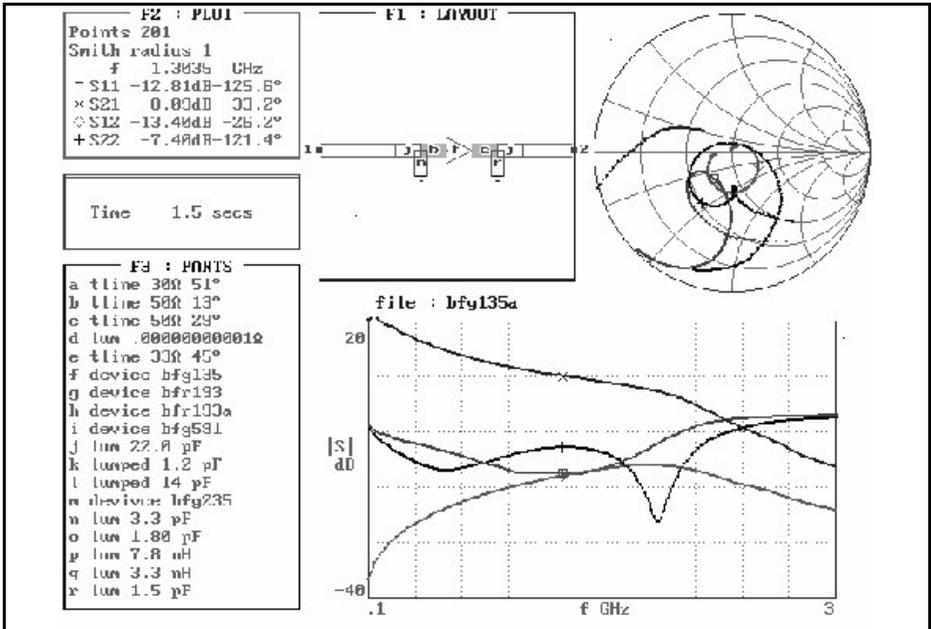


Fig 1 : Matching method one.

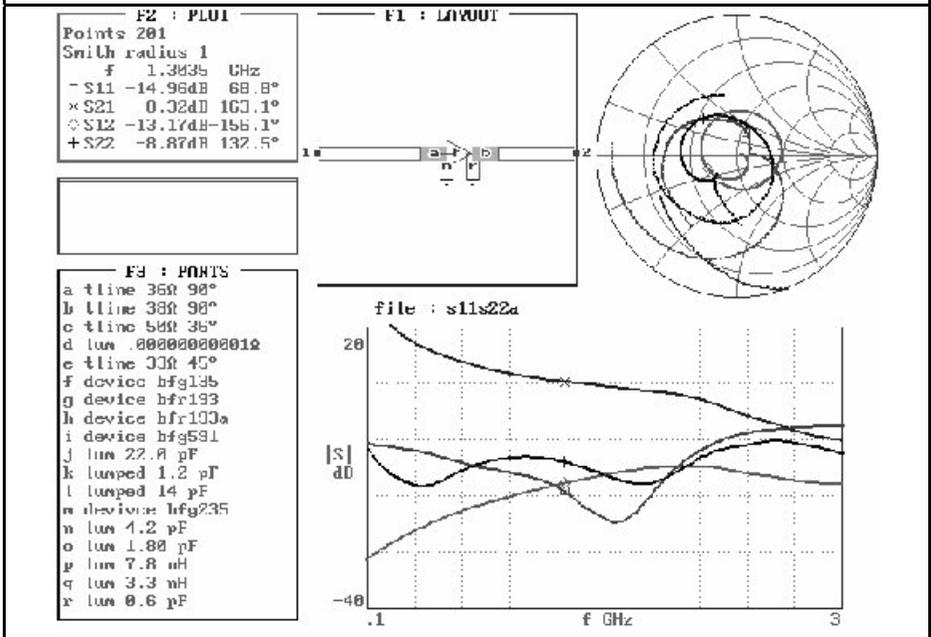


Fig 2 : Matching method two.



ance and $0.57+j0.08$ or $28.35+j4.2\Omega$ “output impedance” in a 50Ω system.

By inspection it can be seen that the device has a $+j$ imaginary input part and the same for the output S22, also S12 under these conditions is not negligible and the result of ignoring it will be demonstrated.

2.1 Matching method one

The first matching method will use Puff to simply determine the length of 50Ω line that will intersect with the normalized admittance contour = 1. Put another way, the line length needed to raise the real part of the impedance to 50Ω but will still leave the line input impedance reactive. Puff can then be used to determine the stub length to produce a reactance of opposite sign at this point hence leaving it purely resistive.

As a personal choice this will be an o/c line and the equivalent capacitance will be provided from a trimmer to allow some adjustment.

This is all standard stuff but Z_{in} and Z_{out} conjugates will treat the 2 port device as having S12 being non existent.

Fig 1 shows that a reasonable match does exist however it is not what you would expect from conjugate matching the input and output, clearly the effect of S12 is apparent.

This match was done using Puff to match the stand alone input and output impedances, and while you could play about and maybe eventually simulate a better match there is a better method.

2.2 Matching method two

The next matching scheme uses basically the same as above but uses a basic program which ignores S12 and uses $1/4\lambda$ lines of the appropriate impedance to match the devices stand alone input and output ports.

Fig 2 shows that this achieves much the same as the previous method but uses

$1/4\lambda$ lines of variable impedance to do the matching and again discrete variable capacitors on the lines to achieve tuning. Whilst the match is better it is still not very good, this again being the result of S12 affecting both S11 and in this case very much affecting S22.

2.3 Matching method three

The third matching scheme takes into account the pronounced effect of S12 in this example on S22. To calculate new values for S11 and S22 under conditions of simultaneous conjugate match the equations in the ARRL book “UHF/Microwave experimenters manual” were used, however they appear in other literature.

The calculations are tedious and it is better to get Excel to do it for you or to write a program for it because it is easy to make mistakes with complex numbers in both magnitude and angle format and in polar coordinate form.

Of course Puff also must use them to produce solutions when reading from a S2P file.

Fig 3 uses the new values for S11 and S22 under simultaneous conjugate match and these are:

$$S11 \quad 0.643 \angle 154.9^\circ \\ \text{or } 11.36 + j10.4\Omega$$

$$S22 \quad 0.266 \angle -88.3^\circ \\ \text{or } 43.95 - j25.0\Omega$$

A considerable improvement in return loss for S11 at -30db and S22 at -36db is seen.

Also although not often thought of in power devices $K=1.07$ and thus the amplifier is unconditionally stable.

Fig 4 shows puff looking into S22 when S11 matched. It is included to help confirm that when the input is matched and terminated in 50Ω , when looking into its output port, S22, the impedance seen is close to that calculated previously. Puff returns S22 as $42.6 - j23.1\Omega$

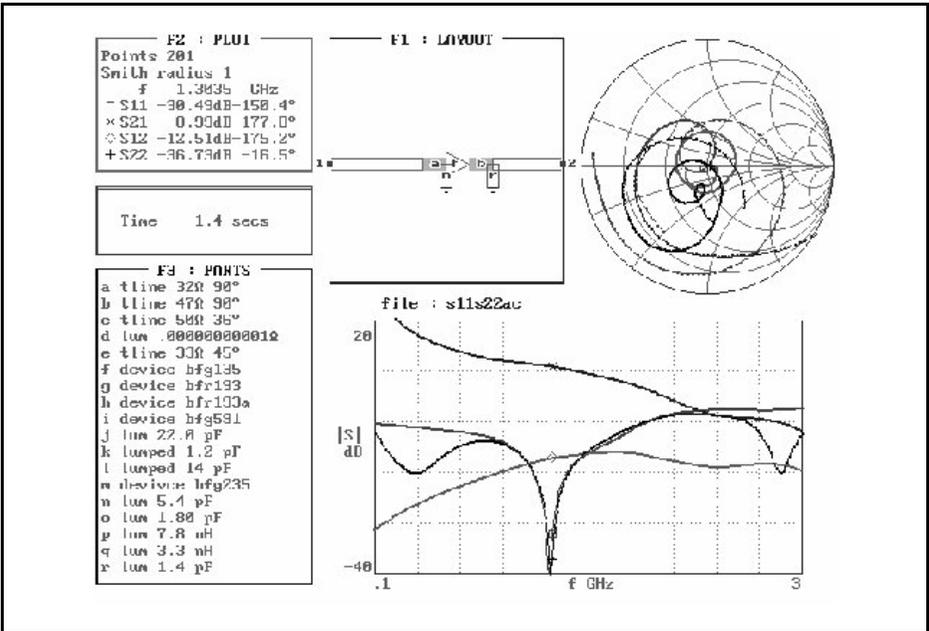


Fig 3 : Matching method three.

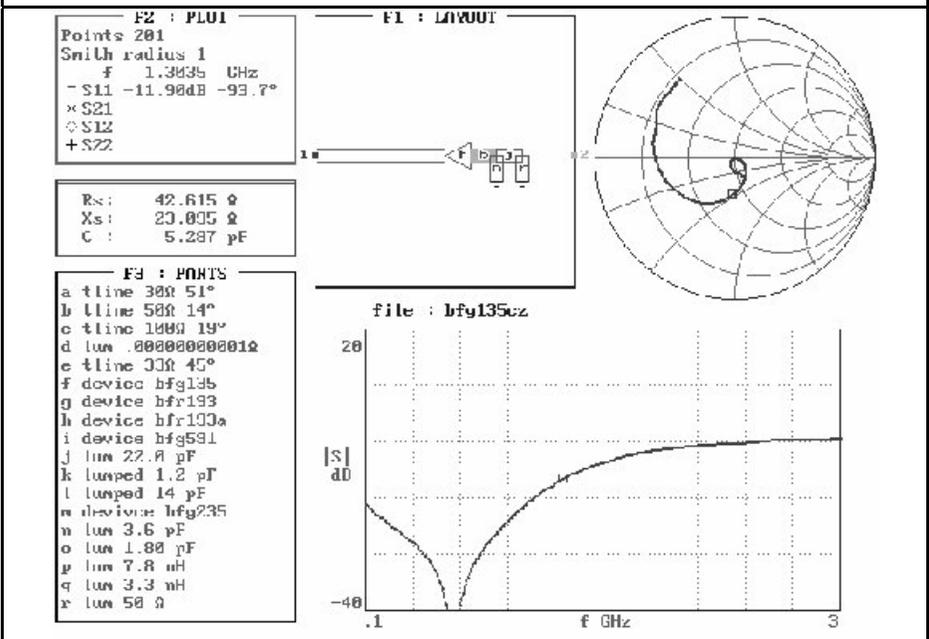


Fig 4 : Looking into S22 when S11 matched.

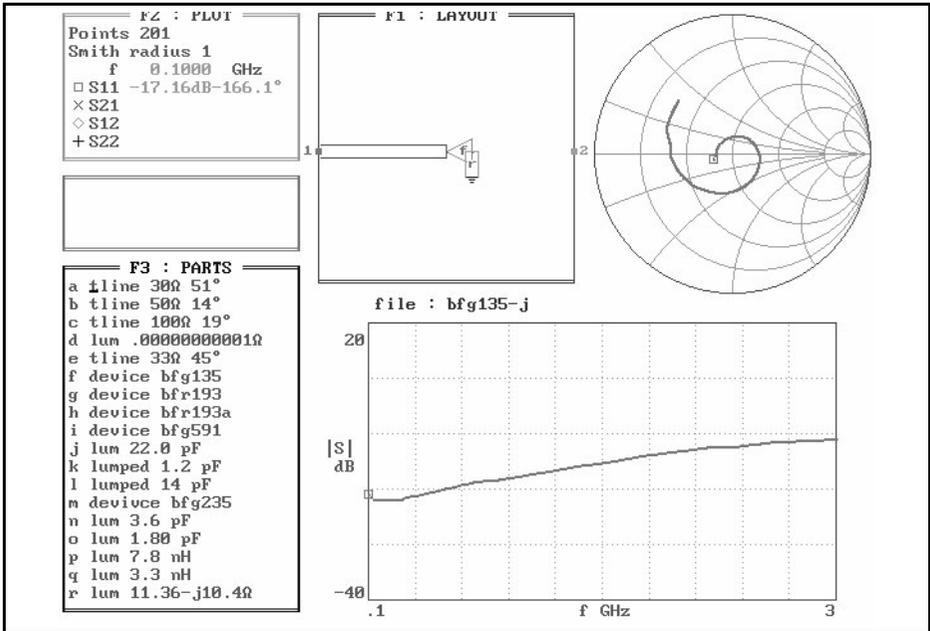


Fig 5 : Looking into output port with new input termination.

which compares favorably with the value of 43.9-j25.0Ω used in Fig 3.

value into account for matching networks.

The same result is achieved by terminating the input in its new complex conjugate i.e. 11.36-j10.4Ω using Puff to look into the output port, it then returns the new S22 value as above, see Fig 5.

3.

Summary

In summary it doesnt matter the way in which you chose to match be it either 1/4λ lines of appropriate impedance, or 50Ω lines of variable length or a mixture of both.

However what is important is the modifying effect of S12 (which is seen in practice by the interaction of the input and output tuning of most solid state amplifiers) and the necessity to take its



Eberhard L. Smolka, DB7UP

In Memorandum of Alois Pendl, OE6AP

A pioneer of VHF/UHF amateur radio



24GHz. He developed and constructed radio components, which he then subjected to comprehensive testing.

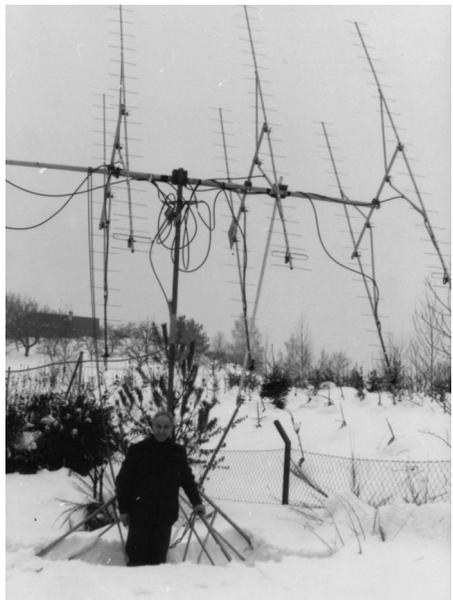
On 3rd May, 1958, Alois, OE6AP, with his fellow amateurs and long standing friends, Emmerich (Emo) Harmet, OE6TH, Rudolf Hirschmann, OE6RH and Heinrich Steinböck, OE6HS, embarked on a bold experiment concerning the first European meteor scatter (MS) reflections. They

He joined the OEVSU in 1938, when he was only 14 years old. He became an airborne radio operator and worked with Wernher von Braun in Peenemünde. He was never to lose his enthusiasm for modern wireless communication, and he gave his heart and soul to amateur radio. His particular interest was in wavelengths from 144MHz upwards, and he gradually progressed right up to





worked 1,623m up, on the top of the Feuerkogel (Salzkammergut), in dramatic weather conditions, with 100 Watts at 144MHz, they used two 17 element antennas, transmitting to SM6BTT in Sweden. The experiment was a success, some time later it was followed by the



first European SSB QSO using meteor scatter (LX1DB and OE6AP).

On 5th May, 1976, he was responsible for the first 2m EME link from Austria (JA6DR OE6AP), followed by a 70cm transmission on 21.11.1980 (OH3TA OE6AP).





He demonstrated tremendous commitment to his hobby of radio; between 1959 and 1963, he was VHF manager for the Austrian amateur radio association, ÖVSV.

He was the initiator, and one of the founders, of "VHF Reports" in 1960.

Surrounded as he was by the Austrian peaks, it came as no surprise that he frequently took part in competitions and range tests in the mountains e.g. on the Hochwechsel, the Zugspitze, the Feuerkogel and countless others.

In recognition of his outstanding achievements in the VHF/UHF/SHF fields, The ÖVSV awarded him a commemorative plaque in 1987.

Alois Pendl, OE6AP, died on 18th March 2001.

Many radio hams from his own area and from further afield will always treasure his memory as will the editorial team here at "VHF Reports".

Eberhard L. Smolka, DB7UP



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Wolfgang Schneider, DJ8ES

Direct mixer for a 1 - 65MHz short-wave synthesiser

A further interesting addition to the 1 - 65 MHz short-wave synthesiser from Bernd Kaa, DG4RBF, is the direct mixer described below. This combines with the second supplementary module, the AD8307 logarithmic detector, to create a spectrum analyser for measurements in the frequency range of 1 - 65MHz. (*The 1 - 65MHz synthesiser has not appeared in VHF Communications, only UKW Berichte. This design is still of interest, if there is a demand the synthesiser project can be translated and appear in a future VHF Communications - Andy*)

1. General

Since the logarithmic detector processes frequencies down to 0Hz with the IC AD8307 (DC), the short-wave synthesiser can be expanded into a low cost spectrum analyser by simple means. The key concept is the direct mixing process.

The HF spectrum to be measured or represented is mixed over the range from 0 to 250kHz, with the 1 to 65MHz short-wave synthesiser acting as the oscillator [1]. The logarithmic detector assembly [2] takes charge of displaying the reading on an oscilloscope; the dynamic range is approximately 70dB.

2. Description of circuit

The central component in this circuit (Fig. 2) is the type SBL-1 ring mixer. The deciding factor here is the frequency response of the HF and LO ports, beginning at less than 1MHz, and the intermediate frequency port going down to direct current (DC). The well known IE500 mixer cant do that!

To match the output from the wobble oscillators at 10mW, the ring mixer is adjusted through a 3dB attenuator at the LO input. The level here, as shown in the specification sheet, should be +7dBm (5mW). The HF input should be wired up to a 10dB attenuator. Thus the maximum input level for the direct mixer assembly can be 1mW (0 dBm). Higher signal strengths generate intermodulation and falsify the reading by causing additional spectrum lines to be displayed.

The intermediate frequency port is followed directly by a 9 pole low pass with a cut off frequency of 250kHz. This gives the band width required so that even rapid measurements on short-wave transmitters can be made. The amplifier stage with $V = 10$ dB guarantees matching to 50 to connect the logarithmic detector assembly with AD8307.

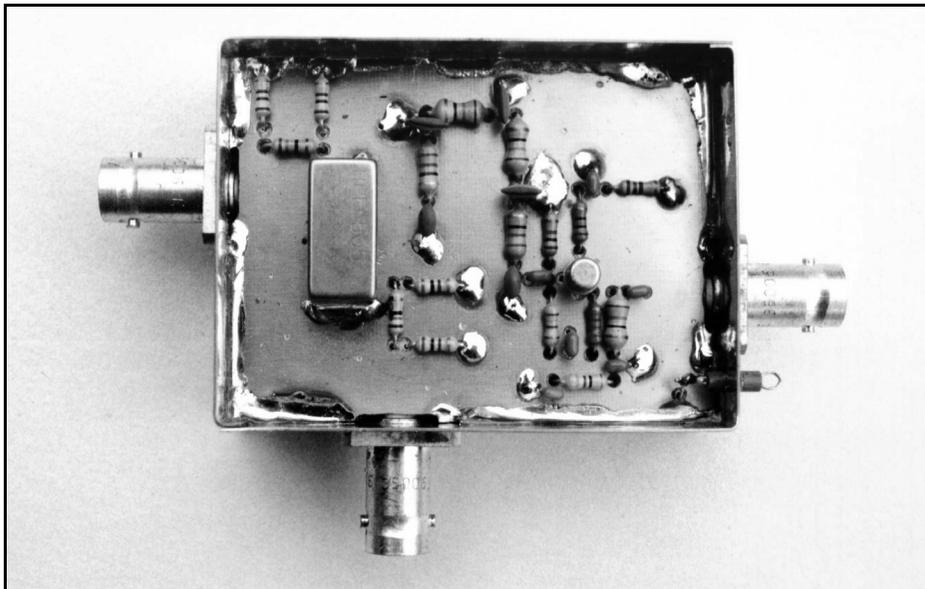


Fig. 1: Direct mixer for 1 - 65 MHz short-wave synthesiser

3.

Assembly instructions

The direct mixer for the 1 to 65MHz short-wave synthesiser is assembled on a double sided epoxy printed circuit board (Fig. 3) with the dimensions 54 mm x 72 mm. For screening purposes, the board is built into a tinfole housing with the dimensions 55.5mm x 74mm x 30mm.

BNC sockets are provided for all connections except the operating voltage of +12V DC, which is fed in by means of a feedthrough capacitor at 1nF.

Before the components are mounted on the printed circuit board, the connections to earth of the ring mixer have to be throughplated with tubular rivets. Then all components such as the ring mixer and the transistors can be soldered in as per the components plan (Fig. 4).

3.1. Components list

1 x	printed circuit board DJ8ES 048
1 x	transistor 2N5179
1 x	ring mixer SBL-1 (alternative SRA-1 or HPF-505)
2 x	choke 33 μ H, axial
3 x	choke 68 μ H, axial
1 x	1nF feedthrough capacitor, solderable
1 x	4mm soldering tag, turned downwards
3 x	BNC flange socket
4 x	full tubular rivets
1 x	tinfole housing 55.5mm x 74mm x 30mm

Resistors 1/4W, RM 10mm, wired

1 x	10 Ω
1 x	18 Ω
1 x	47 Ω
1 x	68 Ω
2 x	100 Ω
1 x	150 Ω
1 x	240 Ω
2 x	300 Ω

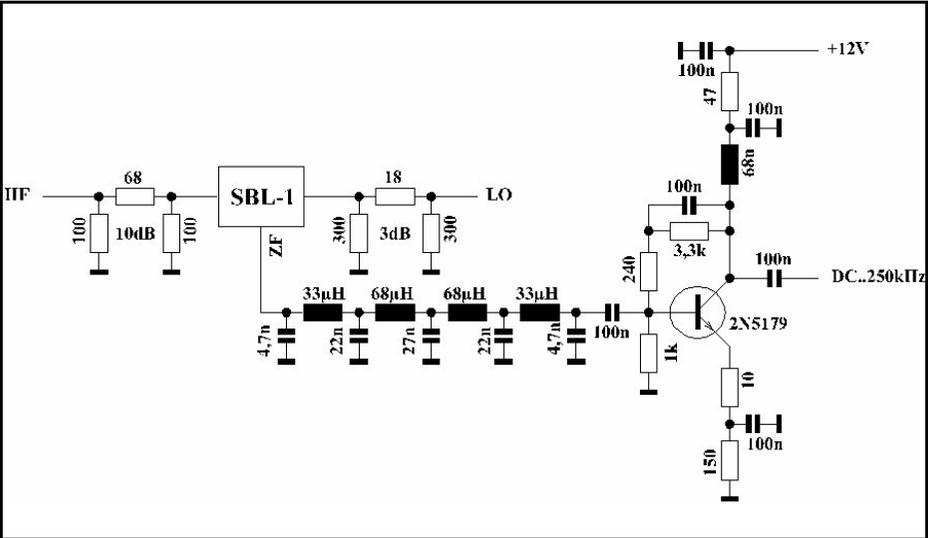


Fig. 2: Direct mixer circuit.

- 1 x 1kΩ
- 1 x 3.3kΩ

Ceramic capacitors, RM 2.5mm, wired

- 2 x 4.7nF
- 2 x 22nF
- 1 x 27nF
- 6 x 100nF

4. Putting into operation

Before putting it into operation for the first time, just check the assembly visually to make sure that the components are correct and that there are no short circuits. With an operating voltage of +12V, the circuit needs a current of approximately 20mA.

The input power at the HF port should

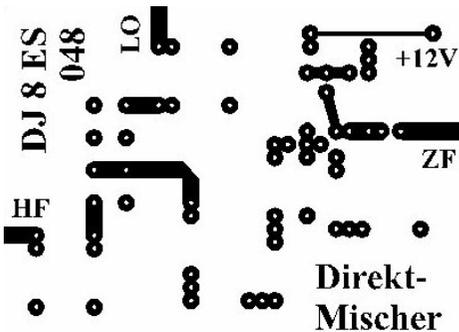


Fig. 3: DJ8 ES 048 printed circuit board layout.

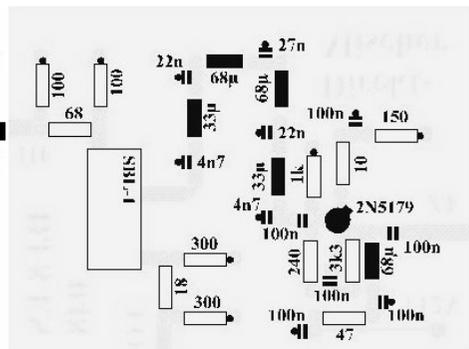


Fig. 4: Component layout of direct mixer for short-wave synthesiser.

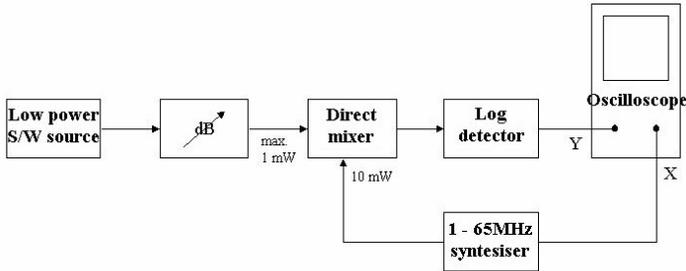


Fig. 5: Measurement rig.

not exceed 1mW (0 dBm). With a higher drive, the ring mixer goes into compression and generates a great many unpleasant intermodulation products. These falsify the reading! So it makes sense to insert a switchable attenuator in front of the direct mixer.

5. Working with the direct mixer assembly

If we combine the short-wave synthesiser for 1 65MHz with DDS at the LO input and the logarithmic detector at the intermediate frequency output of the direct mixers, we now have a spectrum analyser for the range 1 - 65MHz.

For measurements, you need only enter the frequency range to be observed (from/to MHz) on the short-wave synthesiser. No intermediate frequency mismatch is taken into account!

6. Literature

- [1] Bernd Kaa, DG 4 RBF: 1 - 65MHz short-wave synthesiser with DDS, VHF Reports 3/99
- [2] Wolfgang Schneider, DJ 8 ES: AD8307 logarithmic detector for up to 500 MHz, VHF Communications 2/2000 pp 119 - 124



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Andy Barter, G8ATD

International Microwave Handbook

In August 2001 The RSGB approached me about updating the three volume Microwave Handbook, to produce a new single volume book. This was to be in association with The ARRL and would become The International Microwave Handbook.

The existing three volume set handbook was written over a period of ten years by members of the Microwave Committee in the UK and contains an invaluable source of microwave theory, I am sure that many readers of VHF Communications have a set in their collection. Many of the designs had become dated, so my task was to find up to date designs for all of the microwave bands.

The first three months of the project were spent researching for designs from all around the world. Obviously there are articles from The RSGB and The ARRL but I also wanted as wide a range of material as possible. This has resulted in articles from thirty four sources, so there is plenty of interest for everyone.

Once all of the articles were found the fun really began for me. Gathering all of the text and diagrams, with some requiring translation was an enormous task, followed by big DTP job, proof reading and corrections. All of that kept me fully occupied for the next three months but end result is a 474 page handbook packed full of theory, information and

designs for the microwave bands.

The first eight chapters cover, operating techniques, system analysis and propagation, antennas, transmission lines, semiconductors and valves, construction techniques, common equipment and test equipment. Much of this information is taken from books one and two of the existing handbook with updated information on such things as MMICs, semiconductors and decoupling techniques.

The remaining chapters cover each of the amateur bands from 23cm to 24GHz and beyond. The band chapters have up to date designs for pre-amplifiers, transverters, transceivers, power amplifiers, antennas and filters. These designs range from true amateur construction projects to commercially available kits.

The handbook is available from The RSGB or The ARRL, for details see that advertisement on the inside cover page of this magazine.

Andy Barter, G8ATD



Gunthard Kraus, DG8GB

Internet Treasure Trove

KACHINA

Here, among other things, we can find a set of wiring diagrams for the Transceiver 505DSP, which is no longer manufactured. (Obviously you can still obtain residual stocks of this transceiver at favourable prices.) You can also look at some application reports. Address:

<http://www.kachina-az.com>

Synergy

The well known microwave component manufacturer has made a really big effort once again and has completely revised its homepage. So now under a specific heading we find not only data sheets but also appropriate application notes, technical documentation and technical articles (from periodicals). Treasure Trove indeed!

Address:

<http://www.synergymwave.com>

NEC Archive

The website of Ray Anderson, WB6TPU, contains an NEC archive; here, among other things, you can download the Yagi design program YGO3.ZIP, this is also available from the VHF Communications web site.

Address:

<http://www.qsl.net/wb6tpu/swindex.html>

Analog Devices

A big name in the field of IC production, mainly active in analogue circuits, D/A converters, RF-ICs, etc. and also in digital signal processors. The technical documentation has been greatly expanded, and especially the basic tutorials (e. g. on digital signal processors). You can look around for hours now and find more and more new stuff to download.

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you can download some interesting application reports, especially on the subject of "Software Radios" .

Software and Links

Under the address

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you can download a free program for designing mains rectifier circuits. In addition, there are links relating, to hi-fi amplifiers and SPICE.

The Longwave Homepage

Amateur activities in the longwave area are a really new challenge for do-it-yourselfers and for anyone who enjoys fiddling about. And it was only a matter of time until like minded people met up and got organised on the Internet as well. This homepage is the first address anyone interested in longwave reception should visit, as here you can find the technical equipment for getting started, both theoretically and practically. And naturally there are lots of links to similar sites or interesting documents.

Address:

<http://www.lwca.org>

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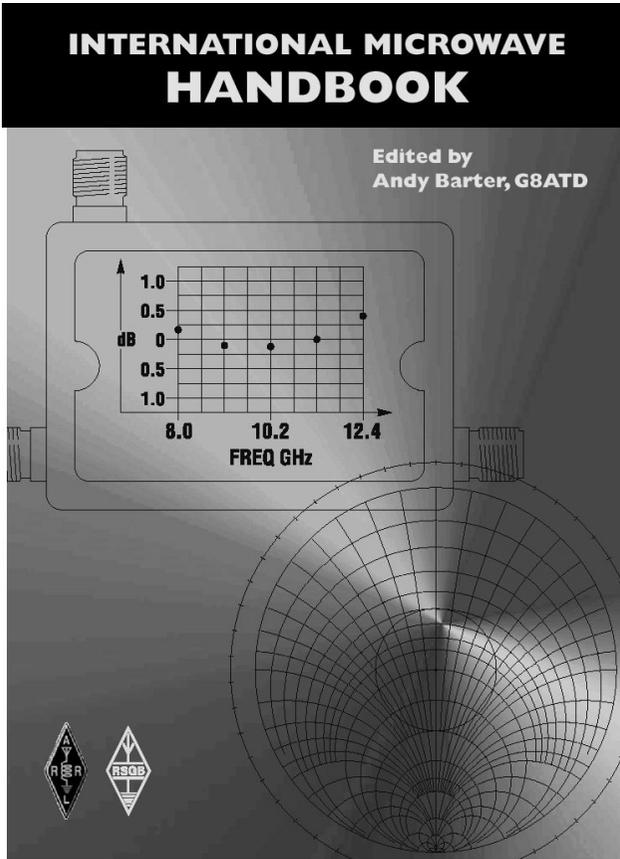
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The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area,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

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With the explosion in consumer electronics using microwave frequencies the opportunity to experiment has never been greater and this book is simply the best guide to the area of microwave radio.

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