

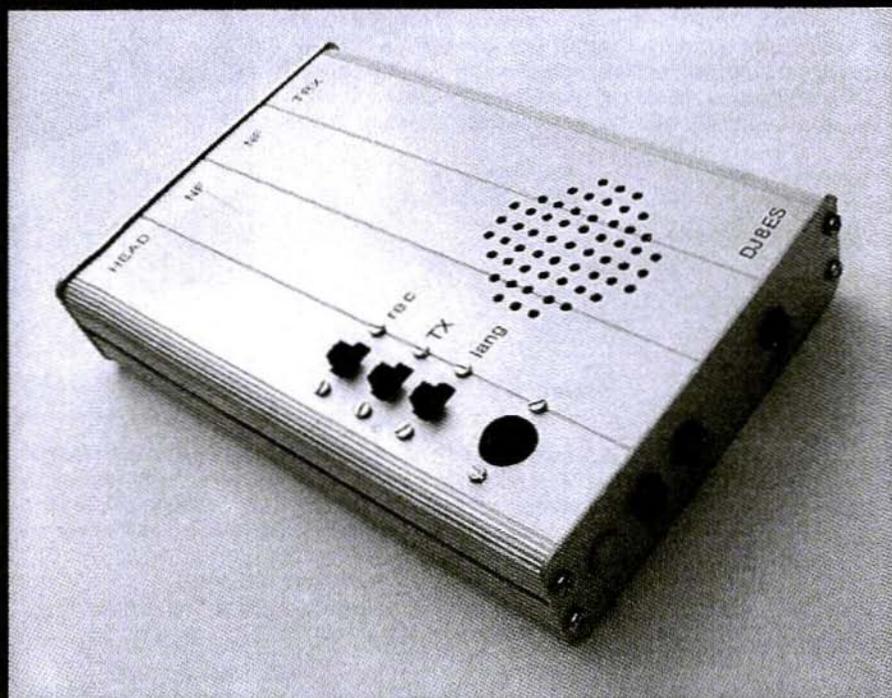


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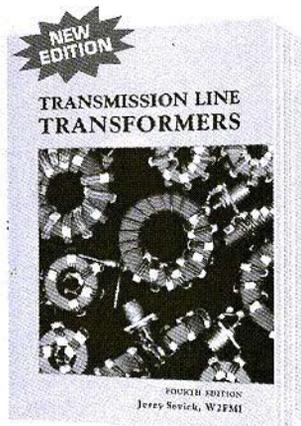
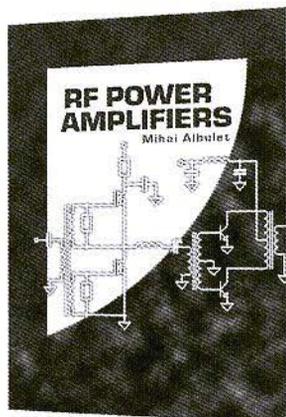
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More interesting articles again in this issue. There are two articles from authors who have not been in the magazine for some time, I still need more articles to broaden the range of subjects covered, please contact me if you have a suitable article.

73s - Andy



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Gunthard Kraus, DG8GB

Modern patch antenna design

Part II

4. Simulation of a complete antenna with SONNET LITE

4.1. Simulated circuits

RF circuit developers today have things easy but there are many options of software to use which makes life more difficult. The theoretical principles for accurate simulation of active and passive circuits up to over 100 GHz are now so precisely understood and have been converted into such efficient programs that there are almost no excuses left if any kind of design does not function as intended.

The design programs available can be split into a number of groups although the boundaries between them are a bit fluid. The following is a brief survey of available software for information purposes.

a) Linear simulators in a given frequency range, which behave as if there were no curved characteristic lines, and no saturation, etc.. We are already using a representative of this kind on a permanent basis: PUFF works with

S-parameters (assumed to be constant). Another product in this group would be Touchstone. These can examine any active or passive structures, the results are as precise as those of the S-parameters used for calculation also the circuit diagram used, including the undesirable dirty effects (such as circuit capacitances or self-inductances and self-capacitances of the components, etc.).

For specialists radio designer there is ARD, which I have also used occasionally. It works with the Y - Parameters and does not convert into S parameters until the closing stages.

b) A fast growing section of the market are programs which make non-linear simulations possible. Thus, for example, we can determine the high-level signal behaviour of an amplifier or the output signal of an oscillator or their noise sidebands. The most reliable and most precise simulations in this area are supplied by the harmonic balanced simulator.

c) There are also non-linear simulators, which are not based on the frequency dependencies in the circuit, but on the time dependancies of the voltages and



currents, thus they work in the time domain. The frequency responses are then calculated from the temporal behaviour using FFT (Fast Fourier Transformation). PSPICE is probably the best-known representative of this group of programs.

d) For examining passive structures (circuits, couplers, antennas....) there is yet another option.

We can try to make a direct calculation of the current and voltage distributions, together with the distribution of the associated electrical and magnetic fields and thus determine all the interesting values of the circuit (hence the name electro magnetic simulator or EM simulator). There are various starting points and methods, but they all run up against the same problem; a gigantic amount of computing power is required for this purpose (due to the many, many integrals and matrix calculations...). This is where we appreciate the advances in PC development, this means that such simulations can now be performed for the first time on the home computer.

Typically these programs divide the structure into a lot of small but equal dominos (the normal correct specialist term is cells), the change effects of these are first analysed and then assembled into a total behaviour.

4.2. SONNET

SONNET is an EM simulator, operating in accordance with the moment method and in addition starts from the known and calculable field distribution in a rectangular box. As soon as the dimensions of the box are made greater and greater, and the lid is finally replaced by free space, we can finally simulate the genuine free space behaviour. The program can also be used for

cavity or a wave guide simulations.

Apart from the high quality of their product, the SONNET company (Link: <http://www.sonnetusa.com>) have also scored a significant success with something else - a market strategy which is simply a work of genius by comparison with the competition. It involves dealing with people who either do not know the program or as private citizens can not afford the equivalent of the price of a posh car for a software package. You can actually download the free LITE demo version from the Internet. This version is not restricted by time but by its efficiency. Apart from a few limitations which will not cause us too much pain initially as we grapple with our puzzles e.g. limitation to two planes (double-sided coated circuit boards). There is one limitation which prevents the analysis of relatively large-scale DIY projects. This is the limitation of the usable operating memory to a maximum of 1 megabyte, but that is where the previously mentioned marketing strategy comes into play. Anyone who wants to use more memory only needs to fill in a free licensing form and post it to the company. Within 24 hours, a response is received, in the shape of an increase to a maximum usable working memory of 16 megabytes, although it is valid only for the computer already in use. This allows you to check out some pretty fancy things such as, for example, our patch antenna or couplers, etc. You should also join the User Forum straight away, and you will then receive the questions or problems of other users automatically by post. You can gain a lot of benefit from the in-depth advice and helpful suggestions from other users and naturally you can also put your own problems to the Sonnet community, or pass on your own experiences.

This free license can also be applied for on behalf of other computers at any time, and I myself have the program installed on my big PC and on the

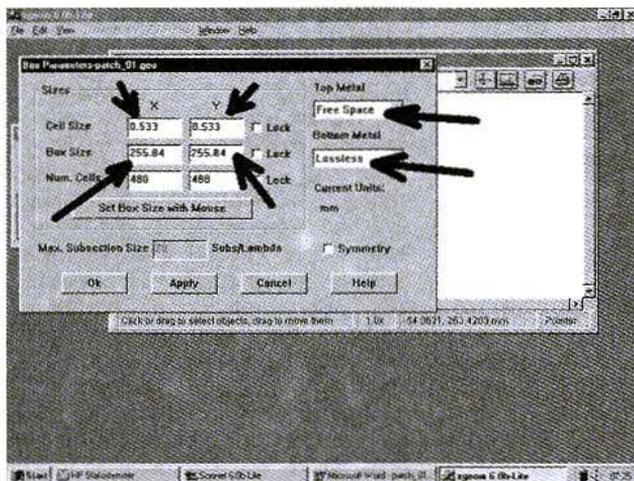


Fig 21 Using the Box Parameters menu during the setup for a SONNET simulation

Notebook. The fact that, in addition to the complete manual of over 500 pages, a really well made and helpful beginners tutorial is also supplied. This makes everything go really smoothly and confirms the good impression (Imagine the surprise, upon first leafing through the thick manual, of coming upon Chapter 15: Antennas and Radiation).

It is also worthwhile making regular visits to the homepage, where you will find, not just the usual FAQ pages, but ever-changing new, useful and helpful additional programs, e.g. for 3D representation of the structure entered or for printing out S-parameter files in the correct Touchstone format after simulation.

4.3. Simulation of patch antenna with SONNET LITE

Once the program has been correctly installed (but with a 16-megabyte license!), we can get going, and only a relatively small start menu is presented to us on the WINDOWS screen. Here we first activate EDIT CIRCUIT, so that we can get to the menu item New Circuit. We have now landed in the

geometry input program xgeom 6.0b-Lite and we see both the sign field and the buttons for the required drawing tools before us.

However, to start with the correct initial conditions must be entered. For this purpose, there follows a summary of the most important rules for a successful antenna simulation (you can find them in Chapters 3 and 15 of the online manual).

a) The length and width of our individual cells should not exceed $1/20$ of the wavelength, otherwise the precision of the simulation will suffer. A further reduction is certainly permissible, but not always expedient, since it means the memory space and the computing time will well and truly increase.

SONNET models a box in a known manner (nothing more than a rectangular tin box, such as is familiar to everyone as a biscuit tin). It is turned into a wave guide for antenna simulation, which is open at the top and terminated with free space, this makes radiation possible.

But now there are specific rules of the game which apply to this wave guide:

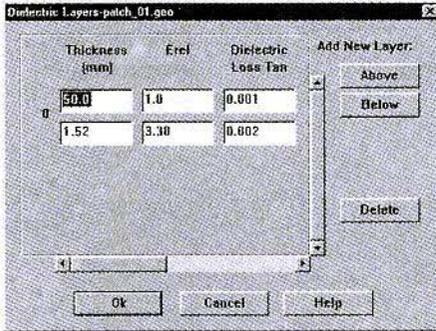


Fig 22 Entering data for the substrate and the air bubble as recommended in the manual

b) So that we do not come into conflict with our lower wave guide limiting frequency (cut off frequency) we should choose at least 1 or 2 wavelengths for the box dimensions (length and width). This is because no wave propagation can take place below the cut off frequency in a wave guide.

c) The radiating edges of our antenna should likewise not come too close to the wave guide walls and consequently should maintain a minimum distance from them of 1 to 3 wavelengths, in order to avoid any kind of unpleasant coupling or reflections.

d) The lid of the box also must have a specific minimum distance between it and the circuit, so that we do not become involved with the stray fields from circuit sections open at the end (head words: fringing and open end extension...). For our patch antenna, an experimental value of approximately half a wavelength is recommended as optimal.

e) As IIDL (top metal) Sonnet offer us three options, namely lossless (an ideal conductor), freespace (this is the famous 377Ω of the free space) and WG load (correct dummy load for the selected wave guide cross-section). The manual recommends the selection free space for patch antenna simulation, as a tried and trusted compromise, and gives the following reasons for this:

Since various modes can be present in the fields in the structure to be examined, a resistance exceeding 377Ω must be selected for a correct adapted load in TE mode. For TM mode, in contrast, a resistance of less than 377Ω should be selected. Thus free space is simply a reasonable compromise which simultaneously enables us to model the energy output correctly as radiation into the free space.

f) We leave the default setting of ideal and lossless metallic conductor for the base of the box (lower side of our circuit board).

And now we can get started on the setup. In the strip at the top edge of the screen, we select Parameters and then Units. There we choose mm.

Now we need the same menu Parameters, but now we want the settings for the BOX (see Fig. 21). For the cell dimensions we should, as far as possible, stay below $1/20$ of the wavelength so for 1,700 MHz, that would mean approximately $1/20$ of 90 mm - so 4.5 mm would be enough. We now cheat a little and take half of the width of our transformation line, i.e. 50 % of 1.066 mm = 0.533 mm for the length and the width of the cell. Thus the width of the $\lambda/4$ circuit will be exactly right, and we can produce a precise model for the patch and / or for the feed. Though this compromise lands us with lots more computing to do.

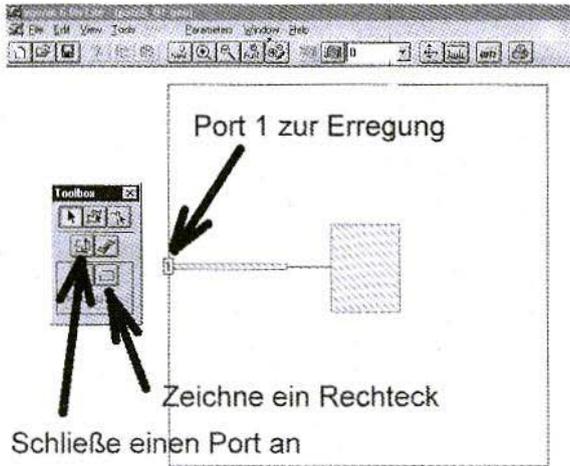


Fig 23 Drawing the antenna structure and connecting to port 1

As regards the dimensions of the wave guide, with our four walls we always remain at least a wavelength away from the antenna structure. I added another 10 % onto that as a reserve and then simply selected approximately the power of two, 256 mm (but all this is not so critical and one has a lot of freedom). We check again to see whether the correct settings have been obtained for the top metal and the bottom metal (free space or lossless) and we save everything using OK.

Then we repeat the procedure and call up Parameters again, but this time we need Dielectric Layers (Fig. 22). In the bottom line, the substrate data is entered for Roger RO4003 material, as per the data sheet, and above this is the air space, with a thickness of 50 mm (approximately half a wavelength). The loss factor of 0.001 for this air bubble comes from a recommendation in the manual, which is intended to lead to more precise results.

Now we can finally draw the antenna structure (see Fig. 23). From the toolbox we take the rectangle draw tool. We left-hand click on the mouse where the top left-hand corner of the patch should be and we then keep the mouse button pressed down and drag until the patch

has the desired dimensions. The dimensions and positions can also be entered directly via the keyboard - see manual. This will not be a completely accurate job, because the rectangle can be increased only by stages in cell units. If the antenna is not exactly in the middle of the wave guide, this can be very easily corrected - left-hand click the mouse in the rectangle area to mark it. Once marked we can left-hand click on the mouse and drag the structure to the required position on the screen.

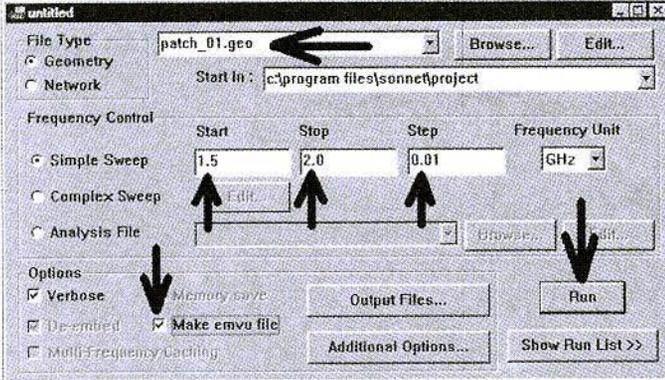
We now draw the transformation line in the same way - first click on the tool button, then draw the line as a rectangle, mark and finally position.

And finally we must play the same game for the 50 Ω - microstrip, though here we have to take one thing into account. In the LITE version, this feed must be taken right up to the side wall, i.e. right up to the wave guide wall, because only there can the entry port be connected to excite the structure and / or analyse the behaviour!

So, when there is really nothing else left to do, we just connect port 1 to this point, first press the appropriate tool button for ADD PORT, then click on the mouse at the desired location, and everything's complete.



Fig 24: Menu for setting the simulation parameters.



Check the Make emvu file tick box to produce a current distribution for the antenna.

Before we go any further, we first have to save everything. We do this in the well-known WINDOWS manner, via File and Save as. Make sure that the file is saved into the Project directory with a reasonable file name e.g. patch_01.geo. Saving files in any other directory leads only to muddle and confusion.

Now things slowly become really exciting, for in the same File main menu we find the line Analyse. We use this to open the menu with the entry parameters for the simulation and we can

pre-set the start frequency, the stop frequency and the step width, in accordance with Fig. 24. We should also check the "make emvu file" tick box, this will show the current distribution over the antenna and the circuits after the analysis. Just another look to check that the correct project is being processed and then press RUN. The program is still just as friendly, and tells us exactly how much operating memory space is needed. This must be less than 16 megabytes otherwise the operation is automatically terminated. We have now

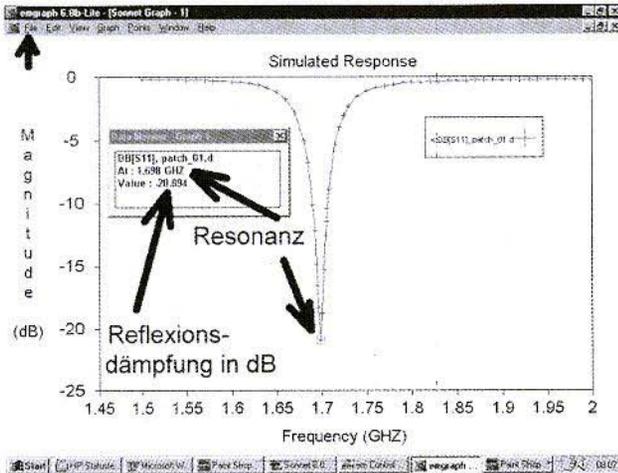


Fig 25 Simulation results for S11 from 1.5 to 2 GHz in Cartesian format

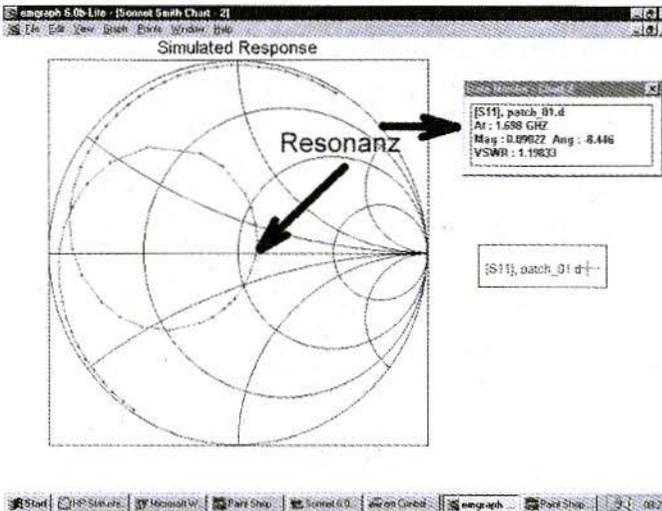


Fig 26: Simulation results for S11 in Smith Chart format

given our Pentium PC two hours work on the desired settings (frequency range 1.6...2 GHz) and we can spend the time doing something else.

At some point, the message job finished will appear on the screen, and we need only press the Open Graph button at the bottom left-hand corner and our S11 appears at once as a rectangular plot (Cartesian display), like the ones we know from PUFF. I just took a brief glance at it and got annoyed about the coarse grading of the display. I immediately ran the simulation again for 1690 to 1710 MHz in 2 MHz steps. Sonnet writes the results of all further simulations as an appendix to the first results file (here: patch_01.d) and displays a warning to this effect with the final results. You just have to keep your nerve and press on regardless. Then Sonnet, like a faithful, obedient hound, finally presents us with Fig. 25. Before we go on to the evaluation, we left-hand click on the mouse twice at a precise point on the curve of S11. In response, a small menu window is superimposed and we can list one value after the other - see PUFF!! - with the help of the cursor keys, and travel along the curve,

with the current value being identified in the diagram by a small rectangle. So the entire picture looks very gratifying, our frequency of resonance, according to Sonnet, is located at 1698 MHz and the reflection attenuation (magnitude of S11) is just a little more than 20 dB. So it would appear that, although the two design routines explained in the previous section took a great deal of time and effort, they were not in vain.

However, one question remains, where has the relevant Smith diagram got to? Lets take another look at Fig. 24, in the top left-hand corner, a coy little arrow can be seen near File. Click on it and then call up OPEN and Smith in the sub-menu. Then we first have to open the Results Data file of our simulation (patch_01.d) and then use the familiar double-click routine on the red curve in the diagram to open up the data window on the screen (please also close the old window for the Cartesian display). Fig. 26 shows all of this, the cursor is at the resonance point, but is indicating, in its friendly way, not only the relevant frequency, the amplitude and the phase of S11, but also the VSWR - first-class service!

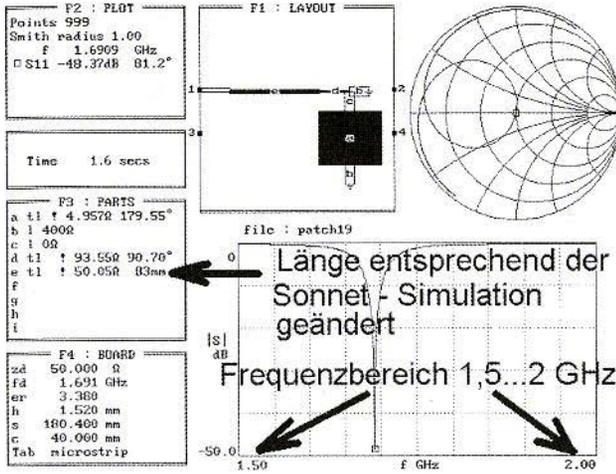


Fig 27: Results from repeated PUFF simulation with revised values.

5. Comparing measurement results from PUFF and Sonnet simulation

Now for the burning question, how accurate are our various designs? With this in mind, we first repeated the PUFF simulation, though now with 82 mm length for the 50Ω feeder, as this was what Sonnet was working with, since there we had to go right up to the wave guide wall. Fig. 27 shows the result, the gradients for the PUFF and Sonnet simulations are identical right up to the area of the resonance. We actually experimented with PUFF until the perfect match was found, while Sonnet maintains that a minimum reflection factor of about 10 % should be expected.

So measurements were made with the vectorial network analyser (hp 8410 with S-parameter test set hp 8745A, sweeper hp 8690 and Polar Display), to determine the gradient of S11 in the frequency range between 1500 and 2000MHz. Fig. 28 shows a photograph of the screen with the polar representation of S11, let us look at some of the details more closely. There is a whole

series of discoveries to be made. Among other things, I came across one silly thing which I'd like to pass on, so that other people don't make the same mistake.

a) The minimum reflection factor shown in the screen photo amounts to approximately 2% (smallest distance of curve from centre). That would be a

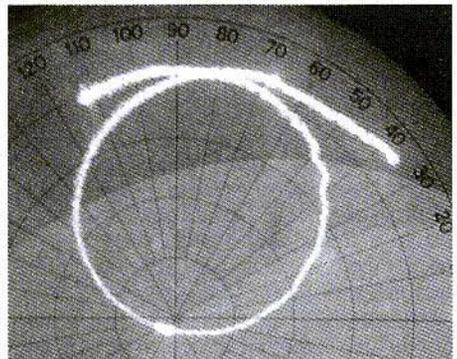
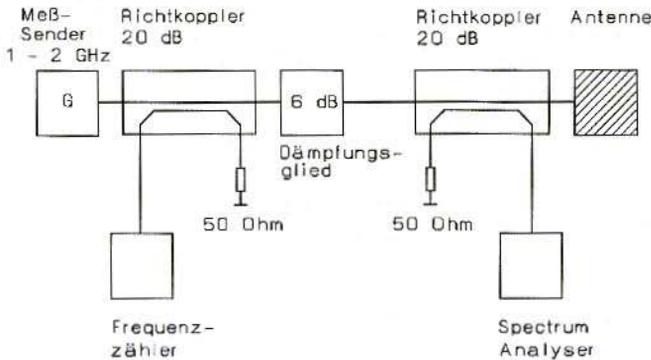


Fig 28: Photograph of display from network analyser measurement showing S11 from 1.5 to 2 GHz with a marker at 1690 MHz.

Fig 29: Revised test measurement layout.



fantastic result, but in subsequent test measurements it always came out as between 7 and 8%. It was a long time and I searched around a bit before the penny dropped. To take this macro-photograph of the screen, the photographer had practically crawled into the test rig, and his stomach had come too close to the antenna. The energy reflected by the antenna and superimposed on the return wave, which was what was used to determine the reflection factor in the network analyser. And the phase would have to match, so that we got an outstandingly good but fake - result. A professional microwave antenna developer who happened to be present just roared with laughter over the discovery. He then showed me how this is put to use in laboratory practice, you slowly bring your right hand nearer to the patch antenna and check whether the resonance point starts to circle round the screen centre point like a propeller. If it does, then the antenna is actually radiating in the desired frequency range and the resonant frequency and the matching are in order. Well well, you never stop learning.

b) But the reflection factor attained by

the real antenna, including the transformation line, is not necessarily anywhere near this 7 to 8 %, but rather nearer to 5 to 6 %, for unfortunately the network analyser also registers all the additional irregularities arising in the practical circuit (e.g. the microstrip - SMA transition on the circuit board or the semi rigid cable switched between the analyser and the antenna circuit board, with its two SMA connections).

c) The 50Ω microstrip connection circuit provided on the printed circuit board is in reality considerably shorter electrically than the value used in the Sonnet simulation (less than 40 mm, as against over 80 mm for Sonnet). This leads to two important differences in the analyser screen compared with the simulation.

Firstly, the curve length for the larger feed length (Sonnet simulation) also appears rotated clockwise.

Secondly, in the Sonnet representation, the additional circuit length causes greater phase displacements, with increasing frequency. This causes curves with a longer feed length to be extended over a larger angular range of the Smith diagram for the same sweep

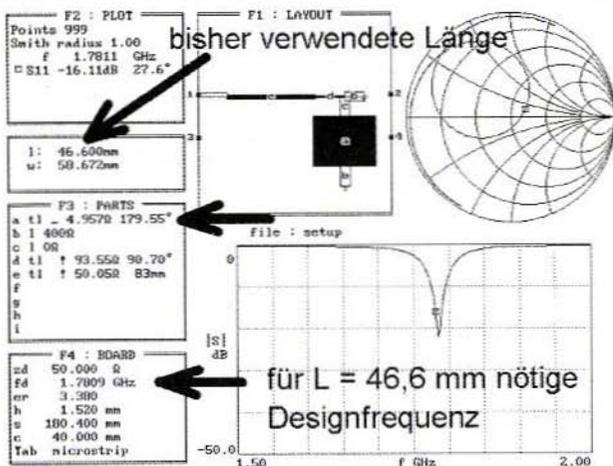


Fig 30: Using PUFF the design frequency is increased to give the patch length of 46.6 mm.

range. This effect is very easy to recognise in a direct comparison of Figs. 26 and 28.

The reflection measurement bridges used with the network analyser can cause reference plane displacement, and thus can compensate these effects. I have used this effect to create the following measurement condition.

The reflection curve measured is valid as a reference point for the input of the SMA connector socket (beginning of Teflon internal insulation) soldered onto circuit board!

Though there is still one little problem, the reflection measurements certainly show up as being very precise when checked with calibrated unmatched dummy loads, in spite of the measurement equipments being already somewhat suspect. So how can the frequency of resonance be tested very exactly and with high precision, but using privately-owned measurement equipment of the generation before last? Also how can the correct patch length be checked along at the same time. So then I did some modifications and assembled a measurement layout as per Fig. 29. Its very simple in principle, the wave fraction reflected from the antenna through the output directional coupler is

determined, with the spectrum analyser serving only as a highly sensitive, selective Voltmeter. The power frequency is measured precisely by means of a further directional coupler and a frequency counter. Now the generator frequency is slowly altered by hand in the area around 1,690 MHz, while the analyser display is observed and the minimum reflection is sought. This was precisely on a frequency of 1,686.6 MHz.

Thus our PUFF/ patch 16 design produces very good matching first go, and only a little follow-up work needs to be done on the frequency of resonance.

The Sonnet frequency of resonance calculation certainly gives a value which is somewhat too high, but a close study of the manual turned up the comment that an error of 1 % should be expected (up to as much as 17 MHz!). Apart from this, in drawing the structure we are stuck with our selected cell size of 0.503 mm as the smallest step width and therefore we don't precisely reproduce the theoretically necessary length $L = 46.6$ mm. This gives us a further uncertainty. The simulated reflection response, in contrast, tallies very well with the measurement, such an overall result is still somehow very

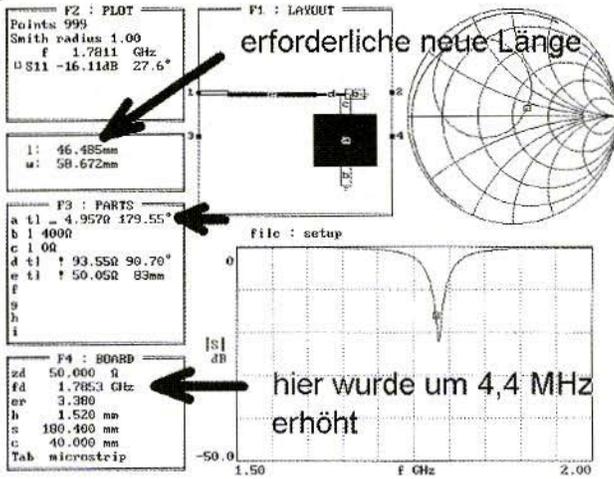


Fig 31: The design frequency is increased by 4.4 MHz to give a new patch length.

satisfying in the end, and is worth the intensive effort.

Lets have a brief look back at the measures that were necessary to complete the development of our antenna.

Part 1: Correction of patch frequency of resonance

For this purpose, lets call up our last PUFF file from Fig. 26 again and copy

it into a new file (In such cases, I simply save it as setup.puf). In the entry line for the 4.997Ω microstrip (which forms the patch), we remove the exclamation mark and then, in field F4, we increase the design frequency until we arrive at the previously used patch length of 46.6 mm (please leave the cursor on the patch line in F3 and key in the equal sign.). Fig. 30 shows this case. The rest is now childs play, for now the value of the design frequency in F4, which is current for this case is simply increased by precisely the devia-

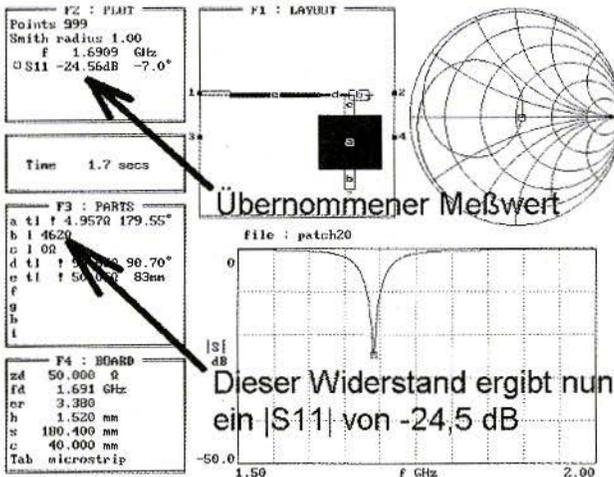


Fig 32: Using PUFF to increase the resistance to give an S11 magnitude of -24.5 dB.

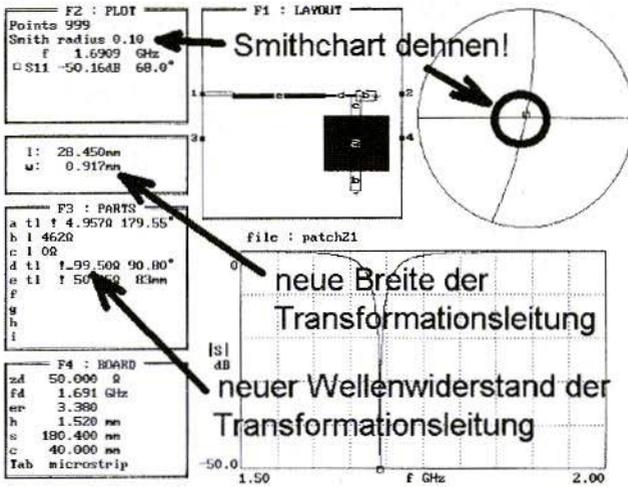


Fig 33: New corrected antenna design with matching corrected.

the new patch length required ($L = 46.48$ mm), see Fig. 31.

Part 2: Correction of matching

Here things become a bit tricky, for we have to choose between two possible paths:

a) We leave the patch dimensions unchanged (apart from the previously discussed tiny length and / or resonance corrections) and we bring about the correct matching by changing the dimensions of the transformation line

b) We leave the matching circuit unchanged and try to ensure correct matching by changing the patch width.

Path a) is easier and can be completed quickly. If we start from the assumption that we have to eliminate about 6 % more of the reflection factor, then we call up the last PUFF simulation onto the screen (it was Fig. 26). In F3, we then increase the resistance connected to the patch edges until an |S11| of approximately -24.5 dB is displayed (for that corresponds to our measured reflec-

tion factor $r = 6$ %). This case (which the network analyser display can be said to reproduce) is shown in Fig. 32, and now we simply correct the wave resistance of the transformation line until correct matching is attained once again, with an |S11| of at least -40 dB (Fig. 33). The changed mechanical dimensions of the transformation line are then printed out, the open end correction is made, and a new layout is drawn up again, with all the necessary changes.

6. Current distribution in the antenna circuit

If, before starting the SONNET analysis, we have activated the option make emvu file, then we can use the corresponding start menu button - View Current - and browse for the file patch_01.jxy. It is in our projects folder and it contains all information concerning the current distribution and / or current densities in our antenna structure. Once it has been called up, it is converted into a very attractive graphics display (Fig. 34). By pressing the marked keys, we call up the various frequency values and, using the

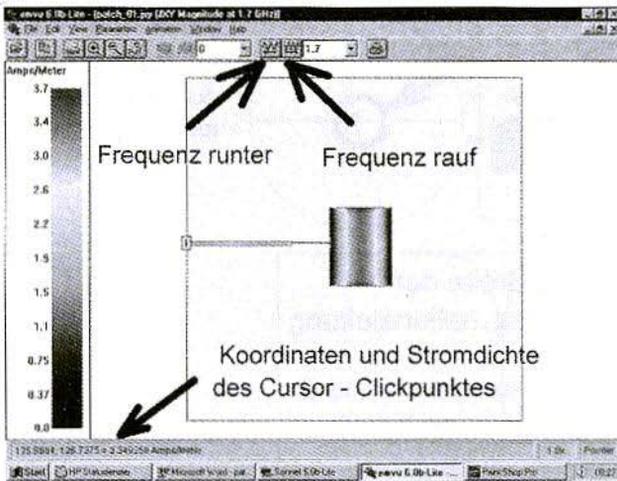


Fig 34: Simulation of the current distribution in the antenna designed using Sonnet.

ture. Once it has been called up, it is converted into a very attractive graphics display (Fig. 34). By pressing the marked keys, we call up the various frequency values and, using the colour scale at the left-hand edge of the image, we can see very precisely where high-level or low-level currents are flowing in our circuit. We can even position the cursor on a specific point in the antenna layout, left-hand click on the mouse, and obtain, not only the co-ordinates, but also the prevailing current density value at this position, superimposed at the bottom edge of the image.

If we take a closer interest in the current distribution at critical points in the circuit (branches, irregularities between thick microstrips or patch edges), then we can simply zoom in on this area using the button with the + - magnifying glass. Naturally, the cell structure of our simulation sets limits for the resolution, but we can easily see where nothing is flowing (there is additional or unnecessary capacitance at this point) or where very high currents occur (higher losses arise with very narrow track structures).

In the Sonnet tutorial there are very precise instructions on how you can insert the individual results for the

various frequencies into a small animation and have it performed in front of you. But apart from the displaying of the frequency dependency of our current densities, there is a further option available, which should also be tried out, just for the sake of interest, the frequency value can be left constant and a time animation can be carried out. To do this, the phase of the voltage applied to port 1 is altered in stages, and we can now, for example, have 36 steps carried out, from zero to 360 degrees of phase displacement. It is very nice to see the well known phenomena of standing waves. It is really impressive and amusing to see how they are formed and where they occur. Particularly informative are the effects on the feeder, when we first consider the correct matching for resonance and then depart from this frequency.

7.

Second Option: Square patch with coaxial feed

The stripline version had functioned well at the first attempt. The $\lambda/4$ transformation line and the 50Ω feeder



now had to be dispensed with, which makes smaller antenna dimensions possible. The antenna is now fed from below with a semi-rigid cable, with the patch itself being used as a transformation line for matching.

7.1. Determining patch data

For this purpose, the Patch 16 program is re-started, and this time key A (for Auto design) is pressed in the main menu. Thus a square patch antenna is drafted with the desired coaxial feed. But first a few preliminary observations:

a) Patch first asks for the design frequency. However, according to the test readings from the previous chapter, we need a corrected patch length of 46.48 mm = 1.83 inches for an actual frequency of resonance of 1,691 MHz. But the simulation results obtained from Patch 16 supplied a frequency of resonance which was too low. So, with the first entry, we immediately start at a rather higher frequency (e.g. 1.695 GHz), to obtain shorter antenna dimensions.

b) Then the program wants to know the desired band width. Here we should call to mind the paragraph from Chapter 3.1 ...a thinner printed circuit board gives the antenna a smaller band width.... That immediately leads us down a somewhat unfrequented path.

In accordance with the band width and the material data keyed in ($\epsilon_r = 3.38$ / loss factor = 0.001), the program will actually calculate the appropriate board thickness!

Thus we propose a specific bandwidth (e.g. $b = 30$ MHz) and have the associated antenna and board data specified. Then we press key e in the main menu and check whether we have already arrived at the desired values.

- Patch length = Patch width = 1.83 inch
- Printed circuit board thickness = 0.06 inch.

Since this will certainly not be the case in the first time, this Autodesign draft step will repeat, using other frequencies of resonance and other band widths,

```

The Resonant Frequency is 1.694 GHz
Qo is 66.9
The Edge Radiation Resistance is 298.28 ohms
Zc of Quarter-wave transformer is 122.1 ohms
Approx. width of the Quarter-wave transformer is 0.021 inches
Length of Quarter-wave transformer is 1.124 inches at the Resonant Freq.

Input Resistance at probe location is 49.72 ohms

The 2:1 USWR Bandwidth is 1.2%
Upper Frequency Limit = 1.704 GHz
Lower Frequency Limit = 1.684 GHz

Directive Gain = 6.5 dBi
Press 'ENTER' to continue: █
  
```

Strahlungs-
widerstand

Erreichte Anpassung
durch Transformation

Fig 35: Results of the Autodesign process for 1698 MHz and 20 MHz bandwidth.

Fig 36: Dimensions for the Autodesigned antenna.

These are the design parameters:

Length (L) = 1.83 inches
 Width (W) = 1.83 inches
 Height (H) = 6.041024E-02 inches
 Dielectric Constant (D) = 3.38
 Loss Tangent (T) = .001
 Feedpoint Distance (F) = .6700109 inches

Do you wish to edit any value (Y/N):

Patchlänge

Patchbreite

Platinendicke

Abstand des Feedpoints
 von der strahlenden Patchkante

until the desired dimensions are obtained.

The final result can be revealed right away, you need a frequency of resonance of 1.698 GHz and a bandwidth of 20 MHz to arrive at the above patch dimensions and the correct board thickness.

Thus we obtain the antenna characteristics listed in Fig. 35, the associated mechanical dimensions and data can be obtained from Fig. 36. After conversion from inches into millimetres, this gives us a length L of 46.48 mm, a width W of 46.48 mm, and a gap between the feed point and the radiating patch edge of 17.02 mm.

It is now very useful to check with PUFF to see how well this draft fits and how, if necessary, it can be improved somewhat further through realistic modelling.

Step 1:

First we use PUFF (see Part 1) to determine the impedance level for a 46.48 mm wide micro-strip line at 6.166 Ohm.

Step 2:

The open-end extension must now be added to the length L = 46.48 mm on

each side. For this purpose, let us cast a glance at Fig. 19 in Part 1. A correction of 50% of the board thickness of 1.52 mm, is associated with the impedance level $Z = 6.166 \text{ Ohm}$ i.e. 0.76 mm. The total length obtained in this way amounts to $(46.48 \text{ mm} + 2 \times 0.76 \text{ mm}) = 48 \text{ mm}$ and it simultaneously represents an electrical length of 180 degrees.

Step 3:

The feed point lay 17.02 mm away from the radiating patch edge. If we again add on the open-end extension of 0.76 mm, we obtain 17.78 mm.

By means of a simple rule of three calculation, we obtain an electrical length of 66.67 degrees for this (if we put this 17.78 mm in a ratio to the 48 mm, the 180 degrees appear).

Step 4:

Now we use PUFF to add together the antennae made from two ideal micro-strip lines, the electrical lengths of which must amount to 66.67 degrees or $(180 / 66.67) \text{ degrees} = 113.33 \text{ degrees}$. At the left-hand and right-hand ends, the radiation resistance calculated by the Patch16 program is connected up (600 Ohms in each case) and the feed point in the centre of the two line

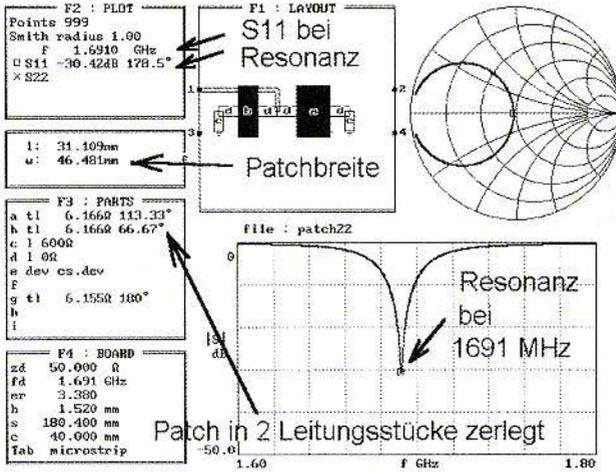


Fig 37: Results of PUFF simulation for S11 using the data from the Patch 16 Autodesign.

sections leads to port 1. If we now start the simulation, we obtain a value for S11 (and thus a reflection attenuation) of approximately -30 dB at precisely 1691 MHz (see Fig. 37).

Step 5:

For realistic modelling, we assign the well-known exclamation mark to each of the line sections listed in F3. If we repeat the simulation, we establish that, apart from the matching, the resonance has also become poorer. So we vary the individual lengths of the two line sec-

tions until everything has been optimised and we obtain a representation corresponding to Fig. 38 on the screen. To do this, please also select a smaller frequency range and a smaller Smith chart radius!

Step 6:

After deleting the exclamation mark and entering the equal sign, we have the mechanical length displayed for each line section. The values must be as follows:

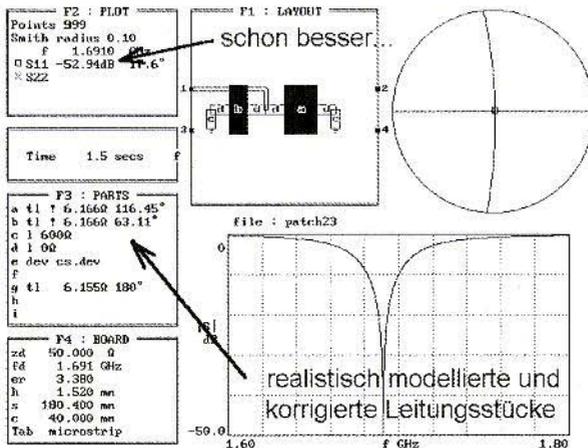


Fig 38: Using PUFF to optimise for best value for S11.

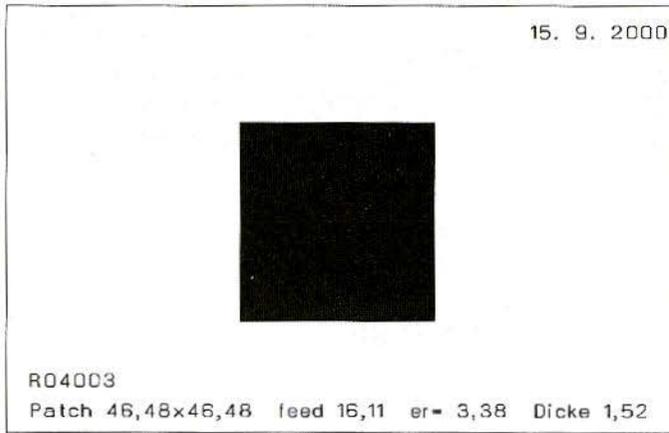


Fig 39: Final layout of the printed circuit board.

- short section: $L = 17.324$ mm
- long section: $L = 31.966$ mm

Step 7:

The sum of these two lengths amounts to $L = 17.324$ mm + 31.966 mm = 49.29 mm. If we compare this with the total length of 48 mm., calculated by Patch16 and open end corrected in Step 3, the PUFF dimensions need to be theoretically shortened to 97.38% (so that the antenna will be tuned when actually constructed).

The same applies to the feed point. Thus the short length values provided by PUFF (17.324 mm) must also be shortened by this factor, which then gives us 16.87 mm. Now we just have to subtract the open end extension of 0.76 mm from this, and thus we arrive at a new position for the feed point, with $(16.87$ mm $- 0.76$ mm) = 16.11 mm, measured from the radiating edge of the antenna in the direction of the centre.

With these mechanical data

- patch length = 46.48 mm

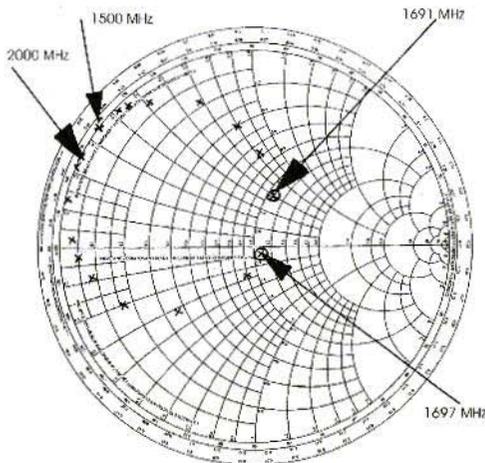


Fig 40: Plot of S11 between 1.5 and 2 GHz.

**Table 1:**

Frequency MHz	Reflection Factor	Phase Angle
1500	1.00	151.0
1600	1.00	142.0
1630	0.98	135.5
1640	0.97	131.0
1650	0.92	127.5
1660	0.89	120.0
1670	0.80	110.0
1680	0.62	97.0
1685	0.50	85.0
1691	0.28	68.0
1697	0.07	-45.0
1700	0.15	-100.0
1710	0.51	-140.0
1720	0.73	-157.0
1730	0.85	-170.0
1740	0.90	-177.0
1750	0.92	177.0
1800	0.98	165.0
1900	1.00	156.0
2000	1.00	153.0

– patch width = 46.48 mm

– feed point distance = 16.11 mm from patch edge

We finally change into the layout program, the result is shown in Fig. 39. For the selected Europa card format of the printed circuit board (160 mm x 100 mm), the feed point was displaced from the patch centre parallel to the longer side of the board, so that a greater distance would be maintained to the board edge for the two radiating edges. The bore diameter for connecting the inner conductor of the feeder is 0.6 mm.

A specimen antenna was prepared in accordance with this model and was

fitted with a section of semi-rigid cable about 7 cm long, together with an SMA plug.

Fig. 42 shows the graph for S11, as it was established by the hp 8410 network analyser in the range between 1.5 and 2 GHz.. A datum level measurement was undertaken first using a wider section of semi-rigid cable with the same length, working without load at the end, with a screwed-on SMA socket. Therefore the measurement applies precisely for the patch / internal conductor soldering joint measurement. The range between 1.5 and 2 GHz was covered by 20 measurement points, so that the step width was reduced in the vicinity of the resonance. The readings can be obtained from Table 1.

A minimal reflection factor of approximately 7 % was obtained here straight away. If we compare the result with both the simulation just carried out and the readings from the Microstrip-Feeding version from Part 1, then several divergences become apparent.

a) Because, with this new version of the antenna, the two microstrip lines (transformation + feed line) with their relatively long electrical lengths, are omitted, the Coaxial-Feeding here does not cause the display to a loop through the frequency dependent phase angle rotations of the graph. So we no longer obtain a beautiful loop (as with Microstrip-Feeding), but only a graph which is not quite fully closed.

b) In comparison with the simulation, the S11 curve in the Smith diagram appears to be rotated it initially looks as if there was another line section in the circuit somewhere, which had been overlooked up until now.

c) Something must have happened to the frequency of resonance as well, for

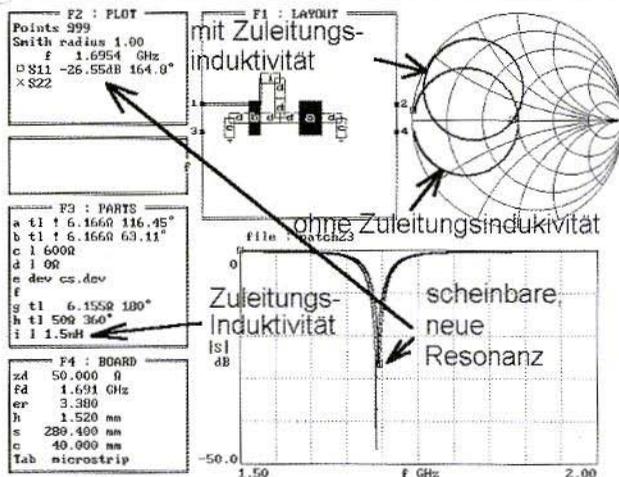


Fig 41: PUFF
 Simulation showing the effect of the bare feeder within the thickness of the printed circuit board.

the frequency mark for 1691 MHz is no longer in the vicinity of the reflection minimum! The new minimal value for the reflection attenuation is now -23 dB at 1697 MHz.

So we have to look for the reason why a divergence of -4 MHz in Microstrip-Feeding (see Part 1) has suddenly become +6 MHz when only the type of feeder has been changed.

Finally only a modest correction was undertaken to the theoretical patch length, in order to obtain precisely the required 1691 MHz - and now we are suddenly way out, with a larger difference in the opposite direction!

The reason does not lie in an incorrect resonator length, but in the fact that the feeder has been fed in from below. This means that the internal conductor travels through the cross-section of the board material in order to get up to the patch. Since there is no external conductor here, we automatically get an irregularity; the bare internal conductor actually acts like an inductance within the thickness of the board. This effect leads to both the measured matching error and the resonance tuning error and brings about the rotation of the curve in the Smith chart. Since, with thin wires, we can roughly reckon on an

inductance of about 1 nH per millimetre of length, it is worthwhile carrying out a short investigation using PUFF to estimate this influence. Moreover Patch16 explicitly warns of this problem if the board thickness selected is too large! In Fig. 41 we can see the corresponding simulation, for which a feed line inductance of 1.5 nH has been assumed. The additional phase angle rotations (in comparison with the ideal feeder) caused by the inductance and the frequency of resonance and matching errors can be very clearly recognised here.

There are three different starting points for solving this problem.

- 1) We select a thinner board material and thus reduce the cable segment without an external conductor.
- 2) As a feed line, we use a semi-rigid cable with a solid copper tube as an external conductor. We then go down to about half the board thickness with a special blind hole borer or spot facer, with a diameter which corresponds to the external diameter of the cable. Through this recessed cable we markedly reduce the cable segment without an external conductor (Fig. 44).
- 3) The extra inductance and its effects

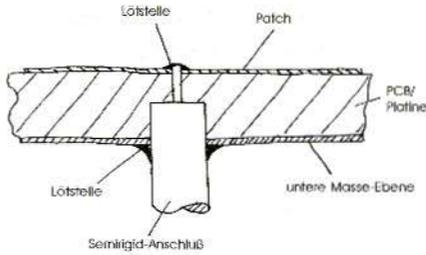
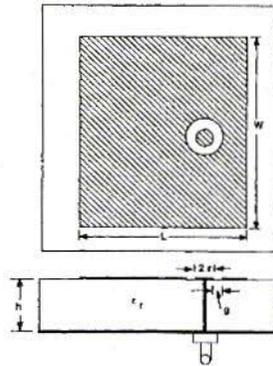


Fig 42: Mechanical design of recessed feeder cable to reduce inductance effect.

are regarded simply as an unavoidable part of the system, and attempts are made to compensate for their influence. It might indeed occur to us to increase the patch length, as this would bring the self-resonant frequency down again. This leads to a capacitive reactive component which compensates for the inductance at the desired operating frequency, which now lies above the natural resonance, as with any parallel resonant circuit. But BE CAREFUL, this is the wrong route to take, as the radiator length of the patch must not be touched.

By operating near the natural resonance, we unfortunately move away from the optimal operating case and antenna efficiency. In this connection, you should cast a glance at the introductory chapter, and especially the note on field lines which look like shaving brush bristles.

For optimal re-radiation, we are dependent very much on the phase difference of exactly 180 degrees between the voltage levels on the two patch edges, when the fields should actually be in phase for the two slot. This condition is



Configuration for inductive reactive compensation for the back-fed probe feed.

Fig 43: Example of ring shaped capacitor taken from reference material [3]

no longer fulfilled if there is any de-tuning i.e. shortened radiator lengths even if this gives us perfect matching at 1691 MHz.

The single permissible route is to compensate for this undesirable inductive series reactance through a type of series capacitance without altering the dimensions of the antenna. This could be, for example, a small SMD capacitor. But where should it be positioned on the printed circuit board? This problem is usually solved by a gap, i.e. a ring-shaped air gap between the soldering joint of the cables internal conductor and the remainder of the patch area (Fig. 43 taken from [3] in Part 1, Page 104). However, things become interesting if, using another simulation from PUFF, we determine the size of the series capacitance required (Fig. 44), it is almost 6 pF. So we have to do a lot of fiddling about with the air gap in order to obtain this capacitance value without simultaneously generating relatively large, unnecessary parallel capacitances to earth at the air gap edges.

In practice then, we see solutions as

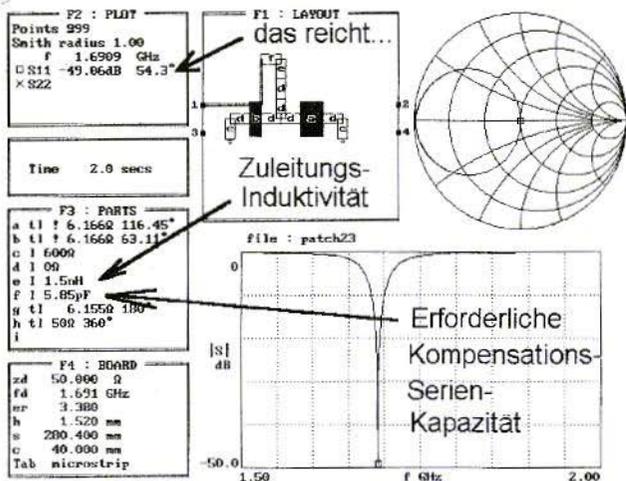


Fig 44: PUFF Simulation to determine the value of the capacitance required.

antennas are built from metallised ceramics or metallised plastics, when this SMD capacitor is integrated into the antenna using a small round chimney at the feed point.

[6]: John D. Kraus, Antennas, bottom of Page 746

[7]: Series of articles on Designing and building microwave circuits with PUFF, Gunthard Kraus, DG8GB, VHF Communications from issue 4/1996 onwards to 2/1999

8. Literature index for complete article

[1]: Friedrich Krug: Microstrip antennas. VHF Reports, Issue 2/ 1985, Pages 113-121

[2]: John D. Kraus: Antennas. Second Edition 1988. McGrawhill International Editions, ISBN 0-07-100482-3.

[3]: P. Barthia / K.V. S. Rao / R. S. Tomar: Millimeter - Wave Microstrip and Printed Circuit Antennas. Artech House Boston and London. ISBN 0-890 06-333-8

[4]: Fa. Mauritz GmbH & Co., P.O. Box 10 43 06, D-20030 Hamburg

[5]: <http://www.rfglobalnet.com> or via search engine <http://www.fireball.de> under all languages.

Then please search for pcaad21.zip or patch16.zip.



Wolfgang Schneider, DJ8ES

Digital Speech Store

A speech store for CQ is a practical aid to radio amateurs, especially in a contest. The CQ is started automatically at the push of a button.

A micro-controller takes charge of controlling the sequences here, including PTT. Thus the number of components in the circuit can be reduced to a minimum.

1.

Description of circuit

The circuit (Fig. 2) for the CQ speech store contains two core elements. The first of these is the ISD 2560 speech memory integrated circuit. The memory can store text lasting a maximum time of 1 minute and repeat it as many times as desired. The memory's contents are retained even if there is no power. A single-pole two-way switch handles the change from record to play.

The ISD 2560 speech memory integrated circuit also requires an external analogue circuit; this provides the filter functions and the level adjustment to the

microphone input of the transceiver. The second essential component is the AT90S2313-10PC micro-controller. The latter is responsible for all sequence control for the transceiver, both as regards the speech memory module and as regards transmit control, through the PTT connection. The external circuit includes an electret microphone which, like the loudspeaker, is directly connected to the IC2.

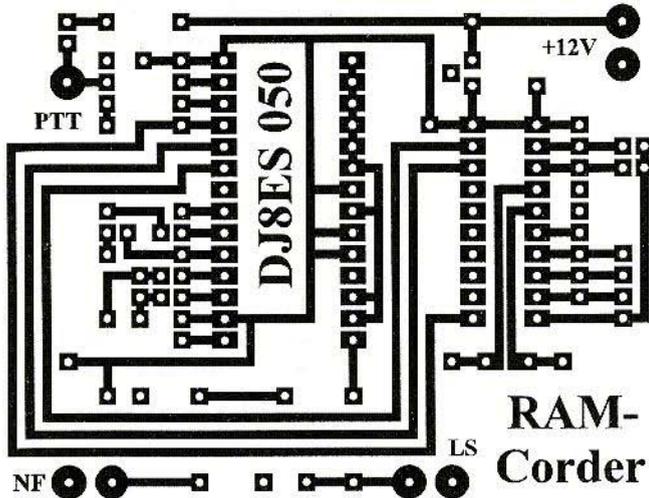
The CQ button starts the output of an individual sequence. This applies both to transmission and to the initial recording of the CQ.

The CQ loop button starts an endless loop for the CQ. Here the interval length between two calls is determined by a jumper. You can select 5, 10 or 15 seconds. During the interval, the PTT connection becomes inactive and the transceiver switches to reception.

The "Stop" button interrupts the endless loop; other sequences are also interrupted through the "Stop" button. Thus a running CQ can be broken off at any time. The "Stop" button is also used to terminate speech recording (initial recording of CQ).



Fig 3: PCB Layout for the Speech Store.



for the CQ loop. Here the desired selection can be made, for example, using jumpers. All control inputs are switched to earth and are thus "low-active".

Before the micro-controller and the speech memory IC can be inserted, the individual operating voltage **must** be measured and thus the integrated 5 V voltage regulator must be tested. When the supply voltage has been applied (+12 V to +13.8 V), this 5 V must be measured at PIN 20 IC1 and PINs 7, 9, 10, 16 and IC2, using a digital multimeter.

2.1. Parts list

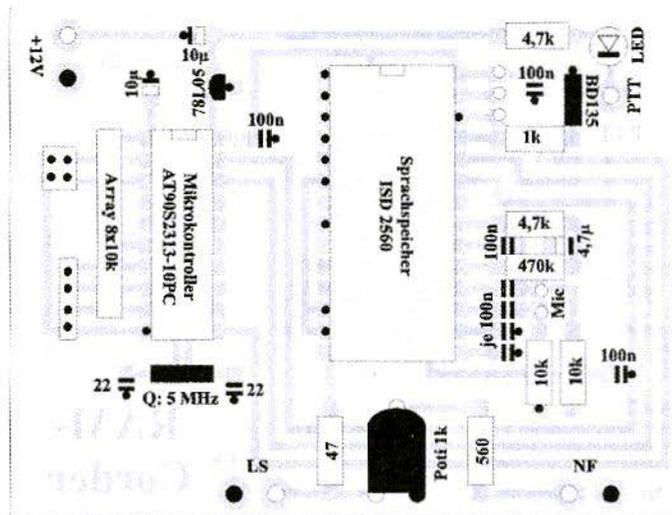
- 1 x AT90S2313-10PC, Micro-controller (IC1)
- 1 x ISD 2560, speech memory IC (IC2)
- 1 x 78L05, voltage regulator
- 1 x BD135, transistor

- 1 x 5.0000 MHz, quartz
- 2 x 10 μ F/25 V, tantalum electrolytic capacitor
- 1 x 4.7 μ F/25 V, tantalum electrolytic capacitor
- 2 x 22 pF, ceramic capacitor
- 8 x 100 nF, ceramic capacitor
- 1 x trimming potentiometer 1 k Ω , horizontal, RM 5/10 mm
- 1 x resistor array 8 x 10 k Ω
- 1 x 0.5 W/8 Ω , loudspeaker
- 1 x electret microphone
- 1 x LED green, low current
- 1 x 20-pin, IC socket
- 1 x 28-pin, IC socket
- 1 x switch, on/on
- 3 x push buttons
- 7 x soldering pins, 1 mm
- 1 x PCB DJ8ES 050

Resistors, 1/4 W, RM 10 mm

- 1 x 47 Ω
- 1 x 560 Ω
- 1 x 1 k Ω

Fig 4: Component layout for the Speech Store.



- 2 x 4,7 k Ω
- 2 x 10 k Ω
- 1 x 470 k Ω

3. Putting into operation

When the circuit has been fully assembled, it can be put into operation immediately. Only the waiting time for the CQ loop and the audio frequency level for the speech output are adjustable. The audio frequency output signal should drive the transceiver to full output, but just avoid overloading it (the key concept here is "band width")! The audio frequency output of the circuit is fed through the microphone socket of the transceiver, parallel to the microphone.

In order to record or save a CQ, proceed as follows:

- Switch in position "Rec"

- Start recording using CQ button
- Speak text (call to all stations) into microphone
- End recording using "Stop" button

The LED light is on during recording. The maximum recording time available is 1 minute.

Note: In recording mode, no other radio equipment should be switched on, otherwise the transceiver automatically switches to transmit through the PTT connection when the CQ button is pressed.

In repeat mode (switch in "Play" position), a button can be pressed to choose between a single CQ or else a CQ loop (endless loop). The desired interval length between two calls can be individually set:

- 5 seconds no jumper

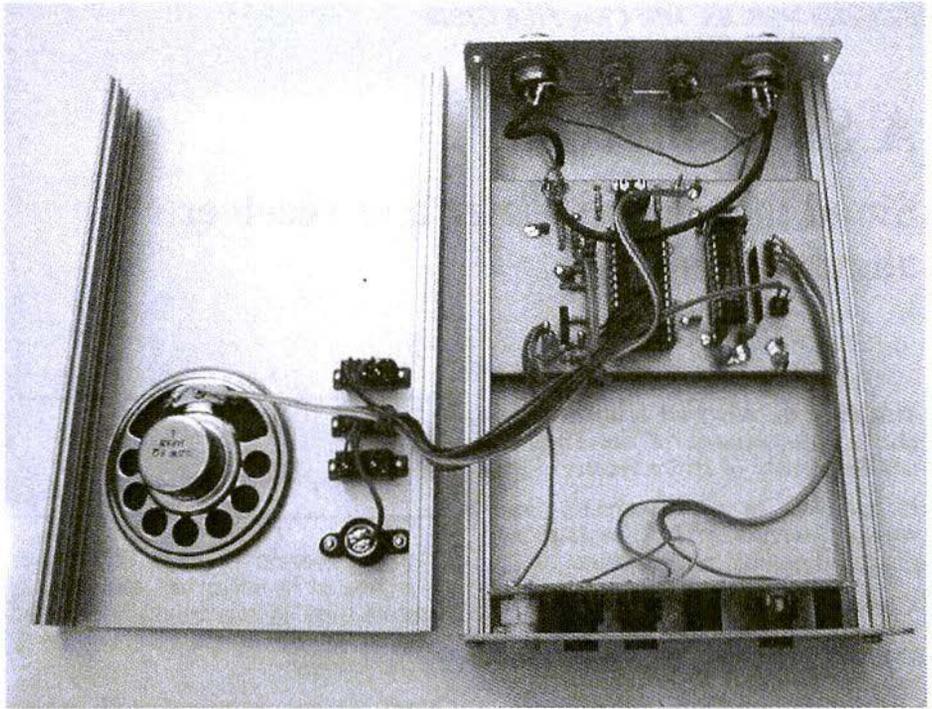


Fig 5: Internal view of the completed Speech Store.

- 10 seconds jumper on connection, pause 10 s
- 15 seconds jumper on connection, pause 15 s

During the speech interval, the PTT connection becomes "high" and the transceiver goes from transmit to receive.

The "Stop" button interrupts the endless loop. Other cycles are also interrupted by using the "Stop" button. Thus a continuous CQ can be aborted at any

time.

4. Literature references

[1] Dipl.-Ing. Max Perner, DM 2 AUO: Low-frequency memcorder with ISD 2560; Funkamateur 12/99, Theuberger Verlag Berlin

[2] Helmut Israel: A chip stores analogue signals; Funkamateur 3/00, Theuberger Verlag Berlin



E.Chicken MBE, BSc MSc CEng FIEE G3BIK

A Sinadmeter

Sinad measuring meter and receiver alignment aid

This easy to construct testmeter simplifies measurement of the 12dB Sinad Sensitivity of an fm receiver.

It connects to the receiver's loudspeaker terminals to automatically display the Sinad measurement .

Power is from an external mains/dc unregulated 12V or 15V adaptor @10mA , or from the receiver's dc supply.

When using the sinadmeter for measurement of a receiver, it must be used in conjunction with an fm signal generator of the type that has a modulating tone of 1kHz with adjustable deviation, and a voltage output control calibrated in microvolts rms . That type of signal generator is fairly standard to those who are involved with fm radio equipment, so should be reasonably available.

A very useful secondary feature of this Sinad meter, is that it can be used as a sensitive response indicator whilst tweaking for optimum sensitivity (maximum quietening) the rf, mixer, and local oscillator stages of an fm receiver . This is of particular advantage for example when retuning an ex PMR receiver for use on the amateur bands.

1.

Sinad Explained

The Sinad-sensitivity of an fm receiver is expressed in microvolts rms , and is derived from an internationally accepted method for determining the sensitivity of the receiver.

Sinad is a ratio expressed in dB, which relates the level of a 1kHz audio-tone output from the loudspeaker, to distortion and noise generated within the receiver, using the formula :-

$$\text{SINAD (dB)} = \frac{\text{Signal plus Noise And Distortion}}{\text{Noise and Distortion}}$$

Note that the term Signal refers to the 1kHz audio-tone output from the loudspeaker, not the rf input signal!

The 12dB Sinad Sensitivity figure which is expressed in microvolts rms (not as Sinad dB), is the accepted standard of Sinad measurement against which to judge the sensitivity of an fm communications receiver. It is defined as that level of microvolts rms (pd) rf signal into the receiver's antenna port , which produces at the audio output a Sinad ratio of 12dB . At this sensitivity the receiver would give an intelligible

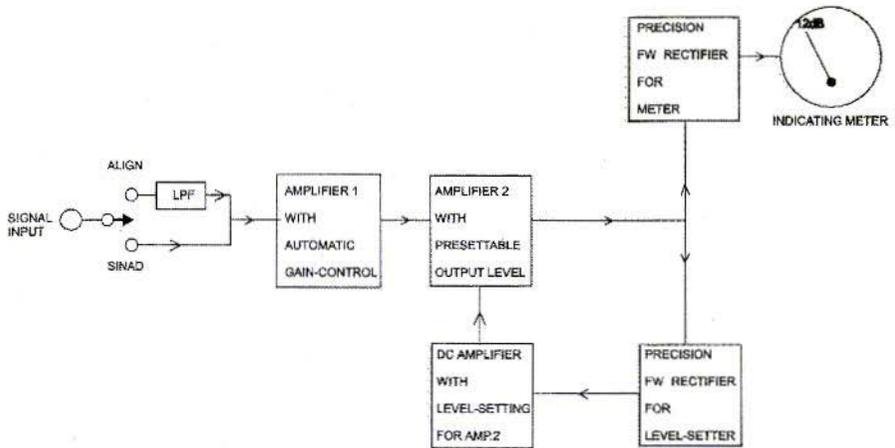


Fig 1: Block Diagram of Sinadmeter

voice signal on an acceptably quiet background. Any modern communications receiver will have a 12dB Sinad Sensitivity in the order of 0.5 microvolt rms or less.

Strictly speaking, the definitive method of Sinad measurement requires that :-

- the rf signal being fed into the receiver's antenna terminal be frequency-modulated by a 1kHz tone with deviation to 60% of the receiver's specified peak deviation;
- the Sinad meter must incorporate a 1kHz band-rejection filter;
- the Sinad meter be connected across the receiver's loudspeaker or an equivalent load;
- the audio output power should be at 50% of rated maximum when the Sinad meter indicates 12dB.

But, for the 5kHz peak deviation typical of today's amateur-band receivers, that

definitive method simplifies to become a 1kHz modulating tone at 3kHz peak deviation from the signal generator. Its microvolts rms level will be in pd because the receiver's input impedance will in normal practice match that of the signal generator.

This design does incorporate the required 1kHz band-reject filter, and for practical purposes, the Sinad meter can simply connect to the external loudspeaker terminals with the volume set to a comfortable listening level.

Bear in mind that (in simplistic terms) for a changing voltage level of tone-modulated rf signal feeding into an fm receiver, the audio level of the receiver's output tone remains sensibly constant, but the noise level changes. As the level of the incoming rf signal *decreases*, the background noise *increases* to the point where the tone-signal is totally swamped to inaudibility. As the level of the incoming rf signal *increases*, the noise level *decreases* until the tone-

signal is on a noise-free background.

That latter feature is exploited in the design of this Sinad meter with its 1kHz rejection filter, to give a near zero reading on the meter in response to a noise-free tone signal from the loudspeaker, and full-scale reading for a noise-maximum inaudible-tone signal. Between these two extremes, the intelligibility of a voice signal from the loudspeaker would vary from excellent to impossible. The meter response is acceptably linear, hence can readily be calibrated in -dB intervals. By assigning a value of 0dB to full-scale deflection, and in the knowledge $-12\text{dB} = \times 0.25$, then for a meter-scale marked 0-100 μA , fsd (100 μA) = 0dB, 1/4 scale (25 μA) = 12dB, minimum-scale (approx 10 μA) = approx. 20dB. Once calibrated, only the 12dB marker is of real importance.

2.

Circuit Description

The Sinadmeter is fully automatic in use. It has one pair of input leads which connect to the receiver's loudspeaker. With its internal +12V dc regulator, it requires a single dc input of 13-15 volts at about 10 mA, which could be obtained from a 12v or 15V unregulated mains/dc adaptor or from the receiver's nominal 13.5V supply.

Figure 1 shows the circuit in block diagram form, and the full circuit detail is given in Figure 2.

The circuit consists of two audio-frequency amplifiers in cascade which feed via a 1kHz reject filter into a fullwave rectifier, to drive an indicating meter calibrated in Sinad dB. Low-cost quad op-amps are used rather than expensive Vogad ics. The first amplifier formed by IC1A and IC1B incorporates its own automatic gain control (agc) circuit. This is used to feed a constant

level of output signal voltage to the second amplifier, from an input signal level of between approximately 100mV and 3V r.m.s. which is taken directly from the loudspeaker. The second amplifier IC1C incorporates a presettable gain/level-set control RV4, which allows a degree of manual control to the overall agc. This combination of automatic and manual gain control ensures a constant preset level of audio signal being fed to the meter's precision full-wave rectifier circuit, irrespective of the level of input signal to the Sinadmeter over the given range, hence allows the meter scale to be pre-calibrated in dB for fully automatic response in use. The level-set control RV4 pre-sets the ac output-level of IC1C, hence that into the IC2D meter-amplifier/rectifier, to give a full-scale deflection on the meter for a totally-noise signal ie the receiver on open squelch with no rf input signal.

The gain-control in each of the two signal amplifiers is similar in principle, in that some of the ac output signal is rectified to be fed back as a dc control voltage to the gate of a field effect transistor. The fet then acts as a variable resistor in the negative feedback loop of the amplifier, to adjust the gain such as to hold the output signal constant at a prescribed voltage level. The gain-control of the first amplifier is fully automatic, but that of signal amplifier no.2 is manually adjustable by RV4. The dc control voltage is obtained by rectifying a sample of the ac output voltage from amplifier IC1C, using the precision f.w. rectifier IC1D and dc amplifiers IC2A, IC2B.

The amplified ac output signal from IC1C is a reasonably faithful reproduction of the audio signal from the receiver's loudspeaker terminal. It feeds via C9 into the 1kHz reject/notch filter formed around IC2C. Meter amplifier/rectifier IC2D is therefore fed only with the noise and distortion, the 1kHz tone content of the combined

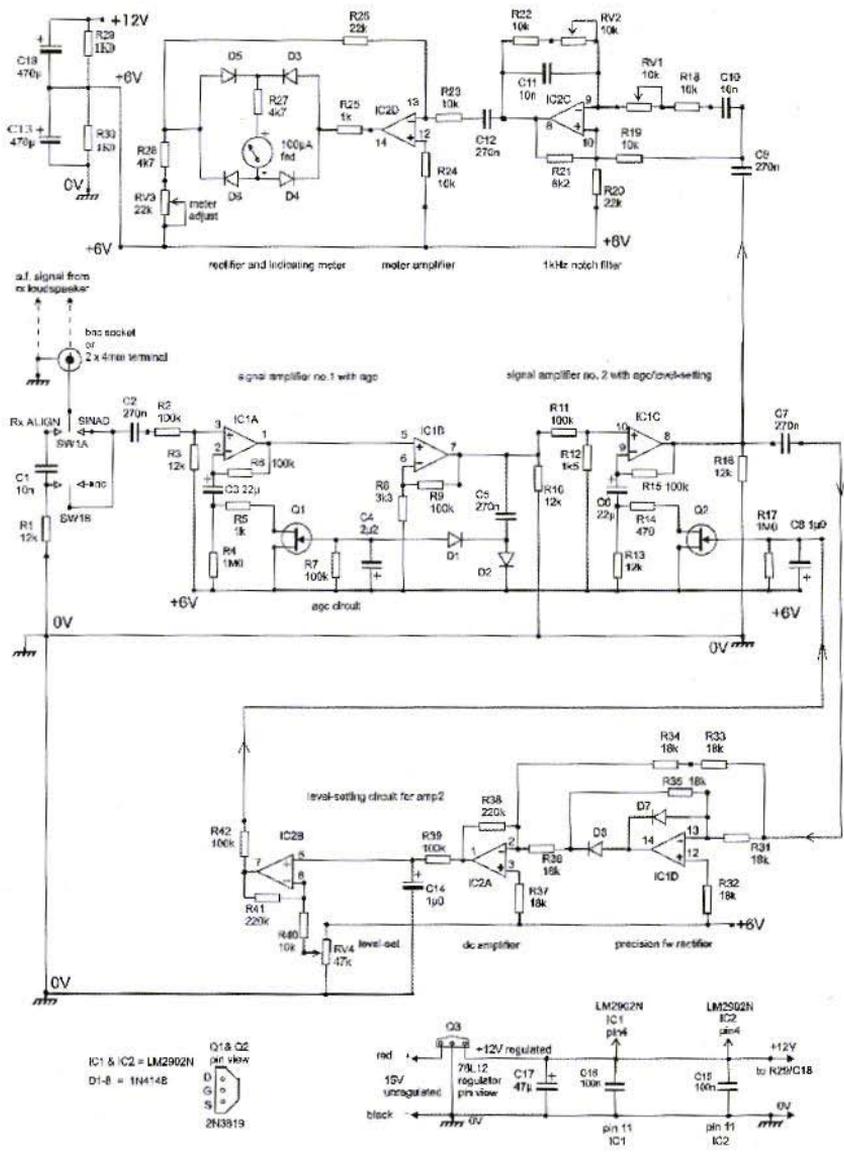


Fig 2: Sinadmeter circuit diagram

removed by the preceding reject filter. Preset resistors RV1 and RV2 tune the notch filter to 1kHz, and are simply adjusted to produce a minimum deflection on the meter, using the 1kHz tone signal from the signal generator feeding into the receiver at about 1000 μ V rms to produce a noise-free tone. Like RV4, this is a once only adjustment.

The precision full-wave rectifier based on IC2D has within its feedback loop, the rectifier-bridge diodes D3,4,5,6, and the 100 μ V dc meter, and the meter-adjust RV3, hence any change to the given component values will affect the meter deflection. RV3 was only included to allow some degree of choice on the available full-scale deflection of meter, other than the specified 100 μ A. It is optional and could be replaced by a fixed 10k resistor, because the level-set control RV4 allows adequate control of full scale deflection on the meter.

Also optional is the input low-pass filter formed by C1/R1, and associated switch SW1ab. The switch has the legends Sinad and Align. In the Sinad position, the low-pass filter is by-passed. The filter is switched into the input signal path when in the Align position. This is to smooth out any slight flicker of the meter needle as it responds to receiver noise, whilst using the meter as a tuning aid when adjusting the tuned circuits of a receiver's rf/local oscillator stages.

3.

Construction

Details are provided for assembly on either copper stripboard (Figures 3&4) or pcb (Figures 5&6). Stripboard is perhaps the most convenient choice for the home constructor. The most critical part of stripboard assembly is the cutting of copper tracks, but the track-cut template of Figure 4 should simplify that task.

The use of a proprietary track-cutting tool is recommended.

Before installing components, carefully check that each track-cut is absolute! This is best done using a powerful magnifying glass, and/or a continuity-test buzzer applied across each cut in turn. The time spent on this routine more than justifies the effort! Be sure also to check for absence of short-circuits between the 0V, +6V, +12V, and +15V rails. Now apply dc and check that the regulated +12V is available. Disconnect the dc supply, insert IC1 and IC2, then set RV1,2,3,4 to mid-position.

For housing the unit, a plastic container would be suitable. Screening of the sinad signal input leads is not necessary.

4.

Setting-up and Calibration

This requires a low-frequency sine-wave source covering 300Hz-3kHz with a variable 0-3V rms output output.

RV1,2,3,4 are at mid-position. Connect the sinad-meter's input leads to the low-frequency source, with its output set to exactly 1kHz at 1volt rms level. Reconnect the power supply to the sinad-meter, and observe some movement of the meter. Adjust RV1 and RV2 successively for minimum deflection on the meter, which should be near the 10 μ A reading. This proves that the 1kHz reject filter is functional. Swing the frequency either side of 1kHz, say to 300Hz and 3kHz, and the meter should rise towards full-scale deflection. With the frequency at 2kHz, adjust level-setting control RV4 to give exactly full-scale deflection ie 100 μ A. Vary the output level of the low-frequency source from zero to about 3V, and observe that the meter deflection remains sensibly constant for input variation from about 100mV to at least 2V rms.

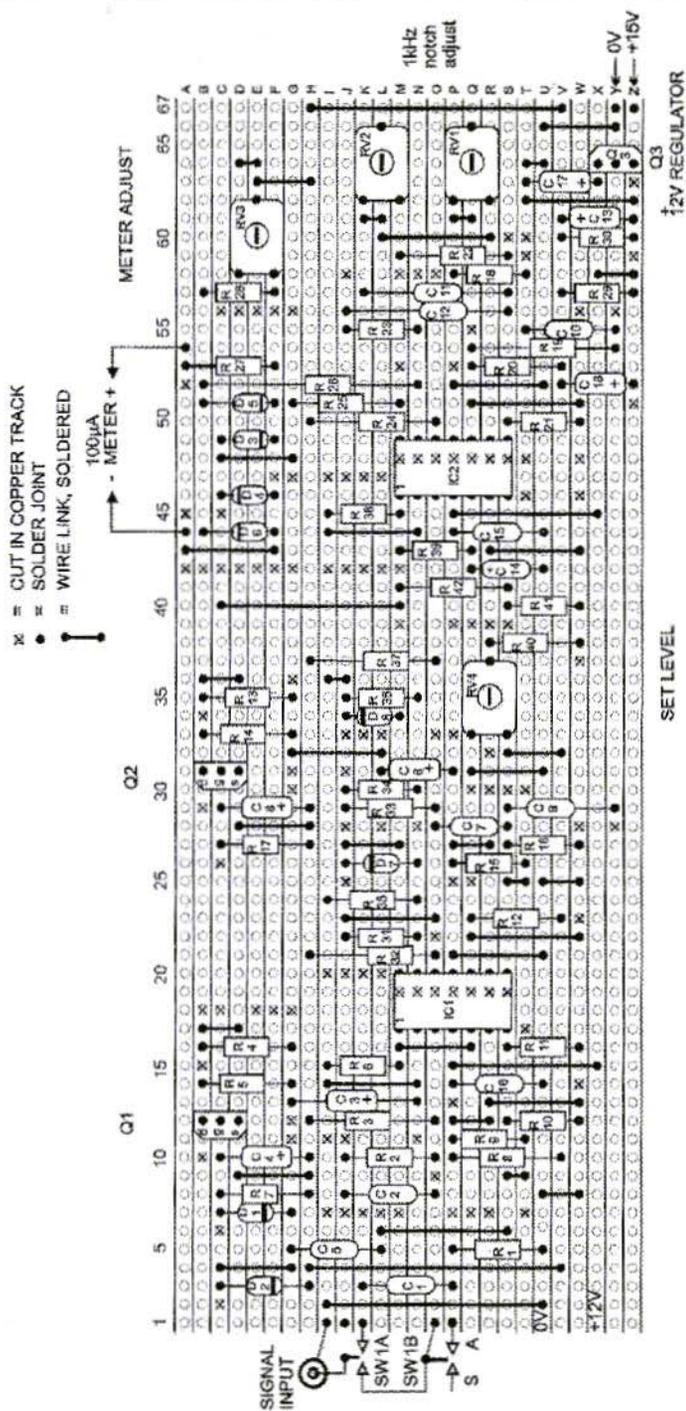


Fig 3: Smadmeter stripboard layout for wiring and components (non copper side)

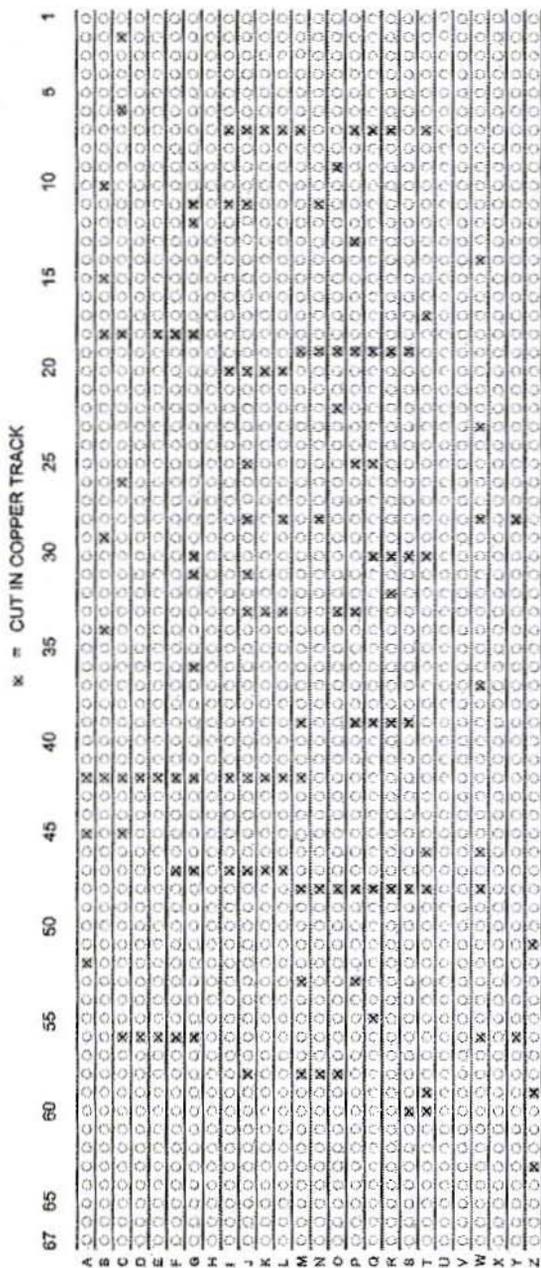


Fig 4: Sinadmeter stripboard track cutting template (copper side)

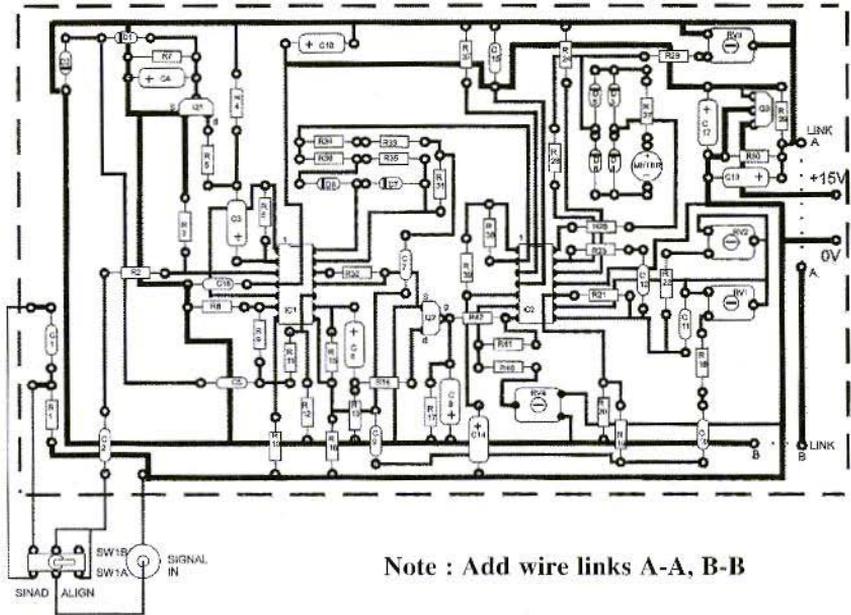


Fig 5: Sinadmeter PCB component side

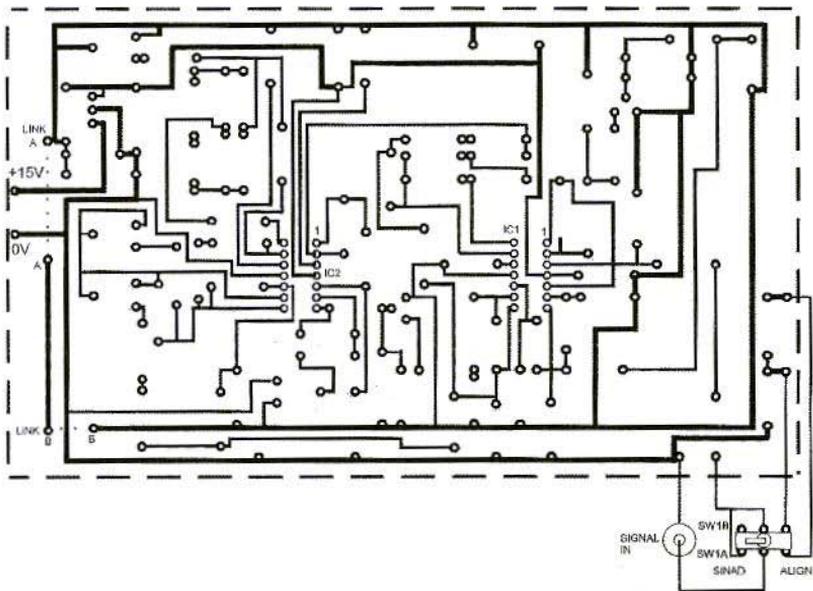


Fig 6: Siandmeter PCB copper side

constant for input variation from about 100mV to at least 2V rms.

Finally, set up an fm receiver with an fm signal source connected to its antenna port. Set the signal generator to the receive frequency, with 1kHz tone-modulation at 3kHz peak deviation, and with its output level at about 1000 μ V rms.

Adjust the receiver volume control to give a comfortable listening level for the 1kHz output tone, which will be on a completely noise-free background.

Connect the sinad-meter's input leads across the receiver's loudspeaker terminals, with the black lead to receiver chassis-potential. The meter should be at or about minimum deflection. Re-adjust RV1 and RV2 to optimise the sinad-meters reject filter to the rf signal-generator's 1kHz modulating-tone.

Disconnect the rf signal source from the receiver, and open the squelch to produce full noise output from the loud-speaker. Readjust RV4 to give full-scale deflection of exactly 100 μ A. This should remain sensibly constant when the volume control is varied.

Meter-set control RV3 is still at mid-position, and there it should remain. As stated earlier, it might only ever be needed if a meter of other than 100 μ A is used. Hence, RV1,2,3,4 are now finally and forever set, and the sinad-meter becomes fully automatic in use.

The only thing left to do, is to mark 12db on the meter-glass at its one-quarter full scale point ie at the 25 μ A position for a 100 μ A fsd meter. This is the 12dB Sinad Sensitivity mark against which all future receiver checks will be made. Remember, the receiver's 12db Sinad Sensitivity is the microvolts rms level from the signal generator that causes the sinad-meter to read 12dB, typically 0.5 μ V or thereabouts.

Parts List

Qty	Item
2	Quad-Opamp ic type LM2902N
8	Diode type 1N4148
2	FET type 2N3819
1	Voltage regulator 12V/100mA type 78L12

Preset resistor, carbon, min., horiz., pin spacing 10mm x 5mm

2	10k	RV1,2
1	22k	RV3
1	47k	RV4

Capacitor, electrolytic, min.axial 16V

2	1.0 μ F	C8,14
1	2.2 μ F	C4
2	22 μ F	C3,6
1	47 μ F	C17
2	470 μ F	C13,18

Capacitor, min.dipped, mylar/polyester film 100V dc

3	10nF	C1,10,11
2	100nF	C15,16
5	270nF	C2,5,7,9,11

Resistor, metal film, 0.25W

1	470R	R4
4	1k0	R5,25,29,30
1	1k5	R12
1	3k3	R8
2	4k7	R27,28
1	6k2	R21
6	10k	R18,19,22,23,24,40
5	12k	R1,3,10,13,16
7	18k	R31,32,33,34,35,36,37



- | | | | | |
|---|----------------------------------------------------------------------------|----------------------|---|-------------------------------------------------------------------|
| 2 | 22k | R20,26 | | miniature, panel mounting |
| 8 | 100k | R2,6,7,9,11,15,39,42 | 1 | Stripboard, 0.1inch, copper, srpb, 170 x 75mm, 26 rows x 67 holes |
| 2 | 220k | R38,41 | | |
| 2 | 1M0 | R4,17 | 1 | Stripboard track-cutting tool |
| 2 | 14pin dil socket | | 1 | Low-cost plastic box with lid, eg Maplin BZ75S |
| 2 | Terminal post, 4mm, small. One black, one yellow | | | |
| 1 | Panel Meter, 100µA dc fsd, calibrated 0-10 or 0-100, eg Maplin 2inch RW92A | | | |
| 1 | Switch, dpdt toggle sub- | | | |



The UK Six Metre Group

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With over 1000 members world-wide, the UK Six Metre Group is the world's largest organisation devoted to 50MHz. The ambition of the group, through the medium of its 60-page quarterly newsletter 'Six News' and through its web site www.uksmg.org, is to provide the best information available on all aspects of the band: including DX news and reports, beacon news, propagation & technical articles, six-metre equipment reviews, DXpedition news and technical articles.

Why not join the UKSMG and give us a try? For more information, contact the secretary Iain Philipps G0RDI, 24 Acres End, Amersham, Buckinghamshire HP7 9DZ, UK or visit the web site.



Hermann Hagn, DK8CI

Radio Astronomy Terms Explained

Those with an interest in radio-astronomy are always asking what the main parameters of an operational radio astronomy installation are. The most important parameters and formulae are listed below.

1.

The antenna

- a) Diameter of parabola: D (m)
 b) Depth of parabola: c (m)
 c) The focal length (focus):

$$f = \frac{D^2}{16c} \text{ (m)}$$

- d) Antenna area:

$$A_{geo} = \frac{D^2\pi}{4} \text{ (m}^2\text{)}$$

- e) Antenna area efficiency η : (normally lies between 0.45 and 0.65). 0.55 can be taken as a good average. Determined by measuring a known radio source.
 f) Effective antenna area:

$$A_{eff} = A_{geo} \cdot \eta \text{ (m}^2\text{)}$$

g) The aperture angle, as in optics, depends on the dimensions (diameter) of the antenna and the wavelength. It gives the angle at which half the power is measured between the points, and is approximately:

$$\text{Angle in degrees} = 70 \cdot \frac{\lambda}{D}$$

2.

Receiver

Noise factor of receiver: in decibels (F) or kT_0

$$F_{(db)} = 10 \cdot \log kT_0$$

or noise temperature of receiver: TE

$$F_{(db)} = 10 \cdot \log \frac{290 + T_E}{290}$$

The value of F(db), kT_0 or TE is determined through measurement.



3.

Flux of a radio source

$$\text{Flux} S \left(\frac{W}{m^2 Hz} \right)$$

Since the flux from radio sources is very low, the value inserted is

$$10^{-26} \frac{W}{m^2 Hz} = 1 \text{ Jansky}$$

The flux parameter SFU (Solar Flux Unit) is used for solar measurements:

$$1 \text{ SFU} = 10^{-22} \frac{W}{m^2 Hz}$$

4.

System noise temperature

In measurements, the noise of the receiver (F, kT₀, T_E) is supplemented by a noise fraction which is caused by the ohmic losses of the antenna, the secondary lobes of the antenna, the radiation from the atmosphere (in front of the antenna), unresolved radio sources at greater distances and the 3k background radiation.

$$T_{\text{system}} = T_E + T_H$$

T_H is the sum of the previously calculated radiation temperature contributions and can be set as approximately 35 K in the range of 1-10 GHz. If a radio source is measured, the system temperature is supplemented by the radiation temperature of the radio source. The signal strength is measured in decibels (db) or in Kelvins (T), if the installation has been calibrated in temperatures, for example:

$$\text{Signal strength (db)} = 10 \cdot \log \frac{T_{\text{sys}} + T_{\text{Aquelle}}}{T_{\text{sys}}}$$

5.

Important relationships

$$S = 2 \frac{k \cdot T_{\text{Aquelle}}}{A_{\text{eff}}}$$

$$k = 1.38 \cdot 10^{-23} \frac{J}{K}$$

(Boltzmann constants)

Factor 2 results from the fact that an unpolarised radiator is being measured with a polarised antenna.

Conversion of effective antenna area into isotropic antenna gain.

$$A_{\text{eff}} = \frac{\lambda^2}{4\pi} G_{\text{iso}}$$

$$G_{\text{iso}} = \frac{4\pi}{\lambda^2} A_{\text{eff}}$$

$$G_{\text{iso}} = G_{\text{Dipol}} + 2.3 \text{ db}$$

Relationship between isotropic antenna gain and system noise temperature:

$$\frac{G_{\text{iso}}}{T_{\text{sys}}} = 0.385 \cdot 10^6 \frac{(10^{\text{db}/10} - 1) \cdot f^2 (\text{GHz})}{S_{\text{Jy}}}$$

Temperature of a radio source:

$$S = 7.35 \cdot T \cdot D^2 \cdot f^2 \left[\frac{W}{m^2 Hz} \cdot 10^{-26} \right]$$

$$T = \frac{S}{7.35 \cdot D^2 \cdot f^2} [K]$$



derived from Rayleigh/Jeans with:

$$S \left[\frac{W}{m^2 Hz} \cdot 10^{-26} \right], T [K], D$$

6.

Literature references

[1] VHF Communications, 2/95 Pp.

112-122, H. Hagn, Reception parameter measurements with radio stars

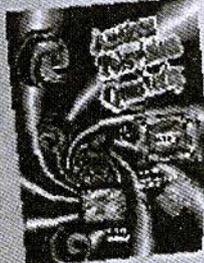
[2] The ARRL UHF/Microwave Experimenters Manual, 7-60; David B. Schaffer, Microwave System Calibration using the Sun and Moon

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Great articles on :

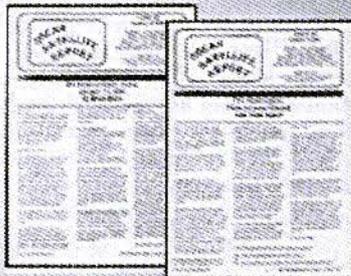
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Hermann Hagn, DK8CI

Astronomical Observations at The German Museum, Munich

Measurement of radio radiation from the Sun during the solar eclipse on 11. 08. 1999 using the radio telescope of the German Museum with an 18 cm wavelength.

1.

Technical data for "Würzburg Giant" radio telescope

<i>Built in:</i>	1943
<i>Application:</i>	radio measurement apparatus until 1945 (radar), from 1946 used as radio telescope in the Netherlands
<i>Weight:</i>	9.5 t
<i>Mirror diameter:</i>	7.5 m
<i>Focus:</i>	1.71 m
<i>Aperture angle:</i>	1.8° (with 18 cm wavelength)
<i>Positioning accuracy:</i>	0.1°
<i>Receiver:</i>	Total Power Radiometer

Antenna exciter: Circular wave guide

Reception frequency: 1680 MHz

Receiver band width: 3 MHz

Effective antenna area: 22 m²

System noise temperature: 90 K (at zenith)

Limiting sensitivity: 10 Jy

(1 Jy = 10⁻²⁶ W/m²/Hz)

2.

Measurement

The signal increase is always measured against the system noise temperature for the declination in question. On 11th August 1999, with a +17° declination, 97.6 K was measured. The mean signal increase for the Moon (at perigee) on 8th 9th and 10th August 1999 was measured at 0.41 dB, which gives a flux of 1212 Jansky.

Signal increase for the Sun on 11th August 1999 before the solar eclipse: 20.25 dB against system noise temperature; which gives us an "antenna temperature" of 10241 K, producing a flux



Fig 1: The "Würzburg Giant"

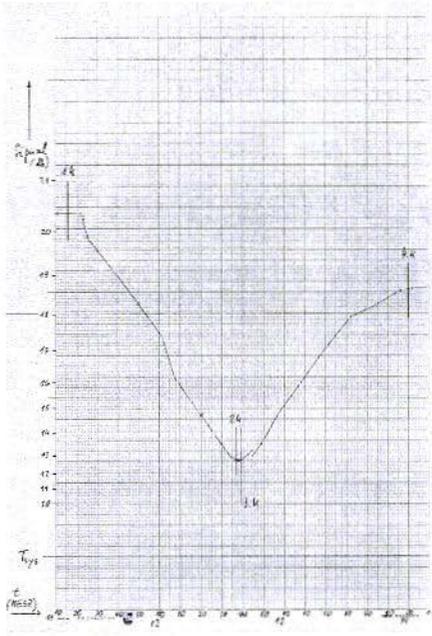


Fig 2 : Signal decrease by Moon cover

of $1.285 \cdot 10^6$ Jansky.

The signal decrease due to covering by the Moon was 7.6 dB, i.e. the signal increase was 12.65 dB against the system noise temperature (Fig. 2). After subtracting the lunar flux, we obtain $0.212 \cdot 10^6$ Jy, i.e. 16.5 % of the total flux.

Still on this subject, there is an interesting early measurement from 12th September, 1950 at 10.7 cm (2800 MHz) by HAGEN, J.P., F.T. HADDOCK and G. REBER: Sky and Telescope 10, 111 (1951) who measured a radio intensity at the minimum (during covering by the Moon) of 17 %.

Ref.: A. UNSÖLD, Physics of stellar atmospheres (see Fig. 4).

Hermann Hagn, Radio-astronomical Observation Group, Ludwig Wagner, German Museum, Munich; Richard Jäger

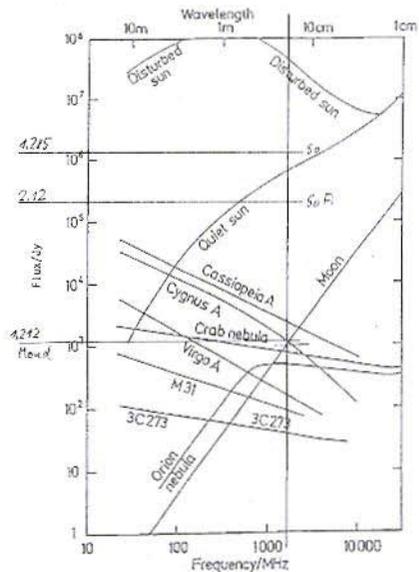


Fig 3 : Flux as a Function of Frequency

XXII. 148. Solar Eclipses

...Sun, about 1948; so here the slowly variable radiation of the plagues not marked by spots has not yet been ruled out. Fig. 237 shows the Cambridge edge darkening measurements for $\lambda 7.9$ m. to 60 cm. At 21 cm and 3.2 cm, the brightness of the solar disc varies only a little, up to about $\Omega \approx 0.7$; we observe a bright ring here at the edge.

An important check on the interferometric edge darkening measurements is permitted by observations of solar eclipses (Table 102). Since in the range of the longer wavelengths the radio Sun is considerably larger than the optical solar disc or the Moon, partial eclipses with a considerable degree of cover are just as useful as total eclipses. For longer wavelengths, the occultation curves are strongly distorted by the disturbed radiation from plagues which may be present. Thus unambiguous results concerning the radiation from the quiet Sun can be expected only at the time of a sunspot minimum and / or in the centimetre wavelengths. The correspondence with the interferometer measurements should be satisfactory here, as a whole.

Date	Amount of maximum phase (in C diameters)	? (cm)	Frequency MHz	Radio intensity at minimum	Ref.
1945 July 9 th	Partial 58% surface	1.25	24000	55%	(1)
1945 July 9 th	Partial	3.2	9400	58%	(2)
1947 May 20 th	1.039	3.2	9400	4%	(3)
1950 Sept. 12 th	1.018	3.2	9400	8%	(4)
1951 Sept. 1 st	0.975	3.2	9350	16%	(5)
1948 Nov. 1 st	0.63	10.0	3000	46%	(6)
1946 Nov. 23 rd	0.59	10.7	2800	55%	(7)
1950 Sept. 12 th	1.018	10.7	2800	17%	(4)
1948 Nov. 1 st	0.63	50	600	51%	(8)
	0.72	50	600	45%	(8)
	0.85	50	600	30%	(8)
1950 Sept. 12 th	1.018	64.5	465	28%	(4)
1951 Sept. 1 st	0.975	178	169	47%	(5)

(1) DICKE, R. H., & R. BERINGER [2236]
 (2) SANDER, K. F.: Nature (London) 159, 506 (1947)
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 (6) PIDDINGTON, J. H. & J. V. HINDMAN: Austral. J. of Sci. Res. (A) 2, 524 (1949)
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 (8) CHRISTIANSEN, W. N., D. E. YABSLEY & B. Y. MILLS: Austral. J. Sci. Res. (A) 2, 506 (1949)

Table 102 - Radio observations of solar eclipses

B) Theory Of Thermal Radio-Frequency Radiation From Quiet Sun

Part I: Decimetre and centimetre range; linear propagation

The thermal emission of completely ionised gases in the radio frequency range, as can easily be seen after Chapter 7, is practically free-free radiation. For the wavelengths in question, we can calculate the radiation emission of an electron (charge e , mass m), which is deflected from an ion with a $+ Ze$ charge on a hyperbolic orbit, in accordance with Maxwells theory, following the well-known work of H.A.Kramers [721]. In this connection, we may once again permit ourselves to simplify matters considerably; since the lower frequencies, quite predominantly at small deflections of the electron...

Fig 4: This text is from a page that was reproduced from a German magazine. It has been translated into English so loses some of its significance - Editor



Bernd Kaa, DG4RBF

Universal Micro-Controller Board, Uniboard C501

This article was originally published in UKW Berichte 3/98 but not in VHF Communications. Following many requests it has been reproduced to help constructors where it has been used.

The forward march of the micro-controller in all areas of electronics shows no signs of slowing down, and so there are more and more electronic circuits which have to be controlled by a micro-controller.

The options that are available using a micro-controller are very numerous, and therefore they can be used to develop high-quality equipment, even in DIY work.

So many voices have been calling for a universal micro-controller board that would be suitable, not only for a specific project, but for many tasks. For requirements needing high-powered computing, the board can be driven with a clock frequency of up to 38 MHz.

Mind you, it should be taken into account that writing the software required will take a relatively long time.

1.

Introduction

Naturally, the description given here of a universal micro-controller board, Uni-

board C501, has topical relevance. There are several projects under way for which such a controller board is required:

- 4-band short-wave SSB/CW transceiver with DD (W Schneider, DJ8ES)
- Double frequency counter up to 4,000 MHz (B Kaa, DG4RBF)
- 144 MHz SSB transceiver with DDS for transverters up to 47 GHz (W Schneider, DJ8ES)
- 1 65 MHz short-wave synthesiser in 1-Hz steps (B Kaa, DG4RBF)

A universal micro-controller board was developed based on the 80C31 micro-controller.

In the development of the micro-controller board, care was taken to ensure that it was not tailored to only one project but was universally applicable (this also applied to I/O ports). Thus some components and expansions are provided for on the printed circuit board which can be fitted or omitted, depending on the application.

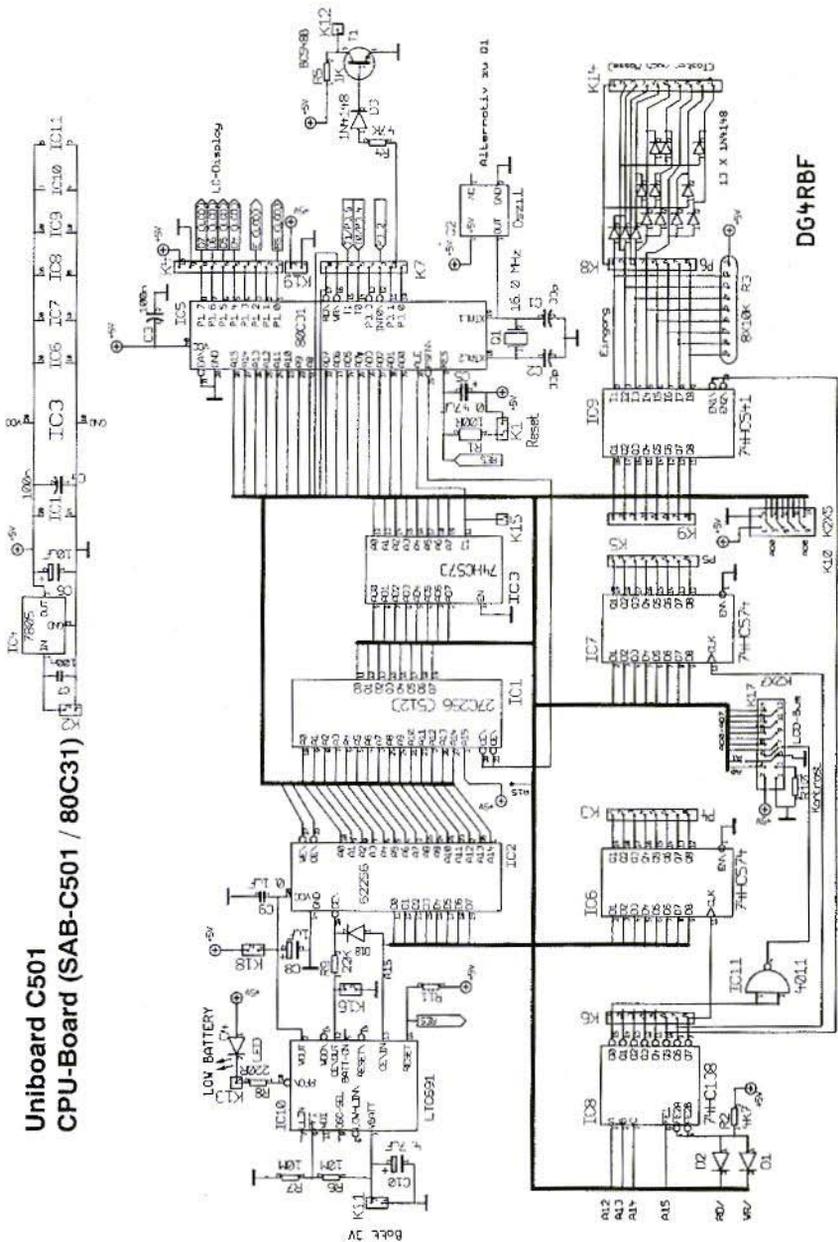


Fig 1: Circuit diagram of Micro-Controller board, Uniboard C501



2.

Circuit Description

2.1. The processor board

The description below refers to the maximum expansion of the Uniboard C501, which is shown in the circuit diagram in Fig. 1.

The board has two output expansion ports, P4 and P5, which were each created using a 74HC574. These outputs can be loaded with up to 20 mA, and are thus suitable for controlling LEDs or small relays directly.

An input expansion port (P6) is fitted with a 74HC541. The inputs are wired with pull-up resistors, so that they can be used as low-active inputs. In addition to this input expansion port, the option for binary coding of the lower 4 bits (P6.0-P6.3) with diodes is also provided for on the printed circuit board. Thus these 4 bits make 10 inputs available at K14. Here, for example, a keyboard (from 0-9) could be connected up, with keying to earth.

Up to 225 input pins could be created in this manner, though the number of diodes and the space required for this would be rather high!

An address decoder present on the printed circuit board (74HC 138) ensures that the data is correctly decoded for the expansion ports. The outputs Q1 to Q4 (which are on K6) are still free and can be used for optional tasks.

An external RAM of up to 256 bytes can be used as a memory. The address line A15 can be used to ensure that there is no address conflict between an external RAM and the expansion port modules.

If the system is operated without a reset IC, diode D18, with resistor R9, should

be fitted to inhibit the RAM if the address decoded includes the A15 line. If a reset IC is used, diode D18 can be omitted, since the A15 address line then directly controls CE/OUT.

The necessary operating software, which has to be developed in accordance with the application, is stored in a type 27C256 or type 27C512 EPROM, as required. There are solder connections on the PCB to select the type of EPROM used. For a type 27C512 EPROM, the line from pin 1 is separated from the + 5 V and connected to the A15 line. For a type 27C128 EPROM, the pin 27 line is separated from the A14 line and connected to + 5 V.

The data bus is available on the K9 pin strip as a single-row connection and at K10 as a double-row connection - for additional expansions and additional boards, the address decoder outputs which remain free. The double-row connection (2 x 5 pins) makes it easy to wire up with pin-and-socket connectors using the IDC connectors.

The ALE signal is available at K15 for future tasks.

The option is also available to use a reset IC (LTC691) on the printed circuit board, which would also take over the clearing the RAM and the data retention. This gives the option for applications which must hold important data in the RAM when switched-off. Provision made for backup battery monitoring.

If the system is operated without a reset IC, K16 and K18 are short-circuited. This causes pin 20 (CE) of the RAM to be earthed and actively switched. Pin 28 of the RAM is connected to the supply voltage.

2.2. The display

The hardware offers two variants for

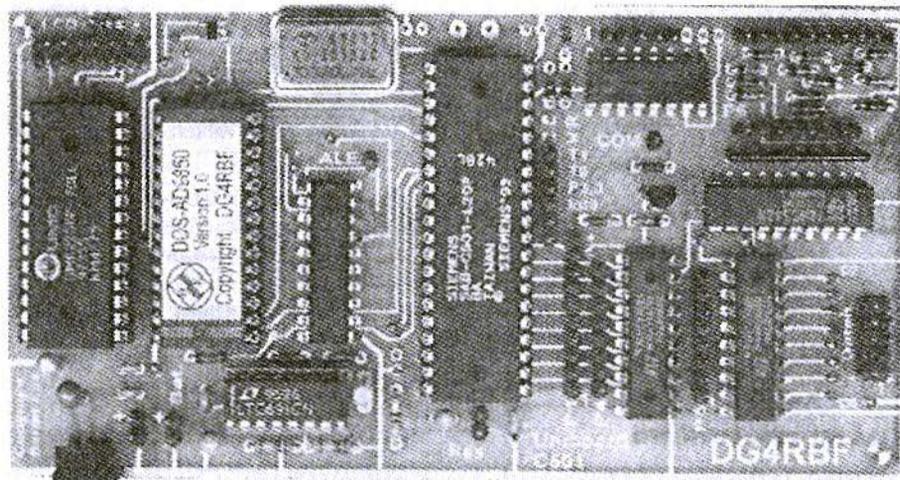


Fig 2: The completed Micro-Controller board

connecting up an LCD dot matrix display.

The display can be operated, for example, in 4-bit mode using port 1, or it is connected directly to the data bus. The 14-pin K17 (LCD) pin-and-socket connector is configured in such a way that a display with a double-row connection can be connected directly with a pin-and-socket connector using an IDC connector.

The enable signal inverter required for the second mode of operation is created using a simple gate (CMOS 4011). The display can be controlled in connection with Q0 of the address decoder, together with the A0/A1 line. The choice of display connection depends on the application and the software.

Pin assignment of K17 for connection to data bus:

Pin 1 = earth

Pin 2 = + 5 V

Pin 3 = contrast (R10)

Pin 4 = RS (A0)

Pin 5 = RW (GND/A1)

Pin 6 = enable (Q0)

Pin 7 pin 14 = D0 D7

The LCD display is connected in accordance with Table 1.

Fig. 3 shows the pin assignment of port 1 (K4 + K19) for the LCD display connection. In order to keep the connection compatible with older projects, the option also exists to use pins 9 and 10

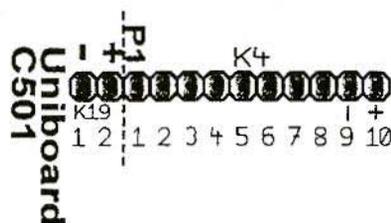


Fig 3: Connections on K4 / K19

**Table 1 : Display connections on port 1**

Display Pin	Meaning	Pin on K4	Port	Comment
Pin 1	Earth	9 (or 1 on K19)	-	GND
Pin 2	+5v	10 (or 2 on K19)	-	+5v
Pin 3	Contrast	-	-	Depends on module
Pin 4	RS	1	P1.0	Register select
Pin 5	RW	2	P1.1	Read write
Pin 6	E	3	P1.2	Enable
Pin 7	D0	-	-	GND
Pin 8	D1	-	-	GND
Pin 9	D2	-	-	GND
Pin 10	D3	-	-	GND
Pin 11	D4	5	P1.4	Data
Pin 12	D5	6	P1.5	Data
Pin 13	D6	7	P1.6	Data
Pin 14	D7	8	P1.7	Data

at K4 for the displays power supply. However, positive and earth at K19 should be used for new projects, since this makes a simpler (parallel) cable feed to the display possible.

2.3. The processor

For normal applications, an 80C31 or 80C32 can be used which is manufactured for clock frequencies up to 16 MHz. For applications requiring high-powered computing, processors can be used of the SAB-C501LXXP type, which operate with a clock frequency of up to 38 MHz.

The SAB-C501 micro-controller from Siemens is an 80C32-compatible controller in a slightly revised format. What is very interesting here is that it is manufactured for various clock frequencies up to 40 MHz.

SAB-C501-LP = 12 MHz

SAB-C501-L20P = 20 MHz

SAB-C501-L24P = 24 MHz

SAB-C501-L40P = 40 MHz

As experiments have shown, the frequency values shown still have a substantial safety margin, since several C501-L20Ps have been tested and all functioned right up to the functioning limit of the micro-controller board (38 MHz). If we assume the same level of quality, the 20-MHz types should also be operated at 25 to 30 MHz.

The option was provided on the CPU printed circuit board to use either a crystal or an external TTL oscillator. However, micro-controller crystals above 20 MHz oscillate only at their fundamental frequency hence the second option. If a TTL oscillator is used, the crystal (Q1) is omitted, together with the two 33 pF capacitors.

A serial output (COM) is also provided for. If the components required for this are fitted, the serial data are available at pin K12.



3.

Micro-Controller Assembly Instructions

IC sockets are used for processors, RAMs and EPROMs. If necessary, the reset IC is also given a socket. The other ICs are simply soldered into the PCB.

Pin 1 of the pin strips is always identified by a (1) on the micro-controller printed circuit board.

The resistor R11, shown as a dotted line, is not used. The latter is not needed unless the reset IC used is not type LTC 691 but type MAX 691 A.

4.

Trouble-Shooting

4.1. Micro-controller

- Check all soldering work with a magnifying glass! (Its usually just a simple soldering error!)

- Are all ICs and diodes inserted the correct way round?
- Is the contrast setting of the LC display correct?
- In most displays, the connection for contrast setting is earthed directly or through a resistor. But there are also displays that require negative voltage to display the text. As a minimum, the bar of the first line must be visible.
- Is the reset capacitor (C5) in order?
- When operating without a reset IC, short-circuit K16 and K18. Should the reset IC be used, then both plug-in links remain open.
- Check the following points with an oscilloscope:
 - Is the ALE signal (square wave signal) present at K15?
 - Do the address and data bus lines have TTL levels? (Intermediate values indicate a short-circuit!)

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As used in many articles in VHF Communications



Carl G. Lodström, EE. SM6MOM & KQ6AX

The Noble Art of De-coupling

1.

Introduction

Much of the electronics built today will require de-coupling of various points for various frequencies. What first comes to mind may be the V+ of IC's. If the IC is an aggressive one, providing fast transitions with a lot of drive capability, the problem is no longer a simple one.

Nevertheless it is treated as if it were! "Ah, let's toss in a 10000 'puff' there! Wait a moment, better make it a 1000 puff and a 'point-one'!" That is: a 1 nF and a 0.1 μ F capacitor.

For one thing, 1 nF or 0.1 μ F will make no difference (to RFI) above 500 MHz and almost none at 200 MHz! The capacitors are already well above their self-resonance frequencies.

The readers of VHF Communications do not have to be reminded about what will happen as clock frequencies are reaching the GHz mark! Even if the signals are not perfectly square, it is reasonable to assume that they are not sinusoidal either. Rise times may be of the order of 100 ~ 200 ps. It is thus proper to consider frequencies up to at least 2 GHz. (As things are going, do you too have the feeling that someone will laugh at this in a few years?!)

2.

Elements

My measurements have shown a 1206 capacitor to have ~ 1.8 nH and a 0805 seems to have ~ 1.5 nH. In some other applications it seemed like the 0805 had ~ 1 nH. Right or wrong, we can probably agree that it will have at least 1 nH and I will use it in the modelling. The Q of the inductance is probably not stellar, let me guess at 30. The Q of the capacitance may well be 50.

The source impedance of the transient generator is probably low. Let us use 10 Ω for this and 100 Ω for the other end, the "load", going to the power supply. We will treat the circuit like a filter, considering S11, S21, and S22 over frequencies to 3 GHz.

3.

Models

For the modelling I have used the Eagleware Superstar program, version 5.2

The first model we may want to take a look at is the one with a single de-coupling capacitor on the V+ pin of the IC (Fig 1). It is imagined here as

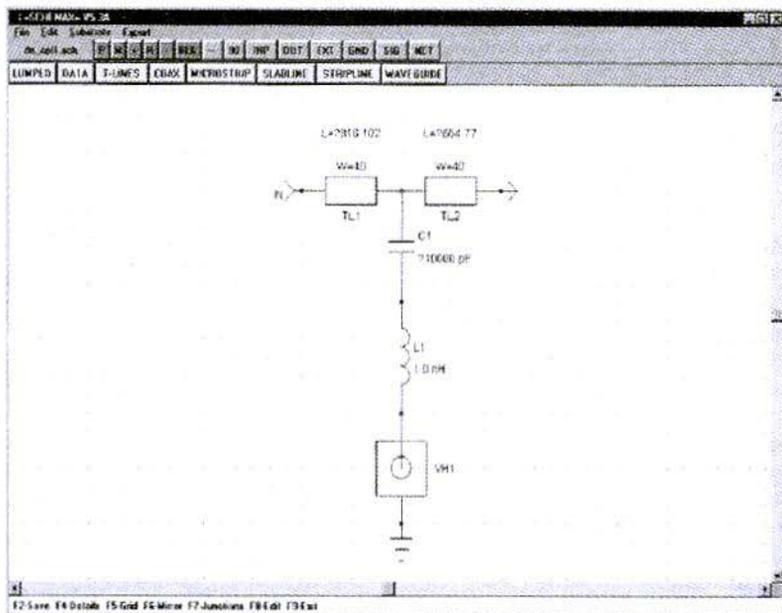


Fig 1: Circuit of the first model

“IN>”

The model, showing a piece of 40 mils (1 mm) wide track to the capacitor, another track to the 100Ω load. Their lengths are mostly inconsequential. The capacitor is pictured with its inductance and it is grounded through a via hole of 24 mils diameter with 1.5 mils metal. The substrate is 32 mils Rogers 4003, with $\epsilon_r=3.38$ nominally, but various substrates would not make any great difference here.

4.

Simulations

On the left diagram (Fig 2) are S11 and S22 with 0 ~ 1 dB vertically. On the right is S21 on a 0 ~ 100 dB scale, showing the efficiency of the de-coupling. Four marker frequencies are set under each graph and their “dB” results

can be read along the bottom.

One set of graphs is dashed and one set is solid. They represent a capacitance of 100 pF and 10 nF respective. Sure, there is a nice attenuation from ~500 MHz to ~30 MHz, but the rest is nothing to write home about. About 20 dB attenuation appears to be the rule, and it is not much of a de-coupling on an intense source of RF.

It appears likely that several stages, compounded, of de-coupling may do better. Let us look at a two stage solution and let SuperStar try to optimise it for -60 dB from 100 ~ 1000 MHz and -40 dB from 50 ~ 2000! (Fig 3)

Well, now we are getting somewhere! The attenuation is below 30 dB up to 1420 MHz! Two dips are visible, corresponding to the self-resonance of the capacitors. The program decided that they better be 587 and 3418 pF respective, with 802 mils from the left and a

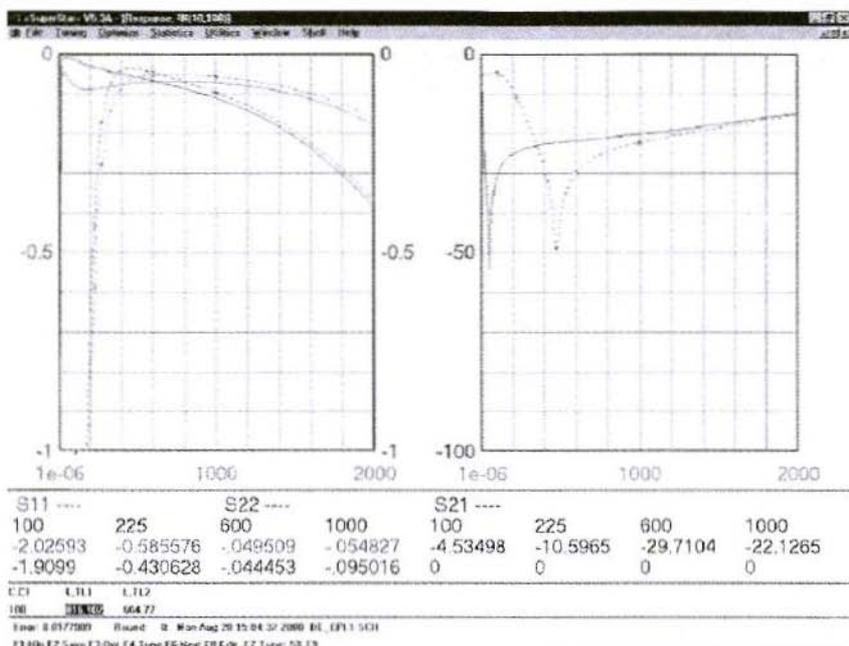


Fig 2: First simulation results

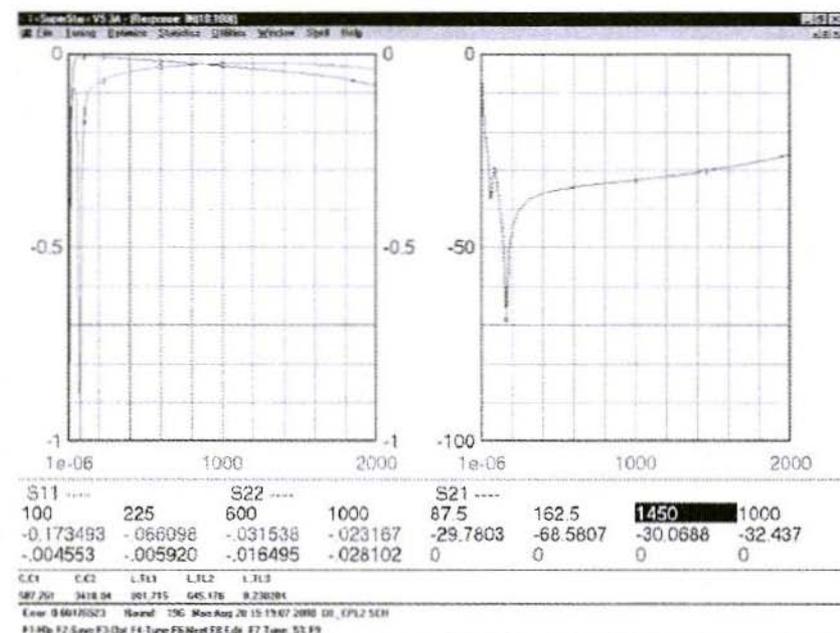


Fig 3: Simulation of a two stage circuit

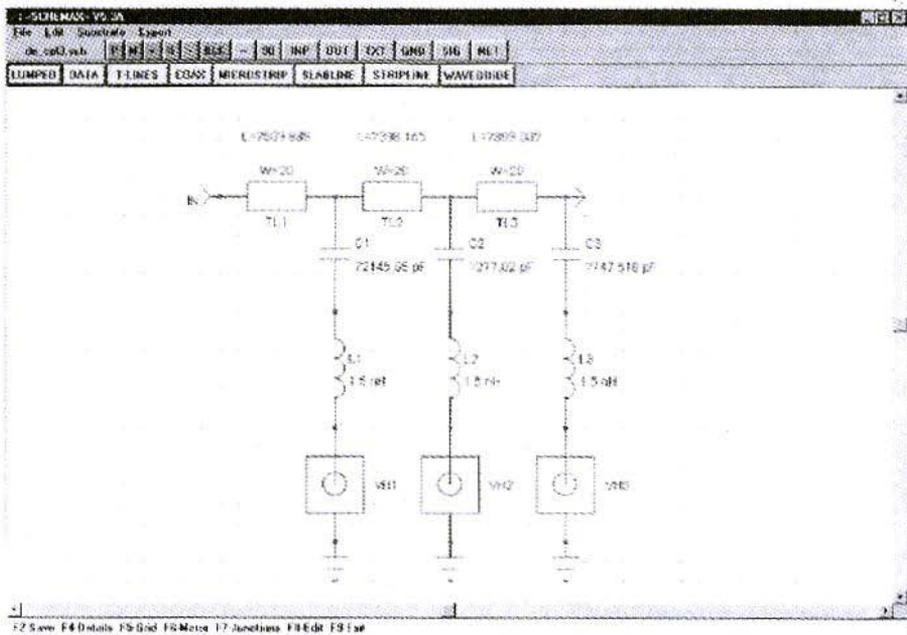


Fig 4: Circuit of expanded circuit

645 mils long lines in-between. The third line, at the right end, to V+, is best when short.

I expanded the circuit to three capacitors, with a 0.5 mm line before each, ran SuperStar on it again for the following configuration (Fig 4)

Indeed, this net shows a very good performance with 50dB from 150 ~ 400 MHz and 40 dB from 80 ~ 2212 MHz. Notice that the max frequency is changed from 2 to 3 GHz! (Fig 5)

On the S11 numbers can we see how pin 8 on the IC finds a near total reflex (<0.1 dB Return Loss) on almost all frequencies although the phase rotates via inductive to near open. (Fig 6)

One possible practical realisation of this net with a Small Outline 8-pin IC and 0805 components.

5.

Conclusions

Use of self-resonant capacitors for de-coupling has been known for most of the radio era. I have a book (that I cannot find just now) from the 1930's. In it is a graph of the "ideal length of de-coupling capacitors vs. frequency". From the values, clearly some 5 ~ 10 nH per cm was anticipated. I am not so sure it was considered in old IF strips and other circuits, but some designers surely knew about it.

As frequencies move up in the VHF and UHF range, so do the number of capacitor values one do no longer have to consider using! The 1 nF capacitor in a 1206 surface mount package will become self resonant at 130 MHz. In a leaded package it will resonate at a much lower frequency. 1 nF can possi-



Dipl. Ing. Wolfgang Borschel, DK2DO

Is Silver-Plating Worth While in RF Applications ?

The question is being asked more and more frequently: Has silver-plating my component actually done any good, or have I gone to all that trouble for nothing? You can't always tell from the measurements, but the scientific proof is unambiguous. The following article is intended to show under what conditions it makes sense to silver-plate material in radio-frequency technology.

1. Principles

Coaxial cable and resonant cavities are the radio-frequency conductors used in communications engineering in which powerful radio-frequency currents can flow. These currents heat the sheath and the internal conductor of coaxial cables and the resonance chamber of resonant cavities. Coaxial plug-and-socket connections come under particular stress. If the radio frequency field is contained, as is always the case in resonant cavities and coaxial cables, it tries to penetrate into or out of the screening material. This displacement increases as the frequency rises. This also influences the phase position of the surface currents in relation to the currents in the

conductor interior, thus an inductive reactance component is generated. The permeability number, μ_r , of the material used thus becomes the second decisive influencing variable for radio-frequency current displacement. The effect is described as the skin effect. It means that the current density and current distribution in the metal is not constant.

2. Material characteristics

In radio-frequency technology only diamagnetic metals ($\mu_r < 1$; e.g. copper) and paramagnetic metals ($\mu_r \sim 1$; e.g. aluminium) are used. Brass is a particularly popular construction material in amateur radio circles, but is controversial because of the zinc component. When the zinc component and other metal fractions are high, these alloys almost tend towards ferro-magnetism. Ferro-magnetic metals such as tinplate and nickel should, not be used, at least not where powerful radio-frequency currents are flowing ($\mu_r > 1$; tinplate: $\mu_r \sim 10$ to 2000; it is important to know that at high frequencies the μ_r value of the tinplate becomes lower!).

Good radio-frequency characteristics are

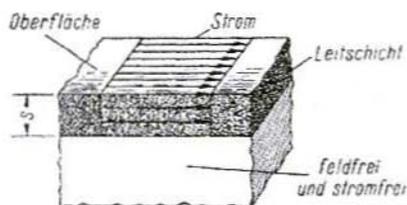


Fig 1: High conductance coating carrying the RF field current with no current flow in the underlying material

incorrectly given to the metal brass (see appendix), whereas these characteristics are worse than the radio-frequency

characteristics of metals such as copper and aluminium. Due to the constituents of zinc, the good characteristics of the copper are lost, and the characteristics of zinc, which are poor anyway, are made noticeably worse in the alloy. Tinplate housings are said to be absolutely unsuitable however if the field strength is low and the frequency is high, there is no noticeable sheath current heat losses. In this case, tinplate housings are thoroughly acceptable.

The radio-frequency characteristics include:

1. Diamagnetic grading
2. Equivalent material conducting layer
3. Electrical material conductance

Special attention should be paid to the fact that the electrical conductance (κ) and the equivalent conducting layer (s) are apparently opposite to one another

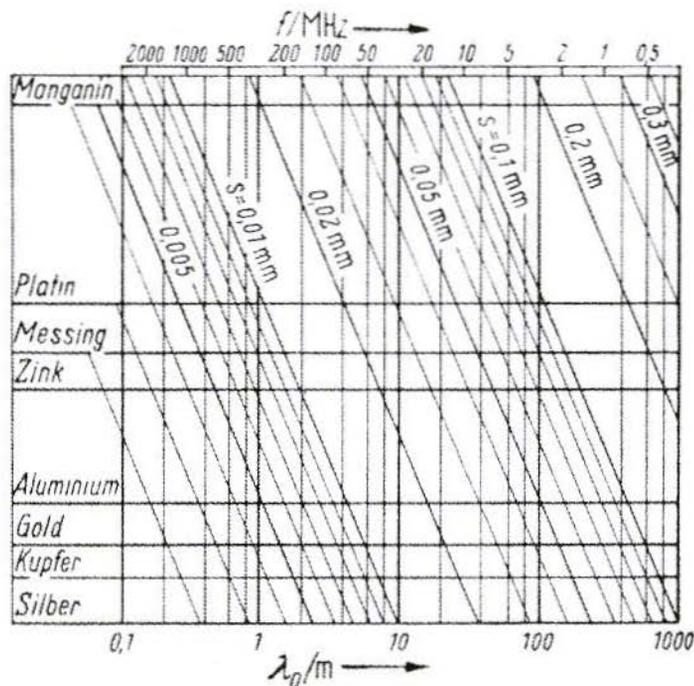


Fig 2: Diagram showing the effects of plating with various materials



Material	$x/\frac{S}{cm}$ (at 20° C)	K_1	$\alpha/^\circ C$
Silver	62×10^4	1.00	4×10^{-3}
Copper	58×10^4	1.03	4×10^{-3}
Gold	45×10^4	1.17	4×10^{-3}
Aluminium	33×10^4	1.37	4×10^{-3}
Zinc	16×10^4	1.97	4×10^{-3}
Brass	12 to 14 $\times 10^4$	2.20	1.5×10^{-3}
Platinum	9.0×10^4	2.62	4×10^{-3}
Magnesium	2.3×10^4	5.20	$\pm 0.01 \times 10^{-3}$
Constantine	2.0×10^4	5.57	-0.05×10^{-3}
Carbon	$\sim 0.02 \times 10^4$	~ 50	$-0/3 \times 10^{-3}$

Table 1: Parameters of various materials

in all metals i.e. a metal which is a good conductor has a low radio-frequency penetration depth and thus a low equivalent conducting layer. The metal layer under the equivalent conducting layer has no current flowing and is field-free. In this inner metal layer, no heat can be generated, and thus no energy can be absorbed from the field.

When metal is coated by a thin conducting layer with a high conductance (Fig. 1) it is shielded from the radio-frequency field hence the problem of eddy currents is reduced. In practice this means that the internal radio-frequency field of a resonant cavity can not penetrate through the metal coating in order to generate eddy currents and thus heat. This energy would be absorbed from the radio-frequency field thus reducing the output obtainable. These

radio-frequency eddy current and heat losses increase exponentially with the internal field strength. In accordance with the diagram below (Fig. 2) a 2 μm thick silver layer at 500 MHz would make the metal layer below it almost completely current free. In non plated aluminium, the current free state would be achieved at 3 μm and in non plated brass it would be more than 5 μm . From Fig. 2 it can be seen that brass has markedly poorer conductance ($\kappa = 12$) than aluminium ($\kappa = 33$).

Copper and silver are shown as close neighbours in the diagram. In addition, both metals also have similar conductance levels (Cu: $\kappa = 58$ S/cm; Ag: $\kappa = 62$ S/cm). It follows from this that silver-plating copper is not worthwhile.

Aluminium lies in the boundary area of

the good conducting layer group (silver, copper, gold, aluminium). But here too, silver-plating would appear to be merely something to please the eye, and to be less important as regards efficiency.

Silver-plating begins to be useful with brass. Moreover, the conductance is already falling away ($\kappa = 12$) but should still be described as good. It is surprising that zinc plate has slightly better radio-frequency characteristics than brass. But what we might at first sight be reluctant to believe has been scientifically demonstrated to be true. Zinc has better conductance ($\kappa = 16$) and, in the 500 MHz range, has an equivalent conducting layer of 4.5 μm .

Manganese ($\kappa = 2.3$) is a metal with extremely poor radio-frequency characteristics; it is mentioned for the sake of completeness. As regards conductance is almost as poor as constantine and would have a penetration depth of 15 μm in the same frequency range.

Particular importance is given to the calculation of the sheath current heat loss for coaxial cables. The conducting layer of the sheath and the inner conductor carry an enormous current density when high levels of power are transmitted, this noticeably heats the cable. For this reason, all manufacturers specify a maximum transferable power at a specific frequency. If the limiting sheath temperature is exceeded, then an exponential rise in heat loss takes effect. The specified attenuation values are then no longer valid.

3.

Appendix - Brass

The main constituents of brass are copper and zinc. Reddish brass has a high copper fraction. In yellow brass, the copper and zinc fractions are roughly the same. If the zinc fraction is even higher, brass becomes almost

white. For various commercial applications, so-called delta metals (brass with a low zinc fraction) and durana metals (brass with admixtures of aluminium, manganese, tin, lead, nickel and iron) are manufactured to secret recipes. Cold working is possible, thanks to these additives. The important value for diamagnetic grading, μ_r , considerably exceeds 1 in all cases. For durana metals, μ_r can be as high as 100. The yellow brass with high zinc fractions, which can be obtained commercially at prices that represent good value, is often made use of in amateur radio projects. This brass is graded with the equivalent conducting layer in the diagram.

4. Literature

- Küpfmüller: Theoretical electrical engineering, Springerverlag
 Kuchling: Physics; tables and laws; Buchverlagsgesellschaft; Cologne
 Meinke/Gundlach: High-frequency technology pocketbook, Springerverlag



Gerhard Schmitt, DJ5AP

Supplement to Article on Low Pass Filter for 2 m and 70 cm

In response to the article by Harald Fleckner, DC8UG in [3] and [4], experiments were carried out with the printed circuit board material RO 4003 from the ROGERS company, and for this purpose the low-pass filters described were constructed to cope with relatively high outputs.

As H. Fleckner has already pointed out in [4], the base material is an outstanding alternative to the relatively expensive Teflon base material. In the frequency ranges that are of interest here, 145 and 435 MHz, no differences could be observed from the results for the Teflon material RT 5870.

Since the importer [5] could make printed circuit board material RO 4003 available at short notice in thicknesses of 0.5 mm and 0.81 mm, the capacitance per unit of area was initially determined again. For the material with a thickness of 0.5 mm, an area of 0.15 cm² gave 1 pF. Thus for a capacitance of 47 pF (filter for 145 MHz) an area of 7.06 cm² was required; for 27 pF, correspondingly, 4.06 cm².

For the 435 MHz filter, the base material used was 0.81 mm thick. An area of 0.243 cm² per pF was determined. When the capacitances were dimensioned, they were slightly increased, which had a positive effect on

the steepness of the filter curves. C1 and C4 gave 13 pF, C2 and C3 = 18 pF. For 13 pF we thus need 3.16 cm²; for 18 pF = 4.38 cm².

The filters for 145 and 435 MHz were again housed in standard commercial tinplate housings, with dimensions of 111 mm x 55 mm x 30 mm. N-sockets with small flanges were used as connections; however, the BNC connectors can also be used for outputs of up to 200 Watts. If the filter is part of a power amplifier, coax cable can be soldered on directly instead of the sockets. It is fed through suitable bores in the faces of the housing, and the cable screen is soldered to the inside of the housing. Teflon cable should be used here, of course!

When the filters had been built into the housings and successfully calibrated, both the filters were tested at 750 Watts. As expected, there was no significant heating after several minutes of operation. Moreover, the values for the attenuation in the transmission band and the filter curves scarcely differed from the earlier Teflon material assemblies. A further advantage of the base material RO 4003 which should be mentioned is the considerably better mechanical rigidity. There is also a price advantage, as against RT- Duroid.

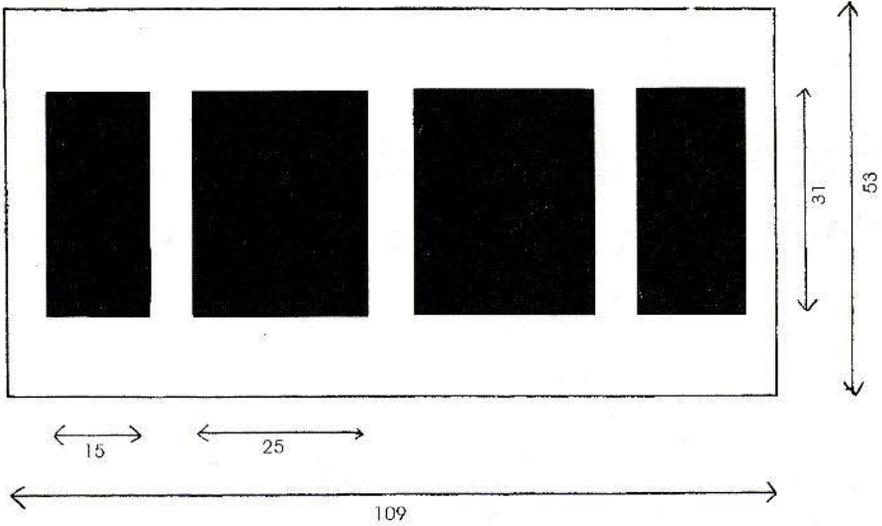


Fig 1: PCB layout for low pass filter for 145 MHz using 0.5 mm thick RO4003 material

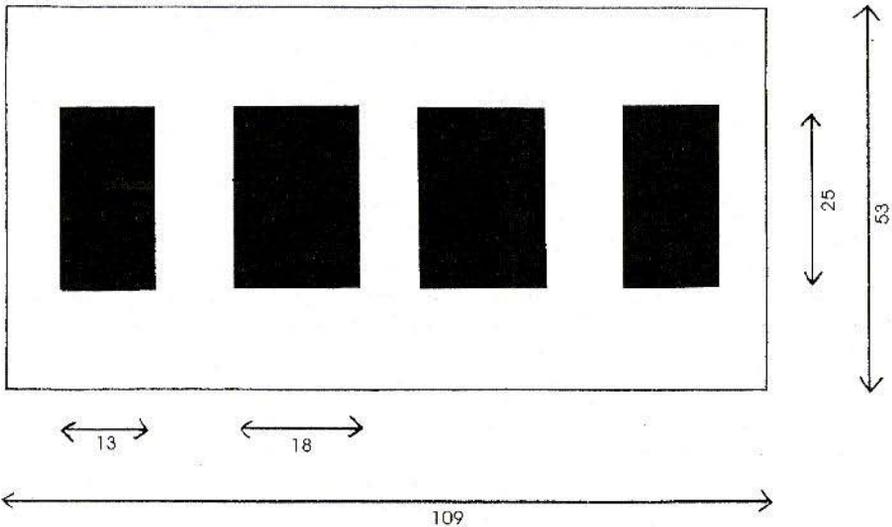


Fig 2: PCB layout for low pass filter for 435 MHz using 0.81 mm thick RO4003 material



To make reproduction easier, one board was developed for 145 MHz (Fig. 1) and one for 435 MHz (Fig. 2) in each case. The 2-m version is assembled on RO 4003 with a thickness of 0.5 mm, the 70-cm version on board 0.81 mm thick.

Literature references

[1] Schmitt, Gerhard, DJ5AP, Low Pass Filter for 2 m and 70 cm; VHF Communications, Issue 3/2000, Pp. 156 - 166

[2] Schmitt, Gerhard, DJ5AP, Low Pass Filter for 2 m and 70 cm, Part 2; VHF Communications, Issue 4/2000, Pp. 232 - 240

[3] Fleckner, Harald, DC8UG, Teflon or epoxy base material ?; VHF Communications, Issue 2/2000, Pp. 90 - 104

[4] Fleckner, Harald, DC8UG, Teflon, epoxy or RO 4000 base material; VHF Communications, Issue 3/2000, Pp. 152 - 155

[5] Data sheet for RO 4003 from the ROGERS company (distribution, DL: Mauritz GmbH+Co. Hamburg).

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Internet Treasure Trove

Synergy

The well-known manufacturer of radio-frequency and microwave components has expanded its product range and (as usual) posted the associated documents on its homepage. Of special interest here are some very well put together short tutorials (in PDF format, ready for download) on the subjects of couplers / modulators and demodulators / mixers / synthesizers / VCO / power dividers.

Address:

<http://www.synergymwave.com>

Trompeter

This firm's homepage offers comprehensive information on radio-frequency plugs and sockets. In addition to catalogues and application notes, there is also an interesting publication. For radio-frequency circuit developers in particular, this publication is indispensable, as it includes all the tricks and tips required with regard to the subject of radio-frequency printed circuit board drafting and the correct ways to connect plugs or sockets with a printed circuit board:

Printed circuit board design guide.

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

Philips

Anyone who has not visited the Philips homepage for some time will be pleasantly surprised, not only is the presentation now very good and user-friendly, but the number of prepared application notes has risen to well above 400. There is now 35 megabytes of information of interest to radio-frequency engineers.

The data referring to RF power amplifiers have multiplied at great speed. Something very recently added is information on such topics as circulators or electro-magnetic compatibility. However, we should explain how to get to where you want to go, for as so often a little game of hide-and-seek takes place:

We go from "Professional Products" through "Semiconductors" to "Product Catalog". And there we finally find "Application Notes".

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

Sonnet Lite / Dr. Mühlhaus

The German agent for Sonnet single-sideband modulation simulation software (for precise information on using the free test version Sonnet Lite, see VHF Communications 2/2001, Page 66) is very active, and has prepared supplementary programs on his homepage to be downloaded free. He has written some of these himself.



His "Sonnet tips and tricks" are very useful, as is the "geo3D" program for the three-dimensional representation of the simulated structure. Just as important is "dconvert" for generating "S-Parameter-Files in Touchstone format" (*.S1P or *.S2P) from the simulation result, etc..

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

Obtaining driver software

You have a new PC and you want to carry on using the hardware you're used to as well. So you soon find yourself faced with the problem that (in the main) existing drivers often do not run on the new machine, because it has more up-to-date operating systems; so you are in urgent need of updated drivers for the new installation.

Where can you get the new drivers you need in a hurry? The address below is the most important one in this context, and you can find practically everything you need concerning current software and suitable links there.

Address: <http://www.heise.de/ct>

Meteosat / Eumetsat

Have you heard? The next generation of Meteosat weather satellites (MSG = Meteosat Second Generation) is on the starting line. The first rocket was due to be launched in October 2000, but this has been postponed until July 2001. So there's still enough time left for anyone who is interested to become familiar with the history of Meteosat, the current technology, the transmission technology for the newcomers (almost everything has finally become digital...) or with all aspects of applications in meteorology or geography. You can also download extremely high-quality material by the

ton, most of which comes under the headings of "Technical & Scientific Literature" or "MSG". Those not wishing to go into the subject so deeply can improve their knowledge now and then by looking at the on-line house magazine "Image".

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

National Semiconductors

No doubt with an eye to the ever-growing mobile phone market, this company has set up its own web site for "wireless communication". With regard to the product range, the greatest interest is aroused by PLLs and synthesiser tuning. So we not only find very good and comprehensive application notes and introductions here, but also software for the easy calculations of appropriate circuits (Name: EasyPLL).

Address:

<http://www.wireless.national.com>

Dayton Hamvention

The best-known American amateur radio exhibition has its own homepage. In live sound: The greatest Amateur Radio Convention in the world!

Here, in addition to the date of the next show in 2001, which lasts for more than one day (18th. 20th. May), you can find everything worth knowing about flights, accommodation, and even the entry prices.

A photo gallery conveys an impression of the gigantic event with its halls, the arena and the unusually extensive flea market within the exhibition space.

Address: <http://www.hamvention.org>



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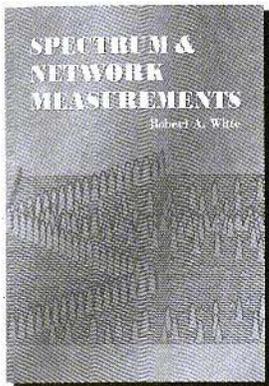
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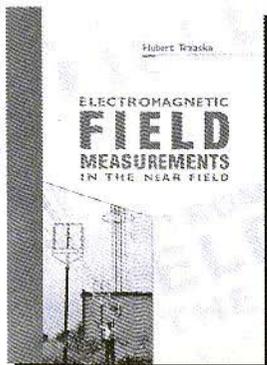
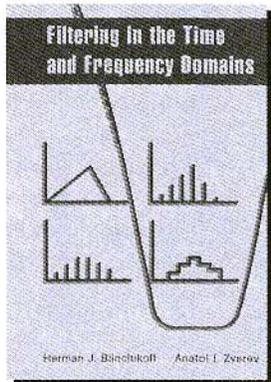
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