



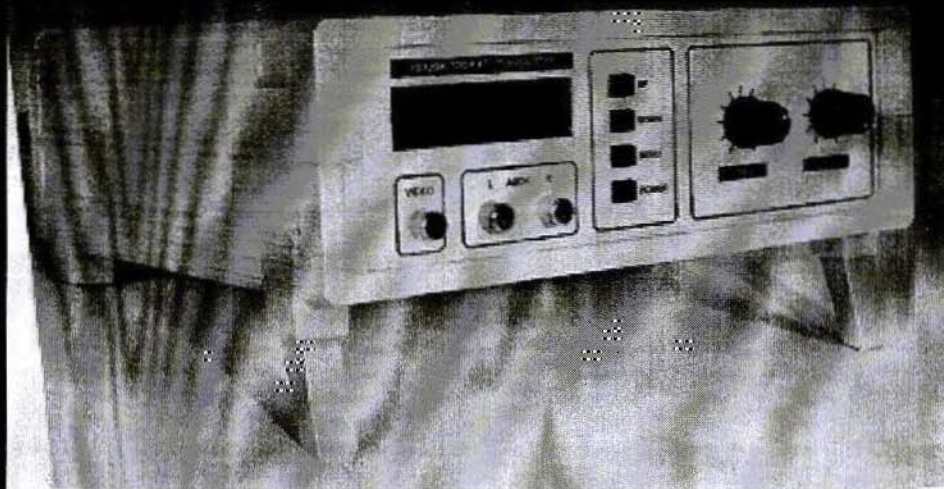
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PE1JOK**



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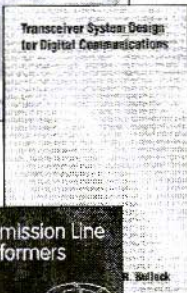
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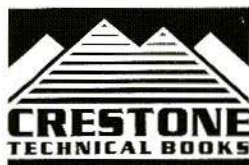
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Henk Medenblik, B.Sc, PEIJOK, Holland

A State-of-the-Art 13cm Amateur Television Transmitter Part-1

1. INTRODUCTION

The growing popularity of frequency modulated Amateur Television has led to an increasing activity on the 23cm band. Unlike other parts of our interesting hobby, ATV is mainly driven by home made designs. This growth and activity is therefore primarily dependent on regular publications of 23cm ATV transmitter designs. However, the attractive 13cm frequency band isn't explored yet by lots of amateurs.

The increasing pressure of ISM activities on this band could lead to the amateur allocations being lost in the future. That's the reason why it seems interesting to get some amateur activity on 13cm. With the publication of a sophisticated design a first step would be taken.

The growing demands for modern components and low cost designs on mobile communication markets like GSM, DECT and PCS, have made it possible

for radio amateurs to create their own professional designs. In this article the first part of a state-of-the-art 13cm ATV transmitter will be presented. This transmitter covers the entire 13cm frequency band in 1 MHz steps.

2. BLOCK DIAGRAM

The block diagram of the 13cm ATV transmitter is shown in Fig.1. It consists of a VCO-PLL unit, a 1.5 Watt power amplifier, video baseband processing and audio subcarrier unit and last but not least a microcontroller board with text inserter.

The microcontroller is responsible for controlling functions of the transmitter. It is possible, for example, to tune the audio carriers separately in 10 kHz increments between 5.5 and 8 MHz. Video polarity is controlled by a software menu and also frequency adjustment of the transmitter is performed with Up and Down tuning keys.

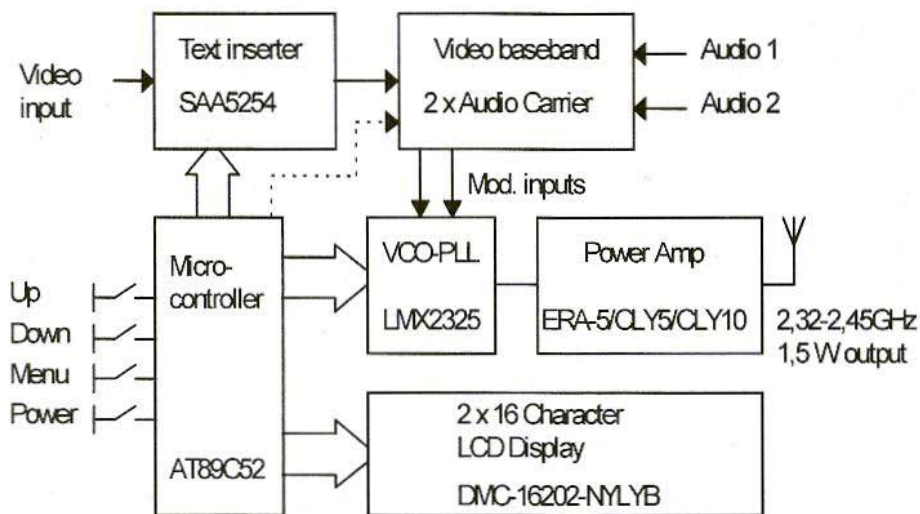


Fig.1: Block Diagram of the 13cm ATV Transmitter

A specially developed text inserter is controlled by the CPU board and is capable of generating a line of text through and inserting it into the composite video signal. The position of this text line and black background is determined by another software menu. The text lines are edited from a special menu and stored in EEPROM.

3. THE VCO

The main part of the transmitter consists of a VCO, which is controlled by a modern PLL IC from National Semiconductor, the LMX2325. Because I have not seen any interesting designs of VCOs covering the 2320-2450 MHz frequency range, I decided to develop one. This resulted in a VCO design which is capable of being tuned between 2250 and 2550 MHz within a 0-12V

tuning voltage range and very good linearity. Above 12 volts the tuning curve becomes non-linear due to non-linearities of the varactor diode. The resulting tuning curve is given in Fig.2.

The schematic of the VCO-PLL unit is shown in Fig.3. The VCO is designed using the negative resistance method. A Common Collector configuration with a certain emitter feedback capacitor will create a negative resistance at the base terminal

A series resonant circuit formed by a transmission line T1, coupling capacitor C10 and varactor D1 will make the total equivalent resistance real and negative at one frequency. This occurs at the frequency of oscillation.

The oscillation process starts due to noise in the circuit. The amplitude of the oscillation will stabilise due to compression effects of the transistor. At this point the total equivalent resistance is equal to zero (or loop gain equals 1).

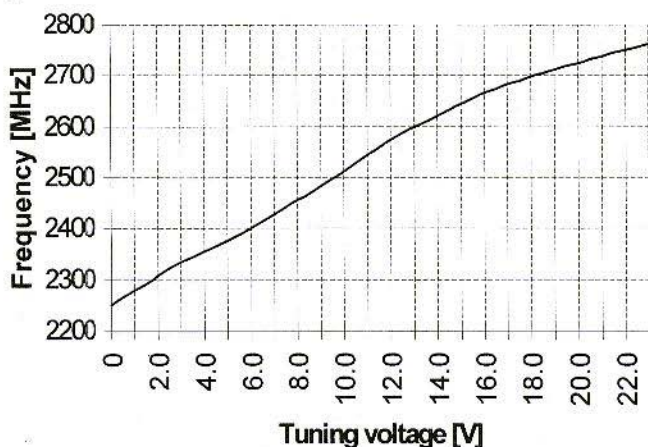


Fig.2:
Tuning Volts vs
VCO Frequency

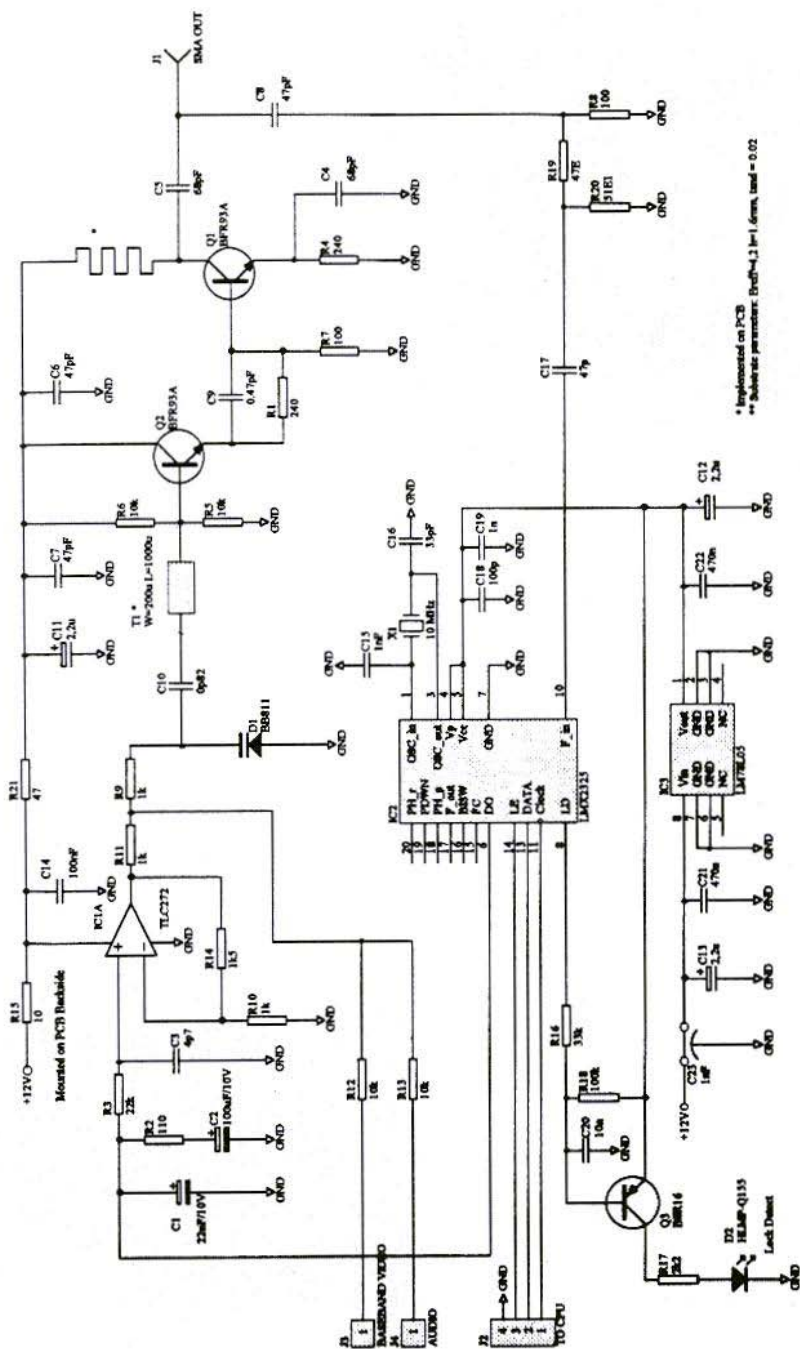
The amount of saturation due to compression of the transistor is determined by the amount of negative resistance during start up. This amount of negative resistance is determined by feedback capacitor value C9 and the internal base-emitter capacitance C of the transistor Q1. The combination of these two are also responsible for the flatness of the negative resistance across the band and thus the flatness of the output power of the oscillator, which is within 1dB.

Maximum output power is obtained at the high end of the frequency band. Most amplifiers have a decreasing gain at higher frequencies so that total cascaded gain will be flat.

Transistor Q2 has two functions, power gain and isolation. Isolation between an oscillator and a load is very important to maintain flatness. This can be explained by the phenomena pulling. Pulling is the change in oscillation frequency due to a variable load. This variable load can be created if the oscillator is loaded by a termination which is driven by a certain coax with a given length.

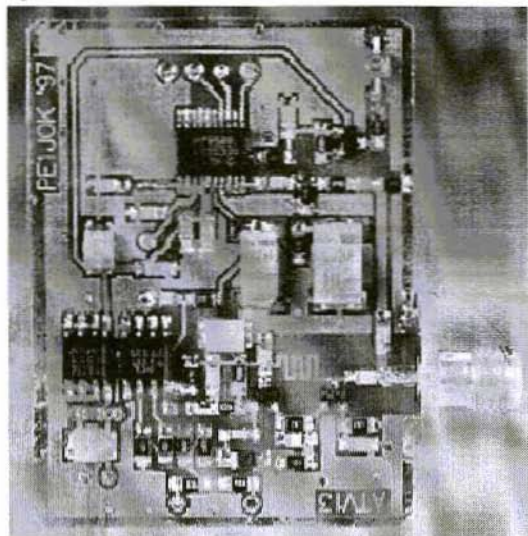
If the characteristic impedance of the coax and the termination are different then a mismatch occurs. This results in a complex impedance which can be seen looking from the output terminal of the oscillator towards the load. Varying the length of the coax corresponds to a constant VSWR circle on the Smith Chart. So the magnitude of the reflection coefficient stays the same but the phase will vary between 0 and 360°. This varying phase is responsible for the change in oscillation frequency if isolation between oscillator and load is bad. The same situation occurs when the coax length stays constant but the oscillation frequency is tuned.

If the coax length is not small compared to a quarter wave length, then a ripple in the tuning curve will occur if isolation is bad. Because we want to modulate the VCO with a broadband video signal and because we want to be able to tune the transmitter all over the entire frequency range, we do not want any non-linearities due to pulling. This can be overcome by an small amount of isolation and short cable lengths between VCO



* Implemented on PCB
 ** Substrate parameters: $\epsilon_{eff} = 4.3$ for 1.6mm, $\tan \delta = 0.02$

Fig.3: VCO-PLL Circuit Diagram



Top side of a completed VCO

and power amplifier (keeping in mind that input VSWRs of most PAs are generally not good).

4. THE PHASE-LOCKED-LOOP

Because the stability of a free running microwave VCO is not good enough, there may be a need for a form of frequency stabilization. This can be obtained by a Phase-Locked-Loop. A disadvantage of a controlling loop is the fact that they all have a certain loop bandwidth. In the case of a phase-locked-loop, disturbances with a frequency lower than this loop bandwidth will be equalized. Frequency disturbances above this loop bandwidth will not disappear.

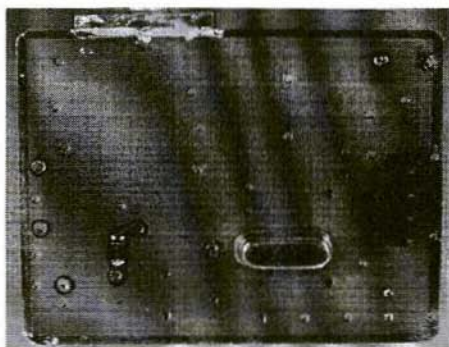
If we want to modulate the VCO with a very broadband signal such as video,

then care has to be taken with the selection of the lower edge of this loop bandwidth. Because video contains very low frequencies (50 Hz) this loop bandwidth must be lower than this frequency. In this design a bandwidth of 30 Hz is chosen.

A disadvantage of a low loop bandwidth is the speed of the loop. With the choice of 30Hz a compromise has been taken between lock time and lowest modulation frequency.

In this design a National Semiconductor LMX2325 IC fulfills the role of the PLL. This IC contains all the components needed to build a phase-locked-loop, a reference oscillator, reference divider, SHF dual modulus prescaler, swallow counter/divider and a low-leakage charge-pump phase detector. All of these functions come in a tiny TSSOP 20 package.

The current output of the phase detector is externally supplied with 5 volts. The loop filter, which consists of some capacitors and resistors, converts this



Bottom side of a completed VCO

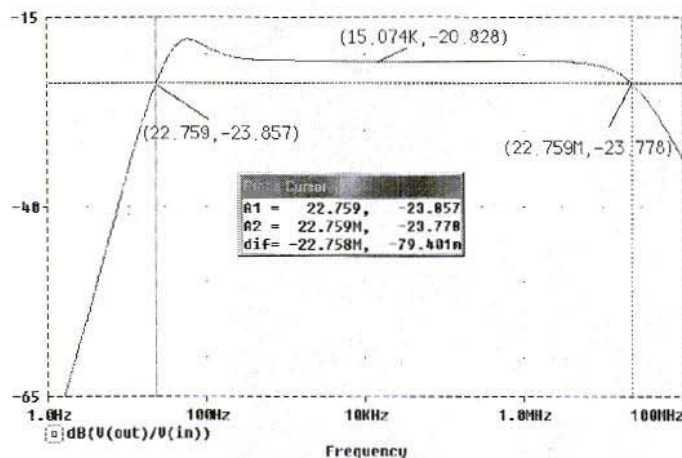


Fig. 4:
Modulation
Response of the
VCO-PLL unit

current into a tuning voltage. This voltage will be limited to 5 volts due to the external phase detector supply voltage. Because the VCO tuning range varies to 12 volt, a voltage amplifier is needed to increase the DC loop gain. This amplifier (TLC272) also creates a low driving impedance towards the varactor input, which results in a broad modulation bandwidth.

In Fig.4 a simulation result of the closed loop modulation response is shown.

The loop filter around the phase locked loop can easily be calculated with formulas given in Nationals application note for the LMX2325. For convenience they are reproduced here in Fig.5.

The structure of this filter in combination with the charge pump output creates a Type II third order system. With the given values of C1, C2, R2, R3 and C3, which are rounded to closest practical values, the loop bandwidth is dimensioned at about 30 Hz.

$$T1 = \frac{\sec \phi_p - \tan \phi_p}{\omega_p} \quad (1)$$

$$T3 = \sqrt{\frac{10^{\text{ATTEN}/20} - 1}{(2\pi \cdot f_{\text{ref}})^2}} \quad (2)$$

$$T3 = R3 \cdot C3 \quad (3)$$

$$T2 = \frac{1}{\omega_c^2 \cdot (T1 + T3)} \quad (4)$$

$$\omega_c = \frac{\tan \phi_p \cdot (T1 + T3)}{[(T1 + T3)^2 + T1 \cdot T3]} \cdot \left[\sqrt{1 + \frac{(T1 + T3)^2 + T1 \cdot T3}{[\tan \phi_p \cdot (T1 + T3)]^2}} - 1 \right] \quad (5)$$

$$C1 = \frac{T1}{T2} \cdot \frac{K_\phi \cdot K_{\text{VCO}}}{\omega_c^2 \cdot N} \cdot \sqrt{\frac{(1 + \omega_c^2 \cdot T2^2)}{(1 + \omega_c^2 \cdot T1^2)(1 + \omega_c^2 \cdot T3^2)}} \quad (6)$$

Fig.5:
Formulae to
Dimension the Loop

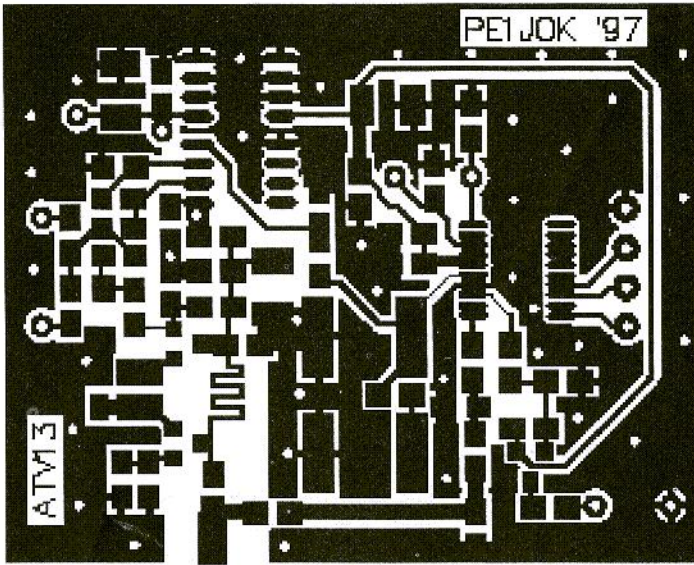


Fig.6a:
VCO Printed
Circuit Board
Layout - Top

Actual size:
47 x 34mm

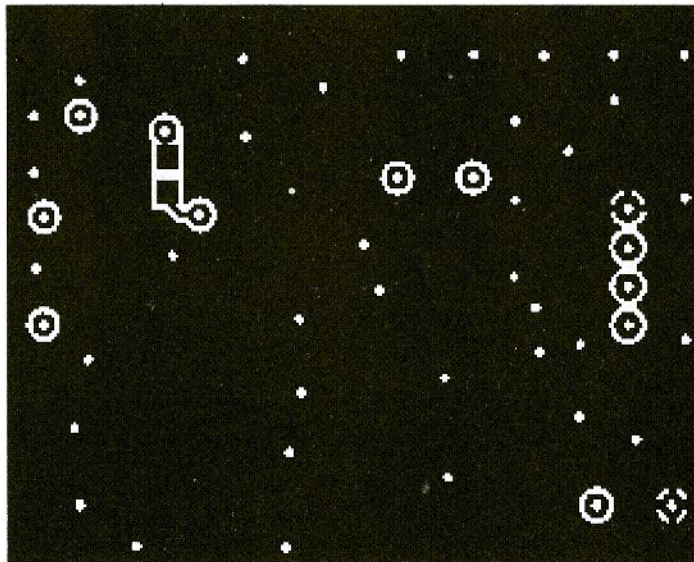


Fig.6b:
VCO Printed
Circuit Board
Layout - Bottom

Actual size:
47 x 34mm

Because the loop gain is very high due to the VCO constant, response of the phase detector and the selected step size of 1 MHz, very high values of loop filter components would result.

To overcome this problem, a very low reference frequency is selected by software. The reference divider is programmed for 25 kHz step size. This causes a high value of the programmable divider and thus a lower loop gain. The lower the loop gain, the

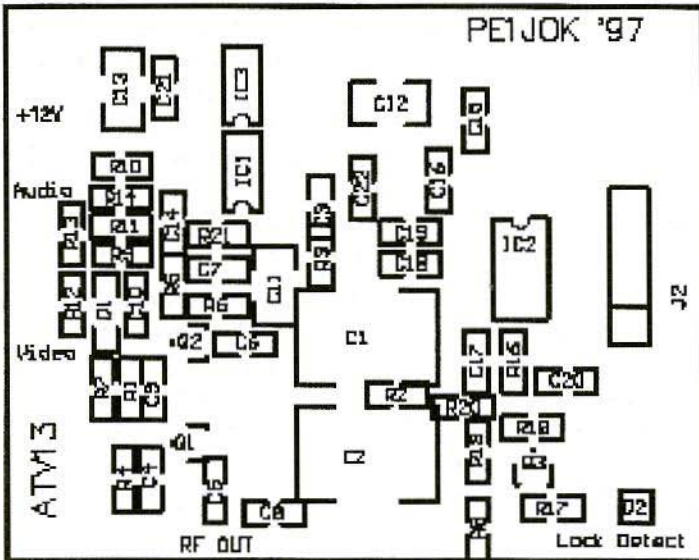


Fig.7a:
VCO PCB
Component
Overlay - Top
 Actual size:
 47 x 34mm

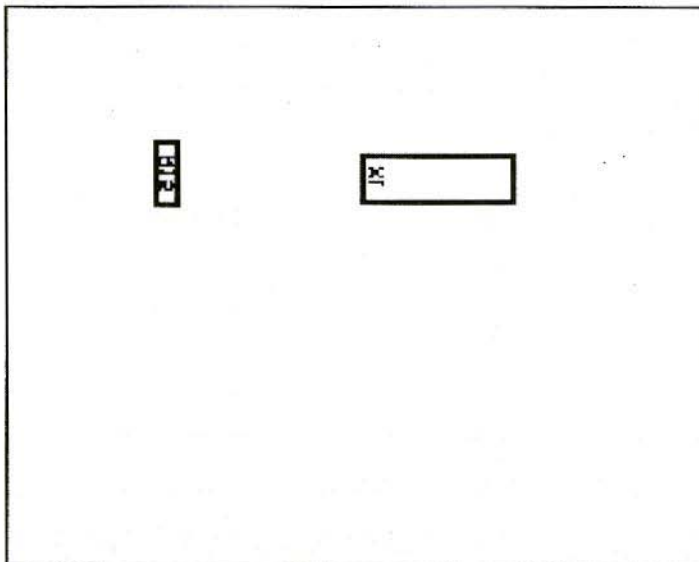


Fig.7a:
VCO PCB
Component
Overlay - Bottom
 Actual size:
 47 x 34mm

more practical component values will result.

The final tuning mechanism is adjusted by software to 1 MHz steps.

5. ASSEMBLY

A primary design goal during development was the implementation on stand-

ard low cost FR4 PCB material with a thickness of 1.6mm. Because the dimensions of the components are small, this resulted in a very small PCB layout and losses could thus be minimised. Fig's.6a and b show the PCB layouts, upper and lower respectively, whilst Fig's.7a and b show the upper and lower component overlays.

Note that a fully metalised bi-layer PCB and parts can be obtained from Spectra BV in Holland, whose telephone number and address are given at the end of this article.

When soldering the LMX2325, too much solder can be simply removed with solder wick.

It is preferable to start by constructing the VCO and buffer. This simple sub circuit can be tested using a high frequency counter. By applying a variable voltage at the varactor input the VCO frequency can be adjusted. A 1 to 8 volt range should result in a frequency range of 2300-2450 MHz. The circuit can be tuned slightly by positioning the varactor on the solder pads. This is equivalent to a change of effective parasitic series inductance and will

therefore result in a change of the frequency range. In worst case situations one can consider incrementing or decrementing the coupling capacitor value C10 (e.g. 0p68 or 1pF instead of 0p82). If this works then the rest of the components can be placed. A metal shielding around the PCB is preferred.

A special board which carries a synthesized audio carrier unit and video baseband part has been developed and the VCO/PLL unit can be placed on this PCB. This board design will be published in the next issue.

To be continued

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Gunthard Kraus, DG8GB

Design and Realisation of Microwave Circuits

Part-5 (cont)

8.3.2.3. Design of 1,700-MHz Stripline band pass, using the calculated values for the standard deep pass

If we start from a standard deep pass, with $n = 5$, and a reflection factor of 10%, we obtain the following filter coefficients:

$$g_1 = 0.97322$$

$$g_2 = 1.37226$$

$$g_3 = 1.80324$$

The Stripline band pass filter here should be designed for an impedance level of 50Ω , with the lower limiting frequency, $f_{\text{bottom}} = 1,680$ MHz and the upper limiting frequency, $f_{\text{top}} = 1,780$ MHz. This gives a band width of 100 MHz.

First step:

Determine the band pass filter coefficient, g_0 , which now contains the limiting frequencies.

$$g_0 = \pi \cdot \frac{(f_{\text{oben}} - f_{\text{unten}})}{(f_{\text{oben}} + f_{\text{unten}})} \quad (1)$$

This gives:

$$g_0 = \pi \cdot \frac{(100 \text{ MHz})}{(3460 \text{ MHz})} = 0,0907974$$

Second step:

Now you can determine the first Stripline pair; it has the same format as the last pair - i.e. no. 5.

First the "standardised impedance value":

$$\frac{Z_0}{k_{0,1}} = \frac{g_0}{\sqrt{g_0 \cdot g_1}} \quad (2)$$

That gives:

$$\frac{Z_0}{k_{0,1}} = \frac{0,0907974}{\sqrt{0,09074 \cdot 0,97322}} = 0,30544$$

EVEN Resistance (= common mode resistance):

$$Z_{\text{even}} = 50 \Omega \cdot \left[1 + \frac{Z_0}{k_{0,1}} + \left(\frac{Z_0}{k_{0,1}} \right)^2 \right] \quad (3)$$

$$Z_{\text{even}} = 50 \Omega \cdot [1 + 0,30544 + (0,330544)^2] = 69,937 \Omega$$



ODD Resistance (= push-pull resistance):

$$Z_{\text{odd}} = 50 \Omega \cdot \left[1 - \frac{Z_0}{k_{0,1}} + \left(\frac{Z_0}{k_{0,1}} \right)^2 \right] \quad (4)$$

$$\begin{aligned} Z_{\text{odd}} &= 50 \Omega \cdot [1 - 0,0544 + (0,30544)^2] = \\ &= 39,3926 \Omega \end{aligned}$$

Third step:

We now come to the next line pair, no. 2 (it is identical to pair no. 4).

$$\begin{aligned} \frac{Z_0}{k_{1,2}} &= \frac{g_0}{\sqrt{g_1 \cdot g_2}} = \\ &= \frac{0,090744}{\sqrt{0,97322 \cdot 1,37226}} = 0,078568 \end{aligned}$$

This gives

$$\begin{aligned} Z_{\text{even}} &= 50 \Omega \cdot [1 + 0,078568 + 0,078568^2] = \\ &= 54,237 \Omega \end{aligned}$$

and

$$\begin{aligned} Z_{\text{odd}} &= 50 \Omega \cdot [1 - 0,078568 + 0,078568^2] = \\ &= 46,38 \Omega \end{aligned}$$

Fourth step:

Central line pair, no. 3.

$$\begin{aligned} \frac{Z_0}{k_{2,3}} &= \frac{g_0}{\sqrt{g_2 \cdot g_3}} = \\ &= \frac{0,0907974}{\sqrt{1,37226 \cdot 1,80324}} = 0,05772 \end{aligned}$$

Solution:

$$\begin{aligned} Z_{\text{even}} &= 50 \Omega \cdot [1 + 0,05772 + 0,05772^2] = \\ &= 53,052 \Omega \end{aligned}$$

$$\begin{aligned} Z_{\text{odd}} &= 50 \Omega \cdot [1 - 0,05772 + 0,05772^2] = \\ &= 47,28 \Omega \end{aligned}$$

Fifth step:

The three line pairs, with their even and odd resistances, are now entered into the F3 component list in Puff, together with the electrical length of 90° . With the complete circuit in field F1, a plot result as in Fig.41 must now be obtained.

We can see that a little "finishing up" is still needed to obtain uniform transmission attenuation waviness which will not be too high.

Obviously such great perfection was not yet possible at the time the original article was written - i.e. in the age of the slide rule and log tables - or else we would have come across the secret simplifications and / or short cuts of the authors of that time.

Naturally today's professional microwave CAD programs, such as Touchstone, SuperCompact, Harmonica, etc. can do everything much better, but of course they cost as much as a mid-range car!

Modern simulation and field analysis methods have taken design in this area to such a pitch of perfection that achieving the manufacturing precision levels for the track structures when the boards are engraved or prepared is like breaking the sound barrier. However, we shouldn't let that stop us, and should continue to experiment a little further with the calculated values. With a little effort and a few changes, we actually arrive at a cycle as in Fig.42.

Special attention should be paid here to obtaining a "ripple" as uniform as possible, for only in this way can you then obtain the lowest transmission attenuation when you plot the curve for

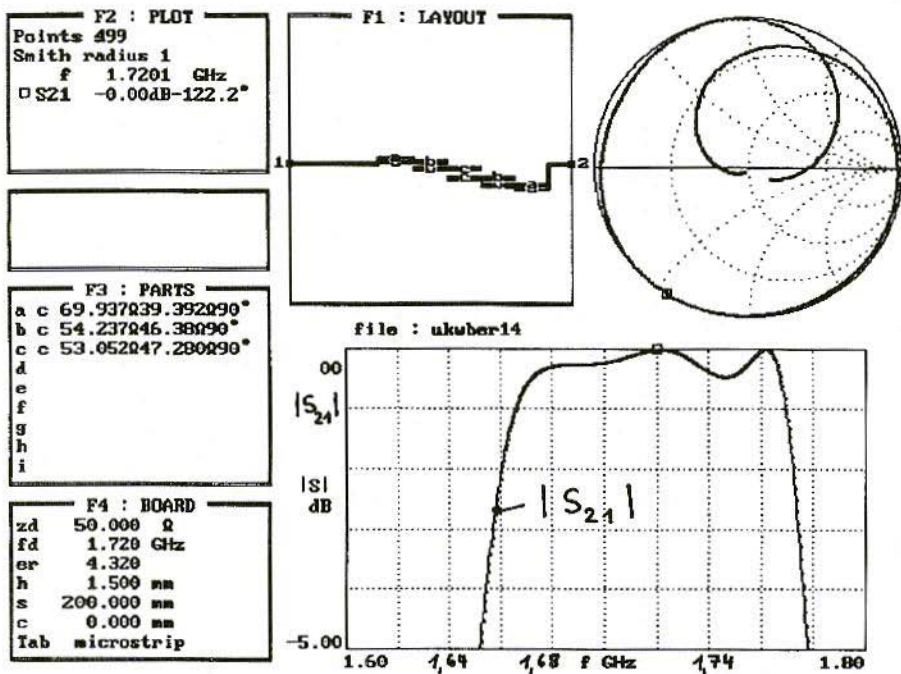


Fig.41: First Simulation of 1,700-MHz Band Pass with the Theoretically Determined Values

the filter with losses!

You have to think about this. If, as an experiment, we increase an even resistance anywhere in the table, the odd resistance should, as far as possible, be reduced by the same value, so that the impedance value of the line pair remains constant. Otherwise, you very quickly lose track.

In the next step, the unavoidable effects of harsh reality must be taken into account again.

So, we load the Puff file into the word processing program again in the usual way, check that all values, such as loss tangent, surface roughness, etc., are still correct, call Puff up again and match the line pair characteristics to reality.

To this end, we must put an exclamation mark in field F3 behind each "c" or "cl". Then the equals sign is keyed in on every line and as long as the electrical value of the component changes, until the resistances indicated in the dialogue field and the electrical lengths correspond to the ideal values of Fig.42.

When this has been done, we still need the mechanical dimensions of the tracks and their intervals, i.e. the "interaction gap" for the board design.

So we temporarily delete the exclamation mark in each component line and we enter the equals sign again. Back to the way things were!

We can now read off the length, width and intervals of the tracks in the

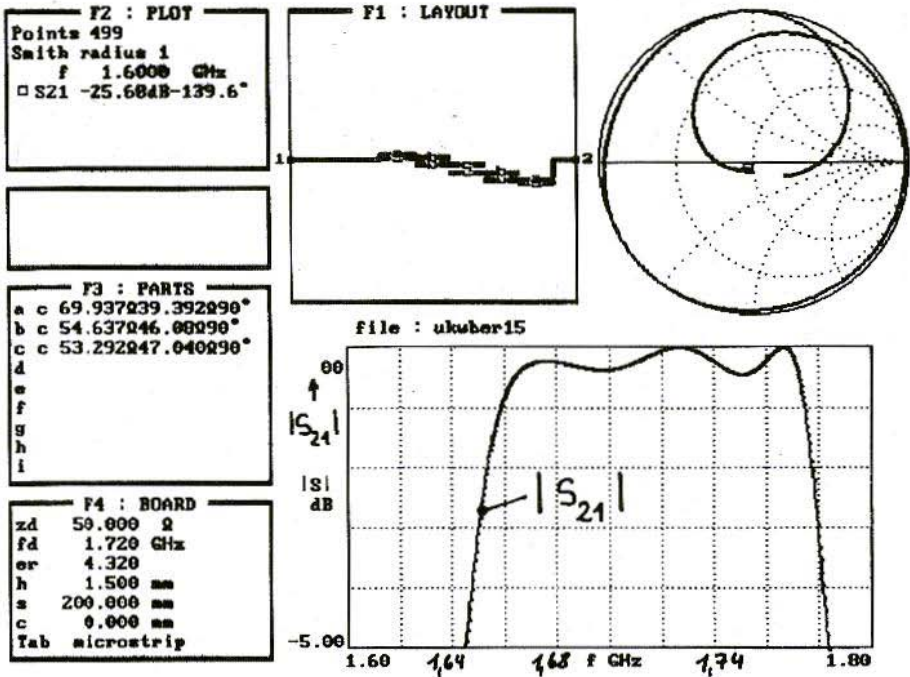


Fig.42: Improved Attenuation Cycle obtained by varying Line Data

dialogue field.

We can thus construct the following table:

(There is also the 50Ω supply line at the

Line pair	Conductor width	Conductor length	Interaction gap
a	2.28	24.95	0.42
b	2.8	24.44	2.2
c	2.82	24.36	2.86

all dimensions in mm

input and output, with a width of 2.88 mm..)

The shortenings at the open ends and/or at the discontinuities between narrow and wide tracks still have to be determined. If we use the method from the

last chapter to do this, we finally obtain the dimensions required for CAD design in accordance with Fig.43. Fig.44 shows the layout created with the Geddy CAD system, using a resolution of 0.02 mm.. The measurement results from the completed board, in a suitably milled aluminium housing with the external dimensions 35 x 135 mm. are entered in Fig.45, for the 1.5 to 2 GHz range, together with the simulation curve. Unfortunately, the measured transmission loss with 7.5 dB at 1,700 MHz is somewhat higher than the simulated result (approximately 6 dB). The pass range curve - as before, in the filter in the previous chapter - is again rather too narrow, whereas the resonator lengths and thus the limiting frequencies match up well.

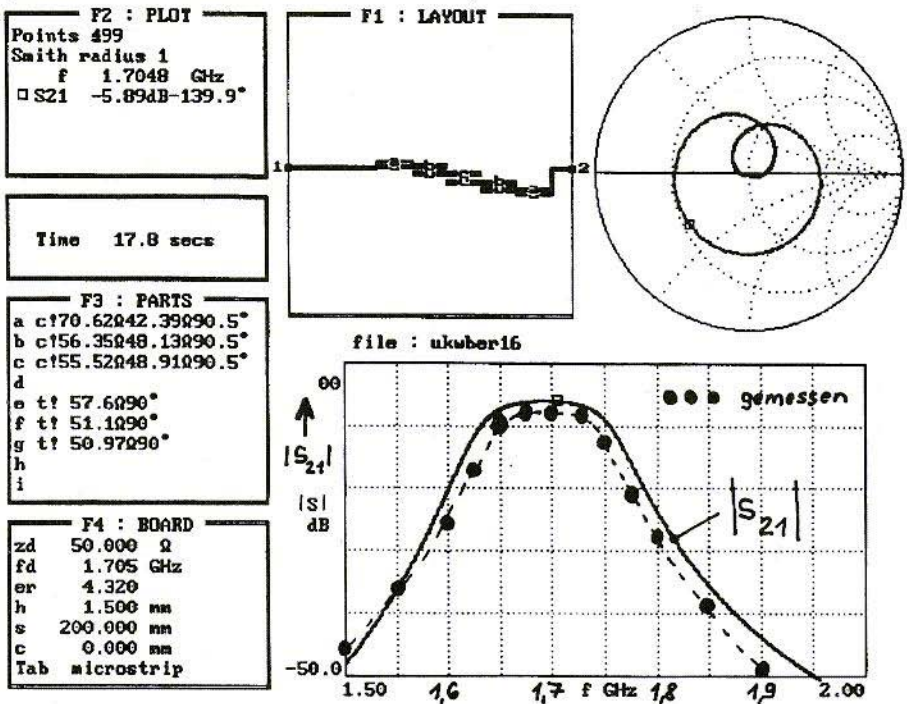


Fig.45: Simulation and Dimensional Results of Stripline Band Pass for
 $f = 1.5$ to 2 GHz and $_S21_ = 0$ to -50 dB
Gemessen = Measured

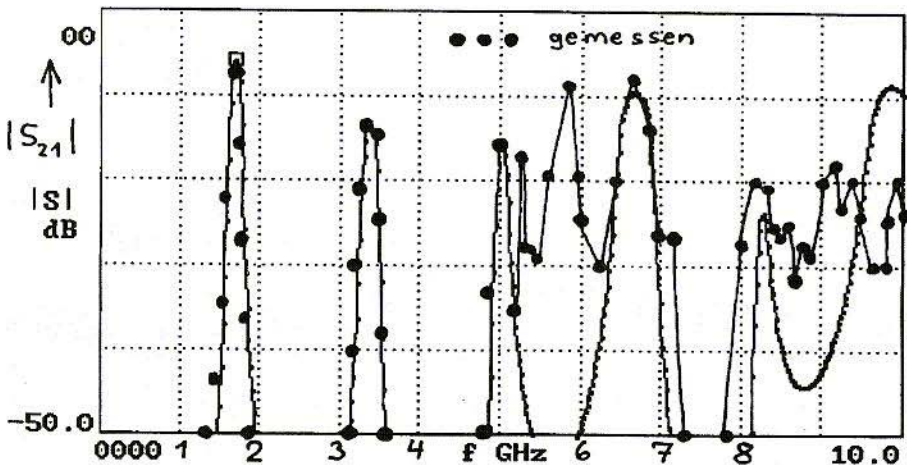


Fig.46: $_S21_$ in Range 0 to 10 GHz; *Gemessen* = Measured

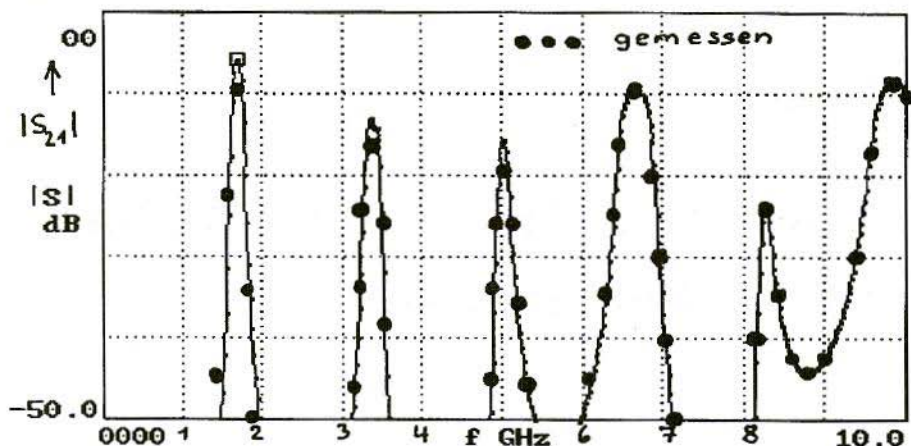


Fig.47: Cavity Resonances Suppressed by Conducting Foamed Material;
 S_{21} from 0 to 10 GHz of 1,700 MHz Band Pass Filter
Gemessen = Measured

The beautifully milled aluminium housing would probably act as a wave guide and / or cavity resonator from 5 GHz. So the inner cavity over the board was lined, in that foamed material sections cut to size, using a conducting foamed material, were stuck to the inside of the cover. This was totally effective - just look at Fig.47! All the dips have disappeared, and the measured results up to 10 GHz are precisely on the simulation curve.

One little thing still needed to be corrected. Clearly the foamed material is a little too thick, and so is too near the Stripline. This shows in the fact that the attenuation calculated in the preceding transmission range at 1,700 MHz, and in the first scatter resonance (i.e. at 2 x 1,700 - 3,400 MHz) has increased by 1 to 2 dB. But this problem can certainly be solved by experiments with varying thicknesses of damping material and comparative measurements.

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 Prentice Hall, New Jersey
 ISBN 0-13-586702-9



Ian Galpin, G1SMD

A Common Date/Time Standard for Amateur Radio

This brief article is to spread the word on a proposal connected with the representation of Date and Time in Amateur Radio.

We write times using 'hh:mm:ss' or 'hh:mm'. Imagine the chaos if some countries used 'ss:mm:hh', others 'mm:ss:hh'. We now use the 24-hour format rather than 12-hour am/pm. The use of Local Time can cause confusion in an International activity such as Amateur Radio. We avoid this by using the UTC Time Zone. All these methods are defined in an International Standard called ISO 8601.

For dates every country seems to do something different. The date 04/01/80 means 4th January in Britain, but April 1st in America, leading to problems in computer programs, on QSL cards, in email, and in newsletters. In a few years time we will have dates like 02/05/03 and 04/09/05 to contend with. ISO 8601 has a solution to these problems. The date is written using the full four digits for the year, with the order Year-Month-Day, and has hyphen separators. A leading Zero is used for '01' to '09'. In this way, a date cannot be confused

with that written in any other format. Some people prefer to retain the Year-Month-Day ordering whilst writing the month as a three letter abbreviation or out in full.

The date is written in any of three easily interchangeable ways:

1980-01-04

1980-Jan-04

1980-January-04.

The ISO standard has already been adopted as ANSI X3.30 in America. In Europe, the standard has been implemented as a 'EuroNorm'. Under the CEN regulations every member state is required to adopt this EN 28601 standard (all of Western Europe, and most of Eastern Europe). This format has been defined in Japan in the JIS X 0301-1992 standard.

ISO 8601 has been in use in Scandinavia, parts of East Europe, and most of Asia for many years, and by astronomers for over 200 years. They realised, long ago, the advantage of working to a common standard, world-wide.

In recent times, IBM has promoted the ISO standard as part of the fix for the



'Year 2000 Problem' with computer systems. The 'full' ISO format solves this problem as well as the UK / US date ambiguity. DOS and Windows already have provision for the Year-Month-Day method built in, as have a number of Amateur Radio programs. The Amateur Radio proposal recommends adoption of the ISO format for all facets of the Amateur Radio hobby: computer programs, log books, QSL cards, email and packet messages, Web Pages, band reports, newsletters, and so on.

The German magazine DUBUS has already adopted the new format, along with 'CQ-TV' published by the BATC; as has the Meteor Scatter software by OH5IY, and the new EME programs by VK3UM. G3SEK is rewriting all his software to cope with the Year 2000 and to adopt the ISO format. G0RUR is using it in his software in 1998. G3RUH has used it for very many years. The ITU are using the ISO format on their Web Page at: <http://www.itu.int/> and in some of their documents; as are the BATC in their Web Pages.

This is the only Internationally agreed format, and I recommend its use to all Radio Amateurs. Use a 4-digit Year and the Year-Month-Day order. We all understand the time '22:44:59'. There should be no problems with the date '1998-03-10', whereas '10/03/98' will always have a different meaning across the two sides of the Atlantic. ISO 8601 replaces 'dd/mm/yy' and 'mm/dd/yy' with 'yyyy-mm-dd'. Also note that Dates are written before Times in the standard.

The full proposal document was printed

in DUBUS 1997-Q1 (Page 83 to 85) and is available on Internet at:

http://www.kirsta.demon.co.uk/iso_8601.htm

Copies have been circulated to RSGB, ARRL, IARU and so on. More information on the Year 2000 and on ISO 8601 issues can be found at:

<ftp://ftp.funet.fi/pub/ham/misc/g1smd.zip>

along with instructions for DOS and Windows to use the ISO format. Also included is an extensive list of sources of further information from magazines and on Internet.

A short general description of the ISO 8601 Date Format can be found at:

http://shell.ihug.co.nz/~hermetic/cal_stud/formats.htm

There are also several files on this subject available from the GB7PFT 'CLIVE' server, and by the 'reqfil' service from several other AX.25 Packet BBS systems in the UK.

My own Web Page is at:

<http://ourworld.compuserve.com/homepages/dstrange/y2k.htm>

and my email address is:

g1smd@amsat.org

This article is based on one that originally appeared in the 1998-March edition (Page 10) of 'Four Metres News' edited by Derek Thom, G3NKS. Thanks to G3NKS email:

4news@g3nks.demon.co.uk

for permission to reproduce that article here. 'Four Metres News' and VHF Communications have now adopted the ISO format.

Carl G. Lodström, SM6MOM/W6

Comments on the Article 'Design and Realisation of Microwave Circuits; Part-4': Issue 4/1997, pp.209-214

I read Gunthard Krauss (DG8GB) article about MMIC amplifier (pp 209 - 214 in the 4/97 issue) with interest. Also with satisfaction, seeing PUFF used again with good result.

Since I have made a good part of my living designing products with this kind of components during the last few years, I have a few thoughts to add: If possible, use 1/32 inch board material above 1 GHz. The radiation from microstrip lines are less on a thinner material. The losses increase, but the only way to cope with that is to go for a better material than FR4. A 50Ω line on 0.031" material is 1.3mm wide.

I fired up the SuperStar Pro program and took a look at the circuit. True, it becomes unstable for inductors over 0.22nH. Just as PUFF showed. While I had it running, I noticed that the S11 and S22 were not all that good. It dawned on me that one possibly could match this better with a few non-50Ω lines on the input and the output. The

quite encouraging result is enclosed. Input and output DC blocking capacitors are not needed to be more than 47pF, and one 0805 of each will fit nicely on a 1.3 mm line.

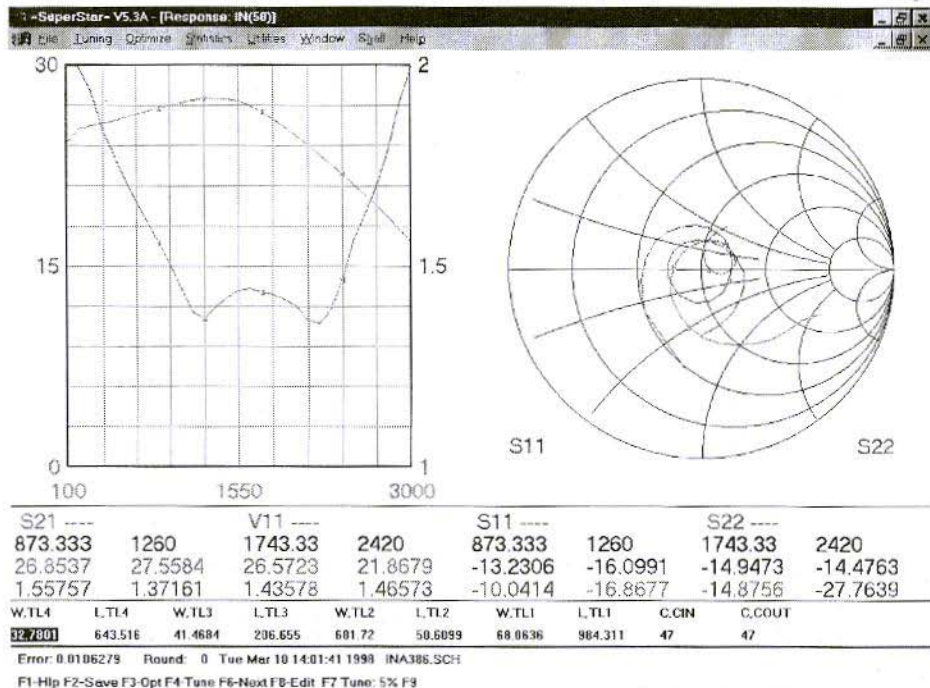
The "STx" on the schematic are the steps in line thickness. Part of the modelling. Line dimensions in the schematic are in "mils", that is 1/1000ths inch. The in/out pins of the amplifier are assumed to be 22mils, 0.56mm, wide.

The in/out of the board are 51 mils, 1.3mm, wide. Here: the dimensions of the four line sections in mm:

Part	Length	Width
TL1	25.0	1.73
TL2	1.43	17.3
TL3	5.25	1.05
TL4	16.34	0.83

It can be seen that these lines improved the match considerably in the 900 - 2400 MHz range.

The awkward TL2 can be replaced by



1.3pF to ground. On the plot, there are four markers on each panel. The first four are 1 - 4 and the last four are 5 - 8. I left in a 0.1nH in the amplifier ground. If it is totally removed, the gain bump is flattened and the match improves even more.

Although maybe not a good idea for boards that are manufactured by the 1000's a month, and stuffed by a robot, there is a way to get the inductance down considerably at the mounting of the amplifier!

Assume double sided board with solid ground on one side and the traces on the other. Now, mount the amplifier on the ground side and let the signal go through with rivets instead! Naturally, you have to cut up two small "islands" for the in/out pins, but the signal cares

not much about this. The ground gets very solid this way.

Two more thoughts are about the feed of DC in Gunthard Krauss circuit. It is a questionable practice to feed the device almost direct (10Ω) with 5V regulated. As temperature varies, the current may vary a lot. The better way is certainly to feed it via a resistor from a "higher source".

Gunthard Kraus has 12V available, and it is easier, cheaper and more temperature stable to burn the power in a resistor! The device drops 5V at 10mA. With 12V in, 7V is to be dropped in a resistor, => 700Ω, or, of course, 680Ω. This gets rid of the inductor as well! The 680Ω can be placed right onto the output line. The shunting of signal in negligible. So, there we save the regula-

tor and it's capacitors while improving the thermal stability. The inductors can be really tricky. A small 56 nH surface mount inductor I have used, but use no more, went through self resonance at 1.2 GHz. You do not want that in your circuit!

For a very interesting and novel treatise on inductors, see the RF Design maga-

zine, around January-98 I think. Randy Rhea, the creator of SuperStar and owner of Noble Publishing indicates that basically, the classical model of an inductor shows an impedance increasing forever with frequency. In reality there are resonances. Randy concluded that the inductor is better treated as a transmission line.

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Richard A. Formato, Ph.D., WW1RF (ex K1POO)

How Good are Genetically Designed Yagis?

1. INTRODUCTION

The Genetic algorithm (GA) is a powerful new antenna design and optimisation tool that is receiving progressively more attention. A natural question is How good are GA-designed antennas?

This note looks at that question by comparing four GA-designed 12-element Yagis. It does not examine how genetic algorithms work because there are many good references (several listed below [1]-[6]). This note emphasises Yagi performance, and how it changes in response to changing the GA parameters.

The GA searches a decision space which is defined by specifying minimum and maximum values for each antenna parameter. Only antennas falling within these limits are allowable solutions to the optimisation problem, which is a

characteristic of genetic algorithms that gives the antenna designer exceptional flexibility. Each element in a Yagi-Uda array has three design parameters: length, spacing, and radius. The min/max range of each parameter can set on an element-by-element basis, but the usual approach is to restrict all elements in the same way (with occasional exceptions as discussed below).

For the arrays described here, the element lengths were restricted as follows (all dimensions are in wavelengths, or waves): reflector, 0.35-0.65 wave; driven element (DE), 0.35-0.60 wave; directors, 0.3-0.6 wave. Element spacing was 0.05-0.5 wave, thus limiting the longest boom to 5.5 wavelengths.

For designs #1-3, the element radius was constant at 0.003369 wave, while design #4 allowed DE radii from 0.001 to 0.0075 wave. The standard Yagi configuration is used in which element #1 is the reflector, and element #2 is the DE.

2. FIGURE-OF-MERIT

The "goodness" of a particular Yagi design is determined by a *figure-of-merit* which is specified by the designer. For the arrays in this note, the FoM was:

$$FoM = \frac{aG - b |Z_o - R_{in}| - c |X_{in}|}{a + b + c}$$

where:

G is the forward gain (zero degrees azimuth) in dBi.

Z_o is the feed system characteristic impedance, in this case 50Ω resistive.

R_{in} and **X_{in}** are the Yagi's input resistance and reactance, respectively.

This FoM is used in [1], which is why it is used here. A GA allows the antenna designer to choose any FoM that reflects the desired balance between various antenna performance parameters. For example, front-to-back or front-to-rear ratio, or maximum side-lobe level could also be included if the designer wished to optimise against these parameters.

The weighting coefficients a, b and c determine the relative importance of each antenna performance parameter. For all designs below, the coefficient a is 40. For designs #1 and 2, b=c=1 (the values suggested in [1]). In design #3, the Yagi input impedance was removed from the FoM by setting b and c equal to zero. In design #4, the input impedance was weighted somewhat more heavily by increasing the coefficients to b=2 and c=3.

3. OPTIMIZED YAGIS

Optimisation was done by YGO (Yagi Genetic Optimiser) [7] which computes antenna performance using the Numerical Electromagnetics Code, Version 2, double-precision (NEC-2D) [8]. Seven segments were used for each array element in the NEC-2D model. Performance of the genetically designed arrays is tabulated in Fig.1 overleaf. Important parameters include the boom length (wavelengths), forward gain (dBi), half-power (-3 dB) beam width (degrees), input impedance (ohms), SWR relative to 50 ohms, front-to-back and front-to-rear ratios (dB//G), and maximum side-lobe level (dB//G).

Designs #2, #3 and #4 are the optimised Yagis, corresponding to the three sets of coefficients b and c, (1,1), (0,0) and (2,3), respectively. Design #1 is a sub-optimal design that appeared in the optimisation run for design #2. It is included to illustrate another important GA characteristic: a genetic algorithm does not produce a single best design, but instead produces a *group* of designs, with each design in the group ranked from best to worst. This feature can be very useful, because even sub-optimal designs may be attractive. Design #1, for example, may fill a real need because of its short boom length and excellent input impedance, even though its gain and FB/FR are worse than that of longer Yagis.

The GA-optimised antennas do indeed reflect different weighting in the figure-of-merit. In design #2, for example, b



Ant #	L	G	HPBW	Zin	SWR	FB	FR	MaxSLL
1	2.53	13.56	35.5	47.9+j0.9	1.05	10.5	10.5	-17.4
2	3.44	15.28	31.0	19.8+j0.4	2.53	20.5	20.5	-16.2
3	3.45	15.86	27.6	5.3+j179	131	13.6	13.6	-16.3
4	3.29	14.53	32.0	48.8-j0.1	1.02	17.7	17.7	-16.4

Fig.1: Performance of Genetically Designed Arrays

and c were chosen to emphasise gain while still controlling the input impedance. The resulting antenna has a gain of 15.28 dBi, a very respectable number, and it is nearly resonant with an input resistance of 19.8 ohms. This moderate resistance level is easily matched, so that this antenna provides very good gain and impedance performance.

By contrast, design #3 was optimised purely for gain, and nothing else. Its G of 15.86 dBi is exceptionally good. In fact, this Yagis gain appears to be slightly above the maximum theoretical gain for a 3.45-wavelength long boom as shown in the ARRL *Antenna Book* [9]. But, because the gain was maximised without any consideration of the input impedance, Zin for this design is a very poor match to 50 ohms. Of course, it is possible to match this antenna, but it is almost certainly not worth the effort in view of designs #2 and #4.

Finally, the FoM coefficients for Yagi #4 (chosen after the results of the two previous optimisation runs were analysed) were selected to improve further the balance between G and Zin. In addition, the DE radius in design #4 was allowed to vary, instead of being fixed at 0.003369 wave. The reason for this change is that the self-impedance of a

centre-fed dipole passes through a maximum as its diameter increases. This effect can be used to advantage to provide a better match to 50 ohms because Yagis tend to be low-impedance antennas. The GA confirmed this expectation by increasing the DE radius to 0.005767 wave.

Yagi #4 exhibits very good performance. Its gain of 14.53 dBi is only about 1.2 dB less than the maximum theoretical gain for the boom length [9]. And, quite significantly, this array provides an almost perfect match to 50 ohms. Feeding this antenna with 50 ohm coax requires only a balun. Completely eliminating any kind of matching network probably makes up for a good part of the 1.2 dB gain reduction. The result is a Yagi that is simpler, less expensive, easier to adjust, probably has greater bandwidth, is less likely to fail than designs with matching networks, and requires little if any maintenance.

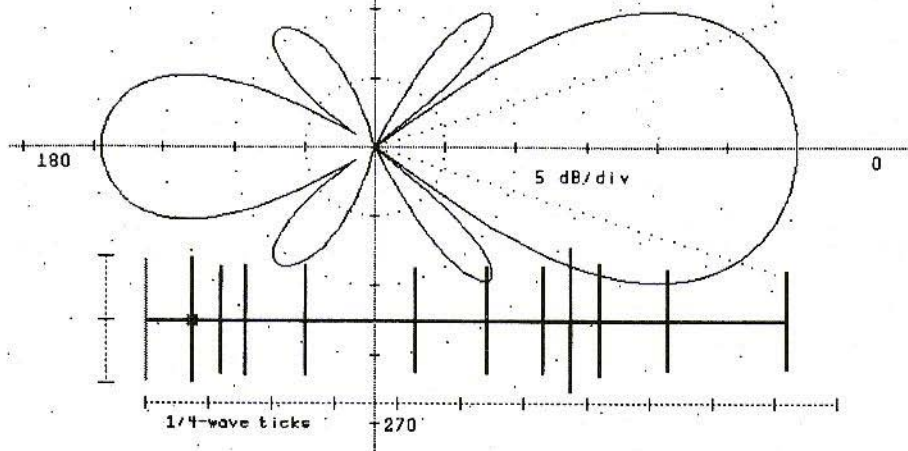
WIJR [10] makes a compelling case that Yagis should be designed to match 50 ohms directly, and there are many amateur and professional antenna designers who share that view. Designs like Yagi #4 show that achieving this objective with good overall performance is certainly possible. It is arguably the preferred approach.

Genetically Optimized 12-el
Yagi E-Plane (ϕ_z) Pattern

Design #1

Chromo #1, Generation #50
Figure-of-Merit = 12.843
Array Length = 2.53 λ_{uln}

Gmax = 13.56 dBi, HPBW = 35.5°
F/B = 10.5 dB, F/R = 10.5 dB
Zin = 47.9 + j.9 Ω SWR = 1.85//50 Ω
SL1 = -17.4 dB, ϕ_z = 49.5°
SL2 = -19.1 dB, ϕ_z = 129°
SL3 = -41.8 dB, ϕ_z = 75°

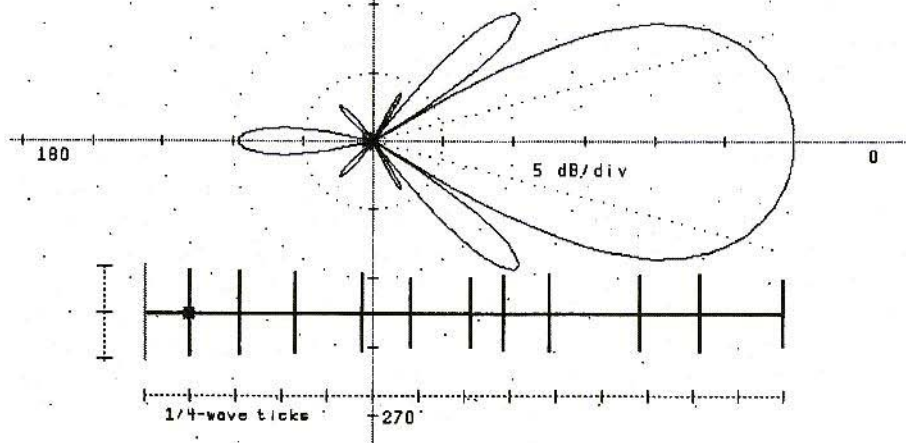


Genetically Optimized 12-el
Yagi E-Plane (ϕ_z) Pattern

Design #2

Chromo #1, Generation #50
Figure-of-Merit = 13.823
Array Length = 3.44 λ_{uln}

Gmax = 15.28 dBi, HPBW = 31°
F/B = 20.5 dB, F/R = 20.5 dB
Zin = 19.8 + j.4 Ω SWR = 2.53//50 Ω
SL1 = -16.2 dB, ϕ_z = 42°
SL2 = -26 dB, ϕ_z = 61°
SL3 = -26.5 dB, ϕ_z = 132°
SL4 = -29.2 dB, ϕ_z = 152°
SL5 = -29.4 dB, ϕ_z = 189°



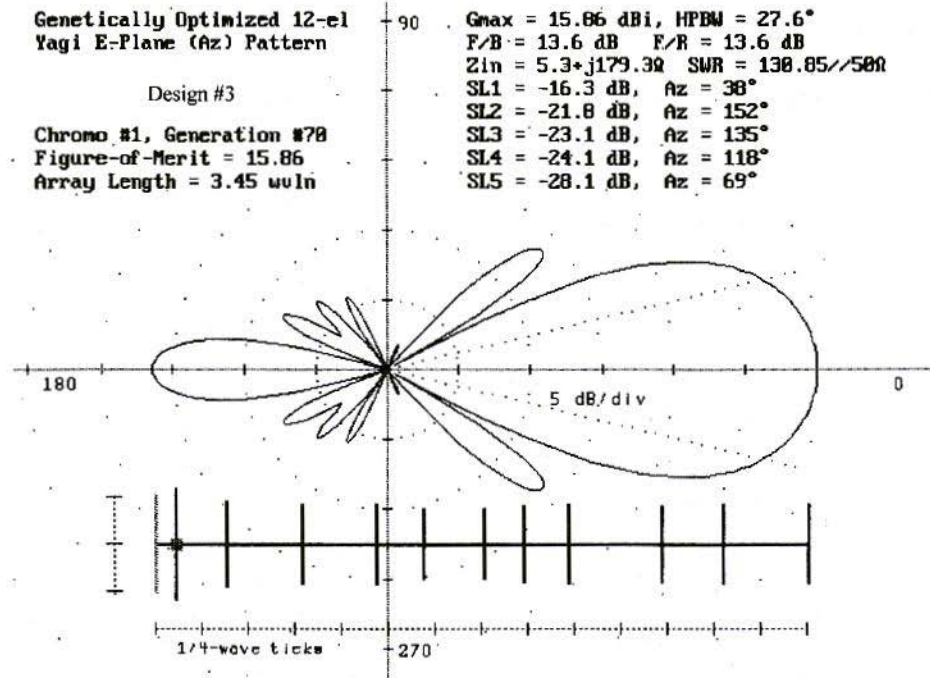


**Genetically Optimized 12-el
Yagi E-Plane (Az) Pattern**

Design #3

Chromo #1, Generation #78
Figure-of-Merit = 15.86
Array Length = 3.45 λ_{u1n}

Gmax = 15.86 dBi, HPBW = 27.6°
F/B = 13.6 dB F/R = 13.6 dB
Zin = 5.3 + j179.3 Ω SWR = 130.85//50 Ω
SL1 = -16.3 dB, Az = 38°
SL2 = -21.8 dB, Az = 152°
SL3 = -23.1 dB, Az = 135°
SL4 = -24.1 dB, Az = 118°
SL5 = -28.1 dB, Az = 69°

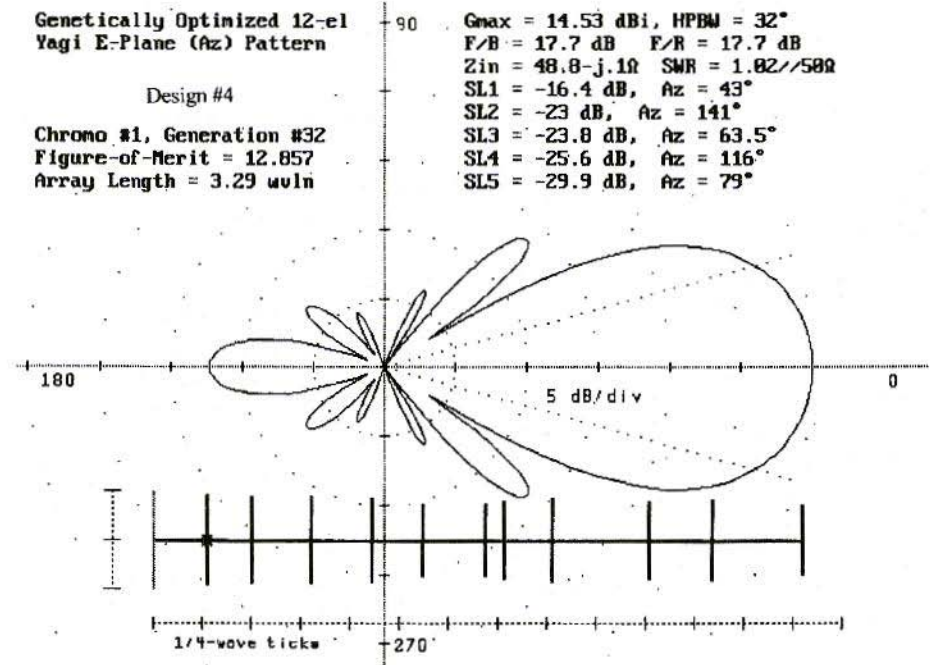


**Genetically Optimized 12-el
Yagi E-Plane (Az) Pattern**

Design #4

Chromo #1, Generation #32
Figure-of-Merit = 12.857
Array Length = 3.29 λ_{u1n}

Gmax = 14.53 dBi, HPBW = 32°
F/B = 17.7 dB F/R = 17.7 dB
Zin = 48.8 - j.1 Ω SWR = 1.02//50 Ω
SL1 = -16.4 dB, Az = 43°
SL2 = -23 dB, Az = 141°
SL3 = -23.8 dB, Az = 63.5°
SL4 = -25.6 dB, Az = 116°
SL5 = -29.9 dB, Az = 79°



El #	Design #1		Design #2		Design #3		Design #4	
	L	S	L	S	L	S	L	S
1(REF)	0.4794	0.0000	0.5006	0.0000	0.5006	0.0000	0.5006	0.0000
2(DE)	0.4853	0.1824	0.4549	0.2388	0.5784	0.1082	0.4588	0.2741
3	0.4165	0.1135	0.4388	0.2671	0.4400	0.2671	0.4365	0.2247
4	0.4329	0.0924	0.4318	0.3006	0.4188	0.4047	0.4341	0.3006
5	0.4341	0.2388	0.4141	0.3641	0.4188	0.3871	0.4259	0.3076
6	0.4047	0.4329	0.3624	0.2582	0.3624	0.2476	0.3647	0.2582
7	0.4224	0.2847	0.3718	0.3218	0.3718	0.3218	0.3718	0.3218
8	0.4176	0.2194	0.4012	0.1806	0.4012	0.2088	0.4012	0.0959
9	0.5671	0.1065	0.4200	0.2424	0.4200	0.2424	0.4212	0.2424
10	0.4388	0.1171	0.4035	0.4876	0.4035	0.4876	0.4035	0.4876
11	0.3906	0.2706	0.4176	0.3218	0.4176	0.3218	0.4176	0.3218
12	0.3847	0.4735	0.3471	0.4559	0.4224	0.4559	0.3482	0.4559

Fig.6: Dimensions for the 4 Designs

The NEC-2D computed E-plane (azimuth) radiation patterns and a scale representation of the array geometry appear in the plots in Fig.2, 3, 4 and 5. It is interesting to note that the element lengths and spacings in the genetically optimised antennas are neither uniform nor tapered as they are in deterministically optimised arrays. The GA-optimised Yagis look different, which is another very interesting aspect of genetic algorithm antenna design. GA-designed antennas are very often counter-intuitive, not necessarily resembling what would be expected. This characteristic of genetic algorithms makes them quite intriguing, because the designer cannot predict what the next best antenna might look like! A good example of this effect is Yagi #3. This array has a very long driven element (nearly 0.58 wave!) with a substantially shorter reflector, while in a conventional design the reflector is always the longest element.

Dimensions (element length, L, and spacing, S, both in wavelengths) are tabulated in Fig.6 for the four designs. All elements are 0.003369 wave radius, except element #2 (DE) in design #4, which has a radius of 0.005767 wave.

4. CONCLUSION

This note describes four genetically designed Yagis that illustrate how effective a GA can be in designing complex antennas. Amateurs interested in antennas are likely to hear more about genetic algorithms, and may wish to learn more about them. The genetic algorithm is a state-of-the-art design tool that is truly revolutionary in nature. Its potential in antenna design is just beginning to be explored, and may lead to some very interesting and unusual antennas.



5.

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- [5] *Antennae Exotica: Genetics Breeds Better Antennas*, Nathan Cohen, NIIIR, *Communications Quarterly (CQ Communications, Inc., Hicksville, NY)*, Fall 1996, page 55.
- [6] *A Genetically Designed Yagi*, R. A. Formato, *VHF Communications (U.K.)*, Vol. 2/97, page 116.
- [7] YGO (Yagi Genetic Optimiser) freeware was developed by Phadean Engineering Co., Inc., P.O. Box 611, Shrewsbury, MA 01545-8611 USA, tel. (508) 869-6077, fax (508) 869-2890, email: phadean@ma.ultranet.com. YGO can be downloaded from the NEC Archive website at: <http://www.qsl.net/WB6TPU> (download file is ygo.zip). Questions concerning the web site should be sent Ray Anderson, WB6TPU, who maintains it as a public service, email: raymond-a@radium.eng.sun.com. Questions concerning YGO should be sent to Phadeans email address.
- [8] NEC-2D is available from the Applied Computational Electromagnetics Society (ACES), Attn: Dr. Richard Adler, ACES Executive Officer, ECE Department, Code ECAB, Naval Postgraduate School, 833 Dyer Road, Room 437, Monterey, CA 93943-5121 USA. NEC-2 download files are also available on the web at: <http://www.qsl.net/wb6tpu/swindex.html>.
- [9] , 17th edition, R. Dean Straw, ed., American Radio Relay League, Newington, CT., 1994, Fig. 29, page 18-19.
- [10] "Yagi/Uda Antenna", Joe Reisert, W1JR, *Communications Quarterly Magazine (CQ Communications, Inc., Hicksville, NY)*, Vol. 8, No. 1, Winter 1998, page 49.

INSTRUCTIONS AND IMPROVEMENTS

SUPPLEMENT TO THE SERIES OF ARTICLES: VHF, UHF AND SHF MEASURING METHODS USING A PC

By Wolfgang Schneider, DJ8ES, published in issues 3/96 - 4/97

Assembly DJ8ES - 029: Interface Card

IC,s should be soldered on the components side, as the printed circuit board is not through-plated.

Assembly DJ8ES - 030: VCO with PLL

Solder SMD resistances with lettering toward track.

Two BB 621 diodes stand vertical in the board.

The circuit diagram shows block C as 0.1 μ F, the components diagram as 1 μ F; you should use 100nF, ceramic.

Fit BC549C rotated through 180° and not as shown in the component overlay, the circuit diagram is correct.

Connect IC pin -1 to earth via a 1.5 M Ω resistor. The circuit will be more stable in operation.

Assembly DJ8ES - 031: Mixer / LO

No bore is provided in the layout for the 100nF capacitor, but one is required.

The 78L09 voltage regulator is correctly drawn on the board, but wrongly in the circuit diagram.

The BFR90 transistors lie in the board. The holes should be drilled correspondingly large.

Insert the hollow rivets from the earth side / components side until they are flush with the track side, and solder on the earth side. No solder should run into

Assembly DJ8ES - 032:**Pin Diode Controller**

D0 and D7 are transposed on the components diagram (Fig. 18). The wiring diagram and the printed circuit board are correct.

All holes on the RK 1 - 12 V relay must be countersunk, using a 2.5mm drill, on the earth side, otherwise there is a risk of a short-circuit.

The value of the SMD resistor on the HF line at the pin controller is correctly stated as 1k5 in the components diagram. The figure of 150Ω in the wiring diagram is wrong.

The MSA sits on the board.

In addition, eight 1nF feedthrough capacitors should be used to decouple the data lines.

The pin diode input could also be better matched through an attenuator.

Assembly DJ8ES - 033: A/D Converter with Detector

Solder earth connections of IC's and terminal pins on both sides of the board.

The figure of 2 x 100nF in the components diagram is correct. The figure in the wiring diagram is wrong.

The relay power supply voltage is given

incorrectly as 12V in the wiring diagram (Fig. 24). The correct value is 5V. However, the board layout and the components diagram are correct.

Assembly DJ8ES - 034:**Direct Mixer for Spectrum Analyser**

A BV5169 can also be used instead of a BV531317.

BV5138 is identified, among other things, by a white dash and red / blue markings.

BV5313 is identified by a combination of figures.

The transistor BC549C is drawn with a metal housing in the components diagram (Fig.34). When it is used with a plastic housing, the flat side of the BC549C faces toward the reed relay.

Instruction for all assemblies in tinplate housing:

Choking the DC lines and data lines in the housing, using chokes or ferrite beads, increases the active and passive immunity to interference.

Any "mucky effects" arising from the software can be eliminated using the entry "Mouse On/Off" in the file HFMESS.CFG.



Dipl.-Ing. Andreas Gerstlauer, DG 5 SEB

Dr.-Ing. Johannes Gerstlauer, DC 9 SX

PUFF 2.1

IMPROVED AND EXPANDED VERSION

Puff has become widely accepted as standard high-frequency CAD software, even in Germany. For many high-frequency developers, it has become a firm fixture in the workshop as a bench tool to assist with development. The 2.1 version introduced here has some remarkable new functions.

1. GENERAL

Puff has become a standard software for the high-frequency CAD domain in recent years, with over 17,000 copies sold world-wide. The current 2.0 version has been on the market since 1991. It incorporated a series of new functions (as against the previous versions, 1.0 and 1.5) which grew out of practical applications.

Professor Dave Rutledge at Caltech Institute is currently working on a Japanese version. Then a new version is

to be produced for use with Windows 95, which will also include the new features in the 2.1 version developed by ourselves and introduced here. In addition to the operator interface (Windows 95), further changes planned relate to expanded high-frequency functions and, above all, to the user manual, which at the moment has unfortunately become very confused.

The reason why Puff is so popular is undoubtedly because it is easy to use. The fact that all the input and output parameters can be simultaneously displayed on the screen, where they can easily be viewed, makes for very efficient operation. In contrast to many other high-frequency simulation programs, only a brief reactivation period is required, even if it has not been used for a long time.

The models used (e.g. for the line configurations) are sufficiently precise for most applications. Provided the limitations of the program are adhered to, its use can go far beyond a mere hobby.



2. CHARACTERISATION OF PUFF 2.1

The Puff 2.1 version introduced here is based on the 2.0 version. Some very interesting additions have been made. In particular, there are these items above all:

The plot routine has been expanded by the addition of a "q" mode (for "quick"). In the quick mode, the computing result is made available only after the complete calculation has been finished. Since the screen display does not have to be altered after each item is calculated, the computing time is considerably shortened, as against the standard plot function.

The quick plot routine is called up, in an analogous way to the standard plot routine, by pressing the "q" key in the plot window (F2). The key combination, "Ctrl-q" works in the familiar way in quick mode too, first displaying the last simulation results computed, and then adding the newly calculated values as well. The functions "p", "q", "Ctrl-p" and "Ctrl-q" can be used in combination.

Probably the most interesting new feature has come out of practical experience. For the complex calculated reflection factors S11, S22, S33 and S44, the elements of the associated series or parallel alternate circuit can now be obtained!

To do this, after the carrying out of the "p" or "q" plot routine, the cursor is then positioned on the desired parameter

Snn in the plot window. Activate the "=" key, and the Rs and Xs elements appear in the message box, together with the associated capacity or inductance value for the series alternate circuit. The Rp and Xp elements are correspondingly displayed, together with the associated C and L values of the parallel alternate circuit, if the Smith diagram has been converted from the impedance representation to the admittance display.

The standard 1 - 2 - 5 gradation has been incorporated into the Smith diagram to provide a more precise interpretation.

Another new factor is the incorporation of a series of additional device files. The complete range of 'dev' files available now comprises: a voltmeter (vm.dev), a voltage-controlled voltage source (vs.dev) with a 20mS steepness, a differential amplifier (da.dev) for amplification 1, and an operational amplifier (op.dev) with an open-circuit voltage amplification of 10000.

All symbols are now uniform in that they have the input on the broad side of the triangle symbol and the output on the pointed side.

All devices are once again described in detail in Table 1.

The program contains some errors which are expressed, for example, in arbitrary program crashes, linked to error messages of the type "Runtime error at...". A whole range of errors of this type has been eliminated.

In some computer installations, the cursor displayed defective behaviour, in that it flashed too frequently. This was combined with occasional "sticking"



Device	Symbol	Circuit	Transfer Function	S-Parameter
vm(eter).dev			$U_2 = 2 U_1$	$S_{11} = 1 \ 0^\circ$ $S_{21} = 2 \ 0^\circ$
vs(ource).dev			$U_2 = \frac{1}{2} U_1$	$S_{11} = 1 \ 0^\circ$ $S_{21} = 1 \ 0^\circ$ $S_{22} = 1 \ 180^\circ$
cs(ource).dev			$I_2 = 20mS \cdot U_1$	$S_{11} = 1 \ 0^\circ$ $S_{21} = 2 \ 180^\circ$ $S_{22} = 1 \ 0^\circ$
da(mp).dev			$U_2 = U_{12}$	$S_{11} = 1 \ 0^\circ$ $S_{22} = 1 \ 0^\circ$ $S_{31} = 2 \ 0^\circ$ $S_{32} = 2 \ 180^\circ$ $S_{33} = 1 \ 180^\circ$
op(amp).dev			$U_2 = 10^4 \cdot U_{12}$	$S_{11} = 1 \ 0^\circ$ $S_{22} = 1 \ 0^\circ$ $S_{31} = 2E4 \ 0^\circ$ $S_{32} = 2E4 \ 180^\circ$ $S_{33} = 1 \ 180^\circ$

Table 1: Available Devices

when the program started. This error has been corrected.

The defective hard copy routine (premature termination of printing, or incomplete print-outs with some printers) has been corrected through the patching of the external supplementary program 'VGA2LASR.COM'.

The system has been matched to various national keyboard layouts. It should now be possible to use version 2.1 for working with all the different keyboard layouts correctly (i.e., as described in the manual). In particular, the key combinations 'Shift-1' to 'Shift-4' were not being correctly handled in the board window (F4) on non-American keyboards.

3. INSTRUCTIONS FOR USE

To use Puff 2.1, call up 'puff'. In the 2.1 version, all the functions are accessible through the corresponding keys as described in the manual.

When using the device files in the low-frequency range, remember that the voltage amplification, S_{nm} , will be displayed multiplied by a factor of 2, corresponding to 6 dB. In the low-frequency range, the reference value at the amplifier input is the generator open-circuit voltage. For the S-parameters, as is usual in high-frequency technology, our reference is the generator output voltage, which is 6 dB lower when match-terminated.

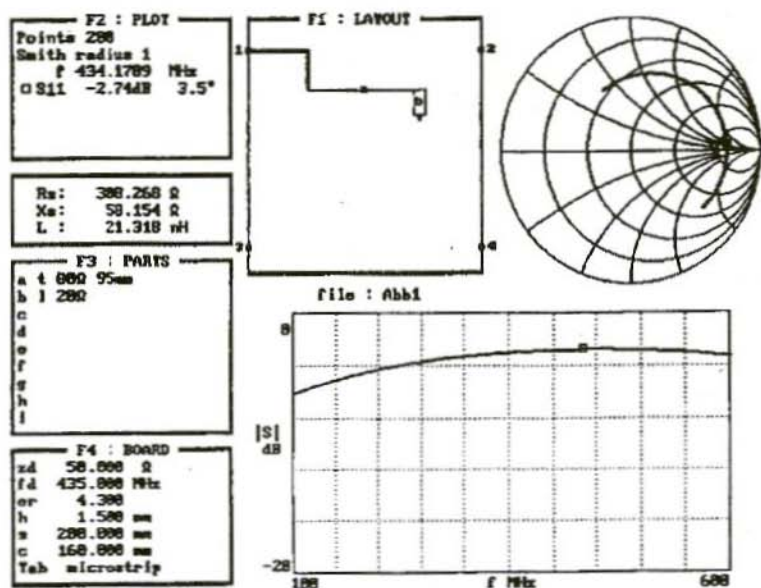


Fig.1: Alternate Series Circuit Elements

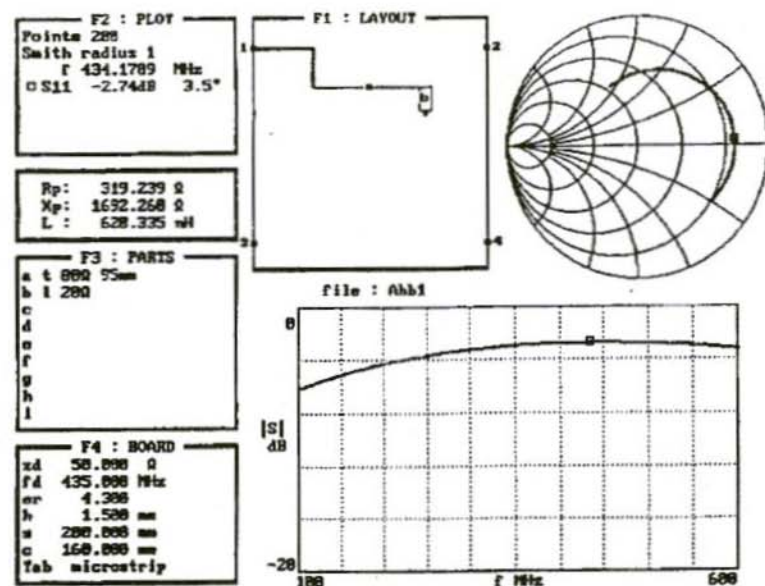


Fig. 2: Alternate Parallel Circuit Elements



If the new model of differential or operational amplifier is being used in the low-frequency range, select a meaningful average frequency, fd, in the board window (F4).

The differential amplifier can be used for potential-free voltage measurement. If the voltage drop is measured at a (small) shunt resistance, the current can thus also be measured.

The impedance or admittance values of the calculated S11 to S44 values are printed out as described in Section 2. After the carrying out of the plot function (using 'p', 'q', 'Ctrl-p' or 'Ctrl-q'), the cursor is positioned on the desired value, Snn, in the plot window (F2). Activate the '=' key. The specification for the Rs and Xs elements belonging to the selected S-parameter for the equivalent alternate series circuit appears in the message window, and the C or L corresponding to the Xs at the operating frequency. Fig. 1 shows the print-out for a simple example.

The corresponding Rp and xp values, and the associated value for C or L for the equivalent alternate parallel circuit are outputted if the Smith diagram is

converted (through the 'Tab' key) from the impedance representation to the admittance display. Fig. 2 shows the associated print-out for the example above.

4. PROTECTED MODE VERSION OF PUFF 2.1

In addition, an experimental protected mode version of Puff 2.1 has been filed in the 'PROTV' sub-directory. However, this is still at the Beta stage.

The protected mode version can be called up from the 'PROTV' sub-directory through 'puffp'.

In contrast to the standard version, the protected mode version actuates the entire main memory of the computer, which means larger circuits can be processed or more items can be calculated for the analysis. The maximum permissible number of plot points has been increased to 1000.

PUFF 2.1 UPDATE

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Bernd Kaa, DG4RBF

HF Synthesizer 5 to 1450 MHz

Part-1

An HF synthesizer which can be added to a radio amateur's measuring equipment is described below.

The appliance takes the form of a high-frequency source with a wide frequency range, which supplies a defined output which is adjustable within certain limits.

1. OBJECTIVES

I wanted to have an independent appliance available which would have some extra amenities for frequency input and operation. When I say "extra amenities", I am referring to operator guidance through LCD dot matrix display, with information on the current possible input range or value and input through a keyboard.

Naturally, these requirements can be met only in combination with a micro-processor, which simultaneously means that suitable software must be prepared (filed in an E-prom). Apart from the drawback of having to spend a lot of

time preparing the special control program, the use of a processor does, of course, also have considerable advantages, which we shall be discussing later.

In the development phase of the synthesizer, it became apparent that there was a wobbling option in addition to the required characteristics. This can be used to amplify the equipment into a slow-scan wobble system, thus making it possible to carry out wobble measurements in the range from 5 to 600 MHz, and from 600 to 1,450 MHz.

2. TECHNICAL DATA

Here is the spec's of the prototype:

Frequency ranges:	5 - 600 MHz & 600 - 1,450 MHz
Min freq raster:	5 kHz
Max. HF output:	30mW
Output control:	1 - 30mW
Level accuracy:	App. \pm 1dB (at 2 - 30mW)
Harmonics:	At 20mW app. 20 - 40dB

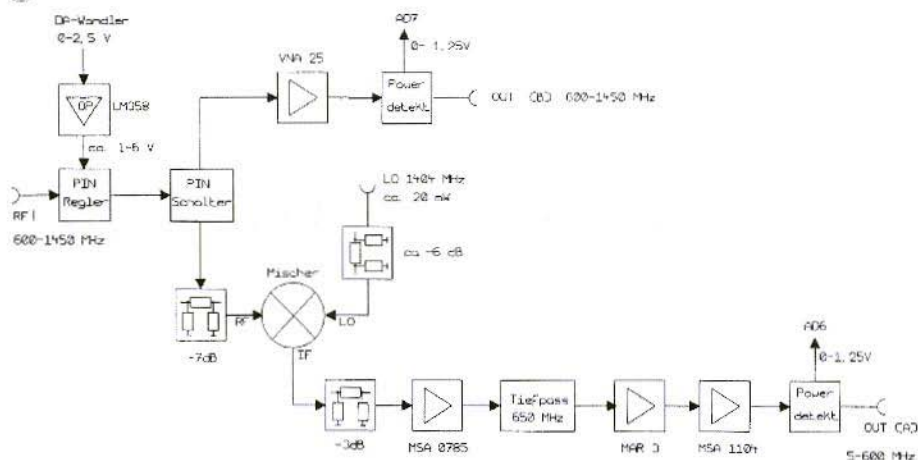


Fig.1: Modular Mimic Display of HF Section of Synthesizer
Wandler = Converter, Regler = Controller, Schalter = Switch, Mischer = Mixer, Ca. = Approximately, Tiefpass = Deep pass, Powerdetekt = Power detector

Band A:	Min. 25dB typ. 35 - 40dB
Band B:	Min. 20dB typ. 25 - 30dB

The frequency precision depends on the reference oscillator used (TCXO) and on the LO.

3. THE HF SYNTHESIZER IN DETAIL

The equipment consists of the following assemblies:

1. Universal micro-controller assembly
2. PLL / HF assembly
3. VCO assembly
4. LO assembly running at 1,404 MHz

Fig.1 shows the modular mimic display of the HF synthesizer.

3.1. The PLL / HF Assembly

The core of the circuit is the modern PMB2306T PLL circuit from Siemens, which was developed for mobile radio applications. The circuit comprises three freely programmable counters: a reference counter (16-bit R counter), an N counter (14-bit N counter), and an A counter (7-bit A counter), the phase detector with charge pump, and a lock detector, and can be programmed by means of a 3-conductor bus. The modular mimic display of the PLL assembly is shown in Fig.2.

The PMB 2306 can process frequencies from 0.1 to 220 MHz directly. The reference oscillator frequency can be selected between 4 and 100 MHz. The frequency data are transmitted serially to the IC in the form of data elements.

So we need "only" write a user-specific program for the functioning of the PLL and the control of the 3-conductor bus.

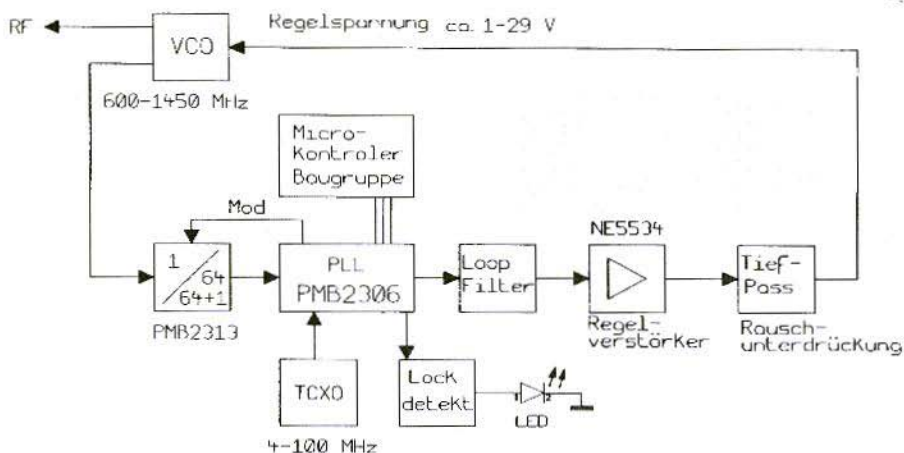


Fig.2: Modular Mimic Display of PLL of Synthesizer

Regelspannung = Control voltage, *Ca.* = Approximately, *Microkontroller Baugruppe* = Micro-controller assembly, *Regelverstärker* = Control amplifier, *Tiefpass* = Deep pass, *Rauschunterdrückung* = Noise suppression, *Lockdetekt* = Lock detector

Since the PLL-IC can process only frequencies up to 220 MHz, a pre-scaler (PMB2314 or PMB2313) must be added for our applications.

3.1.1. Circuit Description of PLL with Pre-Scaler

3.1.1.1. Pre-Scaler

The two pre-scalers, PMB2314 and PMB2313, are dual-mode pre-scalers, with the dividing ratios $1:n$ and $1:n+1$, the mode for which is switched to using pin-6 (MOD). Pin-3 (SW) is used to set a divider ratio of $1:64/65$.

Actually, the PMB2314, which is specified for up to 2,100 MHz, should be used here. As this interesting component was not yet officially available at the time when the synthesizer was developed, I fell back on the PMB2313 counter.

The PMB2313 pre-scaler is specified for 1,100 MHz, and is given a max. frequency of 1,400 MHz in the data sheet. However, the data sheet diagram "PMB2313 dynamic range" (Fig.3) shows that it can function up to and beyond 1,600 MHz if attention is paid to the input level.

Trials carried out by the author have indicated that the PMB2313 is still operating reliably at 1,450 MHz, provided the input level remains in the range between -20dBm and -1dBm.

The signal from the VCO is fed into the PMB2313 pre-scaler at pin-1 (I1), with a maximum level of -10 dBm. If the input power is too high, this leads to engaging problems at the band end of the synthesizer (see above).

Pin-3 (SW) is connected to +5V, which sets a scaler factor of $64/65$. The pre-scaler is actively switched through

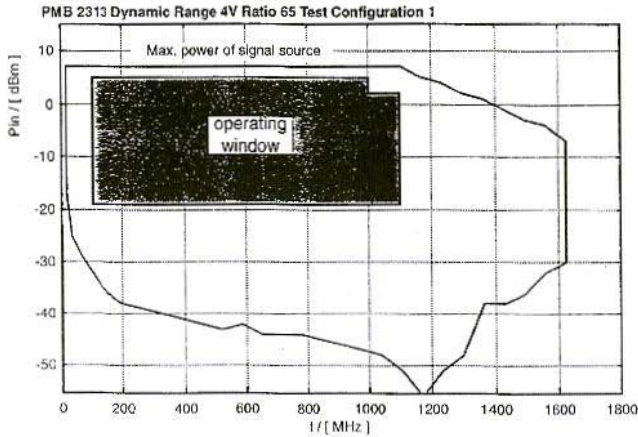


Fig.3:
Dynamic range of
PMB 2313

pin-7 (STB = stand-by). It is also connected to +5V.

The divided signal is available at the output pin-4 (Q), and arrives at pin-8 (FI) of the PLL-IC through the coupling capacitor C4. At pin-6 (Mod), we have the switch-over signal for the dual mode, which is made available by PLL-IC, pin-7 (Mod).

3.1.1.2. PLL-IC Circuit

The signal from the reference oscillator is connected at pin-1 (R1) of the PMB 306 PLL-IC through a capacitor (Fig.4). A small standard TCXO makes the most suitable reference oscillator.

Since the reference oscillator frequency can be selected to be almost anything (depending on the software), there are certainly no procurement problems. The board layout is, correspondingly, universally laid out for three different forms of TCXO housing.

The reference oscillator must, of course, meet the following framework conditions:

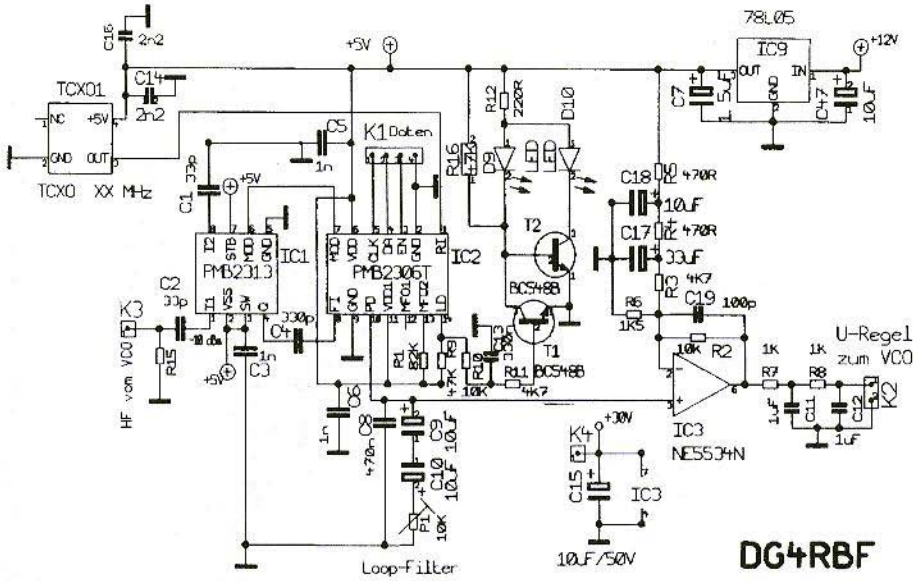
- The frequency must lie between 4 MHz and 100 MHz;
- The frequency must be divisible by 5 kHz;
- The oscillator should operate at a stable frequency (hence the TCXO!), since as the reference it is responsible for the precision of the synthesizer.

All the PLL programming is carried out through a 3-core bus. This consists of the lines "Enable" (pin-3 - EN), "Data" (pin-4 - DA) and "Clock" (pin-5 - CLK), which are linked to the data lines of the micro-controller provided for this through a buffer.

The MOD output (pin-7) controls the dual-mode pre-scaler.

The PD current is set at pin-13 (MFO2) through an 82k Ω resistor. The VCO signal, divided by 64/65, is at the FI input (pin-8).

The loop filter is connected to the PD output (pin-10 = phase detector). It consists of two bi-polar inter-connected 10 μ F tantalum capacitors (C9, C10) or



DG4RBF

Fig.4: Circuit Diagram of PLL Section
U-Regel zum VCO = Switch-over control to VCO
HF von VCO = HF from VCO, Daten = Data

of a 4.7µF foil capacitor, in series with a 10kΩ trimmer. Parallel to this is a 470nF capacitor (C8).

High-quality capacitors must be used here, with a low loss factor, since the loop filter is decisive for the engaging of the PLL and for the control behaviour.

High-quality foil capacitors would be best suited for this. But as these can be obtained only with difficulty in an acceptable size with a value of 4.7µF, the option was provided for of inter-connecting two good tantalum capacitors.

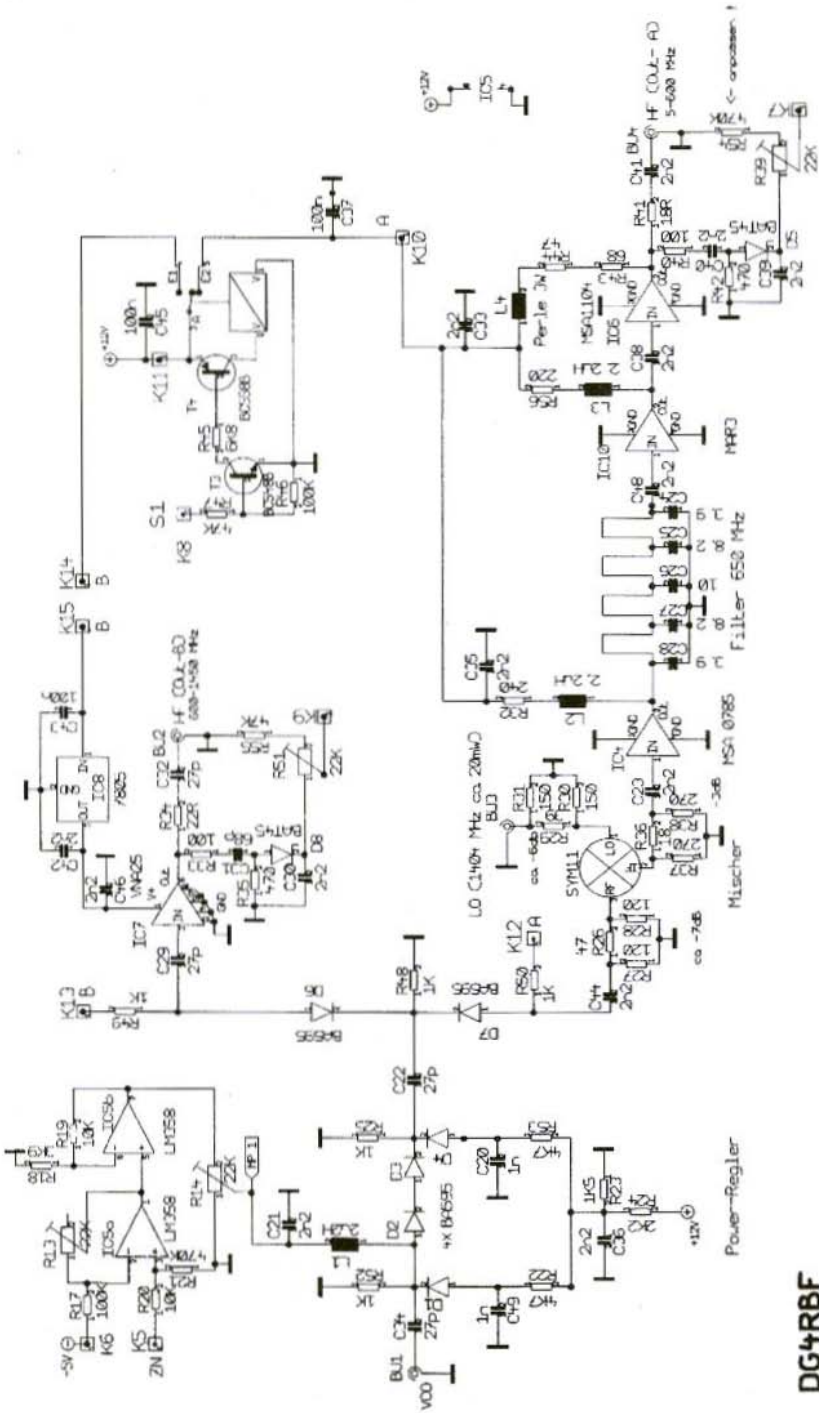
A low-noise DC amplifier with NE5534N (IC3) is wired up after the passive loop filter, and takes the control voltage for the VCO to the 28 - 30 Volts required. The amplification factor

is set through the resistors, R2, R3 and R6.

The + 5V operating voltage is fed through a deep-pass to reduce the noise. There is another deep-pass filter for noise reduction at the operating amplifier output. This deep-pass filter also influences the control behaviour of the PLL, so high-quality capacitors must be used here as well.

The LD signal from pin-14 is used to indicate whether the PLL is engaged or not.

Since the signals are differently polarised on prototypes, depending on the IC used (ES2306 or PMB2306), the option was provided for on the board of evaluating the pulse directly (T4 and LED D1) or inversely (T4, T5 and LED D6).



DG4RB6

Fig.5: Circuit Diagram of HF Section

Power-Regler = Power Control; C.a. = Approximately; Mischer = Mixer; Perle = Bead; Anpassen = Matching



3.1.2. HF Section Circuit

The complete range of the synthesizer is divided into two part-ranges (Fig.5): band (A) - 5 to 600 MHz, and band (B) - 600 to 1,450 MHz.

The range of band (B) corresponds to the direct frequency of the VCO, with the range of band (A) representing the mixed product of the VCO signal and the LO signal (1,404 MHz).

Only the lower mixed product is used, which gives a VCO range in band (A) of 804 to 1,399 MHz.

The software for the synthesizer is laid out so universally that it is even possible to use a different LO frequency in the 1,400 to 1,450 MHz range, if it is divisible by 5 kHz.

The high-frequency VCO signal is fed in at the SMC jack (BU1/VCO).

First comes a PIN-diode controller, which essentially consists of four BA595 diodes. The use of four PIN-diodes gives a higher maximum attenuation and a better control behaviour at low power levels.

The control voltage for the power controller comes from the micro-controller assembly and is connected at point [K5]. The incoming 0 - 2.5V control voltage is provided with an offset in the first OP (IC5a), which can be set using R13. The second OP (IC5b) increases the voltage somewhat, so that it is matched to the advanced control of the PIN-diode controller. The maximum voltage is set through the voltage scaler, R14.

The PIN-diode controller is followed by an HF change-over switch, which was

also created using two BA595 PIN-diodes.

Depending on whether [K12] or [K13] is receiving voltage, diode D7 (band A) or diode D6 (band B) is switched through.

The change-over can also be carried out through a control signal from the micro-controller assembly, which is fed to [K8] (S1). The relay this controls switches the control voltage for the HF change-over switch, and simultaneously the supply voltage for the amplifier stages of band A and band B.

3.1.2.1. Band A Circuit Description

The HF signal arrives at the ring mixer through an attenuator (app. - 6dB in the specimen unit). The mixer used is type SYM 11 from Mini-Circuits, which is comparable to the SMD-C3 ring mixer.

The ring mixer is linked to all three connections with a 50-Ohm attenuator, which creates stable operating conditions for the mixer, and it does a good job.

Make sure that the power at the mixer input (RF) is no higher than is necessary for an output power of 30mW. This prevents unnecessarily high harmonics.

The attenuator at the LO input should be dimensioned in such a way that the level at the mixer is app. 5mW, depending on the output power of the LO.

The signal from the mixer is passed on to the first amplifier stage (IC4, MSA 0785) with an amplification of app. 12dB. This is followed by a 9-pin-deep-pass filter, which provides

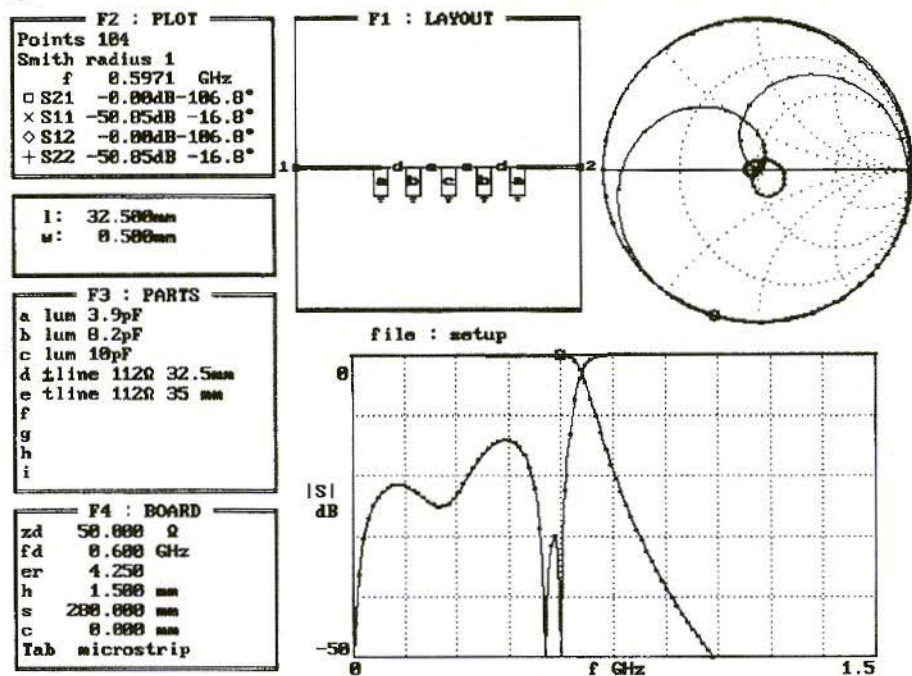


Fig.6: Puff Screen Print-Out of Deep Pass Filter

sufficient attenuation for the upper frequency section of the mixed product and the LO frequency, which also comes through the mixer.

The deep-pass filter was laid out for a limiting frequency of app. 650 MHz, using the micro-wave CAD software Puff (Fig.6).

At the lowest frequency to be expected, 804 MHz ($1,404 - 600 = 804$ MHz), there is already good attenuation, and secondly we can also still be fairly sure that, even if there are some filter tolerances (base material epoxy!), the 600 MHz can still certainly come through.

The signal then passes through two additional integrated broad-band ampli-

fiers, types MAR3 and MS 1104. These ensure the necessary power level.

The operating voltage of the MSA1104 is fed through two resistors wired in series and a choke, which consists of a ferrite bead with three coils of 0.2mm enameled copper wire.

There is a small detector circuit at the output of the MSA1104 for power control.

A few circuit variations and diodes were tested here, and the results obtained with the favourably-priced BAT45 Schottky diode were very acceptable.

At an output power of 30mW, there must also be a voltage of approximately 1.2V at the DC voltage output [K7] for power control.

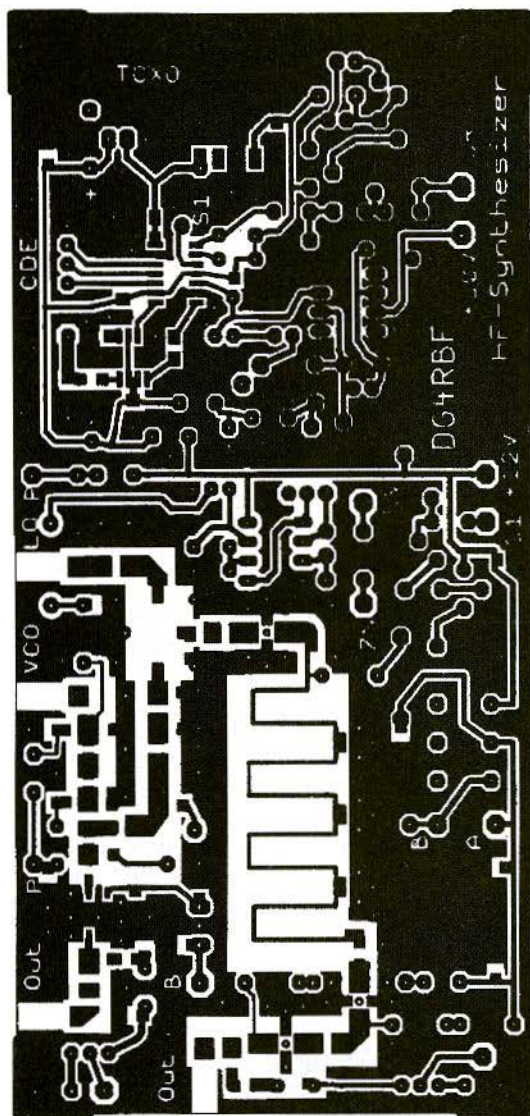


Fig.7: Board Layout of HF Synthesizer

The use of three MMIC's might appear to be going a bit too far at first glance, but there are good reasons for it!

Thus, it would have been possible to use a +7dBm mixer, which essentially is easier to obtain than a high-level mixer.

Moreover, controlling the mixer at low power leads to decidedly fewer harmonics, which means that naturally higher amplification is needed after the mixer. In addition, using two MMIC's as a preliminary stage has advantages if the wobble function of the synthesiser is used.

The combination of the MSA0785 and the MAR3 provides more uniform amplification over the frequency range, and thus gives a smoother amplification line, than if, for example, only one high-amplification MAR8 is used.

3.1.2.2. Band B Circuit Description

This part of the circuit consists of only one MMIC (IC 7). A relatively new, very interesting broad-band amplifier was used here, the VNA25 from Mini-Circuits. This device operates from 500 MHz to 2,500 MHz, with good amplification and a high output power, even in the upper frequency range.

This module has a separate connection for the operational voltage, and can thus be powered direct from a 5 Volt fixed voltage regulator, without a choke. Even the capacitors for the input and output are already integrated into this IC.

The output power reaches the SMC print jack through the small circuit for power detection.

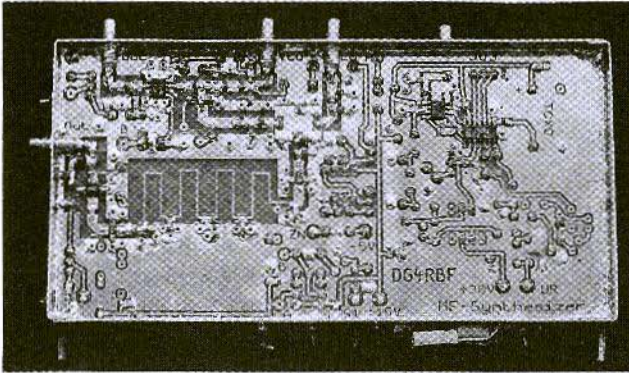


Fig.8:
View of Bottom Face
of HF Synthesizer
Prototype

An MSA1104 did not prove suitable for this amplifier stage!

Some further data on the VNA25 are given in the appendix.

3.1.3. Assembly Instructions

The printed circuit board, unequipped as yet, is prepared first (Fig.7):

- Bores which are not through-hole plated are countersunk on the earth side;
- A rectangular opening (12.7 x 9.5mm) is needed in the board for the ring mixer;
- The SMC jacks are mounted in such a way that the internal conductor lies directly on the corresponding track of the board (Fig.8), which avoids any additional connection inductances.
- 4 rectangular recesses are provided for the SMC jacks (approximately 4 x 7mm);
- 2.2mm bores are provided for the MMIC's: MS 0785 / MAR3, and 4.0mm for MSA1104 (Fig.9);
- The earth through-platings take the form of 1.0 x 2mm or 1.0 x 3mm hollow rivets - use at least 65.

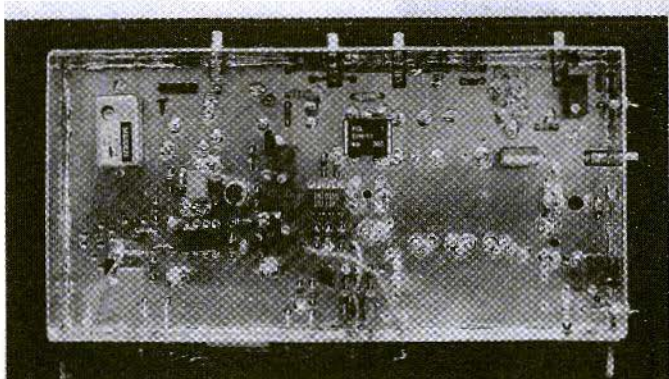


Fig.9:
Components Side of
HF Synthesizer
Prototype



Here it should be pointed out once again that, for high frequencies, hollow rivets provide considerably better through-plating than mere sections of wire. You should also make sure the hollow rivets are definitely well soldered to the upper and lower sides of the board. But they should not fill up with solder.

- Fit the board into the tinplate housing (74 x 148 x 30);
- Make 5mm holes for the SMC jacks.
- Make holes for feedthrough capacitors.
- Solder SMC jacks in. The board is soldered to the tinplate housing all around on the top and bottom faces.

The board can then be fitted, as in Fig's. 10 and 11, with the usual precautionary measures against static charging.

Component list for PLL / HF section:

1 x	TCXO 4 - 100 MHz (e.g. 12.8 MHz)	1 x	78L05, stabiliser
5 x	SMC single-hole panel jacks	1 x	MAR3
1 x	tinplate housing	1 x	SYM11, Mini-Circuits
5 x	feed-through C 1nF	4 x	BC548 B
1 x	10kΩ Cermet trimmer RM 2.54	2 x	1μF, foil / tantalum RM 5
4 x	22kΩ vertical spindle trimmers	1 x	1.5μF, RM 2.5, tantalum
1 x	12V miniature relay	4 x	10μF, tantalum, RM 2.5
6 x	BA595, SMD	1 x	10μF, electrolytic capacitor 50V, RM 2.5
2 x	BAT45, wired	1 x	33μF, tantalum
2 x	LED	2 x	33pF, SMD
1 x	PMB2313T / 2314T Siemens	1 x	330pF, SMD
1 x	PMB2306 T	5 x	1nF, SMB
1 x	NE5534 N, DIL 8	16 x	2.2nF, SMD
1 x	MSA0785	1 x	330nF, ceramic, RM 2.5
1 x	MSA1104	1 x	470nF, plastic, RM 5
1 x	LM358, DIL 8	1 x	100pF, ceramic, RM 2.5
1 x	VNA25, SMD	3 x	100nF, SMD
1 x	7805, stabiliser	2 x	3.9pF, SMD
		2 x	8.2pF, SMD
		1 x	10pF, SMD
		4 x	27pF, SMD
		1 x	68pF, SMD
		3 x	2.2μH choke
		1 x	3mm ferrite bead with 3 windings 0.2mm CuI
		1 x	4-pole pin-strip, 2.54 spacing
			<i>SMD resistors:</i>
		2 x	120Ω
		2 x	150Ω
		2 x	47Ω
		2 x	100Ω
		2 x	18Ω
		1 x	39Ω
		1 x	22Ω
		2 x	470Ω
		2 x	270Ω
		5 x	1kΩ
			<i>0204 wired resistors: 1 x 47Ω</i>



1 x	68 Ω
2 x	220 Ω
1 x	240 Ω
2 x	1k Ω
2 x	1.5k Ω
1 x	2.2k Ω
1 x	3.9k Ω
4 x	4.7k Ω
2 x	470 Ω
1 x	6.8k Ω
4 x	10k Ω
4 x	47k Ω
1 x	82k Ω
2 x	100k Ω
2 x	470k Ω

3.1.4. Commissioning and Calibration

If the micro-controller assembly, the VCO and the LO are functioning, and if all the necessary connections have been made, the PLL / HF assembly can be calibrated in accordance with the following sequence.

3.1.4.1. Calibration of PLL

A frequency meter which should manage app. 50mW is connected to the band B output. An attenuator is provided for if necessary.

Now enter the 600.000 MHz frequency through the keyboard, and keep turning at trimmer P1 (loop filter) until the PLL engages and the LED goes out.

The frequency meter should now display app. 600 MHz. The frequency can be more precisely calibrated at the reference oscillator 9TCXO).

Also check whether the PLL also engages at 1,450 MHz in the upper range.

If necessary, re-adjust P1 a little, until the PLL engages reliably in the lower and upper frequency ranges.

3.1.4.2. Calibration of PIN-Diode Controller

Start the micro-controller in "Test mode / max. H-F power" (switch on and press key 3).

Set R14 to maximum, so that maximum voltage is present at MP1.

- Set synthesizer to 600 MHz.
- Set VCO power to app. 10mW.
- Connect a mW meter at band B output.
- Select menu item "Set Up", and then "Power Calibration".
- If possible, switch the mW meter to a narrow range (app. 1mW) and set the offset (min. voltage) through R13 in such a way that the HF power begins to increase immediately (voltage at MP 1 app. 1.0 to 1.2V).
- Now set the mW meter to a range which can display app. 40mW, and press "Down" key once.
- The (0) on the LCD display will change to (-1), which is equivalent to (255). This means the PIN-diode controller is fully controlled. Now set R14 back until the HF power is just beginning to decrease. At this point, you should have an HF power of app. 30 to 35mW.
- If not, correct the VCO power.

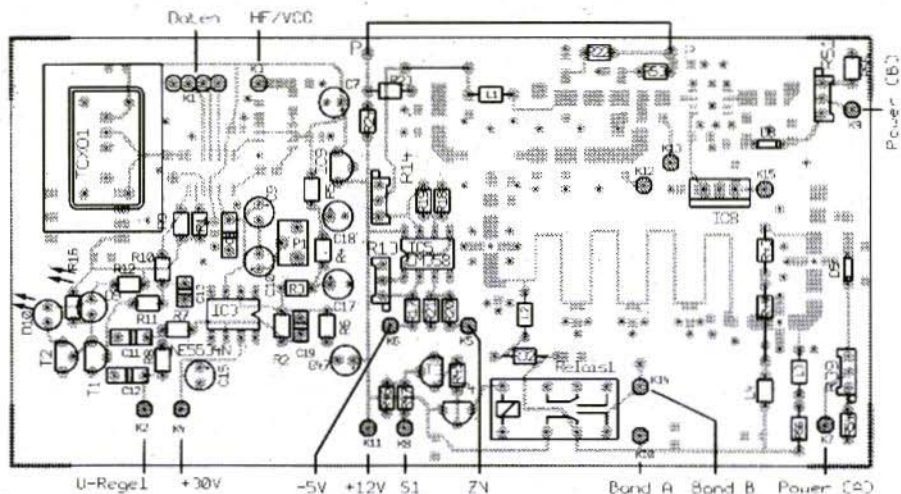


Fig.10: Components Plan of Top Face of HF Section

Daten = Data, U-Regel = Switch-over control

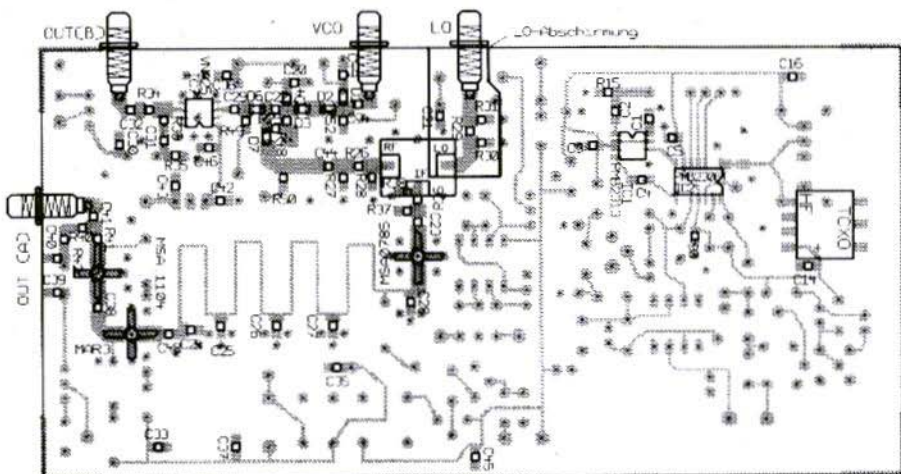


Fig.11: Components Plan of Bottom Face with SMD Components

Abschirmung = Screening

- By switching from the minimum power (0) to the maximum power (-1) you can check the settings again, as the two trimmers have some reciprocal influence on each other.
- This setting allows you to use the full 8 bits (0 - 255) of the PIN-diode controller.

3.1.4.3. Band B Power Recognition (600 - 1,450 MHz)

You are still on the menu item "Power Calibrate". Use the "Up / Down" keys to set the power to 30mW, and then use R51 to set a voltage of 1.20 Volts at [K9]. The "Power Calibrate" program has so far been used only to switch the PIN-diode controller into a specific state. The calibration itself will be carried out a little later.

Now leave the "Power Calibrate" program by pressing the star key until you reach the end of the calibration program (* = "Do not store value, or simply switch the micro-controller off and then back on.

3.1.4.4. Band A Tuning (5 - 600 MHz)

Check whether a reading of at least 30 to 35mW is obtained throughout band A. Use "Test Mode / Max. HF" to do this. If necessary, adjust attenuator at mixer input.

3.1.4.5. Band A Power Recognition (5 - 600 MHz)

Enter a frequency in band A - for example, 300 MHz - call up "Power Calibrate" again, and repeat the process set out in Section 3.1.3.3., but with the difference that the voltage is set at [K7] with R39.

This completes the hardware calibration. We now need only run through the two calibration programs ("Power Calibrate" and "Frequency Response, and then the HF synthesizer is ready for operation.

3.1.4.6. Power Calibration

Connect up an mW meter, select a frequency in band B - 800 MHz, for example - and then start the "Power Calibrate" program.

Now you just need to set the specified power, using the "Up / Down" key, and confirm it with [# = Return].

The program runs through power stages from 0.5 to 30mW.

If you press the star key [*] instead of [#], the value is not stored.

In this way, you can re-calibrate individual values only.

3.1.4.7. Frequency Response Calibration

Here a specific frequency (5 - 1,451 MHz) is pre-set and selected by the program. You need only set the specified power value, which is 10mW, using the "Up / Down" keys, and confirm it with [# = Return].

If you press the star key [*] instead of [#], the value is not stored.

In this way, you can re-calibrate individual values only.

3.1.4.8. Fault Finding

What do you do if the PLL does not engage? Here are some tips:

- Check the soldering of the SMD IC with a magnifying glass.
- Check the voltage (5V) at the PLL and the pre-scaler.



Spektralanalysator DB1NV, Version 2.06 (DG4RBF) PC - Plot

Messobjekt: HF-Synthesizer (DG4RBF) 10 MHz/5mW

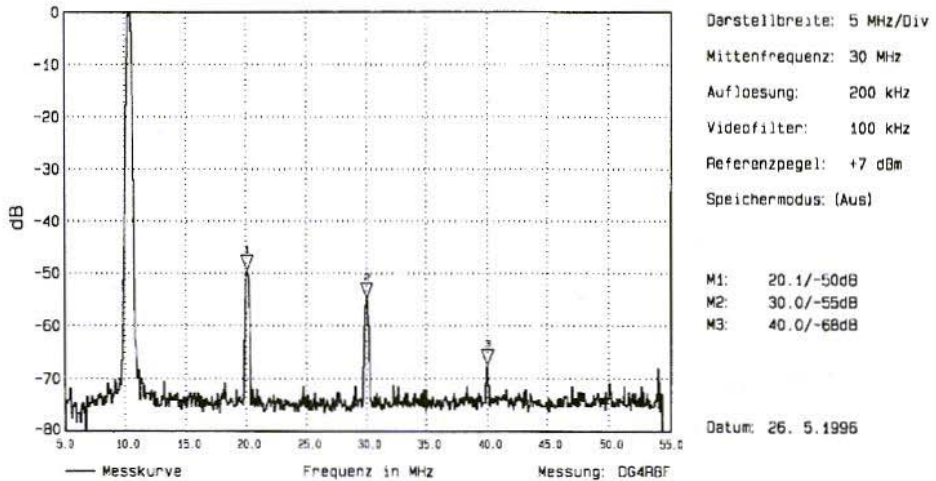


Fig.12: At 5mW Output Power, the First Harmonic is Attenuated by 50dB
Spektralanalysator = Spectrum analyser, *Messobjekt* = Test object,
Darstellbreite = Representation width, *Mittenfrequenz* = Medium
 frequency, *Auflösung* = Resolution, *Referenzpegel* = Reference level,
Speichermodus = Memory mode, *Aus* = Off, *Messkurve* = Measurement
 curve, *Frequenz* = Frequency, *Messung* = Measurement, *Datum* = Date

- Check the data lines (3-conductor bus) with an oscilloscope.
- Set oscilloscope to (2mS / DC / 2V) and check the following signals:
- [EN] = High level, briefly becoming "Low" if frequency changes
- [DA] If frequency changes, a few bytes scurry over the screen (but they can not be synchronised by the oscilloscope!)
- [CLK] Short positive spike pulses!
- Check pre-scaler output (pin-4) with frequency meter or oscilloscope (is VCO frequency / 64 + 1)
- In engaged condition, there should be a DC voltage of app. 0.1 to 5V at (pin-10) of the PLL IC (depending on the frequency)
- R1 (82k) can perhaps also be varied (PD current setting).
- If the PLL fails to engage in the upper frequency range only:
- Check whether the max. control voltage is sufficient to set 1,450 MHz at the VCO (app. 29V)
- The scaler may be receiving too much power at [K3] (max. - 10dBm)

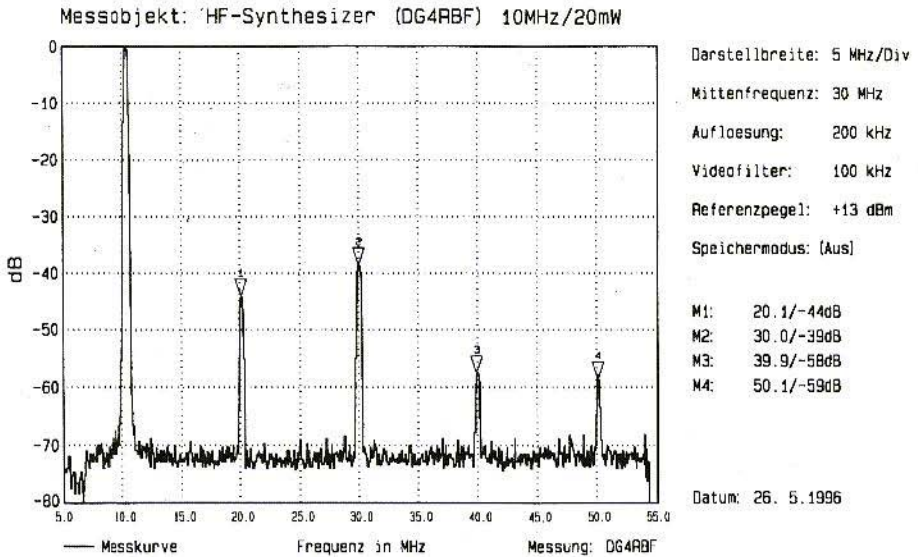


Fig.13: At 20 mW Output Power, the First Harmonic is still Reduced by 44 dB
Messobjekt = Test object, *Darstellbreite* = Representation width,
Mittenfrequenz = Medium frequency, *Auflösung* = Resolution,
Referenzpegel = Reference level, *Speichermodus* = Memory mode,
Aus = Off, *Messkurve* = Measurement curve, *Frequenz* = Frequency,
Messung = Measurement, *Datum* = Date

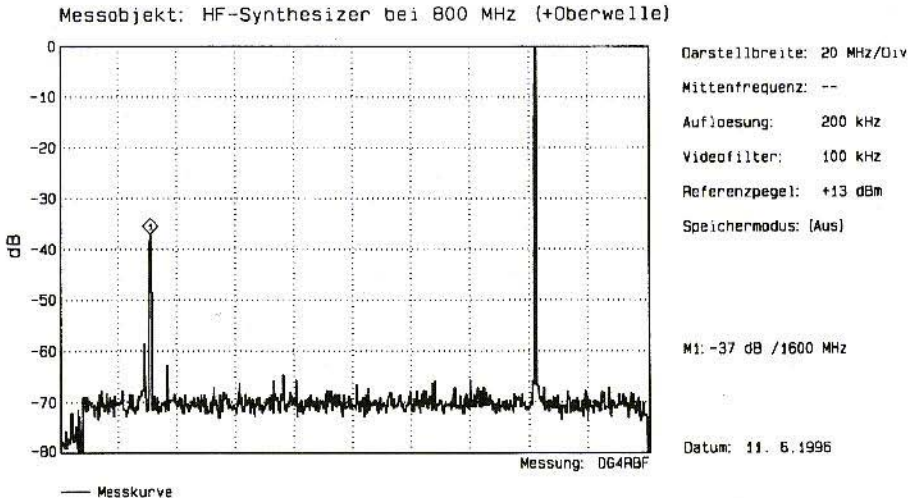
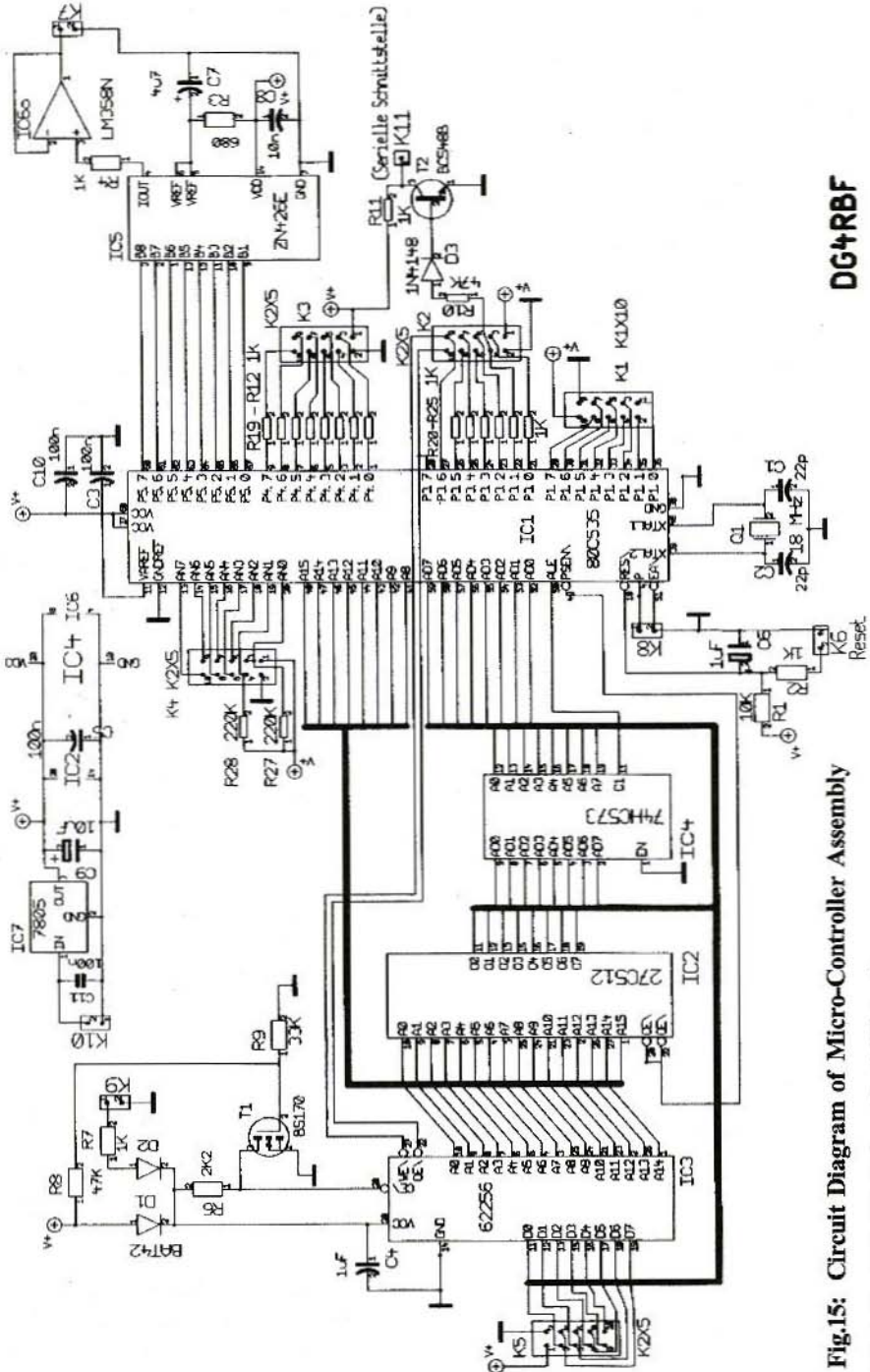


Fig.14: In Band B, the Harmonic is F2 at - 37 dB



DG4RBF

Fig.15: Circuit Diagram of Micro-Controller Assembly
Serielle Schmittstelle = Serial Interface



3.1.4.9. Harmonics

To obtain good harmonic attenuation, it is important that the monolithic amplifiers (MMIC's) are operated at the correct working current. An improvement can often be brought about by varying the working current. The pre-resistors have been optimised in this respect.

The values for the pre-resistors (R bias) should therefore be adhered to. But anyone who has a spectrum analyser can naturally optimise further.

In band A, the output power has a considerably greater effect on the harmonics, as the main decisive factor here is how strongly the ring mixer is controlled.

If, for example, you need only 5mW output power at 10 MHz, you have outstanding harmonic attenuation. The harmonic (F2) is then attenuated by 50dB at 20 MHz and (F3) is attenuated by 55dB at 30 MHz (Figs. 12 and 13).

Fig.14 shows harmonic F2 at 1,600 MHz, with -37dB for a frequency in band B (800 MHz). The signal quality, which is acceptable throughout, is shown further on, in Fig.39 (*Part-2 in issue 3/98*).

3.2. Micro-Controller Assembly (CPU board)

The micro-controller assembly (Fig.15) is responsible for all the control operations relating to the HF synthesizer, the keyboard and the LC display. A standard circuit is used here, with an 80C535 micro-controller from the 8051 family.

This type was selected because it has

enough ports for this application, and an A/D converter with eight switchable inputs is already integrated.

The 80C535 is an 8-bit computer - i.e., the data values of the data bus have a width of 8 bits, so 8 lines are needed (D0-D7).

The 16 bit wide address bus (A0-A15) addresses the program memory.

The high byte of the address (A8-A15) is outputted directly, whereas the low byte (A0-A7) is intermediately stored in a memory, the address latch (74CH573).

This is necessary, since the data (D0-D7) are at the same pins of the processor as the low bytes (A0-A7) of the address. The data and the low byte of the address are then selected in a time multiplex procedure. So a maximum of 64 kbyte program memories can be addressed using the existing 16 data lines (A0-A15).

The assembly also contains an EPROM with 512 kbits (= 64 kbytes), of which about 70% is used for the synthesizer's operating program.

All the variables are stored in a 256-kbit RAM, together with the values determined during calibration. So that the data are not lost when the synthesizer is switched off, the RAM is buffered with a 3V lithium battery (e.g. BR-2/3A).

Transistor T1 monitors the operational voltage, and is responsible for the necessary switching between data buffering and the active condition of the RAM.

This transistor is a self-locking MOSFET, which becomes conductive at approx. 3 Volts gate voltage, and the "Chip

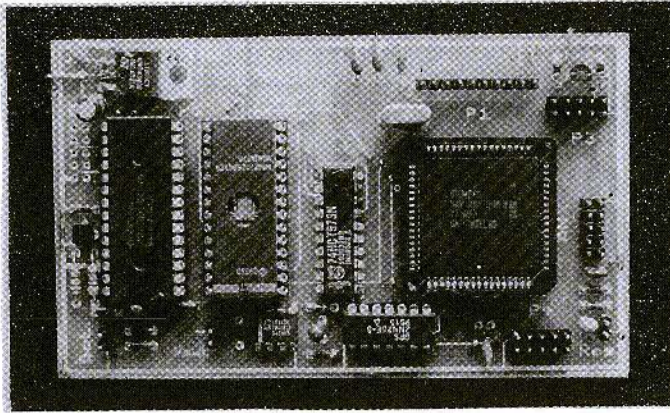


Fig.16:
Prototype of
Micro-Controller
Assembly

Enable" circuit, C/E, of the RAM is earthed, and is thus wired up in the active condition.

If no operational voltage is fed in, the transistor locks and the voltage from the lithium battery switches the RAM into stand-by mode.

The 16 MHz version should be used as micro-controller, as we are working with 18.432 MHz here.

Several processors of the SAB 80 C 535 N 16 type were tested in the prototype trials (Fig.16), and they all functioned at 18 MHz without any problems.

To be continued

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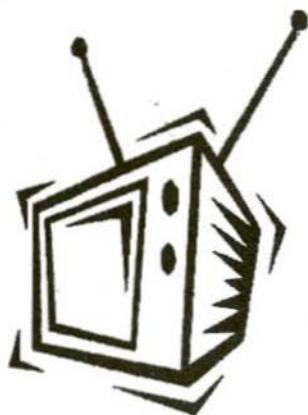
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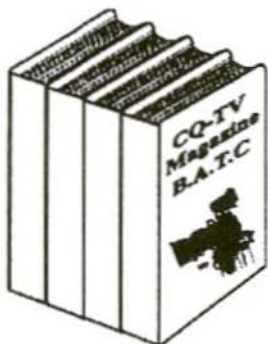
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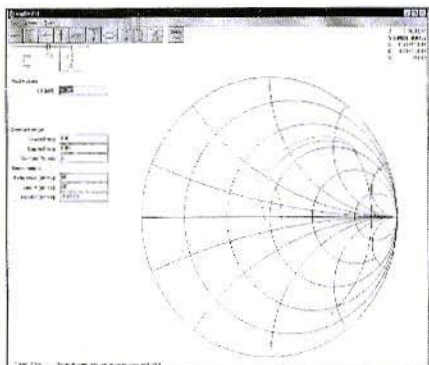
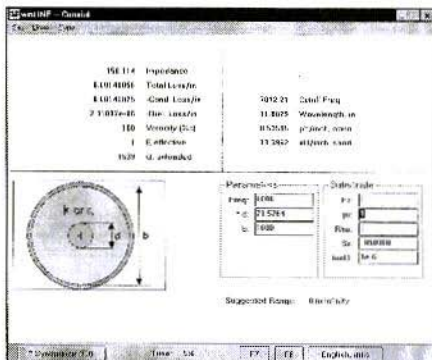
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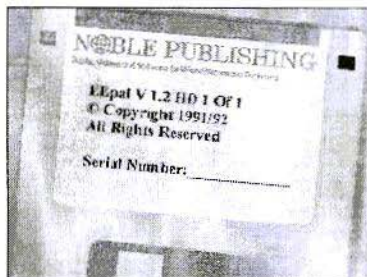
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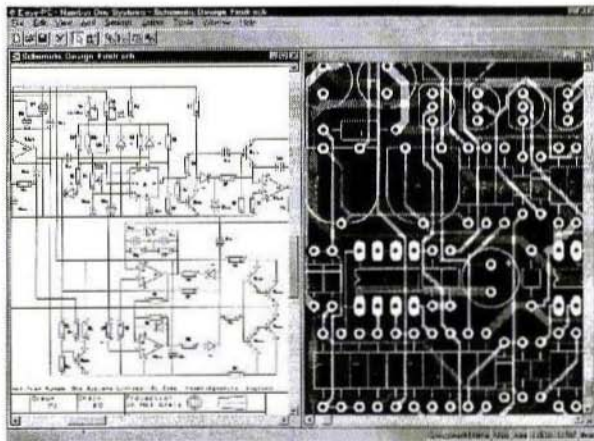
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