VHF
COMMUNICATIONS

Pre Divider (:10) up to 5GHz
Alexander Meier, DG6RBP

Contents

Very interesting articles again with some good constructional projects. From feedback with this years subscriptions most readers prefer the constructional projects.

One of the most popular projects over the past few years has been the spectrum analyser by Matjaz Vidmar. This is a very ambitious project with only a set of printed circuit boards available. I am interested to hear from anyone who has successfully built this project since it would be good to do an article with hints and tips for the construction and testing of the unit. I have also been asked to start a web page devoted to the project so that "would be" constructors can locate some of the more difficult to obtain parts.

Thanks to UKW Berichte the very popular CAD software PUFF is available again, I will be holding a small stock, if you want a copy please use the web site or fax to place an order.

73s - Andy

Wolfgang Schneider, DJ8ES

Frequency Generator (Wobbler) to 4GHz

The following article describes a frequency generator with a wobble function for the frequency range from 10 MHz to 4 GHz. The output is approximately +10dBm (10 mW) with a ripple of less than ±1.5 dB over the whole range. It is split into two overlapping ranges. Automatic switching between the two frequency ranges is not absolutely necessary for use in amateur radio and was therefore not included in this design on cost grounds. This also applies to automatic level correction.

1.
The Design

The frequency generator/wobbler for up to 4 GHz is a useful aid in measurement and calibration work in the amateur radio field. The technical requirements for the design have been restricted to the absolutely necessary. The block diagram (Fig. 2) shows the functional blocks used at a glance, together with their interaction. The core of the circuit is a YIG oscillator. The maximum tuning range of the module used here lies between 1.6 and 4 GHz. The lower frequency range, between 10MHz and 1.8GHz, is covered using a mixer. This has an associated oscillator, a low pass filter and an amplifier. In both frequency ranges, the output is approximately +10dBm, with a ripple of less than ±1.5dB. Better values can not be obtained without a regulated output amplifier, but this was deliberately discounted with for the reasons explained above.

The wobbler is controlled using a micro controller. This sets YIG oscillator according to the frequency range, and generates the saw tooth tuning voltage required for the wobble. A two line by 16 character liquid crystal display is used to display the current mean frequency and the span. A 10dB directional coupler is inserted at the radio frequency output of the YIG oscillator. This additional output is for measurement purposes (e.g. frequency counters for calibration) and if applicable for the connecting up of a PLL control circuit for narrow band applications. This is, however, optional and is not covered in this article.

2.
The YIG Oscillator

The great advantage of YIG oscillators is obvious: a YIG oscillator can be electronically tuned over a very wide frequency range, with a very linear relationship between the tuning voltage (and / or current) and the frequency. The response shown in Fig. 3 shows the frequency

Fig 1: Frequency generator (wobbler) to 4GHz

response of the YIG oscillator in the prototype over a range from 1.6 to 4 GHz.

Just a brief note on the theory behind a YIG oscillator: the abbreviation "YIG" stands for Yttrium Iron Garnet. This is a synthetic crystalline ferrite material, consisting of yttrium and iron (Y3Fe6O16). If such a crystal is surrounded by a magnetic field and in addition radio

frequency energy is fed in through a magnetic loop, the element begins to resonate in the radio frequency range. The frequency of resonance is directly proportional to the strength of the magnetic field.

For the practical construction of a YIG oscillator, the magnetic field is generated through an electromagnet. The frequency of resonance is determined by the current
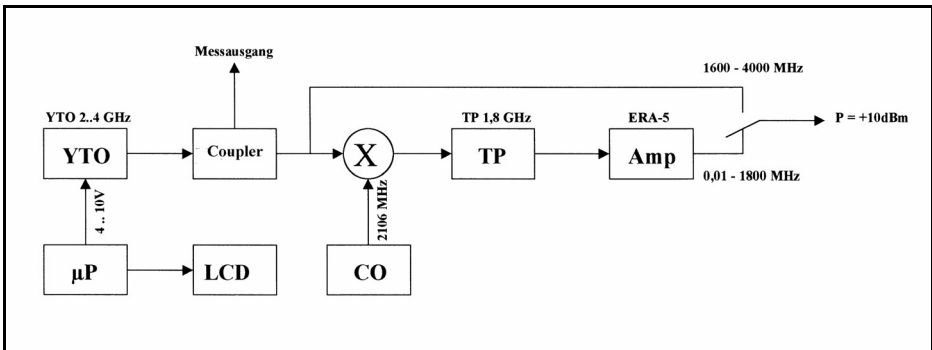
Fig 2: Block diagram of frequency generator.

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Frequency	Input
1600	4,09
1700	4,32
1800	4,55
1900	4,79
2000	5,00
2100	5,23
2200	5,46
2300	5,69
2400	5,93
2500	6,16
2600	6,39
2700	6,62
2800	6,85
2900	7,08
3000	7,31
3100	7,53
3200	7,77
3300	8,00
3400	8,23
3500	8,46
3600	8,68
3700	8,92
3800	9,14
3900	9,38
4000	9,62

Frequency response of YIG Oscillator
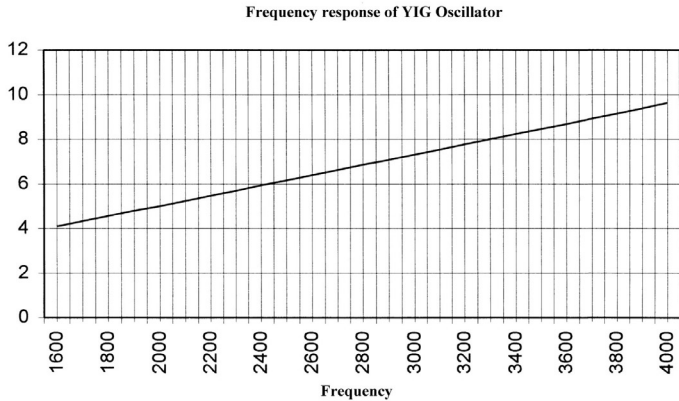
Fig 3: Frequency response of YIG oscillator.

in the magnet coil, and displays linear behaviour. The coils are responsible for the relatively high weight of YIG oscillators. YIG oscillators are available covering the frequency range between 0.5 and 50GHz, in segments. The tuning range usually stretches over one or more octaves. Equipment of this type, in the range of interest for amateur radio between 2 and 18GHz, can be obtained at reasonable prices in surplus stores.

The YIG oscillator used here originates from the HP141T HP spectrum analyser (radio frequency assembly HP8355) and has the following operational data:

- Operating voltage +20 V / 70 mA
  -10 V / 7 mA

- Frequency range 1.6 - 4 GHz

- Output 25 mW ±1 dB

- Coil current 50 - 120 mA

The frequency is set using a micro
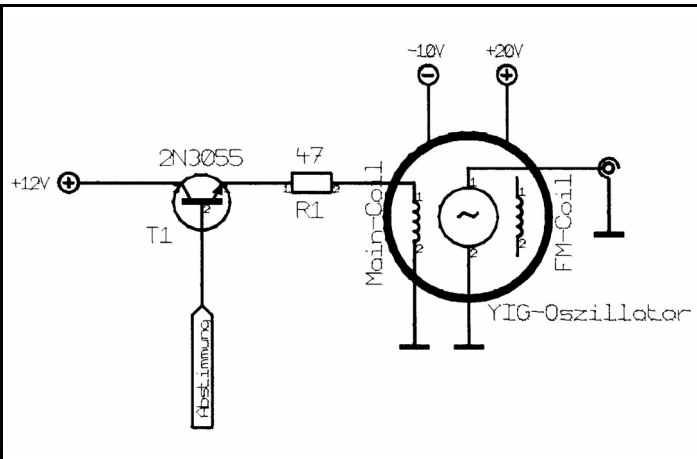
Fig 4: The driver for the YIG oscillator.

Fig 5: Picture of YIG oscillator.

controller with an integrated D/A converter. Since this can not generate the relatively high current required for the tuning coil (main coil), the transistor circuit with a 2N3055, acts as a driver (Fig. 4). It is important that both the transistor and the subsequent series resistance (47Ω) should be thermally stable. Therefore the 2N3055 is mounted on a cooling surface of generous dimensions. A 25 W type was selected for the 47Ω resistor, which was mounted directly on the mounting plate (base plate).
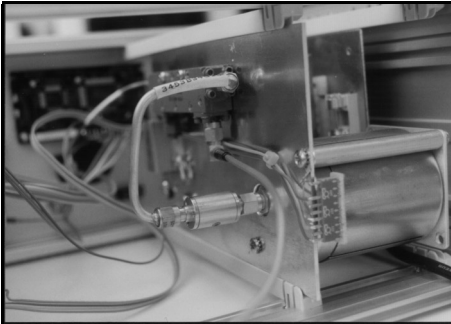
A 3dB attenuator is fitted directly to the output socket of the YIG oscillator with SMA connections (Fig. 5). This isolates the oscillator from the subsequent circuits. The output, including all losses in the attenuator, the directional coupler, the change over relay and the cable, is almost exactly 10mW.

3.
Mixer With Oscillator, Low Pass Filter and Amplifier

3.1. Mixer assembly

For the lower frequency range up to 1.8GHz, the output frequency of the YIG oscillator must be mixed down. The mixer used is an ADE-42MH from Mini Circuits, which is operating in the specified frequency range. In this SMD module the LO and RF ports are specified for the frequency range 5 to 4,200MHz and the intermediate frequency port for 5 to 3,500MHz.

The mixed signal is fed through a 21 pole low pass filter using stripline technology. This suppresses both the image frequency and the signal from the local oscillator at 2,106MHz. Fig. 7 shows the very good frequency response of the low pass filter. Next comes a single stage amplifier using an ERA-1 MMIC which provides a high degree of amplification of approximately 19dB over a wide frequency range. The output level of the mixer assembly is approximately +10dBm (10 mW).

3.1.1. Assembly tips for the mixer module

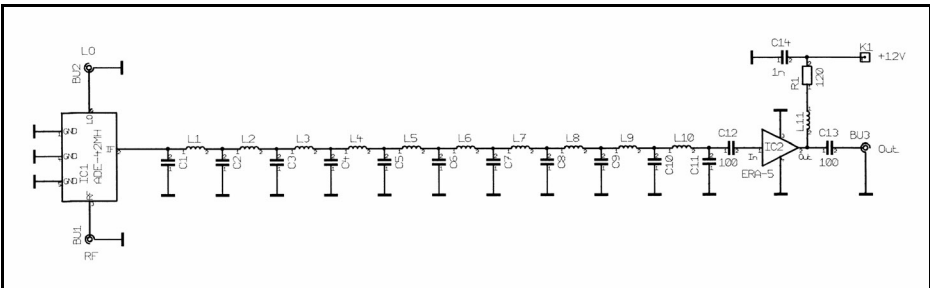A double sided copper coated circuit board made from epoxy material with dimensions 53.5 mm. x 146 mm, was

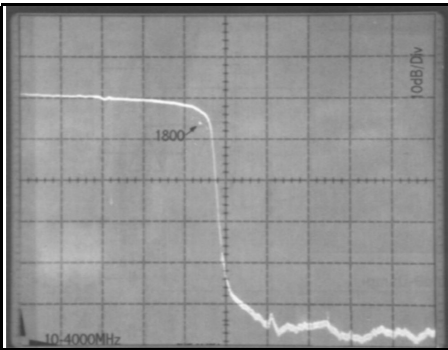**Fig 6: Circuit of mixer with oscillator and low pass filter.**

Fig 7: Frequency response of 21 pole filter.

designed for the mixer assembly. First the printed circuit board, still without the components, is soldered into a suitable standard tinplate housing (55 mm x 148 mm x 30 mm). Then the ADE-42MH ring mixer, ERA-5 MMIC and the two 100pF coupling capacitors are soldered in on the copper side. The 120Ω resistor and the 1nF blocking capacitor are soldered in on the component side.

The L11 choke consists of 2 turns wound onto a 2mm former being one lead of the 120Ω resistor. The position of the choke is thus at a distance of approximately 1mm above the printed circuit board.

The supply voltage (+12V) is fed through a feedthrough capacitor (1nF, solderable) in the housing wall. At this supply voltage, the current consumption of the circuit is approximately 60mA. Three SMA connectors are used for radio frequency connections, input (LO and RF) and / or output.

3.1.2. Component list for mixer assembly

IC1   ADE-42MH, ring mixer (Mini-Circuits)
IC2   ERA-5, MMIC (Mini-Circuits)
L1-L10 Stripline, printed
L11   2 turns on 2mm - see text
C1-C11 Capacitor, printed
R1   120Ω / 0.6 W, RM 10 mm
C12,C13 100pF, SMD 0805
C14   1nF, EGPU, RM 2.5 mm
1 x   DJ8ES 053 PCB
1 x   Tinplate housing, 55 mm x 148 mm x 30 mm
1 x   1nF, feedthrough, solderable
3 x   SMA flanged socket

3.2. Oscillator for 2,106MHz

A tried and tested design from Michael Kuhne (DB6NT) acts as a local oscillator. The circuit (Fig. 10) has been slightly modified from the original (oscillator for transverter for the 13cms band).

The quartz oscillator runs at 117.0MHz using a U310 FET. This signal is fed through various frequency multiplier stages to reach the desired frequency of 2,106MHz. A tripler (BFR93A) to 351MHz, a further tripler (BFG93A) to
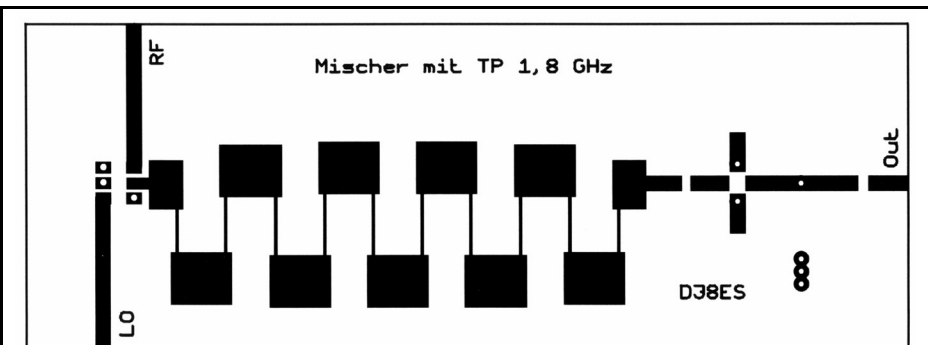

Fig 8: Printed circuit board for mixer DJ8ES-053.
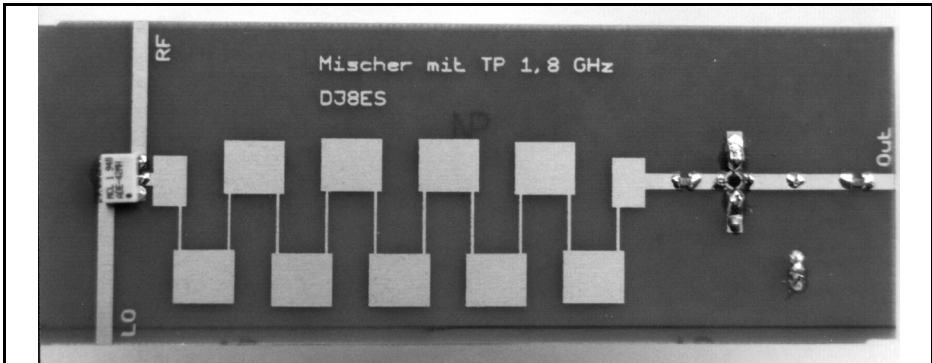
Fig 9: Photograph of the prototype mixer.

1,053MHz and a frequency doubler (BFG93A) to 2,106MHz. The output level here is approximately 2mW.

3.2.1. Assembly tips for local oscillator equipment

The oscillator for 2,106MHz is built on a double sided copper coated epoxy printed circuit board with the dimensions 34 mm x 53.5 mm. The printed circuit board is soldered into a suitable tinplate housing (37 mm x 55 mm x 30 mm).

The DJ8ES 054 printed circuit board is assembled with the mainly SMD component on the track side (Fig. 13). Only the wired components such as the U310 FET, the 117.0MHz crystal and the filters, are inserted in the usual way from components side (fully coated side, Fig.

14). The component drawings show all the details with regard to the position of the components, etc.

The supply voltage (+12V) is fed through a feedthrough capacitor (1nF, solderable). An SMA socket acts as the output for the radio frequency.

3.2.2. Components list for local oscillator

T1	U310, FET
T2	BFR93A, SMD transistor
T3,T4	BFG93A, SMD transistor
IC1	78L09, SMD voltage regulator
Q1	Crystal 117.0MHz, HC-18U, Series resonance 7th overtone
L1	BV5061, 0,1µH, Neosid
L2	5HW-35045A-365, helix filter
L3	5HW-100090A-1010, helix

Fig 10: Circuit diagram of 2106MHz oscillator.

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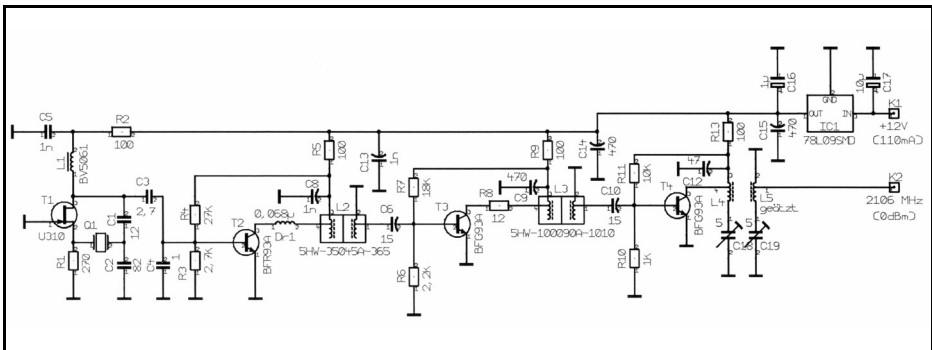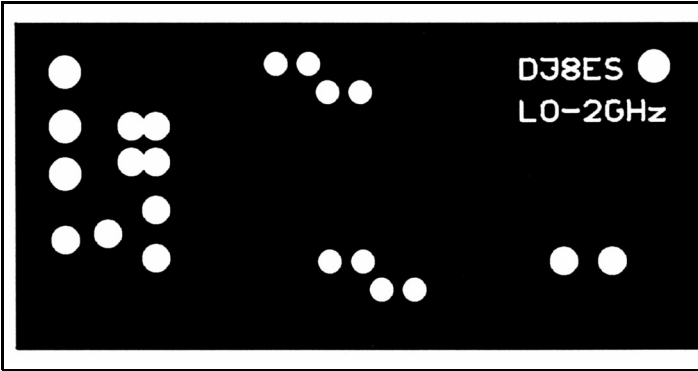Fig 11: Oscillator printed circuit board, ground plane side.

	filter	1 x	82pF
L4,L5	Stripline filter, printed	3 x	40pF
C18,C19	Trimmer 5pF, green, Sky	3 x	1nF
1 x	DJ8ES 054 PCB	1 x	1μF, tantalum
1 x	Tinplate housing, 37 mm x 55 mm x 30 mm	1 x	10μF, tantalum
1 x	1nF, solderable, feedthrough		
1 x	SMA flanged socket		

3.2.3. Putting oscillator assembly into operation

all remaining components in SMD format 1206 or 0805:

1 x	12Ω
4 x	100Ω
1 x	270Ω
1 x	1kΩ
1 x	2.2kΩ
1 x	2.7kΩ
1 x	10kΩ
1 x	18kΩ
1 x	27kΩ
1 x	2.7pF
1 x	12pF
2 x	15pF
1 x	47pF

After a final visual check of the fully assembled oscillator assembly for placement errors or solder breaks, it can be put into operation for the first time. The output is determined using a power meter suitable for this frequency range. The maximum current consumption at +12V is 110mA. However, this value is not reached until setup is complete.

There must be + 9 V at the voltage regulator output (IC1). The voltage drop through resistor R2 should be approximately 0.7V. The crystal oscillator can be tuned, the core of the oscillator coil L1 is slowly rotated while the voltage
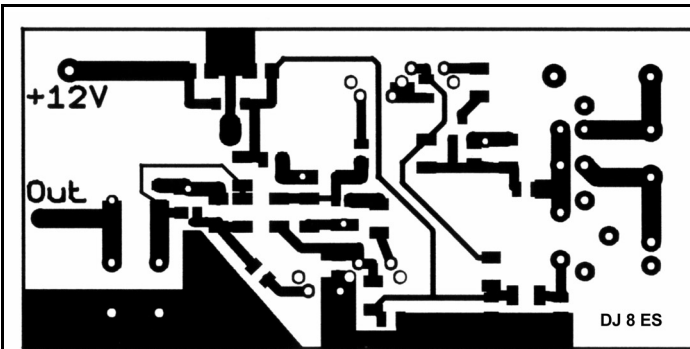
Fig 12: Oscillator printed circuit board, track side DJ8ES-054.
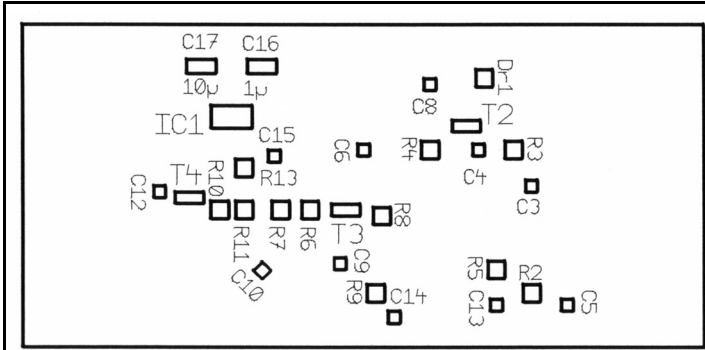
Fig 13:
Component layout of oscillator printed circuit board, track side.
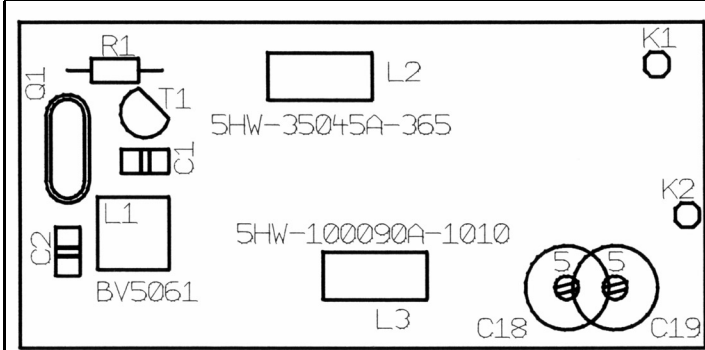
Fig 14:
Component layout of oscillator printed circuit board, ground plane side.

drop is measured across resistor R5. As soon as the oscillator starts to, this voltage drop increases rapidly, reaching a maximum value of 3.4V. The optimal setting for the coil core, is slightly below the maximum on the slow rising slope.

Next the tripler (T2, BFR93A) is tuned to 351MHz. To do this, the voltage drop across R9 is used. Using reciprocal tuning of the two circuit helix filter L2, an unambiguous maximum reading (approximately 3.4V) can be obtained. The same applies for the tripler to 1,053MHz. Here the voltage drop across R13 (3.4V) acts as an indicator. This time L3 should be tuned (again reciprocally). The doubler to 2,106MHz should be tuned by
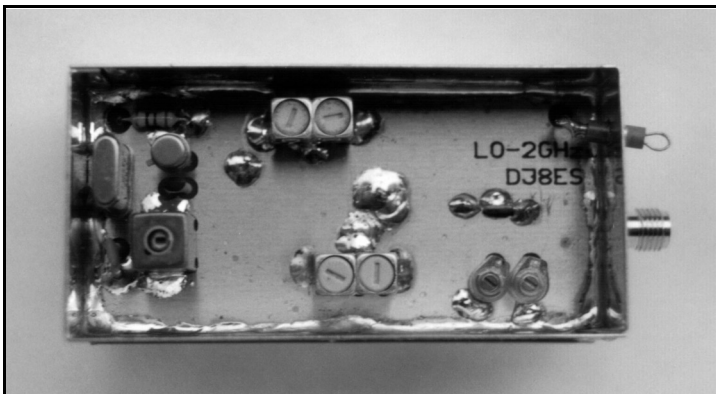
Fig 15: Prototype of the 2106MHz oscillator.

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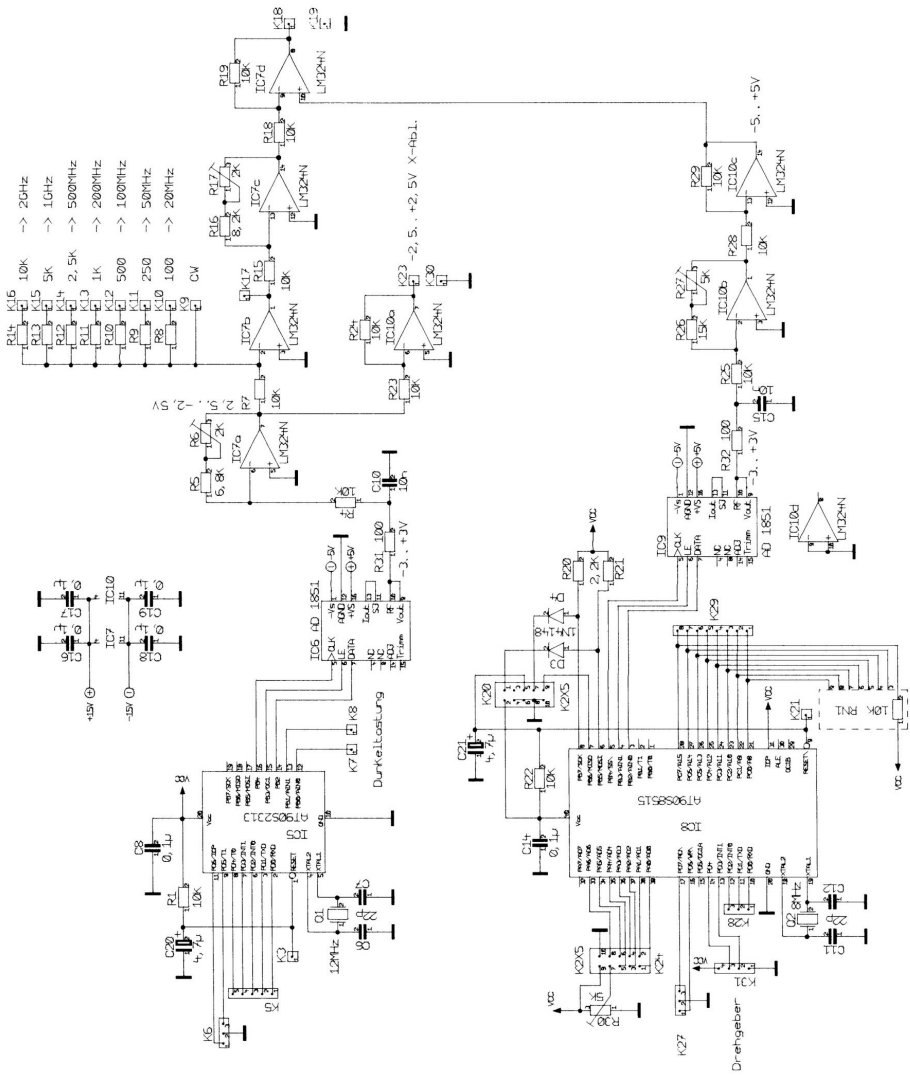
Fig 16: Circuit diagram of micro controller.
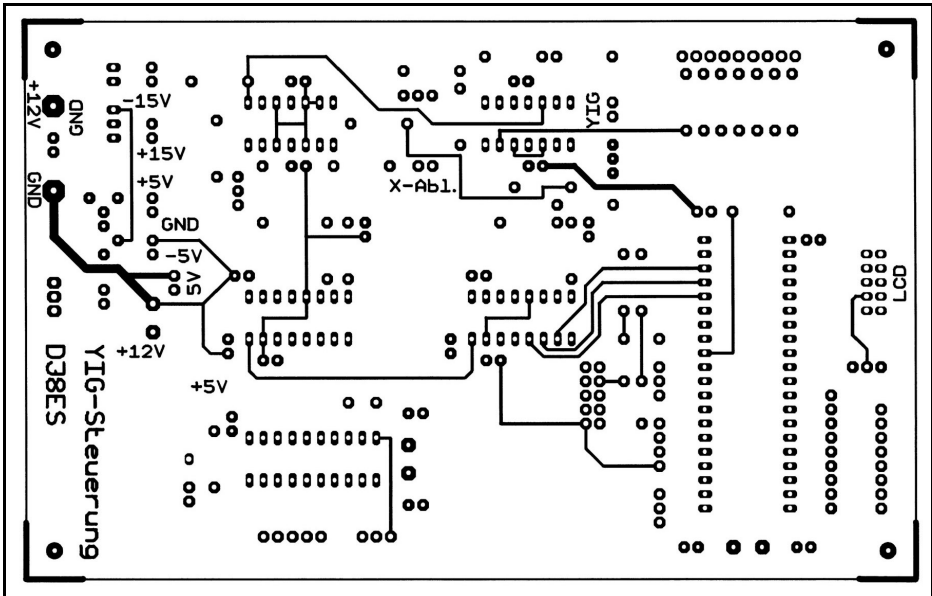
Fig 17: Printed circuit board for micro controller, top side DJ8ES-055.

means of the two 5pF Sky trimmers (C18, C19). Here the output of approximately 2mW should be obtained.

4.
Micro controller control

The micro controller circuit (Fig. 16) is divided into two functional parts. A µP AT90S2313 (IC5) micro processor generates the saw form tuning voltage for wobbling. This processor also generates the blanking signal during the flyback on the display. The second micro processor, a µP AT90S8515 (IC8), sets the mean frequency through a shaft encoder and controls the LC display.

The circuit requires a single positive supply voltage of +12V. Other voltages required are generated on the circuit board:

- ±15 V for the operational amplifier with IC1 (NMA1215S),

- ±5 V for the analogue/digital converter with IC2 (78L05) and IC3 (79L05) and

- +5 V for the micro-controller and the LC display with IC4 (7805).

Each micro controller generates a 16 bit digital word that is converted by the digital to analogue converter to give the required voltages for the frequency setting. Downstream amplifiers provide level adjustment to standard values. Thus, for example, the output for the horizontal deflection is defined on a display screen (e.g. oscilloscope) as ±2.5 V. For precise calibration, 3 precision trimming capacitors are provided. In the last operational amplifier, the two analogue signals are combined and used to drive the YIG oscillator.

The amplification of the operational amplifier IC7b can be switched in stages to set the span. The values given for the resistors in the feedback loop allow a span of 2GHz, 1GHz, 500MHz, ... right down to 20MHz. The CW switch posi-

Fig 18: Printed circuit board for micro controller, bottom side.

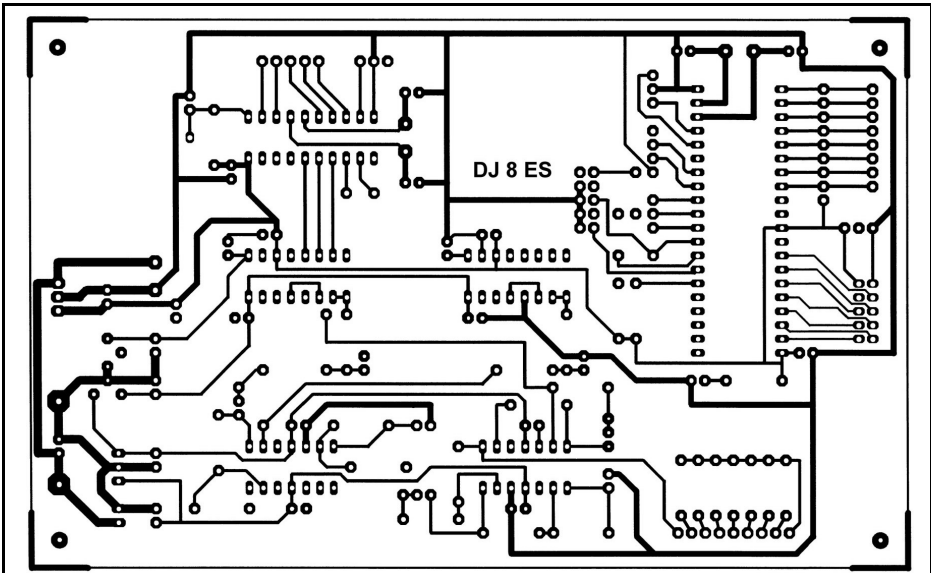tion makes standard signal generator operation possible at a frequency that can be set by means of the shaft encoder. The sweep width is not used in this mode. The frequency stability is better than 100kHz over the entire frequency range. Any improvement for genuine narrow band measurements (quartz filters or the like.), can be made using an additional frequency control circuit with a PLL.

To set the wobble speed, 5 connections are taken from IC5 to a socket strip (K5). For faster wobble speeds, one of these connections can be switched to earth. Inverted blanking signals are available for flyback with K7 and K8. K28, pin 1 makes it possible to switch between the upper and lower frequency ranges on the liquid crystal display and pin 3 makes it possible to switch between normal operation and frequency calibration.

For the settings lower frequency range and frequency calibration, the connection pin in question should be switched to earth. Pin 2 is not used in the current software.

4.1. Assembly instructions for micro controller module

The micro controller module occupies a double sided coated epoxy printed circuit board, with the dimensions 100 mm x 160 mm (European standard size pc board) (Figs. 17,18). In contrast to the oscillator assembly, only wired components are used, and no SMD parts are used. The components layout is shown in Fig. 19. The micro processors and A/D converter should be mounted using IC sockets. It is more sensible for all inputs and outputs to use plug connections. The LC display is connected to K24 as per list:

Pin 1:   11, DB4
Pin 2:   12, DB5
Pin 3:   13, DB6
Pin 4:   14, DB7
Pin 5:   6, EN
Pin 6:   4, RS
Pin 7:   3, VEE (LCD drive)
Pin 8:   5, R/W
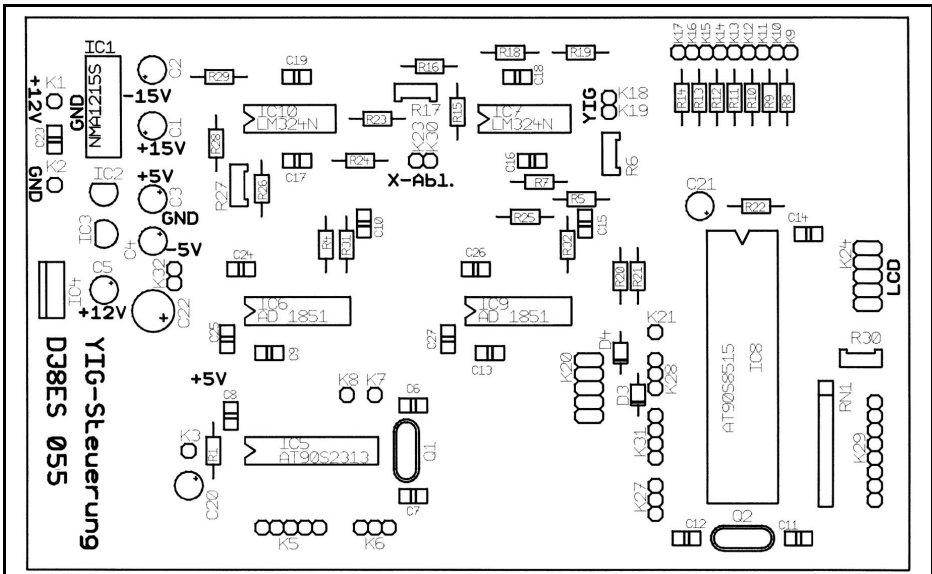Pin 9:   2, VDD (+5V)
Pin 10:   1, VSS (GND)

Fig 19: Component layout for micro controller printed circuit board.

The 5kΩ trimming potentiometer (R30) adjusts the contrast of the liquid cryatal display. For illuminated displays, a stabilised +5V voltage is available on K32. The connection usually goes through an external 12Ω resistor.

4.2. Component list for micro controller module

IC1	NMA1215S, voltage converter
IC2	78L05, fixed voltage regulator
IC3	79L05, fixed voltage regulator
IC4	7805, fixed voltage regulator
IC5	AT90S2313-10, micro controller
IC6,IC9	AD1851, D/A converter
IC8	AT90S8515, micro controller
IC7,IC10	LM324, operational amplifier
Q1	Crystal, 12 MHz, HC18-U,
Q2	Crystal, 8 MHz, HC18-U,
R6,R17	2kΩ precision spindle trimmer
R27,R30	5kΩ precision spindle trimmer

Capacitors:
1 x	1000µF, RM 5mm, electrolytic capacitor
5 x	10µF, tantalum electrolytic capacitor
2 x	4.7µF, tantalum electrolytic capacitor
4 x	22pF, RM 2,5 mm, ceramic
2 x	10 nF, RM 2,5 mm, ceramic
13 x	100nF, RM 2,5 mm, ceramic

Resistors, ¼ W, RM 10 mm:
2 x	100Ω
1 x	2.2kΩ
1 x	6.8kΩ
1 x	8.2kΩ
10 x	10kΩ
1 x	15kΩ

Precision resistor, ¼ W, RM 10 mm:
1 x	100Ω
1 x	250Ω
1 x	500Ω
1 x	1kΩ
1 x	2.5kΩ
1 x	5 kΩ

1 x	DJ8ES 055 PCB

4.3. Putting into operation with setup of micro controller module

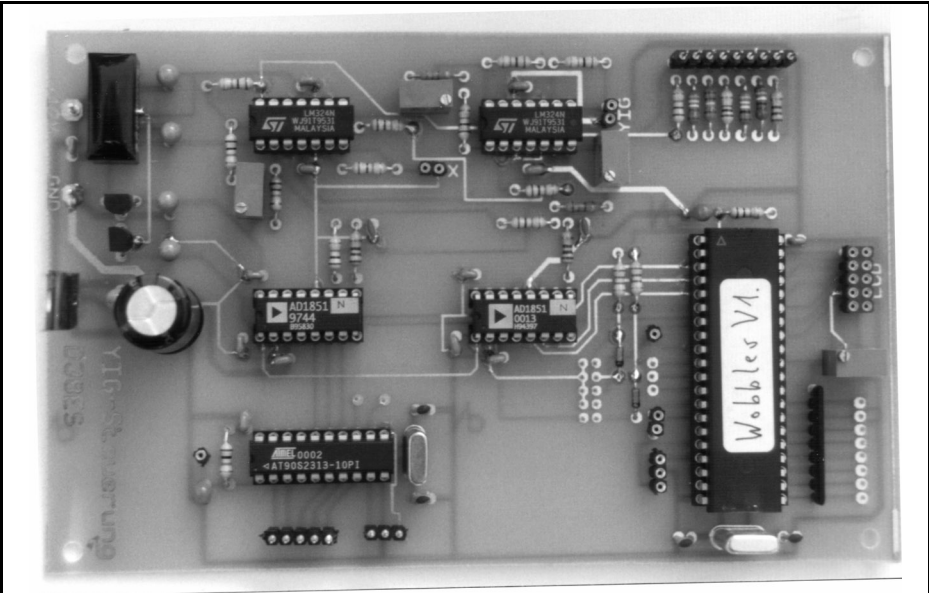When the micro controller module is

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Fig 20: Prototype of micro controller board.

switched on, the tuning voltage generated through the FC AT90S2313 (IC5) starts automatically. The waveform and voltage level can be monitored at connection K23. Positive or negative blanking pulses, each with a pulse length of 10ms, are fed to K7 and K8. A low level at K6/pin 3 is used to switch the micro controller to the lower value. Using potentiometer R6, this minimum value is set to precisely 2.50V. A low level at K3 (Reset) re starts the sawtooth.

A low level at K6/pin 1 is now used to switch the micro controller to the upper value, +2.50V can be monitored at K23. In the event of slight discrepancies in either of these values, the operational voltage, either ±15V for the operational amplifier (IC1, NMA1215S) or ±5V for the analogue/digital converter, is not symmetrical from IC2 (78L05) and IC3 (79L05). In this case, the same voltage level, positive and negative, is set at K23 using R6. The testing and adjustment of the tuning voltage for the mean frequency of 5V to +5V at IC10/pin 14 is

carried out in a similar way. This time, the micro controller 90AT8515 (IC8) is switched to the upper or lower maximum value using K27 and calibrated with the precision trimmer R27. The reset connection is available on K21.

To monitor the mean frequency of the YIG oscillator, the sawtooth is switched to CW, i.e. no span, by connecting K9 to K17. The micro controller is switched into calibration mode (K28/pin 3 to earth). Following the reset of the micro controller, the mean frequency is adjusted and programmed. In the prototype, this is the setting value was 22446 for the D/A converter, which corresponds to a frequency of 2,800MHz for the YIG oscillator. 6.85V is now applied at pin 18, or precisely half the voltage, 3.425 V at IC10/pin 14. Fine adjustments can still be made to the frequency using the shaft encoder.

All voltage values are taken from the characteristic of the YIG oscillator shown in Fig. 3. The setting value for the mean frequency is calculated as follows:

Setting value = 32768 * Tuning voltage / 5.0 V / 2
= 32768 * 6.85 V / 50 V / 2
= 22446

To set the span, the shaft encoder remains set to the mean frequency. The span is set to 2GHz (connect K16 with K1). The mean frequencies must be set through the micro controller IC5 and the trimming potentiometer R17; with K6 / pin 3 low minus 1GHz ,and with K6/pin 1 low plus 1GHz , for the mean frequency of the upper maximum frequency. In the specimen apparatus, the following values are valid for this setting:

• 1,800 MHz = 4.55 V

• 2,800 MHz = 6.85 V

• 3,800 MHz = 9.14 V

This gives a voltage range of 9.14V 4.55V = 4.59V for a maximum sweep width of 2GHz. For the final programming of the micro controller IC8 (AT90S8515) by Frank Peter Richter, the settings just obtained are important. The correct display for the frequency can now be programmed for the upper and lower frequency range in the micro controller.

The values for the specimen apparatus are:

• Oscillator frequency: 2,106 MHz

• Mean frequency YIG: 2,800 MHz

• Shaft encoder setting:
22464 (rounded off, must be divisible by 64)

• Frequency interval per step: 0.13333 MHz per shaft encoder step

If this design is copied using a YIG oscillator which differs considerably from the one in the prototype, then the rate of rise of the tuning voltage and thus the amplification of IC7c must be adjusted, if applicable. This results in a re calculation of resistor R16. In the proto-type, the following values apply to this resistor:

R16 = (R15 * dispersion / 5 V) - (R17 / 2)
= (10000Ω * 4.59 V / 5 V) - (2000Ω / 2)
= 8180Ω

Thus for R16 a standard value is selected (E24 series) of 8.2 kΩ.

5.
Operational experience and prospects

This frequency generator (wobbler) up to 4GHz has already proved its worth many times both as a test transmitter and as a wobbler (e.g. calibration of band filters) in the VHF/UHF and SHF frequency ranges. The principle of this apparatus is already being used in a further development (spectrum analyser up to 1.8GHz or as slimmed down version up to 500MHz).

Suitable detectors must be used for this frequency. DIY products, even in coaxial formats, can frequently be used only up to frequencies of 2GHz. Beyond this, the input impedance of the detector is far beyond 50Ω and does not allow any reliable measurements to be carried out.

If a standard oscilloscope is used as a display screen, the blanking of the fly-back, if applicable, can not be accomplished without modification. One conceivable plan is to short circuit the detector DC voltage with an FET (e.g. BS170) using a small additional circuit. Alternatively, a micro controller with a delta voltage is also available instead of the sawtooth tuning voltage. However, because of the electrical behaviour of the YIG tuning coil due to the back electromotive force, this only works for slow wobble speeds.

At this point, I would like to express my

sincere gratitude to Frank Peter Richter (DL5HAT) for the development of the software and for volunteering to prepare individually programmed micro controllers. In this connection it is conceivable that additional functions could be implemented into the software. Any suggestions along these lines will be particularly welcome.

6.
Literature

[1] Bert Kehren, WB5MZJ Microwave components, Proceedings of 45th Weinheim VHF Congress

[2] Michael Kuhne, DB6NT, 13cm linear transverter; DUBUS 3/1993

Dipl. Ing. Detlef Burchard, Nairobi

A Simple Procedure for Measurement

An example of the AD606 as a logarithmic amplifier

In the field of RF measurement technology, there are some very good and very accurate procedures, although they are not necessarily what everyone can achieve.

However, the properties of components used in radio technology can be measured using relatively simple means, as is described in the example below.

1.
Introduction

There has already been an article in this magazine by Jirmann [2] on the good characteristics of the AD 606 analogue device (Fig 1 shows the pin connections for the AD 606), with evidence obtained from the use of an RF shielded, high power signal generator and a precision calibration measurement circuit. Not every radio amateur has such equipment available. However, anyone who has, as a minimum, an oscilloscope, can use it for methods described in [1] (The article will be placed on the VHF Communications web site to assist understanding of this text - Ed.), as is demonstrated below.

It proved not to be very easy to get hold of an AD 606, but one was obtained,

though it took a certain amount of time and trouble.

2.
Measurement procedure

The physical principles have already been described in [1]. The full measurement circuit is shown in Fig. 2. A few special features have to be kept in mind if the measurement is to be carried out successfully at first go.

- The logarithmic AD 606 intermediate frequency amplifier IC has a relatively wide input voltage range, as is specified in the data sheet. In order to investigate it and make the most of it, the generator should be able to supply an initial output of +15dBm. To obtain this, the voltage on the switching transistor must be increased to approximately 50V. Not every BF 199 can withstand this voltage, and there may be some background noise. So in case of emergency try another unit or use another type of transistor.

- The limiter outputs of the AD 606 may not be left open circuit. Nor may they be connected directly to +5V. In both cases there are feedback effects, which can distort the shape of the
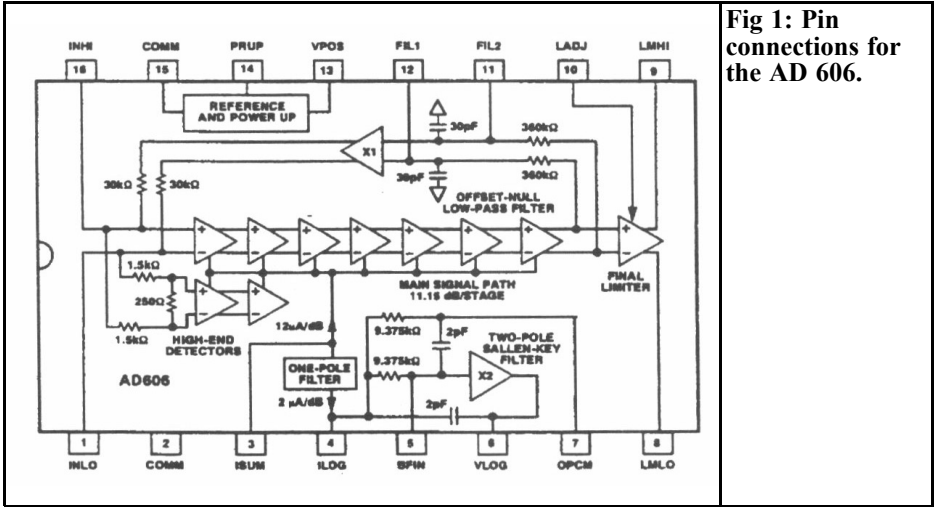
Fig 1: Pin connections for the AD 606.

VLOG output curve below -60dBm. A 51 load resistor reduces these feedback effects to a minimum.

• Ringing prevention resistors are needed on the output pin 6 and on the supply voltage pins 13/14.

• The circuit should be kept as compact as possible to ensure wide band-width and gain of the IC is preserved.

• The Amidon toroidal core is mounted behind a screen made from printed circuit board material or tinplate, so that the coupling to the AD 606 is minimal.

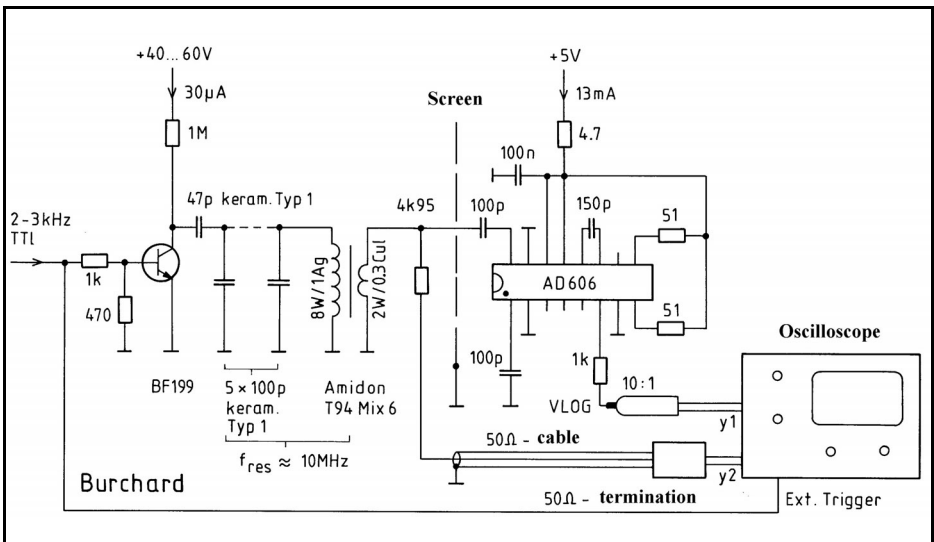• The probe on the input pin 16 is very sensitive and might act as an antenna

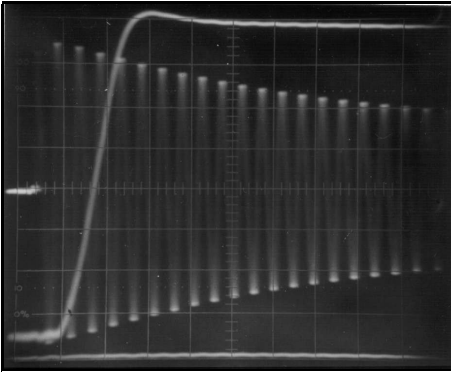**Fig 2: Circuit diagram of the measurement system.**

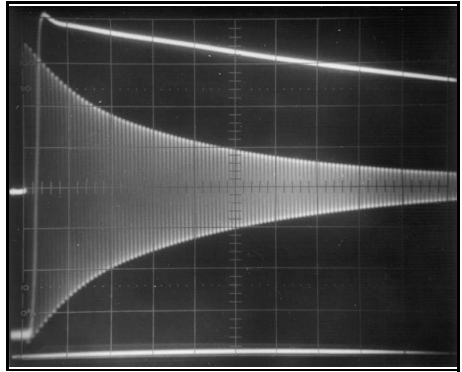**Fig 3: Output display with Y1 and Y2: 500mV/div. X: 200nS/div.**

Fig 4: Timebase extended to give 1/10 amplitude (-20dB), X: 2dB/div.

for interference. So miniature resistors are used (4.95kΩ = 3.65kΩ + 1.30kΩ from the E96 range), with 50Ω coax cable and a 50Ω termination gives a scale factor of 100:1. This does not need any capacitive compensation up to and above 100MHz.

• There should be no fall off in the frequency response of the oscilloscope below 10 MHz.

Let us now look at the details of the procedure and at the curves that are obtained. We need only to adjust the initial voltage by varying the supply voltage of the BF199 to +15dBm = 1.25 Veff = 3.5 Vss.

3.
Understanding the outputs obtained

The logarithmic output voltage, VLOG of the AD606, goes through a low pass filter. As Fig. 3 shows, the time taken for the voltage to build up to the correct value is 400ns, which requires four oscillations of the decay process. It can be estimated that the first decibel of the decay process is thus incorrectly represented. Consequently, if the decline takes place one hundred times more slowly, the distortion for the remaining curve amounts to only an additional 1/100 dB.
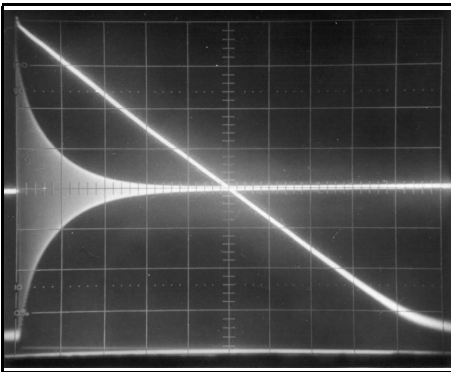

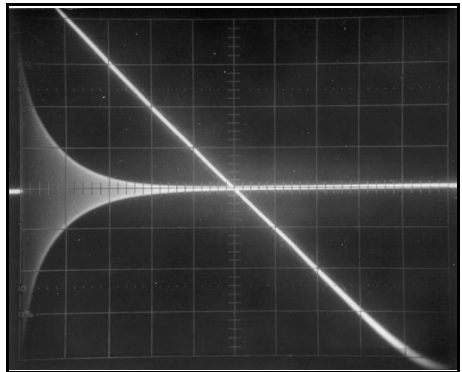**Fig 5: Timebase five times slower, X: 10dB/div.**

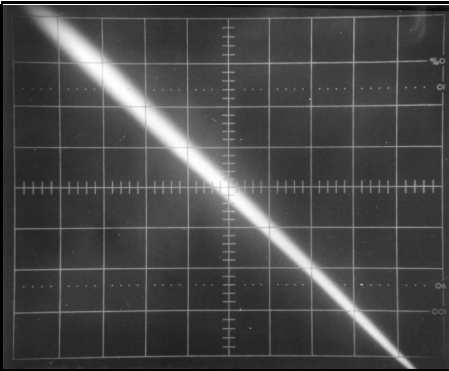**Fig 6: Y1 gain adjusted to centre -45 degree slope.**

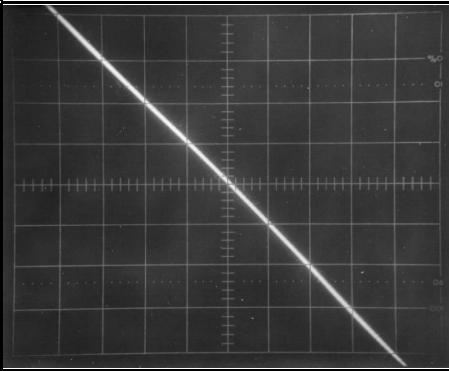**Fig 6a: Amplification of X and Y channels increased and display centred on top part of curve, +15bis -5dBm.**
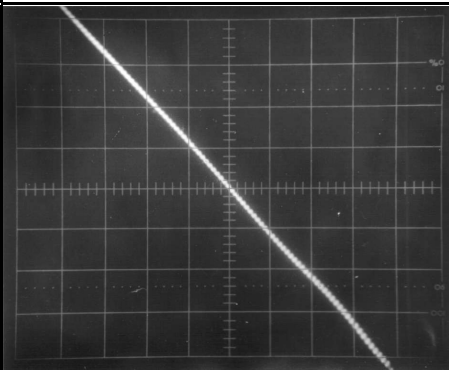

Fig 6b: Display of -25bis -45dBm

Fig 6c: Display of -55bis -75dBm

Now the time base sweep of the oscilloscope is reduced slowly in uncalibrated mode, until the decaying oscillation at the end of the screen has decreased to one tenth of the initial amplitude. The time axis now shows 2dB/section (Fig. 4). It is switched to being five times slower, using the step switch, and we obtain Fig. 5 with 10dB per section. It can be seen at a glance that the VLOG curve displays the desired relationship above approximately 90dB, although the resolution is not yet high enough to check the manufacturers data.

The Y1 amplification is now adjusted in uncalibrated mode, in such a way that the curve goes down to below -45 degrees in the central section and we obtain Fig. 6.

Subsequently, the X and Y amplifications can be further increased in known stages, and the section of the curve that is of interest can be brought onto the screen. For Fig. 7, both axes have been extended by a factor of 5 and every small departure from the ideal 45 degree line becomes visible. In Fig. 7a, the line at high levels can be seen. The manufacturer gives no details on the response, +15bis + 5dBm. It can be seen that the curve runs at an angle slightly greater than 45 degrees, but remains far below the guaranteed deviation of 1.5dB. The slight ripple effect can also be seen. In the central section (Fig. 7b), the line is very flat. A slight sagging of perhaps 0.2dB can be recognised. Finally, in Fig. 7c, we can see that the curve at -75dBm is approximately 0.5dB high, i.e. here it slightly deviates from the typical accuracy limit of 0.4 dB.

4.
Summary

As a whole, the AD606 has excellent characteristics. In practical applications, people will be happy to use the additional unspecified input voltage range between

+5 and +15dBm. If you work within a 50Ω system, then a 10dB voltage amplification can be brought about with a 1:3 step up transformer, which is mounted in front of the AD606 and connected up at 450Ω. The input resistance of the component lies between 500 and 2500, the secondary load resistor must be individually matched to the individual AD606. However, that is certainly reasonable for a dynamic range amplified to 90dB.

5.
Literature

[1] D. Burchard (1993): Logarithmic converters and measurement of their characteristics; VHF Reports, issue 3/199 3, Pp. 140-154; Verlag UKW-Berichte, Baiersdorf and VHF Communications 3/1994 Pp 174 - 190

[2] J. Jirmann (1995): A high-precision logarithmic intermediate-frequency amplifier. VHF Reports, issue 2/1995 Pp. 97-101; Verlag UKW-Berichte, Baiersdorf and VHF Communications 3/1996 Pp 185 - 188

Alexander Meier, DG6RBP

Pre Divider (:10) up to 5GHz

A frequency counter is part of the standard equipment in almost any radio-frequency laboratory, but the frequency range usually goes up to no further than 1.3GHz. Although almost all measurements are carried out within this range, we nevertheless often wish we could measure higher frequencies as well. As an alternative to purchasing an expensive microwave counter, there is the option of expanding the range of an existing piece of apparatus with an external pre-divider.

1.
Circuit description

There are only a few components in the circuit of the pre-divider for frequency counters, Fig. 1 shows the wiring diagram. A similar pre-divider was presented a few years ago in [1]. Since the Plessey SP 8910 divider IC used has not been obtainable for some time, the project has tended to be forgotten. In the meantime, this IC has been brought back, and is currently available (once again) in a modern SMD housing from Zarlink [2]. So what could be more obvious than to develop a new frequency divider using this IC?

At the pre-selector input of the circuit there is a very broad band ERA-1 amplifier from Mini Circuits [3]. It amplifies the input signal up to 5GHz with approximately 11 to 12dB, before the signal is fed to the actual divider, U2, at PIN 2.

Like other dividers, this one also oscillates without an input signal, in which case approximately 550MHz can be measured at the output. On some dividers, this oscillation can be suppressed by means of a resistance between the input pin and earth, but this was not successful here.

The resistor R3 provides for an output impedance of approximately 50 Ohms, and the output level is approximately -10dBm. But the required input level represents a greater problem. The curve in Fig. 2 shows how high the minimum level must be for the divider to function satisfactorily. We can also see that the divider can still be used over 5GHz. Be careful the input levels are not too low! Fig. 3 shows what happens at the output, with an input frequency of 1GHz, if the input level (27dBm) is too low, an input frequency of 2GHz is faked! Figs. 4 and 5 in contrast show the spectrum at the output with the correct input levels. Input levels that are too high should likewise be avoided.

The supply voltage for the divider is stabilised with a fixed voltage regulator (U3). The one selected here was in a TO-220 housing.
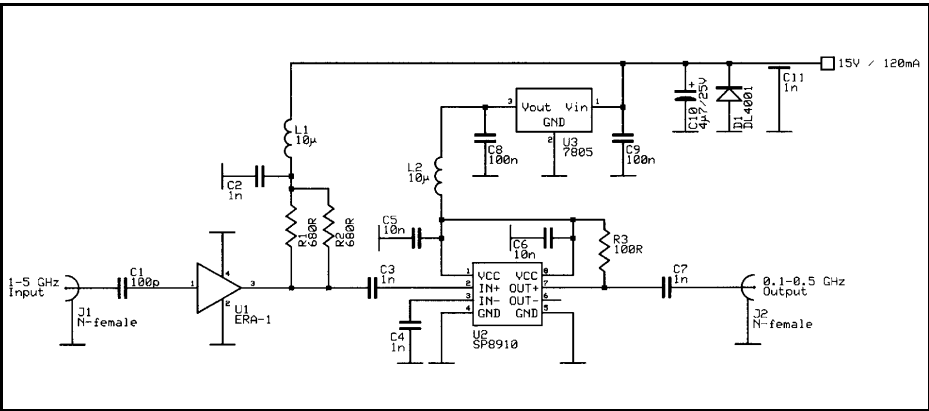
Fig 1: Circuit diagram of 5GHz pre divider.

2.
Circuit assembly

The circuit is assembled on a 45mm x 30mm Teflon printed circuit board (Fig. 6). This has a gap at one corner for the voltage regulator, U3.

When building the printed circuit board in accordance with the component layout (Fig. 7), pay particular attention to mounting the amplifier U1. Its earth connections must be connected to the earth side of the printed circuit board by the shortest path, or it will have a tendency to oscillate due to the feed inductances arising.

Earth connections using through plating are not successful unless there is sufficient through connection. So another method is used here, which is always successful. The MMIC is sunk into a hole ($\varnothing$ 2.3mm) in the board itself. The earth connections are bent downwards and soldered flush with the earth surface.
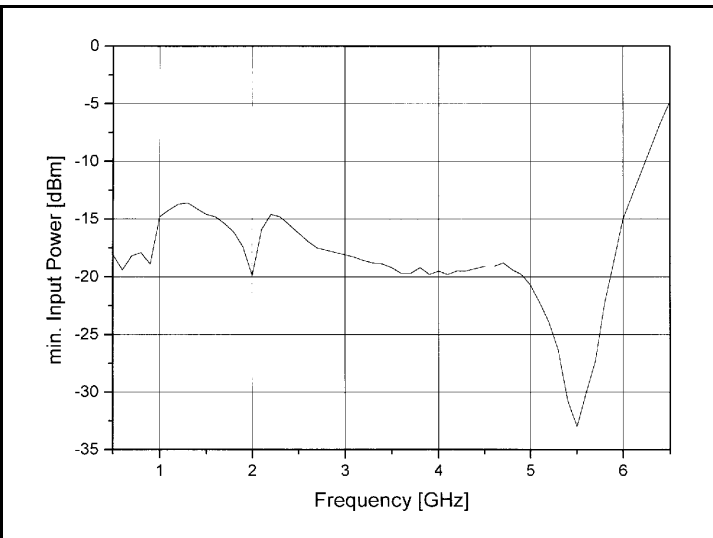

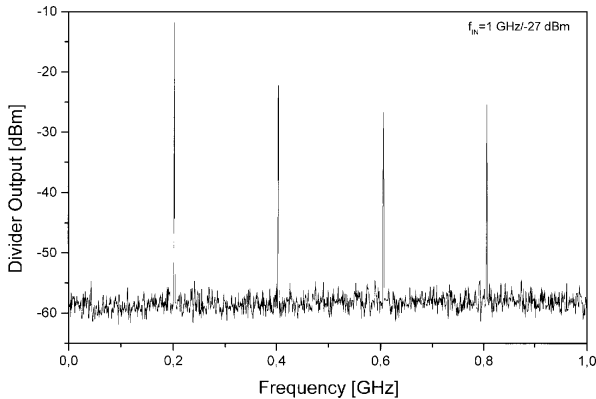**Fig 2: Minimum input level for divider to function correctly.**

Fig 3: Output of divider when input level is too low (1GHz, -27dBm).

In the same way, the inputs and outputs are bent upwards and soldered to the tracks.

When the board has been fully populated (except for the voltage regulator), the flux residues are cleaned off the earth side and it is screwed into the milled aluminium housing. Then the voltage regulator and the connectors can also be fixed and soldered on.

Before the housing is screwed down, the top of the board should be cleaned again and the circuit should be tested. You should also make sure that U1 is not oscillating!

2.1. Parts list

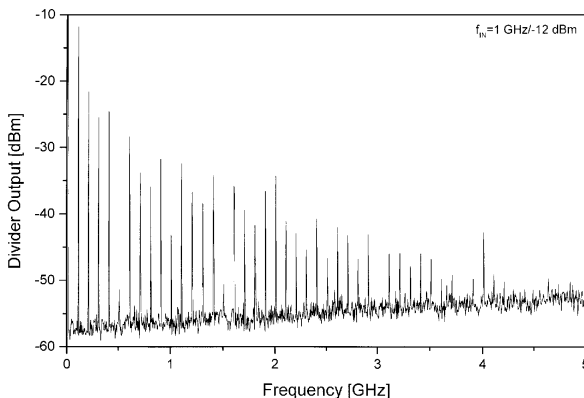|	
U1	ERA-1 (Mini-Circuits)
U2	SP8910 (SO8, Zarlink)
U3	7805
C1	100pF, 0805
C2,C3,C4,C7	1nF, 0805
C5,C6	10nF, 0805
C8,C9	100nF, 0805
C10	4.7µF/25 V, SMD
C11	1nF DF capacitor, M3 Actipass AFC-102P-10A
L1,L2	10µH, SMD

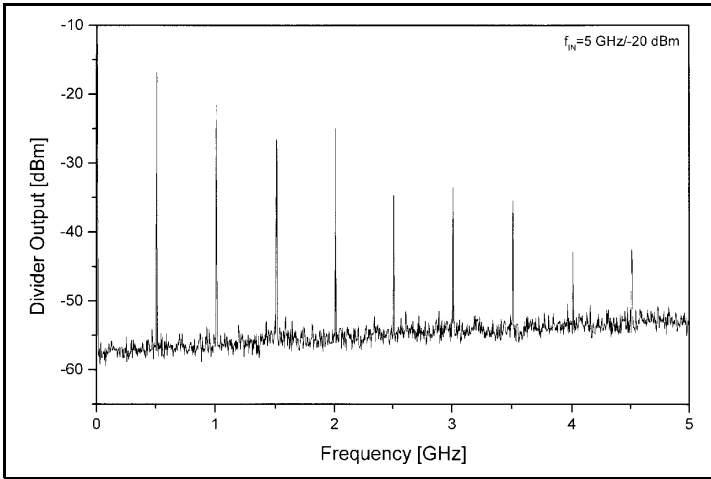**Fig 4: Output of divider when input level is correct (1GHz, -12dBm).**

Fig 5: Output of divider when input level is correct (5GHz, -20dBm).

R1,R2	680R, 1206
R3	100R, 1206
2 x	N-flanged sockets, small 4-hole flange
1 x	Teflon PCB, DG6RBP 002, through plated
1 x	Aluminium housing
1 x	Soldering lug 3.2 mm for C11 (earth connection)

3.
Technical data

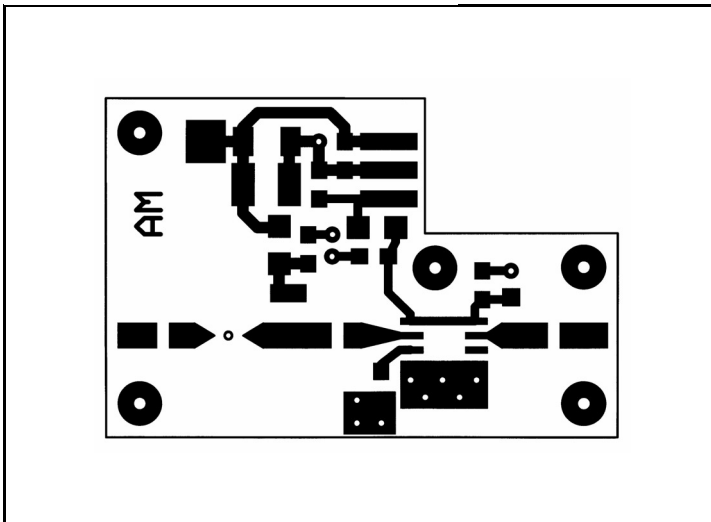| Input freq. range | 1 to 5 GHz |
Output range	100 to 500 MHz
Divider factor	10
Input level approx.	13 to 7 dBm
Output level approx.	10 dBm
Input/output	N socket
Supply	+15 V, 120 mA

Fig 6: Printed circuit board layout for 5GHz pre divider.
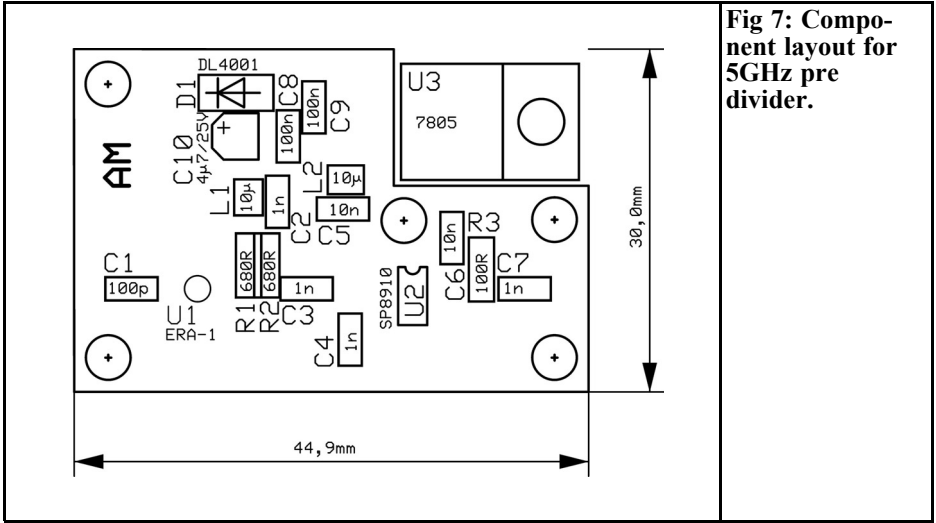
Fig 7: Component layout for 5GHz pre divider.

4.
Literature

[1] Dr.-Ing. J. Jirmann und Michael Kuhne: Measurement aids for the UHF amateur, VHF Reports 1/93, Verlag UKW-Berichte, Baiersdorf and VHF Communications 4/1993 Pp 207 - 213

[2] Data sheet SP8910, Zarlink, www.zarlink.com

[3] Data sheet ERA-1, Mini-Circuits, ww.mini-circuits.com

Fig 8: Photograph of completed 5GHz pre divider.

The UK Six Metre Group

www.uksmg.org

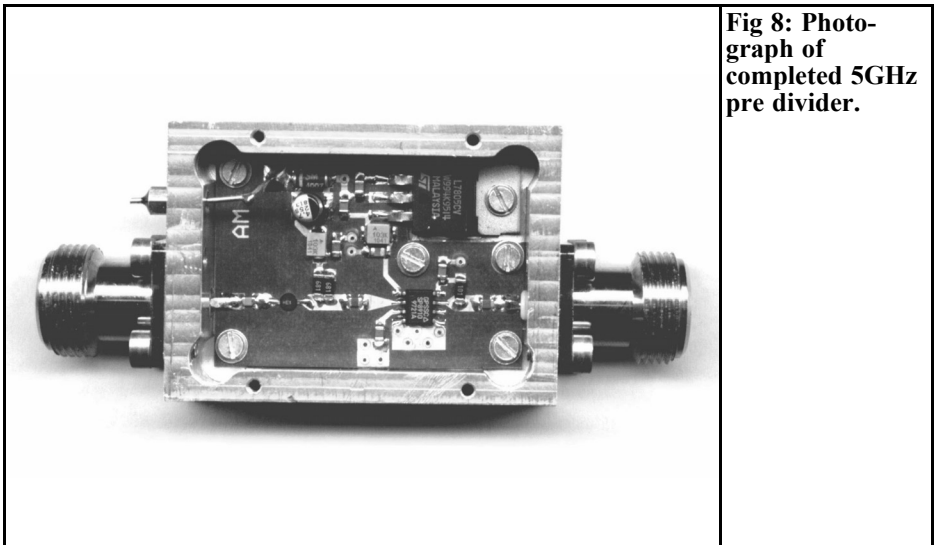**With over 1000 members world-wide, the UK Six Metre Group is the world's largest organisation devoted to 50MHz. The ambition of the group, through the medium of its 60-page quarterly newsletter 'Six News' and through it's web site www.uksmg.org, is to provide the best information available on all aspects 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.

Gunthard Kraus, DG8GB

Internet Treasure Trove

Stanford Microdevices / Sirenza Microdevices

Here you have to make a quick re-adjustment, the well known firm of Stanford has suddenly changed its name. Nevertheless, its activities remain unchanged. They still make new low noise amplifiers, mixers, gain blocks, etc for the frequency range between 0 and 10GHz.

And there are naturally lots of application notes and other interesting data to download.

Address:

http://www.stanfordmicro.com

Flex-PDE

The Internet has some new EM simulation software, known as Flex-PDE for 2D and 3D analyses in the microwave range. This is especially suitable for the extremely high frequency range between 10 and 100GHz. Those interested can download the free test version.

The program is so universal that it can also be used to investigate many other phenomena, e.g. currents in fluids, heat conduction and distribution, chemical processes, etc. The homepage also lists some other interesting documents. The textbooks, in particular, are first class, and worth the price (e.g. $12 for Fields of Physics).

Address:

http://www.pdesolutions.com

Linmic

Again, something similar, but centring on what is (according to the publicity) a combination of EM simulators and layout orientated CAD software packages which is unique in the world. Here too there is a demo version for testing.

Address:

http://www.linmic.com

Metelics

Here you can, find out all about microwave diodes, i.e. Schottky diodes, PIN diodes, tunnel diodes, varactor diodes, etc. Naturally, there are also comprehensive catalogues, data sheets and application notes.

Address:

http://www.metelics.com

Marki Microwave

Never heard of them? Well, anyone whos looking for fast doublers, mixers, multipliers or converters at reasonable prices for the frequency range between 0 and 40GHz in SMD format should take a glance at this page.

Address:

http://www.MarkiMicrowave.com

Rf Nitro

This site has nothing to do with artificial fertilisers or explosives. It deals with the cutting edge of MMIC development in GaAs or GaN technology. A wide range of products for the microwave range and some excellent application notes practically compel you to visit this site.

Address:

http://www.rfnitro.com

National Instruments

A company manufacturing and marketing measurement technology hardware and software on such a large scale (think of LabView, for example) is naturally a real Treasure Trove for those with relevant interests. Here we find not only data sheets and test software CDs but also separate document packages for users and developers. Also on offer are tutorials on various subjects and product groups.

You could spend hours finding more and more items of interest. The documentation on FFT, in particular, is outstanding!

Address:

http://www.ni.com

Mini Circuits

Well known manufacturer of active radio frequency components including, among others, mixers and MMICs such as, for example, the Era range.

Many comprehensive data sheets.

Address:

http://www.Mini-Circuits.com

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Gunthard Kraus, DG8GB

Modern Design of Stripline Low Pass Filters

The subject of low pass filters is covered through a comparison of designs, with the assistance of contemporary development tools based on PC programs. The special features of Micro Stripline filters are also referred to.

1. Introduction

The current situation and recent progress in relation to the design of stripline band pass filters have already been discussed here The present article is devoted to low pass filters. In the first section, the work is done using the tried and tested microwave CAD program PUFF, while in the second section the design work is repeated using the most recent student version of ANSOFT Serenade. The results can thus be compared with one another and the progress made in microwave CADs and in circuit simulation can be evaluated Finally, this subject is rounded off with measurements on the assembled specimen circuits.

2. Filter design using PUFF

2.1. Design principles for stripline filters

The filter type chosen is the Chebyschev low pass (for more detailed information on the selection of filters of a suitable type and grade, see the article referred to in [1]).

The following values are given:

Filter grade n	**Number of poles = number of components = 5**
π format	**(Shut capacitors with series inductors)**
Impedance Z	**50Ω**
Maximum ripple of S21 in transmission range	**0.1dB**
Cut off frequency	**1,700 MHz**

The components required are calculated in accordance with these parameters, using the tried and tested filter program "fds.zip" (Fig. 1). We therefore need two capacitors, each with 2.147pF, a capacitor with 3.67pF and two coils, each with 6.42nH.

Now we use the well known method for implementation (Fig. 2):

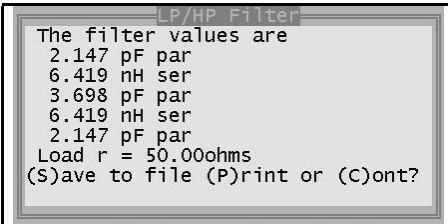The capacitors are created using short but

Fig 1: Filter design using fds.zip for cut off frequency of 1700MHz.

Fig 2: Schematic of low low pass filter.

very wide stripline sections with an impedance of less than 20Ω. This ensures that the inductive fraction of the line plays no great role. Very thin, short lines (with Z = at least 100Ω) are used to produce the inductances, provided the capacitive fractions are still sufficiently low.

For successful design, stick to the following rules, which originate from practical experience in filter construction:

The electrical lengths of the line sections should lie between 10 and 30 degrees when the cut off frequency of the filter is used as the design frequency of the line.

For the "capacitors", the impedance level of the line sections used for a 50Ω system should not exceed 20Ω. For the "coils", select lines with at least Z = 100Ω.

For the "wide capacitor lines", the width to length ratio should not exceed 8.

For the "thin inductance lines", stick as close as possible to a minimum conductor width of approximately 0.2 mm as the lower limit. This results in a high inductance and thus short line lengths. Unfortunately such extremely thin tracks are notable for their high losses; the lines therefore have to be made slightly wider, which makes it easier to manufacture the printed circuit board.

But we must be clear about one thing: Replacing genuine capacitors and coils with line sections does not immediately lead to a perfect solution, since it is well known that any line consists of inductive and capacitive parts. In addition, trans-
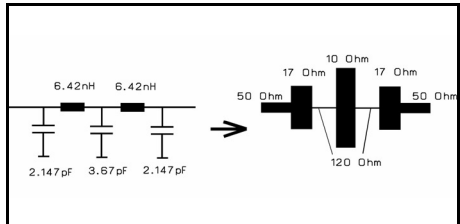
formed lines also have resistances, the source resistance and the load resistance. Consequently discrepancies are possible on the first simulation run, thus optimisation is indispensable.

2.2. Determination of line data

Use a text editor to open the SETUP file for PUFF, in order to enter and then store the design frequency, fd, together with the data for the board material used (Rogers R04003, with a thickness of 32 MIL). They are shown in bold in the following file extract:

\b{oard} .puf file for PUFF, version 2.1d}

d   0   {display: 0 VGA or PUFF chooses, 1 EGA}

o   1   {artwork output format: 0 dot-matrix, 1 Laser Jet, 2 HPGL file}

t   0   {type: 0 for microstrip, 1 for stripline, 2 for Manhattan}

zd 50.000 Ohms {normalizing impedance. 0<zd}

fd 1.7 GHz {design frequency. 0<fd}

er 3.380 {dielectric constant. er>0}

h   0.813 mm {dielectric thickness. h>0}

s   100.000 mm {circuit-board side length. s>0}

c   100.000 mm {connector separation. c>=0}

r   0.010 mm {circuit resolution, r>0, use Um for micrometers}

a   0.000 mm {artwork width correction.}

mt 35 Um {metal thickness, use Um for micrometers.}

sr 5.000 Um {metal surface roughness, use Um for micrometers.}

lt 1.0E-0003 {dielectric loss tangent.}

cd 5.8E+0007 {conductivity of metal in mhos/meter.}

p   2.000 {photographic reduction ratio. p<=203.2mm/s}
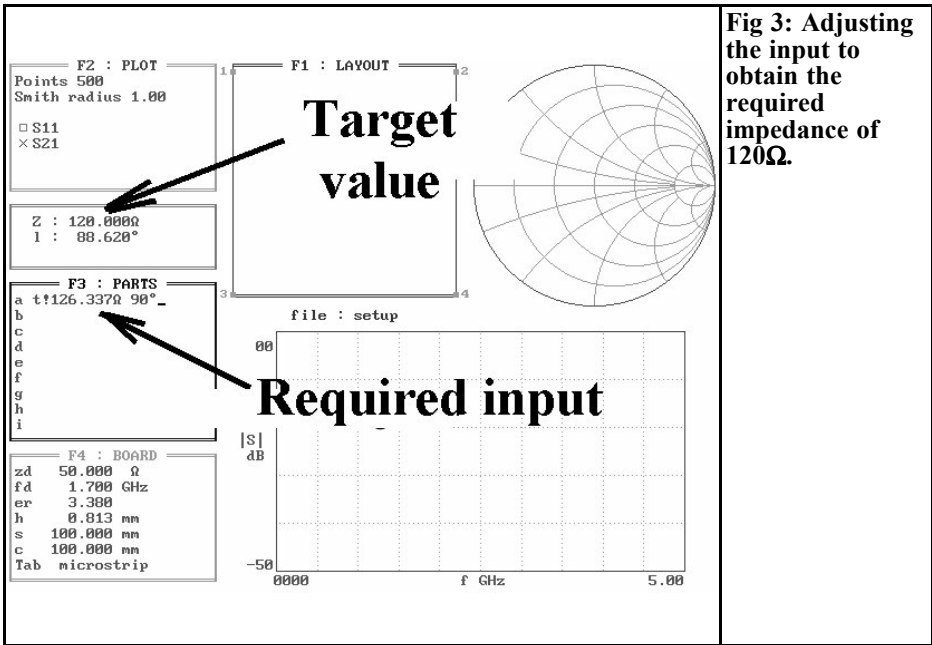
m 0.600 {mitering fraction. 0<=m<1}

Fig 3: Adjusting the input to obtain the required impedance of 120Ω.

Next use a width of 0.25 mm for the "inductance line section", thus there is no great problems in producing the printed circuit board.

Then PUFF is started to determine the impedance levels for the conductor widths. This is done in the following manner. A real model transmission line is entered in field F3, with an impedance level of, e.g., 100Ω and an electrical length of 90 degrees. The "real modelling mode", as is well known, is invoked using the exclamation mark after the letter "t". If the equals sign is now pressed, then the actual impedance level and the actual electrical length appear in the dialogue field. The entry value is altered until an actual impedance level of 120Ω appears in the dialogue field. As Fig. 3 shows, it is necessary to increase the input value to 126.337Ω? to achieve this. If we now delete the exclamation mark and press the equals sign again, we see the mechanical data for this line. Fig. 4 shows that the width of 0.25 mm is well attained.

In order to add the two outer 2.147pF capacitors, we use lines with an impedance of Z = 17Ω, whilst for the central capacitor (3.67pF) a wider line with Z = 10Ω makes better sense. If the line is too long, this unfortunately makes the attenuation worse at very high frequencies, which also accounts for the recommendation for an electrical line length of 10 to 30 degrees. We use PUFF for these two lines as well, to determine the necessary conductor widths. We should also include the 50Ω feed lines to connect the filter structure with the connection sockets.

If we keep strictly to the methods referred to, we finally obtain the entry values in accordance with Fig. 5. If we finally remove the exclamation mark in each line and enter the equals sign, we can compile the first little table:

Line	50 Ω	10 Ω	17 Ω	120 Ω
Conductor width (mm)	1.84	14.62	7.94	0.25

We now continue using simple approximation formulae to determine the required line lengths. Here we start with the narrow lines serving as coils (w =
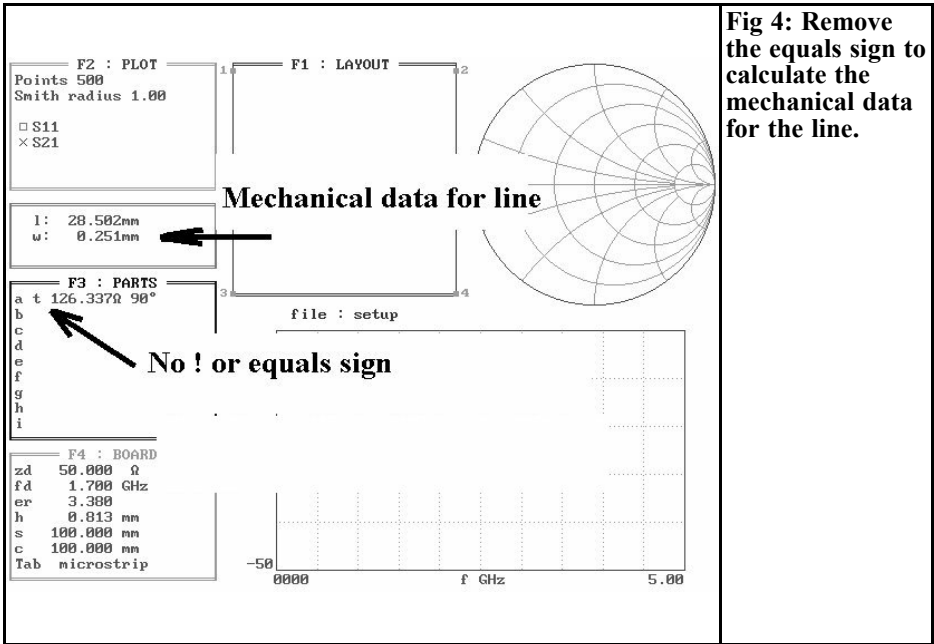
Fig 4: Remove the equals sign to calculate the mechanical data for the line.

0.25 mm), where:

$$\frac{I_{L-Line}}{\lambda} = \frac{f \cdot L}{Z_L}$$

ZL is the selected impedance for the "inductive line", f is the actual frequency and L is the required inductance.

For f = 1700MHz, L = 6.419nH and ZL= 120Ω we obtain the following standardised length:

$$\frac{L_{L-Line}}{\lambda} = \frac{1700 \cdot 6.41}{120} = 0.090935$$

Fig 5: Real inputs for 120Ω, 17Ω, 10Ω and 50Ω lines.

Since a complete wavelength corresponds to an angle of 360 degrees, this line section will have an electrical length of 0.090935 x 360 degrees = 32.74 degrees. We obtain the associated mechanical length through the well known procedure using PUFF:

Put an exclamation mark behind "t", enter the length, and continue to correct until this value is finally obtained when the equals sign is entered (Fig. 6). Now delete the exclamation mark and press the equals sign again to obtain the figures for the mechanical dimensions. This gives a physical length of 10.53 mm.

Summary of complete line data for inductances with a design frequency of 1,700 MHz:

Z	**120 Ω**
Electrical length	**32,74 Degree**
Mechanical length	**10.53 mm**
Conductor width	**0.25 mm**

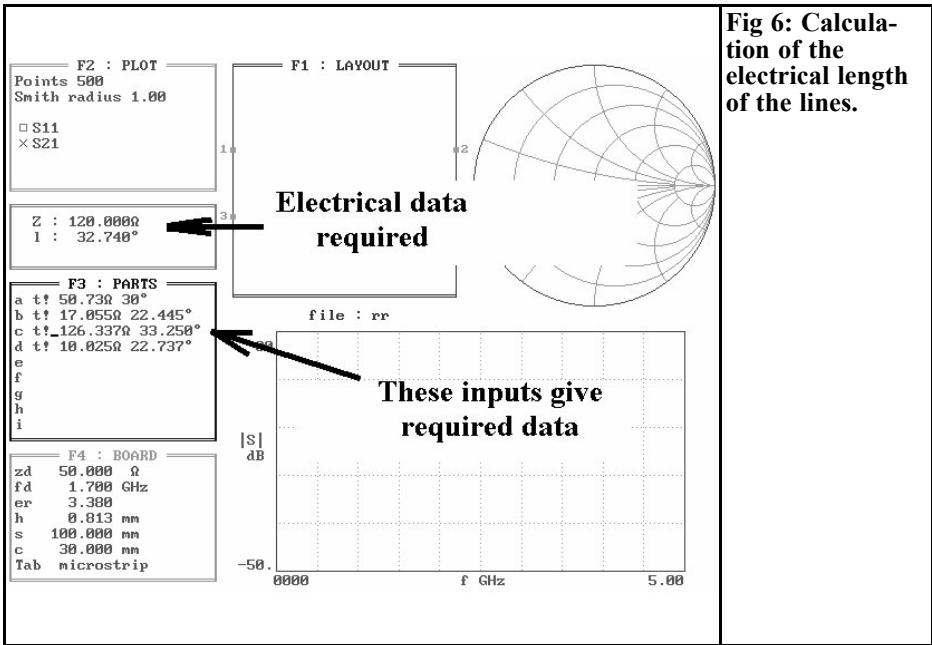The capacitor line calculation requires the following approximation formula:

Fig 6: Calculation of the electrical length of the lines.

$$\frac{I_{C-Line}}{\lambda} = f \cdot C \cdot Z_C$$

where Zc is the impedance of the capacitor line.

For the first and third capacitors with C = 2.147pF, at 1,700MHz we obtain:

$$\frac{I_{C-Line}}{\lambda} = 1700 \cdot 2.147 \cdot 17 = 0.062$$

That gives an electrical length for the line of 0.062 x 360 degrees = 22.32 degrees and in addition PUFF supplies a conductor length of 6.31 mm with a conductor width of 7.94 mm.

Summary of full line data for first and third capacities for a design frequency of 1,700MHz:

Z	17 Ω
Electrical length	22.32 degrees
Mechanical length	6.31 mm
Conductor width	7.94 mm

It is frequently necessary to repeat this run for the central capacitor of 3.698 pF:

$$\frac{I_{C-Line}}{\lambda} = 1700 \cdot 3.698 \cdot 10 = 0.082866$$

That gives an electrical length for the line of 0.062866 x 360 degrees = 22.63 degrees and a conductor length of 6.29 mm with a conductor width of 14.62 mm.

Summary of full line data for central capacity for a design frequency of 1,700MHz:

Z	10 Ω
Electrical length	22.63 degrees
Physical length	6.29 mm
Conductor width	14.62 mm

2.3. Determination of substitute data for stripline steps

In the existing circuit, a repeated change (Impedance Step) is carried out between wide and narrow stripline sections. But each individual step acts as an irregularity, the approximate effect of these can be determined by using a substitute circuit made from a series inductance and a case capacitance (like a simple low pass filter!).
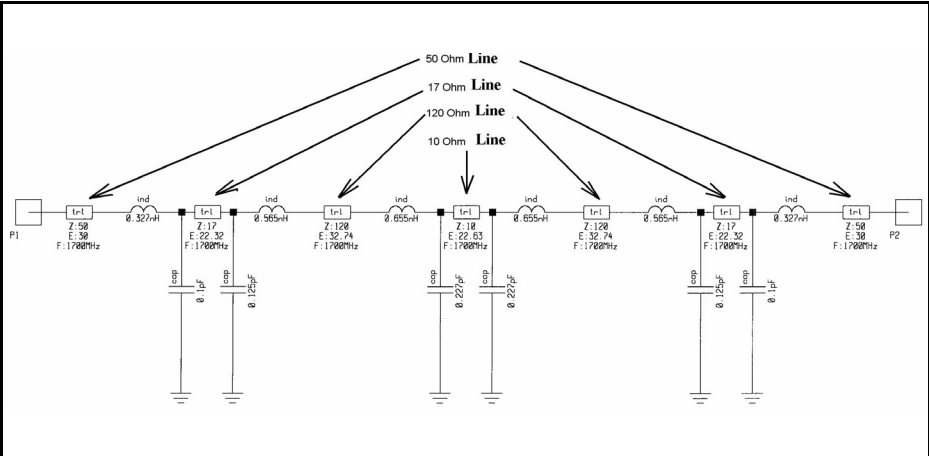
Fig 7: Circuit diagram of low pass filter produced using ANSOFT Seranade, showing impedance steps.

If the simulated circuit is to correspond precisely to the parameters (as far as possible) without a lot of reworking, and at the same time be capable of use, we can not avoid the need for these circuit extensions. But they are already in existence in a modern microwave CAD program as modules and the correction is thus relatively simple. PUFF is unfortunately not that well developed yet, so the calculation formulae from [4] are used.

In order to specify these substitute components, we require not only the impedance values and conductor widths of each line section, but also the associated effective dielectric constants. These constants are determined and used by PUFF during its calculations, but unfortunately they are not output. So here we have to outwit the program:

First step:

Use a pocket calculator to work out the free space wavelength for the frequency of 1,700MHz

$$\lambda = \frac{c}{f} = \frac{3.10^8}{1700} = 176.47mm$$

Second step:

Increase the inputs on all four lines in field F3, one after another, until the real lengths are precisely 360 degrees when the equals sign is pressed (i.e. one wavelength). Then delete the exclamation mark in each line and enter the equals sign instead. This will give you the corresponding physical lengths. They are:

Line	50 Ω	10 Ω	17 Ω	120 Ω
Conductor length	108.496	99.998	102.004	115.785

Third step:

Using the following simple rearranged short cut formula,

$$\varepsilon_{eff} = \left(\frac{\lambda_{air}}{\lambda_{Line}} \right)^2$$

All the effective dielectric constants can now be determined using the pocket calculator:

Line	50 Ω	10 Ω	17 Ω	120 Ω
Conductor width	1.84	14.62	7.94	0.25
ε_{eff}	2,646	3,114	2,993	2,323

All the step components can now be specified using this data. The series inductance at each step can be deter-
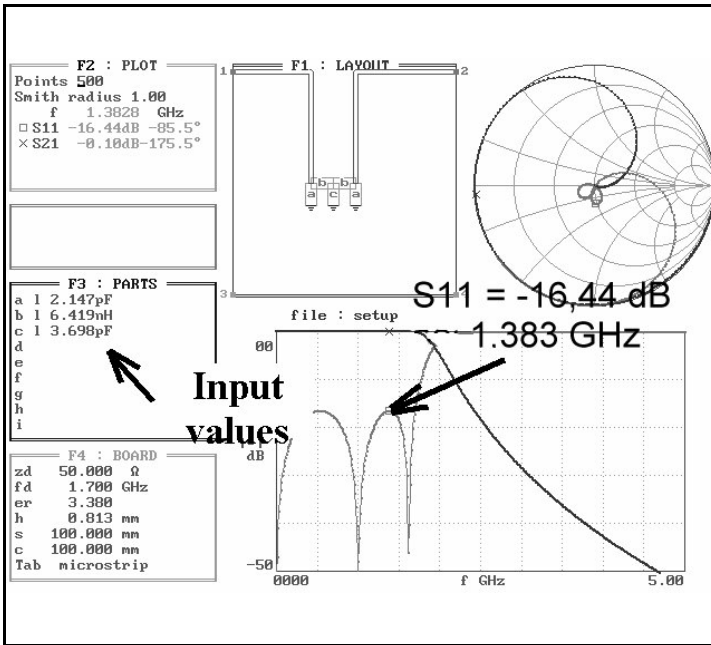
Fig 8: Results of simulating ideal filter.

mined using the following approximation formula:

$$L = 0.000987 \cdot h \cdot \left(1 - \frac{Z_1}{Z_2} \cdot \frac{\sqrt{\varepsilon_{1eff}}}{\sqrt{\varepsilon_{2eff}}}\right)$$

where: h is the board thickness in µm, Z1 the impedance level of the wide line, Z2 the impedance level of the narrow line, ε_{1eff} the effective dielectric constant of of the wide line on the board, ε_{2eff} the effective dielectric constant of the narrow line on the board. The inductance is in nH.

Thus we can once again compile a table:

Transition from	50 Ω to 17 Ω	17 Ω to 120 Ω	120 Ω to 10 Ω
Inductance	1.84 nH	14.62 nH	7.94 nH

It is somewhat more laborious to calculate the case capacitance arising:

$$C = 0.00137 \cdot \frac{\sqrt{\varepsilon_{1eff}}}{Z_1} \cdot \left(1 - \frac{W_2}{W_1}\right) \cdot h \cdot \left(\frac{\varepsilon_{1eff} + 0.3}{\varepsilon_{1eff} - 0.258}\right) \cdot \left(\frac{\frac{W_1}{h} + 0.264}{\frac{W_1}{h} + 0.8}\right)$$

where: Z1 = Impedance of the wide line, ε_{1eff} = effective dielectric constant of wide line, W1 = conductor width of wide line, W2 = conductor width of narrow line, h = board thickness in µm (here: 0.813 mm = 813 µm). The capacity is in pF.

With the above data, we obtain a second table using the pocket calculator:

Transition from	50 Ω to 17 Ω	17 Ω to 120 Ω	120 Ω to 10 Ω
Capacity	0,1 pF	0,125 pF	0,227 pF

When drawing the simulation wiring diagram in the next chapter, please bear in mind that:

series inductances are always connected directly to the narrower line section, whilst the shunt capacitances are always mounted parallel to the wider line section!

The circuit diagram that applies to the simulation, with all the irregularities is shown in Fig. 7. It was drawn using the circuit editor of ANSOFT Serenade and
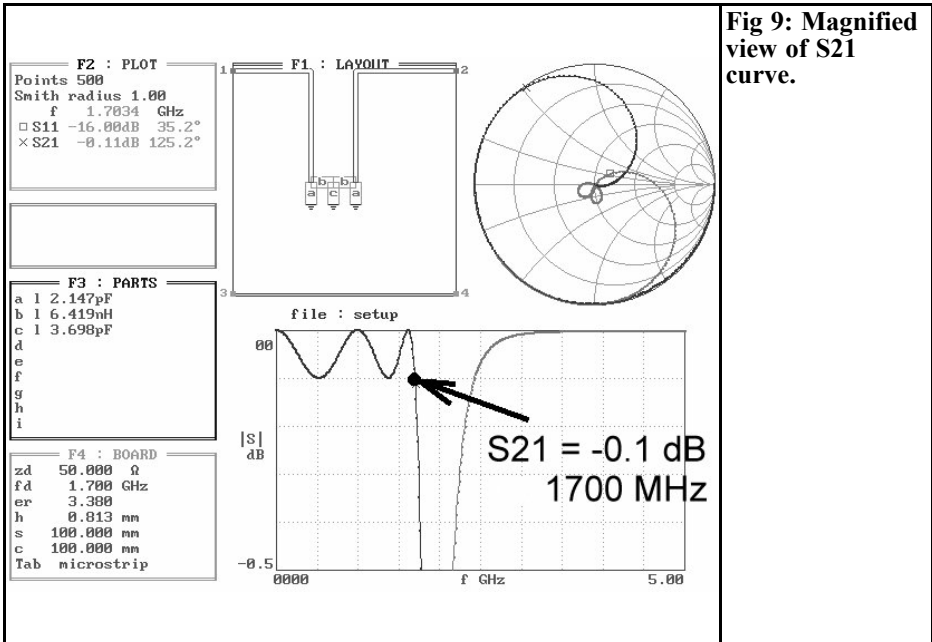
```
F2 : PLOT
Points 500
Smith radius 1.00
    f   1.7034  GHz
□ S11 -16.00dB  35.2°
× S21  -0.11dB 125.2°

F1 : LAYOUT

file : setup

F3 : PARTS
a l 2.147pF
b l 6.419nH
c l 3.698pF
d
e
f
g
h
i

|S|
dB

S21 = -0.1 dB
1700 MHz

F4 : BOARD
zd   50.000  Ω
fd    1.700  GHz
er    3.380
h     0.813  mm
s   100.000  mm
c   100.000  mm
Tab  microstrip
```

immediately gives an impression of the layout of this modern program. But the number of components required is not very far off the maximum of 25 components that can be used with the free student version of ANSOFT Serenade.

2.4. Simulation and optimisation of circuit

First we have to simulate the ideal low pass filter made from coils and capacitors, in order to have the optimisation target visible. The result can be seen in Fig. 8, where special attention should be paid to the precise heights of the two "camels humps" of S11. S11 rises right up to -16.44 dB since it is associated with the S21 ripple of -0.1 dB). If this value is obtained again following the optimisation of the stripline circuit, the ripple is also tuned, being approximately -0.1 dB in the transmission range. It is increased merely by the attenuation, that rises with the frequency, due to the printed circuit board losses and the skin effect, etc.

The ideal curve for S21 is shown in Fig. 9 on a greatly expanded scale. From now

on, these two sheets should be kept next to the PC for the work that follows.

We now enter all the components required for the circuit in field F3. Make sure you dont forget the exclamation mark for all line sections. See Fig. 6 for the precise data for the lines. Once everything has been entered, the circuit is correctly assembled in F1, then switched to F2 and simulated. The simulation result from S11 and S21 in the frequency range from 0 to 5GHz and the amplitude range from 0 to 1dB for monitoring the ripple is shown in Fig. 10. In Fig. 11 the amplitude range between 0 and -50dB has been selected in order to check the two S11 humps. It can be seen that the impedance leaps have the most important effect, and also that the cut off frequency has fallen by almost 400MHz.

But dont worry: optimisation is not especially difficult, for you merely have to balance out the influence of the additional inductances:

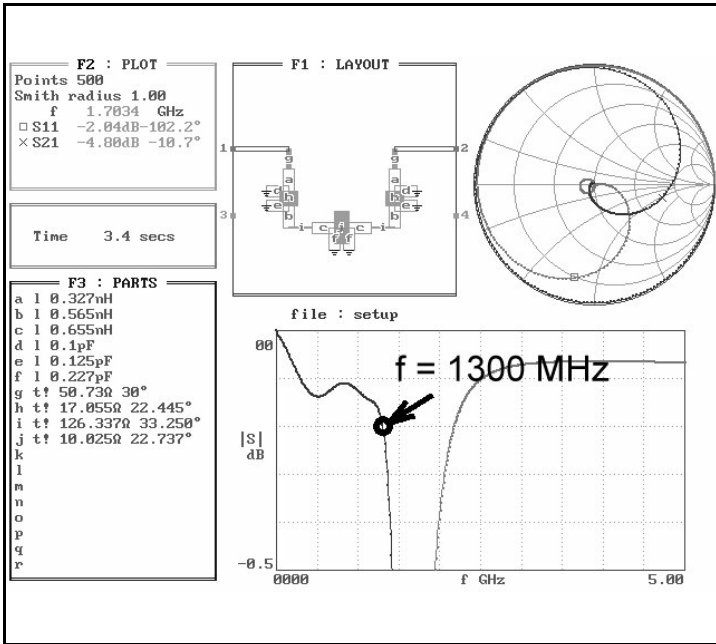- The camel hump at the bottom right hand corner can be lifted by means

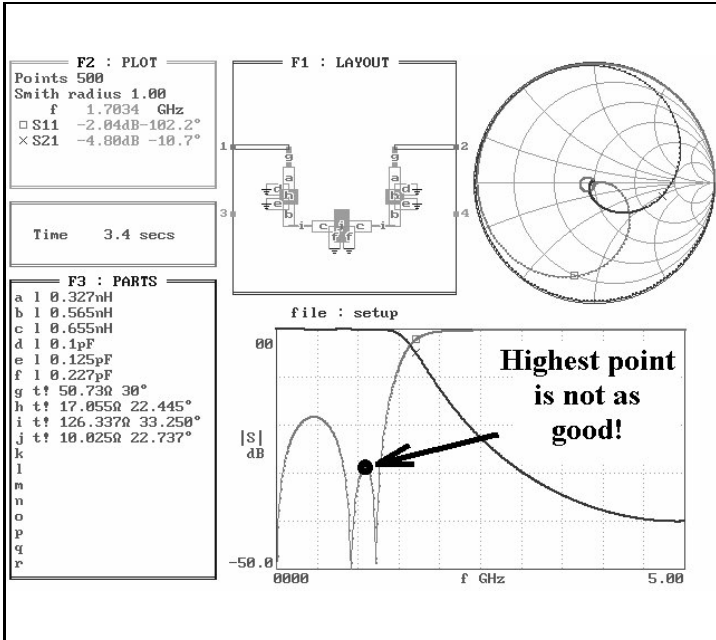**Fig 10:**
Simulation results from 0 to 5GHz.

Fig 11:
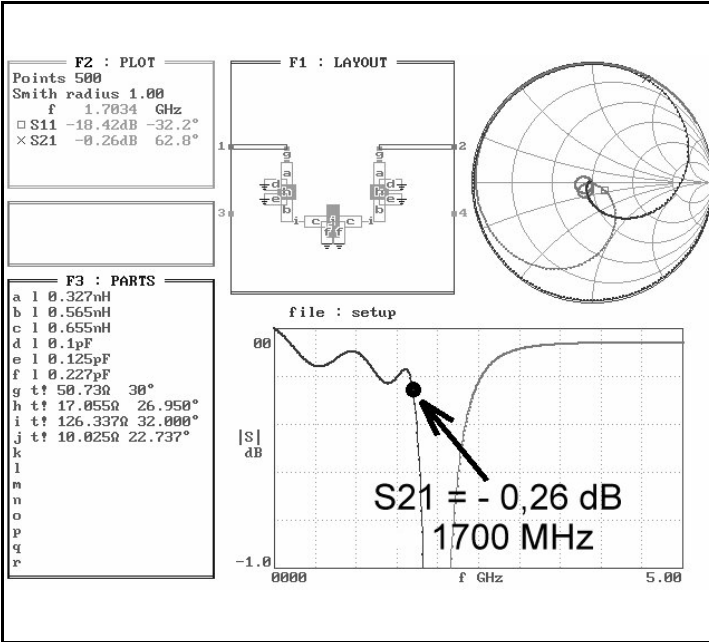**Expanded view of 0 to 5GHz results.**

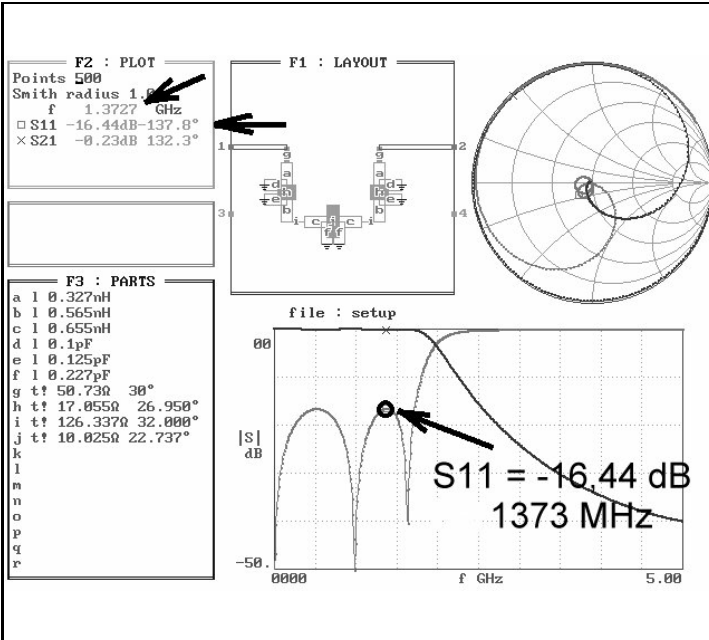**Fig 12: Results showing S21 curve after optimisation.**

Fig 13: Results showing S11 curve after optimisation.

of an extension of the two external 17Ω line sections.

- Then, if necessary, the coils (the two thin 120Ω line sections) are slightly shortened, and in this way the two humps become the same height (target = -16.44dB).

Once this has been done, the display range for S21 and S11 is restricted to 0 to 1dB and simulated again. The pre set ripple value of 0.1dB can then be very easily recognised on the fundamental attenuation, which rises as the frequency increases. Yet the cut off frequency of the filter is still much too low. We therefore simply increase the design frequency in field F4 until, at 1,700MHz, we can observe the sharp transition from the transmission range into the filter attenuation band. This is at fd = 2,100MHz in this case. Once again, we convert this to the display range between 0 and 50dB using a linear plot diagram and the precise height of the two humps is checked with the help of the cursor. Any small discrepancies present are eliminated in accordance with methods a) and b). The results can be seen in the two diagrams, Figs. 12 and 13.

Finally, another table can be compiled, in which the definitive mechanical and electrical data for the individual line sections are entered after optimisation. These are not needed for the draft layout, but they can be used as a way in to monitor this PUFF design in the second part, using ANSOFT Serenade.

Make sure that:

- The design frequency has been

increased to 2,100MHz and that the matter of the open end extension (the increase in the line length caused by projecting field lines) has also been dealt with automatically through the insertion of the impedance steps.

- The pre set mechanical lengths and widths for the line sections may therefore be transferred directly into the printed circuit board layout without any further correction! See Table 1 below.

2.5. PUFF versus HARMONICA

Using PUFF we can switch, by means of a simple exclamation mark, from ideal modelling (entering impedance level and electrical length at design frequency) to real modelling (with physical length and width, together with all the line losses).

ANSOFT unfortunately separates these two options completely and even uses different graphic symbols in the two cases. So separate wiring diagrams and projects must also be drawn up for it!

Here we can go straight to the simulation of physical dimensions in the first run. The original PUFF simulation is investigated (housing cover not taken into account). In the second run a space of 13mm between the board and the cover is used which simulates typical apparatus and its influence determined.

2.5.1 Simulation check of PUFF design with SERENADE

Serenade is started up, a new project is set up and then the circuit is drawn. Assistance should be obtained from Fig.

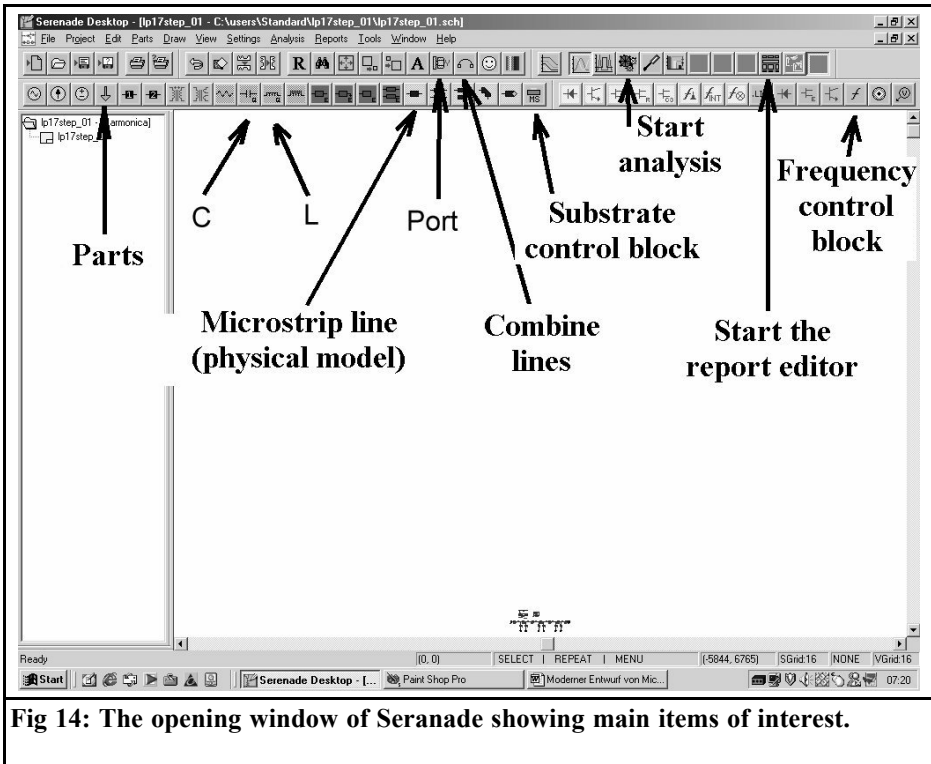| Table 1: Pre set mechanical dimensions.				
	50Ω line	10Ω line	17Ω line	120Ω line
Width	1.84mm	14.62mm	7.94mm	0.25mm
Length		5.09mm	6.15mm	8.20mm
Electrical length in deg. at 2100MHz		22.63	26.8	31.51

Fig 14: The opening window of Seranade showing main items of interest.

14 here. It shows the buttons behind which the various components and elements are hidden. The correct entries in the substrate screen can be seen in Fig. 15 (board and housing data), where the unit mm should not be forgotten, for all dimensions. Fig. 16 shows the circuit ready for simulation, including the frequency module for the sweep from 0 to 5GHz in 5MHz steps.

As soon as all this has been done, the analysis button (button with the little gear on top) can be pressed. At first nothing can be seen, until the turquoise / grey Report Editor button is pressed, this gives the option of outputting the results in diagram form. S11 and S21 are shown in dB as set up by the method shown in Fig. 17. Fig. 18 shows, on a greatly enlarged scale, the result of the simulation. Both the shape of the S21 curve and the ripple cut off frequency entered (with attention being paid to the fundamental

filter attenuation!) tally completely with Fig. 12, the simulation using PUFF. This can also be seen in Fig. 19, here the results are represented without magnification and thus the shape of the S11 curve can be checked. A comparison with Fig. 13 gives a big boost to our trust in PUFF.

2.5.2. Simulation using housing data

To do this we first call upon the Stripline Calculator TRL85 and calculate the corrected physical line data for an interval of 13 mm between board and cover. The first step in doing this is naturally to prepare the line and board targets required:

First use the line data for the simulation using PUFF at a design frequency of 2,100MHz (see Table 1) and then the technical data for the printed circuit board. The interval referred to between
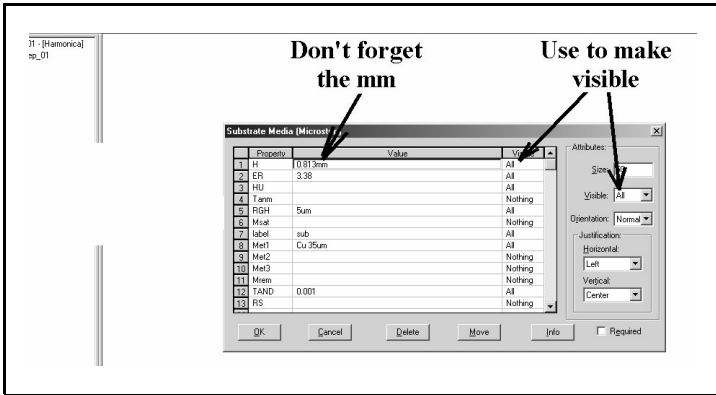
Fig 15: Entry of substrate data.

the cover and the board is added here as well (= marked in bold):

Material:	Rogers R04003
Board thickness:	H = 32 MIL = 0.813 mm
Dielectric constant:	er = 3.38
Loss factor:	TAND = 0.001
Copper coating:	TH = 35 μm
Surface roughness:	RGH = 5 μm
Distance between	
cover and board:	**HU = 13 mm**

The determining the new values for the first line section (Z = 17Ω, electrical length = 26.8 degrees at 2,100MHz) and, once again, the steps required to get there are shown in Fig. 20. If we repeat that for the other lines, then we finally obtain the

results table (Table 2).

Anyone who compares the two tables (with and without housing) is immediately struck by the fact that including the cover distance referred to affects only the wide line sections, their conductor width is reduced. This is easy to understand, since some additional electrical field lines are pulled up by these large plates. This increases the capacity and must be corrected, whereas for the short "fips", showing the widths of 120Ω and 0.25 mm, scarcely anything is noticeable.

Now the previous HARMONICA project is re started and every physical line value in it is checked and / or corrected in
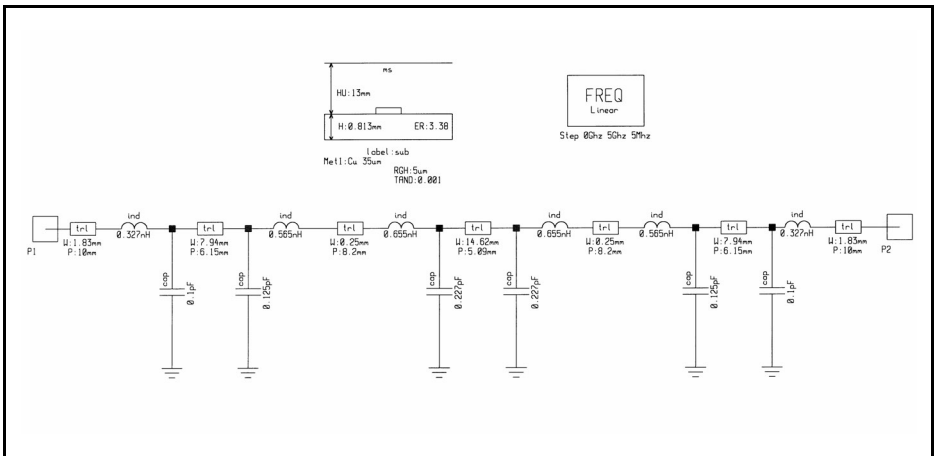
Fig 16: Circuit diagram ready for simulation.
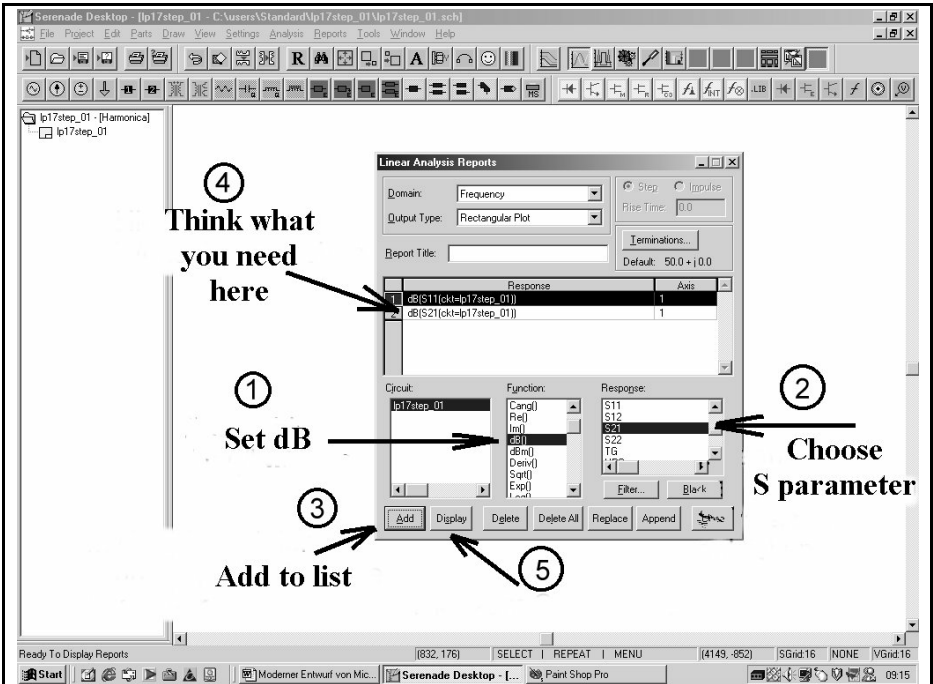
Fig 17: Setting up reports editor.

accordance with the table.

The new entry "HU = 13 mm" for the substrate control block is very important here, and should not be forgotten!

If we now start the simulation and extract and magnify S21 (Fig. 21), all comment becomes superfluous: see Fig. 12, see Fig. 18.

One further tip: in diagrammatic representations, the right hand mouse button conceals not only the zoom functions, but also the options for the insertion of fixed

or changing data markers, together with the output of the associated curve value at the selected frequency in small additional windows. You should try out these options yourself!

Table 2:				
	50Ω line	10Ω line	17Ω line	120Ω line
Electrical length in deg. at 2100Mhz		22.63	26.8	31.51
Width	1.83mm	14.53mm	7.90mm	0.25mm
Length		5.07mm	6.13mm	8.20mm
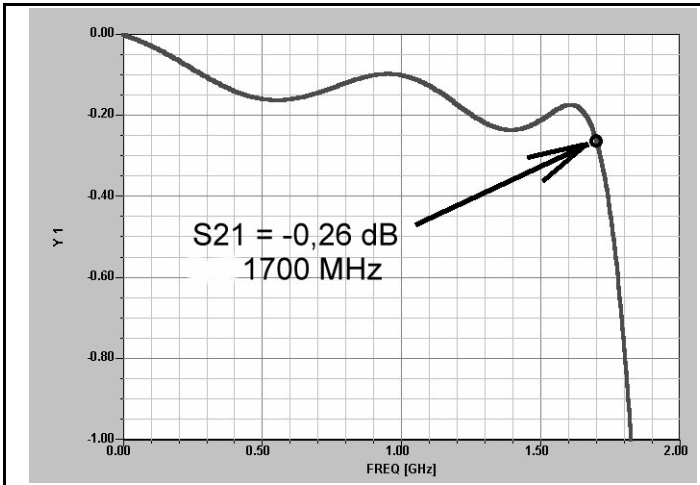
Fig 18: Results of simulation showing S21 curve.

3.
Repetition of design using ANSOFT Serenade

3.1. Assembly of filter circuit

For this, go right back to the beginning and begin as if you knew absolutely nothing about the preceding design using PUFF. Just avoid duplicating work by taking all the data already determined which are valid for both designs from Sections 2.1 to 2.3.

First step:

Use a filter program to determine the coils and capacitors required for the corresponding LC low pass. Fig. 1 gives the following values for this: 2.147pF, 6.419nH, 3.698pF, 6.419nH, 2.147pF

Second step:

Use the approximation formulae provided to determine the data for the line sections for a design frequency of 1,700MHz. (Do you remember ? We didnt get involved with optimisation with PUFF until we reached a frequency of 2,100MHz, so we ignore that here too). In accordance with the selected individual impedance level, the following values were obtained from Table 3.

Third step:

Use the TRL85 calculator to determine the physical dimensions of the individual line sections, and also include the cover distance of 13 mm in the list of given printed circuit board data:

Material:	**Rogers R04003**
Board thickness:	**H = 32 MIL = 0.813 mm**
Dielectric constant:	**$\varepsilon r = 3.38$**
Loss factorr:	**TAND = 0.001**
Copper coating:	**TH = 35 μm**
Surface roughness:	**RGH = 5 μm**
Distance between cover and PCB:	**HU = 13 mm**

The associated simulation (Synthesis) for the first line section, with 17Ω at the

Table 3:				
	50Ω line	10Ω line	17Ω line	120Ω line
Electrical length at 1700MHz		22.63	22.32	32.74
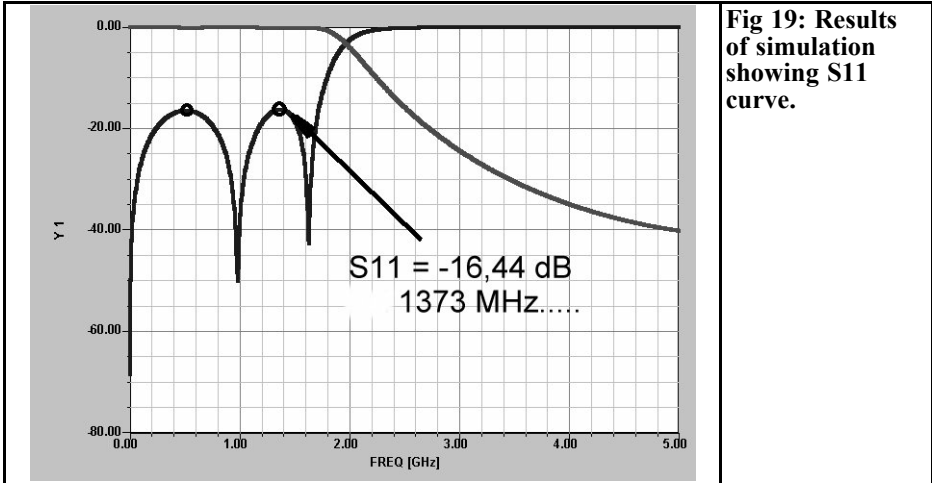
Fig 19: Results of simulation showing S11 curve.

Fig 20: Setting up data for simulation taking distance to cover into account.
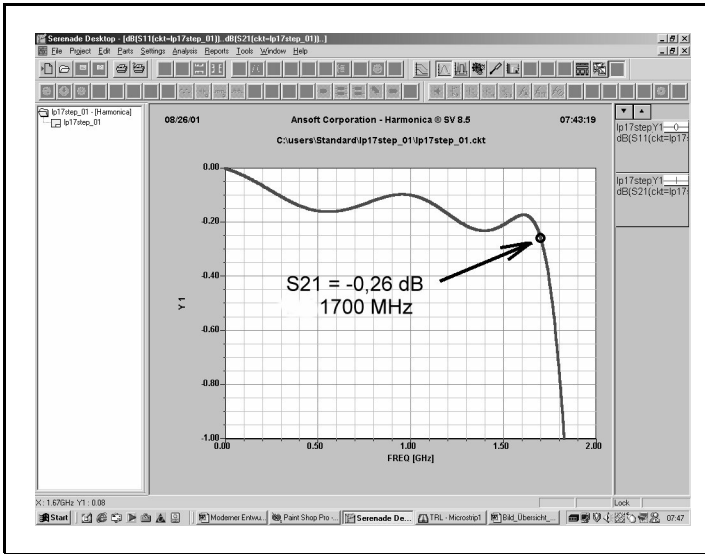
Fig 21: Results of simulation taking into account distance to cover, showing S21 curve.

frequency of 1,700MHz, shows Fig. 22. The remaining results have been entered directly into Table 4.

Fourth step:

Now we must tackle the irregularities by altering the conductor widths (Impedance Steps). Unfortunately ANSOFT was a bit mean here and blocked this option in the student version (and who just happens to have $ 20,000 to spare for the full version?). So we have to go back to the formulae in Section 2.3 and determine the substitute components using the pocket calculator. Fortunately, the conductor widths with and without the housing cover do not differ very much, so the component values can be transferred from the section referred to (quite literally, its only in the third decimal place that there is a small difference!):

Transition from	50 Ω to 17 Ω	17 Ω to 120 Ω	120 Ω to 10 Ω
Capacity	0.1pF	0.125pF	0.227pF
Inductance	0.327nH	0.565 nH	0,655 nH

Fifth step:

A new project is set up and then the circuit is assembled, using the physical line model. But for the length specifications of the lines we enter not values, but variables, to make optimisation easier. The following classification applies here:

• 17Ω line: Line1

• 120Ω line: „Line2

• 10Ω line: Line3

Here we access the variables block

Table 4:

	50Ω line	10Ω line	17Ω line	120Ω line
Electrical length in deg at 1700MHz		22.63	26.8	31.51
Width	1.83mm	14.53mm	7.89mm	0.25mm
Length		6.27mm	6.31mm	10.52mm
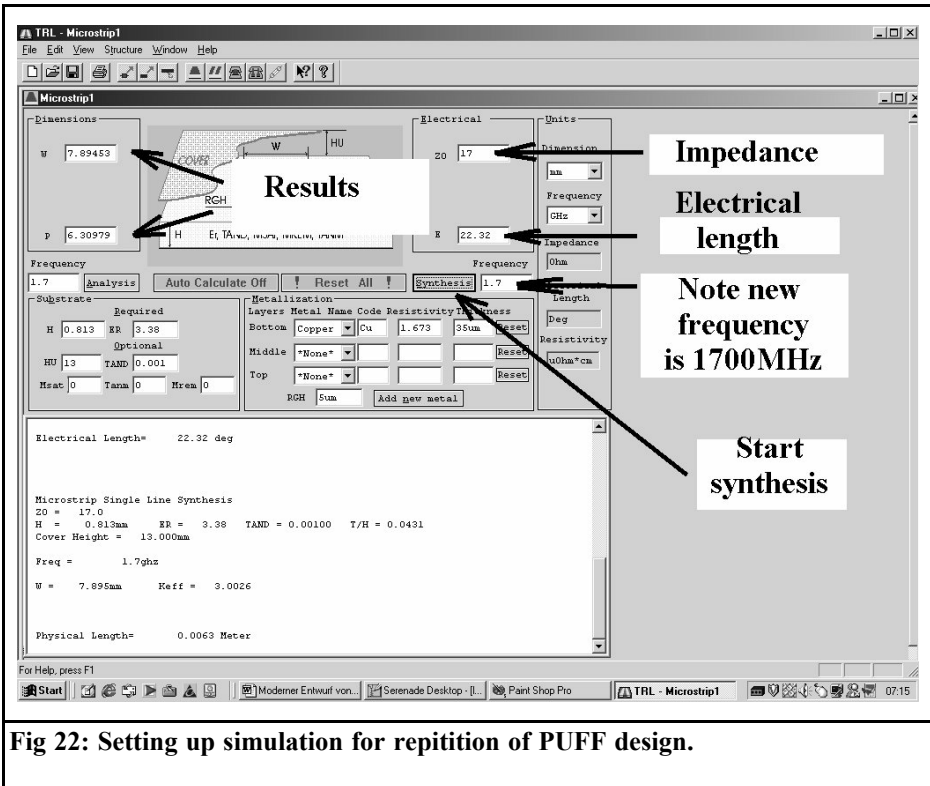
Fig 22: Setting up simulation for repitition of PUFF design.

through the "Parts" button, followed by "Control Blocks" and finally "Variables". Double click on this block to show the entries required. The lowest limiting value permitted, the target value and the upper limiting value for optimisation must always be entered between question marks (Fig. 23). In Fig. 24 we can finally see the complete circuit, ready for simulation, together with a substrate control block, a frequency control block, a variables block and an optimisation block.

Sixth step:

Now the simulation is carried out and then the results are analysed (Fig. 25 and Fig. 26). The result is exactly the same as the simulation with PUFF from Figs.11 and 12. Because of the components causing irregularities, the ripple cut off frequency has fallen to 1,300MHz and the

two camels humps in S11 are of different heights.

3.2. Adjusting components for optimisation of design

Here we basically proceed as for the design with PUFF. In the first run, we vary the length of the first and last line sections, together with that of the narrow lines acting as inductances. The tuning mode is available for such purposes. The following conditions apply:

• Extending Line1 lifts the right hump, S11, in particular.

• Shortening Line2 lifts the left hump, S11, in particular.

In this manner, we can bring the maximum values of the two S11 camels humps to a value of -16.44 dB. Finally, we increase the cut off frequency by
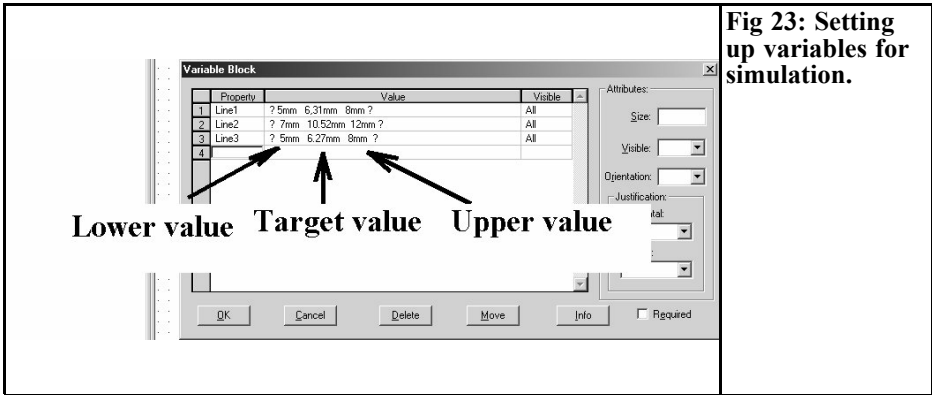
Fig 23: Setting up variables for simulation.

means of simultaneous reduction of all line lengths by the same factor to the required 1,700MHz.

Now we begin, for example, with the two outer 17Ω line sections and their length Line1. First we call up the circuit diagram onto the screen and execute the following checklist exactly (Fig. 27):

• Check that the Accumulate button has been activated.

• Click on the Tune button, which calls up the Tune menu.

• Double click on Variables Control Block in the circuit diagram to

ensure that the three variables (Line1 to Line3) are transferred into the tuning list. Then click on Line1 to select it.

• Switch over to Sweep

• Delete the tick for Step Width in %

• The start value (6.5mm selected) , the final value (8mm selected) and the step width of 0.5mm can be entered in the corresponding fields.

• We are ready, we can press Tune. But first, you should call up the window showing the representation of S11 and S22 to the foreground,
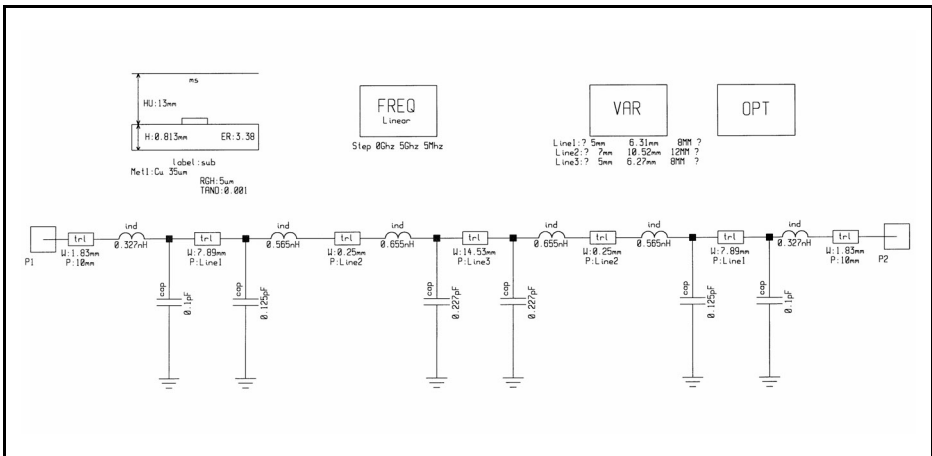
Fig 24: Completed circuit ready for simulation.
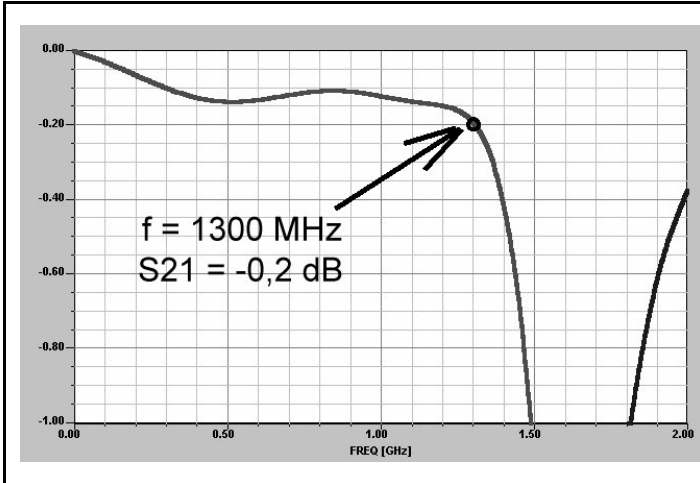
Fig 25: Results of simulation showing S21 curve.

this is very useful for looking at individual simulations and classifying them.

We can see immediately from Fig. 28 that a value of Line1 = 7.5 corresponds almost exactly to the required specifications. It is thus entered into the list as a new target value by means of double clicking on Variables Control Block and then the simulation is repeated. This time we are dealing with the precise value of the present limiting ripple value, which can be taken from Fig. 29 as S21 = -0.26 dB at 1,380 MHz. This is 320MHz too low and so we use our pocket calculator

in order to shorten all line lengths by the same factor. The relationship which applies is:

$$I_{NEW} = \frac{I_{OLD} \cdot 1700}{(1700 + 320)}$$

and thus we obtain the following values, which are immediately entered into the variables control block again:

- Line1 = 6.31 mm
- Line2 = 8.85 mm
- Line3 = 5.28 mm
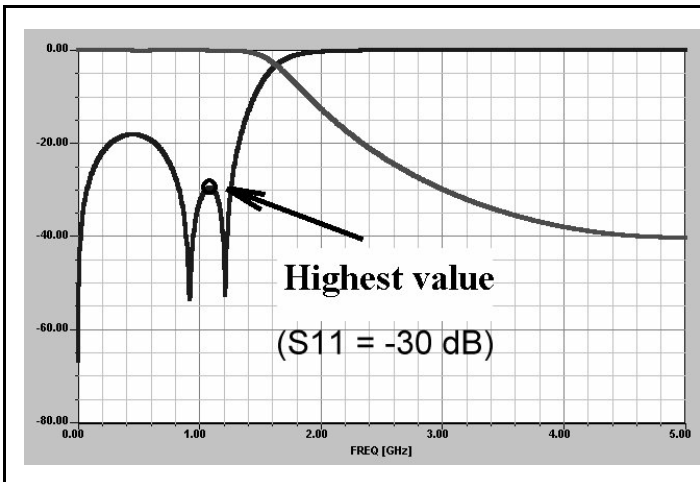
Fig 26: Results of simulation showing S11 curve.
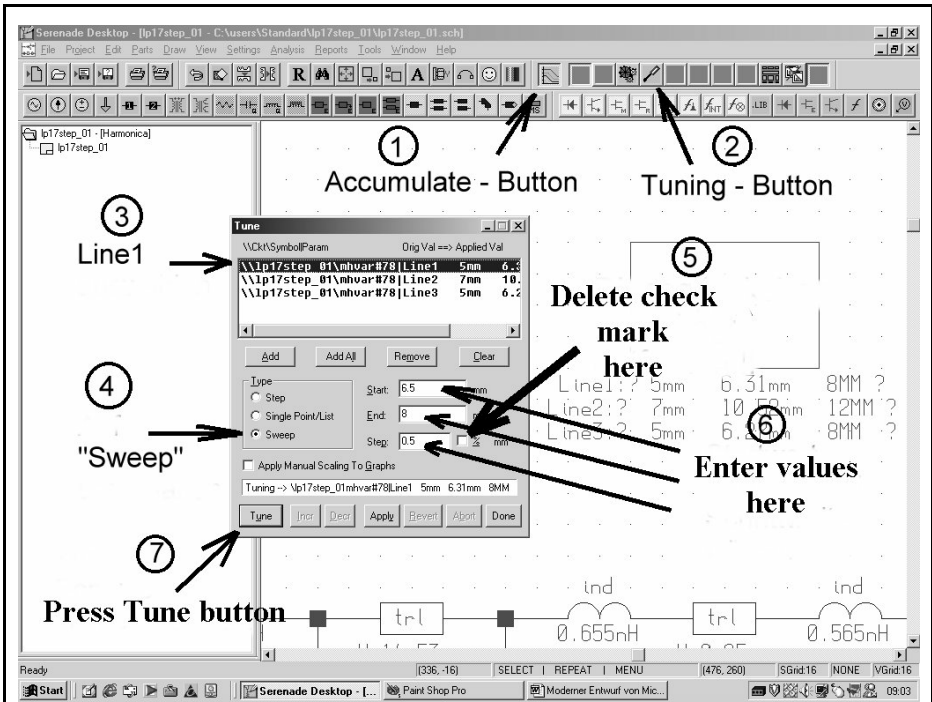
Fig 27: Setting up for tuning process.

A new simulation immediately provides information on the effect of these changes. It can be seen that here we have a cut off frequency of 1,620MHz and that a further correction of 80MHz is required. (One striking factor here is that the design frequency of the lines is now obtained at 2,100MHz, which corresponds exactly to the design using PUFF).

So now we have the following line
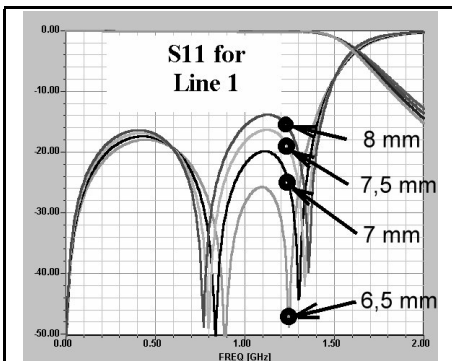
Fig 28: Results of tuning showing S11 curve.
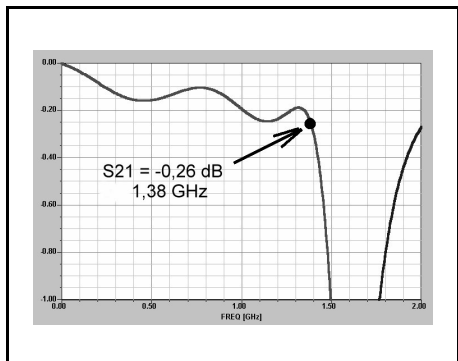
Fig 29: Results of tuning showing S21 curve.
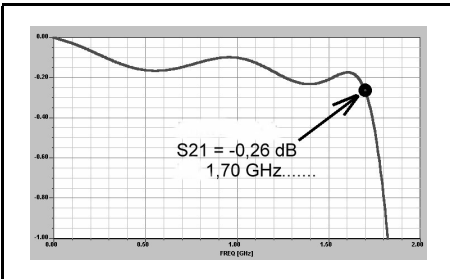
Fig 30: Final simulation results showing S21 curve.

lengths for the next stage:

• Line1 = 6.07 mm

• Line2 = 8.52 mm

• Line3 = 5.08 mm

If we use these for a simulation, we already have a cut off frequency of 1,670MHz. Yet we should look at the shape of S11 again, for these corrections have changed the hump amplitudes again, by 0.5dB. So Its clear what must be done next:

First correct the shape of S11 through Line1 and (if necessary) Line2, until you arrive at the desired hump amplitude of approximately -16.5dB. Then check the magnified shape of S21 to find the current value of the limiting ripple frequency (rated value : -0.26dB at 1,700MHz) and change all the length lines again, if necessary, by the same correction factor. If you are not satisfied, repeat this procedure again. Here, as a check and as an example, is the optimisation result for the printed circuit board design (Table 5).

Fig. 30 shows the associated final simulation of S21 as a proof of the successful

design. In Fig. 31, on the other hand, we can see a sweep from 0 to 10GHz. It shows the required shape for S11 and demonstrates what should never be forgotten in relation to these microstrip line filters, the attenuation rises to a maximum and then falls again (the technical expression is scatter resonance). In principle, this is unavoidable when line sections are used in filters, and can, for example, be improved by inserting a further low pass filters with a higher cut off frequency in series.

Finally, it is very interesting to compare this design with the one using PUFF from the first part of the article. It can actually be seen that the differences in the line lengths for all designs amount to only a few hundredths of a millimetre, and it is therefore certainly the time to prepare some test boards. Their measured data (with and without housing) will then decide on any further steps necessary to make the design ready for production. Normally, an accuracy of approximately 2% can be obtained in the first stage by applying the design principles of this article.

3.3. The Optimisation Option using HARMONICA

There is also the option of carrying out this optimisation entirely by a program. To this end, for specific frequencies or frequency ranges, a minimum value or a maximum value can be entered for several parameters.

It would also be entirely conceivable to forecast as many points as possible on the S11 curve and / or the S21 curve for this filter. Then carry out about 1,000 tests, and thus to get the computer to sort out the problem, which in certain circum-

Table 5:				
	50Ω line	10Ω line	17Ω line	120Ω line
Width	1.83mm	14.53mm	7.89mm	0.25mm
Length		5.08mm	6.14mm	8.20mm

Fig 31: Sweep from 0 to 10GHz showing scatter resonance.
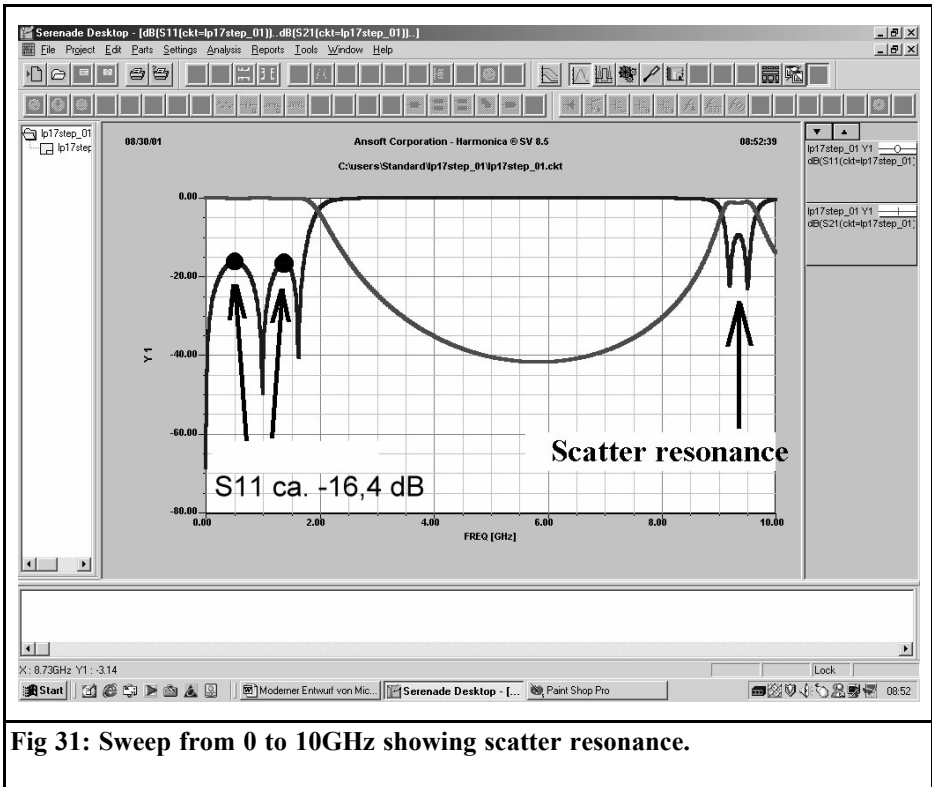
stances can be done within hours.

Unfortunately a maximum of 3 optimisation object are possible using the student version, and that just isnt enough. A program with so few basic points to work on quite often bends the filter curves in unpredictable or impermissible ways during the optimisation. So there would have to be a lot more objects for this to work successfully. If anyone would like to know more about this all the same, this is how to start on your way:

First step:

This would be to enter the minimum value, the target value and the maximum value for the three line lengths in the variables block. They must be entered between question marks. Luckily, that has already been done some time ago.

Second step:

Double click on the Linear Optimisation button (=black yellow red target) and open the sheet with the target objectives (Fig. 32) and enter the following conditions correctly:

Objective 1:

Make sure that in the frequency range between 1MHz and 1,700MHz the amplitude of S21 does not fall below -0.26dB.

Objective 2:

Make sure that in the frequency range between 1MHz and 1,500MHz the amplitude of S11 always remains below -16.5 dB.

Objective 3:

Make sure that at 1,700MHz precisely an amplitude of -0.26dB can be measured at S21.

Tip: Please avoid the value Zero MHz

Fig 32: Setting up objectives for optimisation.
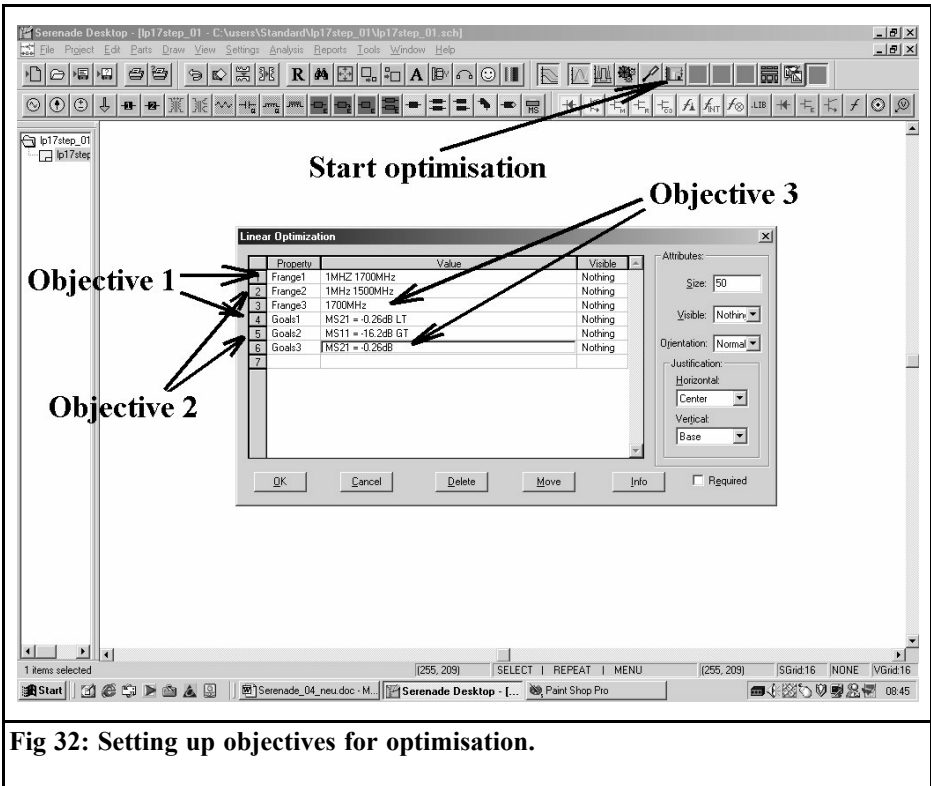
when entering frequencies, since this will lead to an error message!

Then close the Objective menu and press the button to start the optimisation.

Third step:

Please follow the sequence listed in Fig. 33 exactly and to do this call up the (blue) diagram with the simulation results for S11 and S21 into the foreground for observation. Then close the window again with Close. Hopefully everything has gone all right as described, and a usable result has been obtained.

4.
Measurement results

Both design methods produce a prototype built into a milled aluminium housing, fitted with SMA flanged sockets, and finally measured with a modern network analyser. The results are presented in the original untouched condition, for we are actually trying to make sure that the simulation can produce something that agrees with reality.

Fig. 34 shows the measured transmission range for the design using PUFF (amplitude range between 0 and -1 dB). Several things immediately strike one by comparison with the simulation:

The first Chebyschev ripple has a higher amplitude than the second
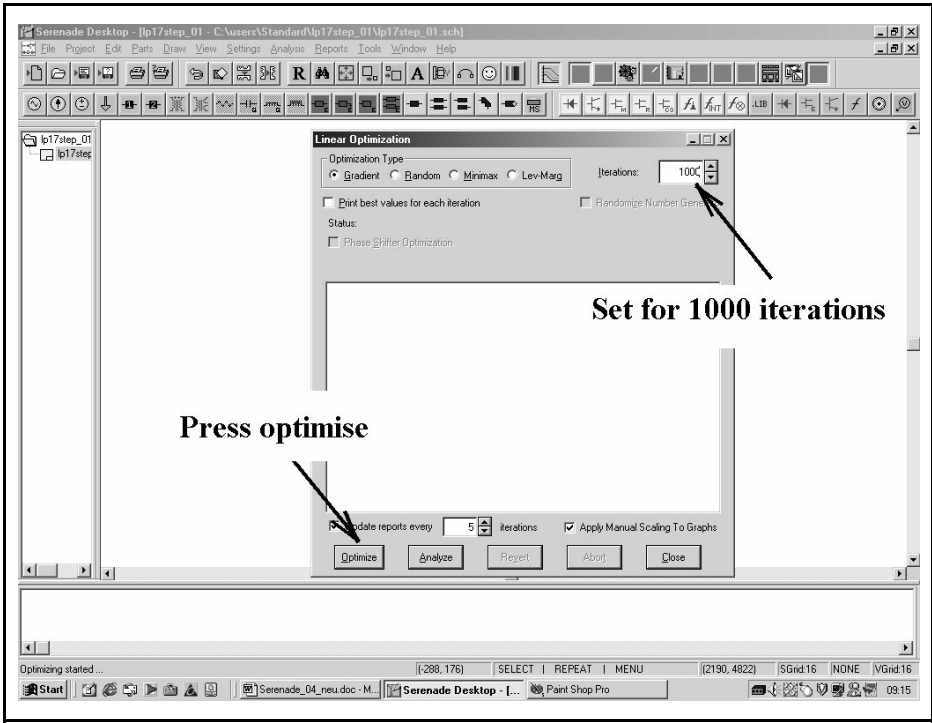
Fig 33: Final set up to satrt optimisation.

The cut off frequency, at 1.76GHz, is a good 3 % higher than forecast

The fundamental filter attenuation is markedly lower than was expected

Now lets look at these points in order:

In Fig. 35 we can immediately see the reasons for the unequal ripples observed in a). The two camels humps of S11 are not the same height! (If we look at the literature again, we see where the root cause lies> The formulae for the replacement components for the impedance steps are no longer accurate enough when there are extreme differences in the widths, such as 0.25mm to 14.6mm. Observant readers will realise at once that the lengths of the two external capacitor lines must be increased somewhat for correction purposes. This automatically lowers the cut off frequency, so that both objections, a) and b), have

disappeared afterwards.

The low fundamental filter attenuation is also interesting. It has its origins in the fact that the loss factor decreases markedly as the frequency falls, which is why a comprehensive data sheet was requested from the German importer. This contains an interesting diagram, showing the insertion attenuation of a microstrip line for various frequencies, in which various ROGERS materials are used one after another (Fig. 36). If we analyse these curves very precisely, for the R04003 material, then we can come to the following conclusions:

The loss factor $\tan\delta$, that is 0.002 at 10GHz, will fall by a factor of approximately 5.3 if the operating frequency is reduced to 1.5 to 2GHz.

If we now repeat the simulation with the new value of $\tan\delta = 0.002 \ / \ 5.3 = $
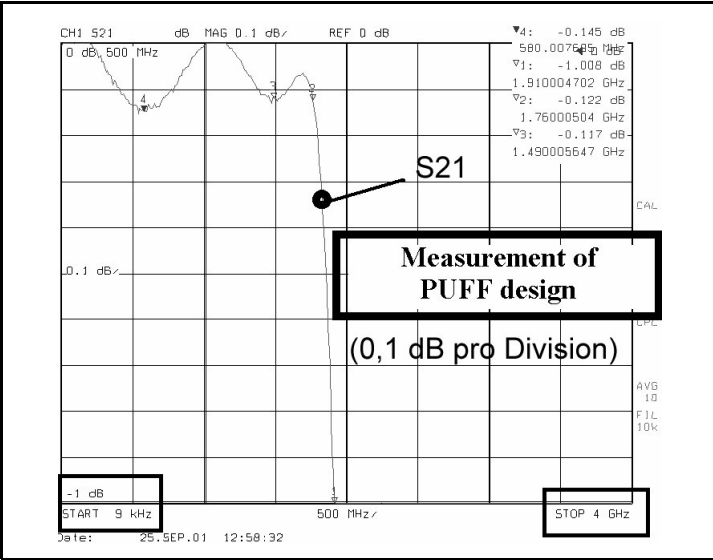
Fig 34:
Measurement
results on PUFF
design.

0.00038, then we obtain, very accurately, the attenuation observed during the measurements. (From now on, this must naturally be taken into account in future designs).

There is thus almost nothing more to say with regard to Fig. 37 and Fig. 38, for the ANSOFT design which supplies only a slightly lower ripple in the transmission range and smaller discrepancies for the camels humps of S11. Both the essential shape and the fundamental attenuation and the cut off frequency show scarcely any difference from the design using PUFF, so that what has just been said is also true for any further procedures using the ANSOFT printed circuit board.

The old adage from developers experi-
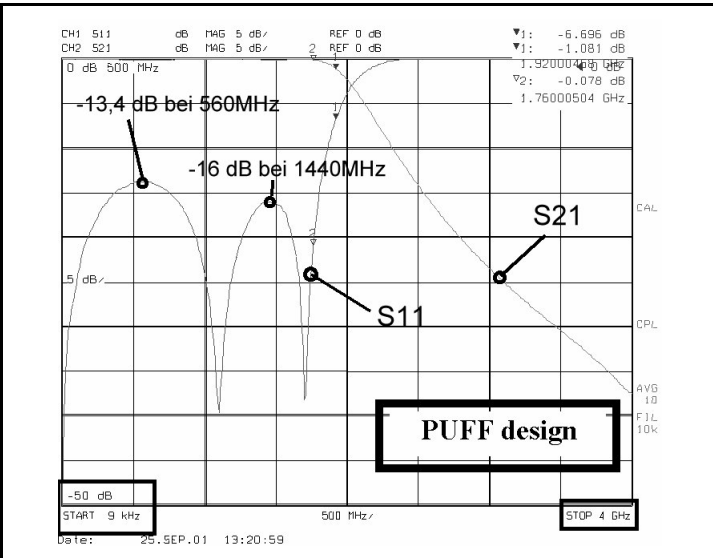
Fig 35:
Measurement
results on PUFF
design showing
unequal humps.
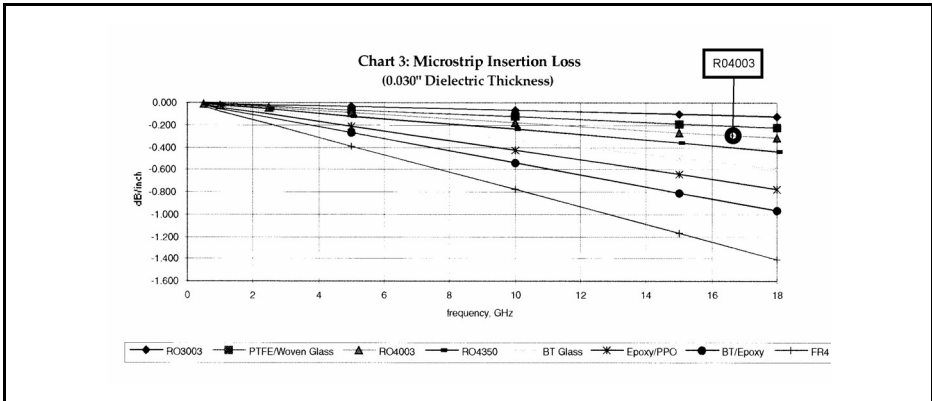
Fig 36: Data sheet on Rogers material.

ence is also confirmed in this case, that its not until the third board design that youre ready for full production.

5.
Summary and Evaluation

PUFF and ANSOFT HARMONICA are equally suitable for filter design. Those who prefer to trust their instincts and their experience as regards the optimisation which is required, or who want to test out spontaneous ideas as fast as possible, or who work with such programs only occasionally, are usually going to achieve their objectives faster using PUFF.

However, as soon as housing influences are brought in, or if many parameters are varied, or if various zoom functions are used, if markers are superimposed or if all WINDOWS options are to be used,
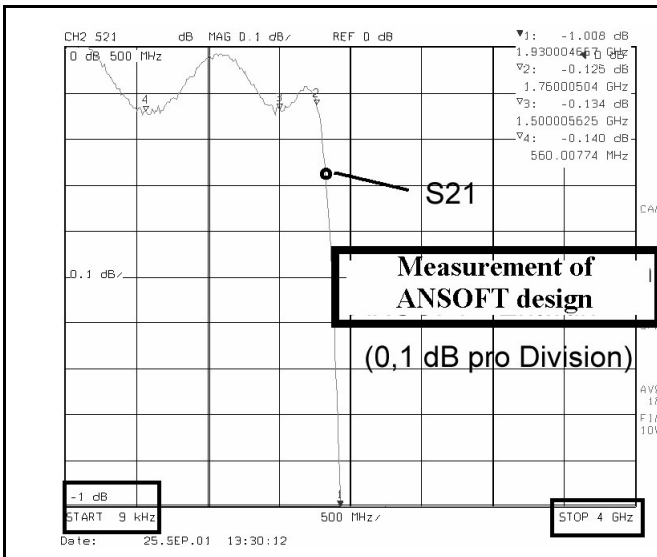
Fig 37: Measurement results on ANSOFT design.
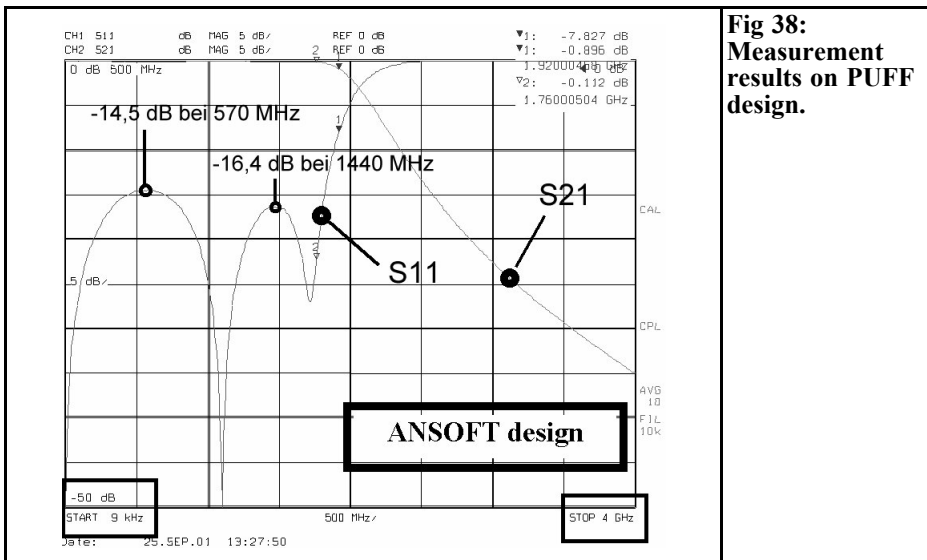
Fig 38: Measurement results on PUFF design.

you are naturally much better off with the student version of Serenade. It is simply a very modern development tool, accurately tailored to the practical needs of radio frequency developers. If only those all important options currently blocked were also accessible, then there would be no question at all about what to use.

6. Literature

[1] Gunthard Kraus: Modern Design of Stripline Band-passes VHF Reports, issue 2/2001

[2] English. Original Manual for PUFF

[3] Manual for ANSOFT-Serenade, (included in program for download)

[4] K. C. Gupta, Ramesh Garg, Inder Bahl, Prakash Bhartia: Microstrip Lines and Slotlines. Artech House Publishers Boston and London 1996. ISBN-Nr.: 0-89006-766-X

[5] David Pozar: Microwave Engineering. John Wiley and Sons, New York 1998. ISBN Nr.: 0-471-17096-8

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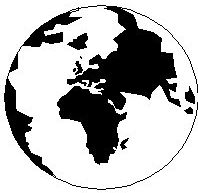
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