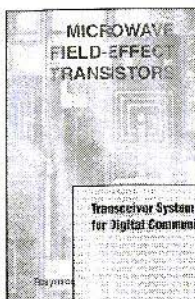


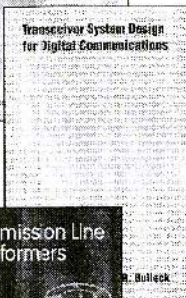
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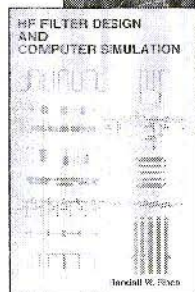
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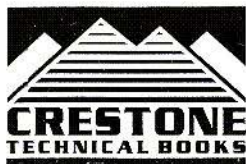
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Another New Year is upon us and a rather notable one at that - 1999 - the end of the Millennium (depending on your point of view of course!).

Krystyna and I would like to take this opportunity to wish you all a Happy and Prosperous New Year and welcome you as subscribers for 1999.

Mike



KM Publications, 5 Ware Orchard, Barby, Rugby, CV23 8UF, UK

Telephone: (0)1788 890365; INT: +44 1788 890365; FAX: (0)1788 891883

Email: michael.j@vhfcomm.co.uk

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



Gunthard Kraus, DG8GB

Design and Realisation of Microwave Circuits

Part-6

9. MICROSTRIP DEEP PASSES

9.1. Output Deep Pass after Mixer

Following the extensive band pass chapter, we now have to develop the two deep pass filters in the circuit

a) The intermediate frequency - i.e. the difference frequency between the input signal and the oscillator signal - should be screened out after the mixer. Thus the total frequency - the oscillator frequency including its second and third harmonics, as well as the total range of reception coming from the amplifier - must be suppressed. For this purpose, a deep pass is inserted here, with a limiting frequency of approximately 1 GHz.

b) To reduce the interference spectrum arising in the mixer, among other harmonics from the oscillator, there is a further deep pass between the oscillator and the mixer, with a limiting frequency of 1.7 GHz.

Both deep passes should be executed using stripline technology, with short sections of line with lengths of less than 10% of the wavelength as a replacement for discrete capacitors and coils as their extremely tiny values would lead to problems at higher frequencies with their natural resonances, or with precise calibration in series production. Stripline deep passes, however, have considerably greater dimensions to make up for this.

The following points apply here:

a) Wide, short sections of line with low impedance levels have high capacities, the inductances of which can still be neglected for short line lengths.

b) Short sections of line, as narrow as possible, with high impedance levels display a very small, even more insignificant capacity, and are therefore used as inductances here.

Of course, anyone using a board material with a dielectric constant which is too high - e.g. 10 - will have a nasty surprise here. Realising the inductances creates very big difficulties, since the

conductor widths can not be narrowed down indefinitely, and therefore the self-capacitance is always too high. It is best to stay below $\varepsilon_r = 5$.

It should also be pointed out that capacitors can be made effective in stripline technology with fewer problems than inductances. For this reason, we use the "low-coil" model of the already well-known Tschebyscheff standard deep pass, with $n = 5$, and a maximum reflection factor of 10%.

The filter coefficients were:

$$a_1 = 0.97322$$

$$a_2 = 1.37226$$

$$a_3 = 1.80324$$

By using these standardised values, we can, once again, implement any limiting frequency whatever very rapidly, and then easily specify the required component values.

If we initially start from the limiting frequency $f_g = 1$ GHz, we need only repeat the calculation from the previous chapter:

These component values should now be

$$\begin{aligned} Y_{C1} &= \frac{0,97322}{50 \Omega} \Rightarrow \\ \Rightarrow 2 \pi \cdot f_g \cdot C_1 &= \frac{0,97322}{50 \Omega} \Rightarrow \\ \Rightarrow C_1 &= \frac{0,97322}{2 \pi \cdot 1000 \text{ MHz} \cdot 50 \Omega} \Rightarrow \\ \Rightarrow C_1 &= 3,097 \text{ pF} \end{aligned}$$

$$\begin{aligned} X_{L2} &= 1,377226 \cdot 50 \Omega \Rightarrow \\ \Rightarrow 2 \pi \cdot f_g \cdot L_2 &= 1,377226 \cdot 50 \Omega \Rightarrow \\ \Rightarrow L_2 &= \frac{1,377226 \cdot 50 \Omega}{2 \pi \cdot 1000 \text{ MHz}} \Rightarrow \\ \Rightarrow L_2 &= 10,92 \text{ nH} \end{aligned}$$

$$\begin{aligned} Y_{C3} &= \frac{180324}{50 \Omega} \Rightarrow \\ \Rightarrow 2 \pi \cdot 1000 \text{ MHz} \cdot C_3 &= \frac{180324}{50 \Omega} \Rightarrow \\ \Rightarrow C_3 &= \frac{180324}{2 \pi \cdot 1000 \text{ MHz} \cdot 50 \Omega} \Rightarrow \\ \Rightarrow C_3 &= 5,739 \text{ pF} \end{aligned}$$

replaced by the sections of line. The relative length, l_c , of a short, wide line section, which is to act as a capacitance, is determined according to the following formula:

$$\frac{l_c}{\lambda} = f \cdot C \cdot Z_c$$

where "f" is the operating frequency and Z_c the impedance level of the (wide) line used. We normally use values of between 10 and 20 Ω for this impedance level.

A value as low as possible should be selected as this gives shorter line lengths, since the shorter the section of line is, the further the unavoidable "scatter resonances", with their attenuation interruptions in the filter attenuation band, slide away from the intended limiting frequency. True, this increases the conductor width, but its absolute limit is extended only by the width of the board or the screening housing.

Hence the experiment here with $Z = 10 \Omega$. For capacitor C1 this gives:

$$\begin{aligned} \frac{l_{C1}}{\lambda} &= 1000 \text{ MHz} \cdot 3,097 \text{ pF} \cdot 10 \Omega \Rightarrow \\ \Rightarrow \frac{l_{C1}}{\lambda} &= 0,03097 \end{aligned}$$

To work with Puff, we need this electrical length in degrees. To do this, we merely need to multiply the value just obtained by 360, and thus obtain:

$$l_{C1} = 0.03097 \cdot 360^\circ = 11.15^\circ$$

The same procedure is applied to the capacitor C3:

$$\frac{l_{C3}}{\lambda} = 1000 \text{ MHz} \cdot 5,739 \text{ pF} \cdot 10 \Omega \Rightarrow$$

$$\Rightarrow \frac{l_{C3}}{\lambda} = 0,05739$$

$$l_{C3} = 0,05739 \cdot 360^\circ = 20,66^\circ$$

Now we need "coils" again. In the narrow, high-impedance sections of line for the relative line lengths for producing an inductance, we have:

$$\frac{l}{\lambda} = \frac{f \cdot L}{Z_L}$$

Z_L represents the impedance level of the selected circuit for realising inductances. Here the conductor width should be selected to be as narrow as possible, and should go up to the practically realisable limit of 0.1 to 0.2mm.

Only in this way do we obtain the highest inductance values for the shortest circuit lengths, with the advantage of scatter resonances which are further away and higher filter attenuation.

Naturally, these extremely narrow paths increase the transmission loss in the pass band somewhat, but this is not important except for filters, which are used in transmitter high-level stages. Here there can certainly be problems - e.g. with heating up of the paths.

An impedance level of 135Ω is therefore selected here, which gives us:

$$\frac{l_2}{\lambda} = \frac{1000 \text{ MHz} \cdot 10,92 \text{ nH}}{135 \Omega} \Rightarrow$$

$$\Rightarrow \frac{l_2}{\lambda} = 0,08088$$

In degrees:

$$l_{1,2} = 0,08088 \cdot 360^\circ = 29,12^\circ$$

The design procedure just discussed is summarised once again in Fig.48.

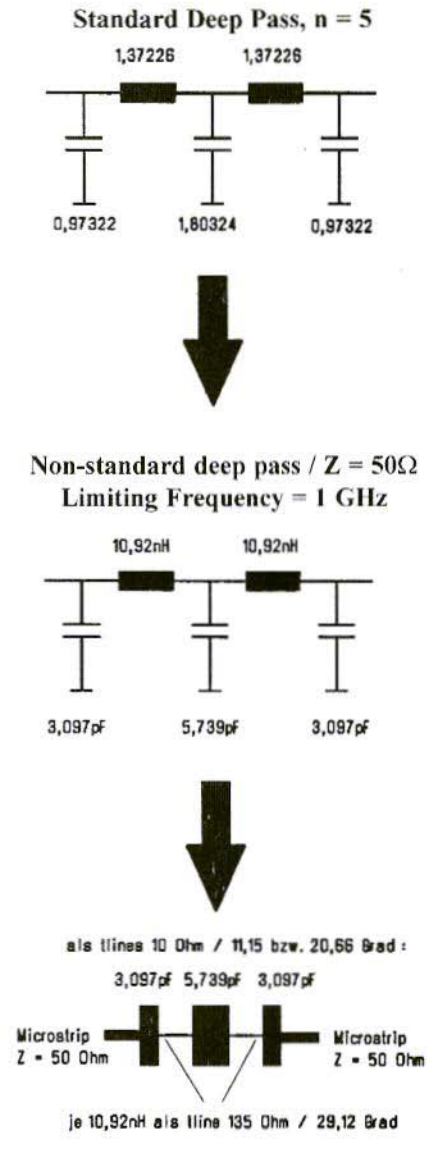


Fig.48: The Path from the Standard Deep Pass to the Microstrip Circuit

*Als = As, Grad = Degree,
Je = Every*

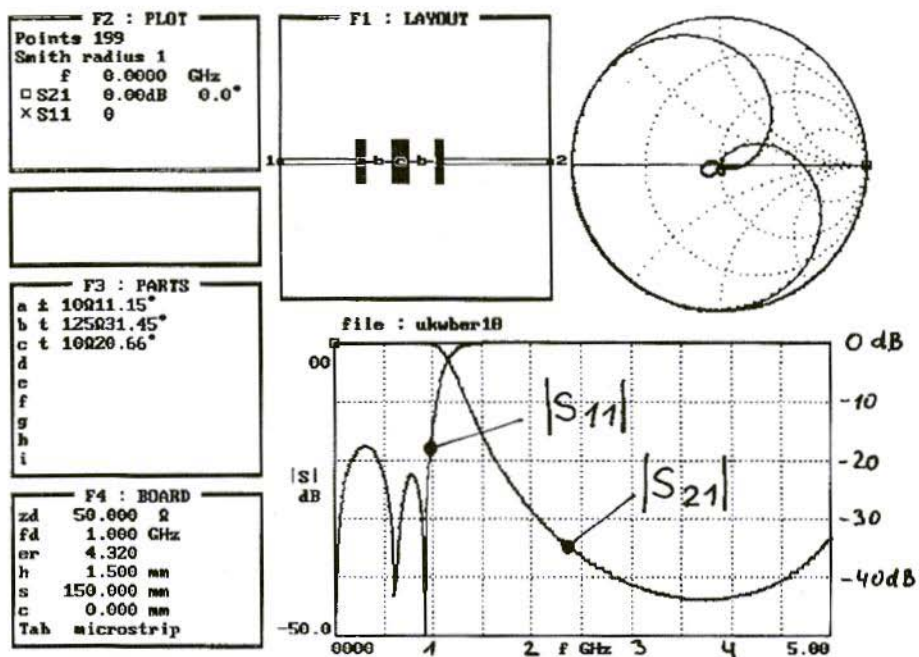


Fig.49: First Simulation of 1-GHz Deep Pass in Microstrip Technology (ideal circuit)

And now for the simulation using Puff. We enter the strip lines in field F3, prepare the layout in F1 and call up a plotting procedure. The inputs required are: frequency range 0 to 5 GHz, value range for $|S_{11}|$ and $|S_{21}|$ from 0 to -50dB. The result is shown in Fig.49, which naturally requires some explanation:

The values of the minimum input reflection attenuation $|S_{11}|$ are not equal, as between the poles, so that the procedure looks somewhat inelegant. This is simply due to the fact that we can unfortunately already observe the capacitive fraction in the "coils" - i.e. the wafer-thin lines - and we have to do something to counter this.

The solution is quite simple. We carefully reduce the length of the central line section "c" (which represents capacitor C3) in the component list F3 until the "unevennesses" have been eliminated. Check after each change by calling up a new plot.

If the minimum reflection attenuation is still rather too poor, we must also reduce to some extent the line sections "a" at the beginning and end of the filter, which form capacitors C1 and C5. A final correction at C3 then gives approximately the 20dB minimum reflection attenuation required.

If we continue, working with advanced modeling - i.e. with lossy sections of line, placing exclamation marks behind

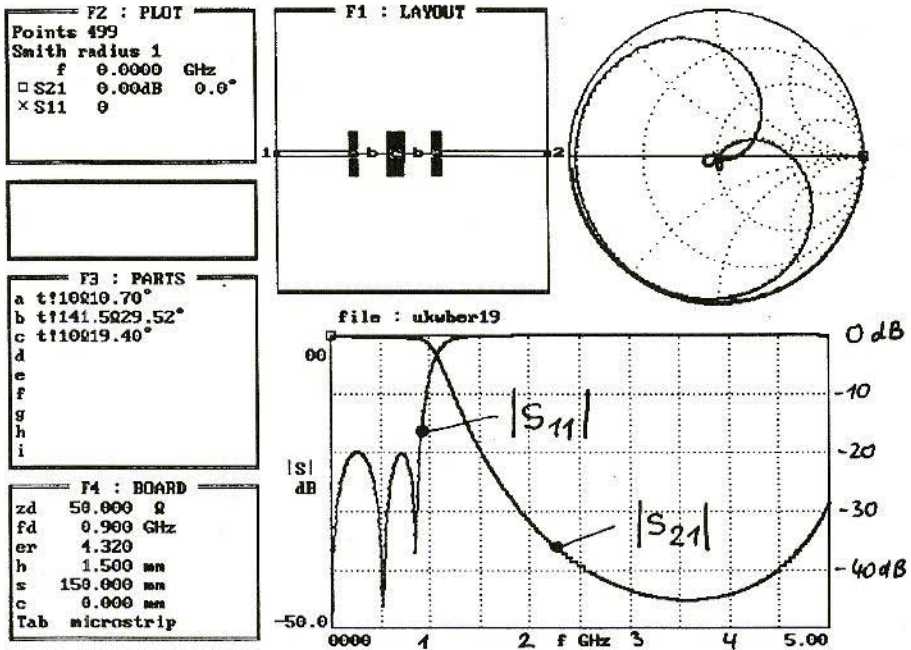


Fig.50: GHz Deep Pass with Losses following fine correction

the “line” values again, and correcting the earth until the ideal values are displayed again, we obtain Fig.50.

Here the limiting frequency has just been decreased a little further to obtain at least 20dB filter band attenuation at the oscillator frequency of 1,600 MHz. We have also made sure that the third harmonic of the oscillator frequency (4,800 MHz) was also sufficiently suppressed. This is ensured if app. 20dB of attenuation is present there as well.

Now it's the well-known procedure again: the open-end shortening of the wide line sections is determined again, the narrow lines are lengthened by these abbreviations, feed lines are connected with $Z = 50\Omega$, and the layout is designed. The diagram in the Puff manual (page 37) already mentioned

gives us a shortening of $0.47 \times$ board thickness for 10Ω and $\epsilon_r = 4.32$. For a thickness of 1.5mm, this gives, fairly accurately, 0.7mm.

The necessary CAD table is then as described by Table 1.

The 50Ω feed requires an additional width of 2.86mm.

Note:

Different open-end corrections are necessary for line section “a” (for C1 and C5), for it is connected with the “inductance” on one side only. The tiny width of 0.22mm at “b” need not be taken into account, as against 23.68mm, so on this side the complete shortening applies of $0.47 \times h = 0.47 \times 1.5\text{mm} = 0.7\text{mm}$.

But the width of the 50Ω line connected



Section	Z	CAD-Width	PUFF-Length	open-end Correction	CAD-Length
a	10Ω	23.68 mm	4.98 mm	see special note	see special note
b	125Ω	0.22 mm	16.16 mm	+2 x 0.7 mm = +1.4 mm	17.56 mm
c	10Ω	23.68 mm	9.02 mm	-2 x 0.47 x h = -2 x 0.7 mm	7.62 mm

Table 1

to the other side is 2.86mm. So here we need correct only by:

$$\begin{aligned} \Delta l &= 0,7 \text{ mm} \cdot \left(1 - \frac{W_1}{W_2}\right) = \\ &= 0,7 \text{ mm} \cdot \left(1 - \frac{2,86}{23,68}\right) = 0,62 \text{ mm} \end{aligned}$$

Thus, for both the first and last sections of line, we obtain a length of 4.98mm - 0.7mm - 0.62mm = 3.66mm, with a width of 23.68mm. Fig.51 shows the layout thus created.

As a comparison, we also investigated the influence of the foamed material glued to the underside of the housing cover to suppress the wave guide effects and cavity resonances.

Fig.52 shows the transmission loss simulation and measurement result for the frequency range from 0 to 1 GHz.

To sum up, we can say the following:

a) The limiting frequency measured, which was rather lower than the theoretical value, and the rather high degree of ripple for the attenuation in the transmission band can be traced back, for the most part, to the fact that the dielectric constants in the FR4 circuit board material are dependent on the frequency, which means the component

values obtained are not precisely what they should be.

This made an immediate impression. At low frequencies, ϵ_r was still 4.8, and had fallen to 4.3 at 1.7 GHz. If we call Puff up again and check through which

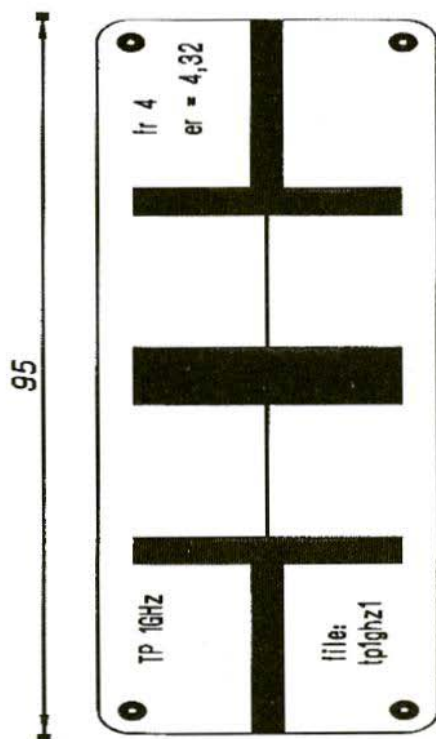


Fig.51: Layout of 1 GHz Deep Pass

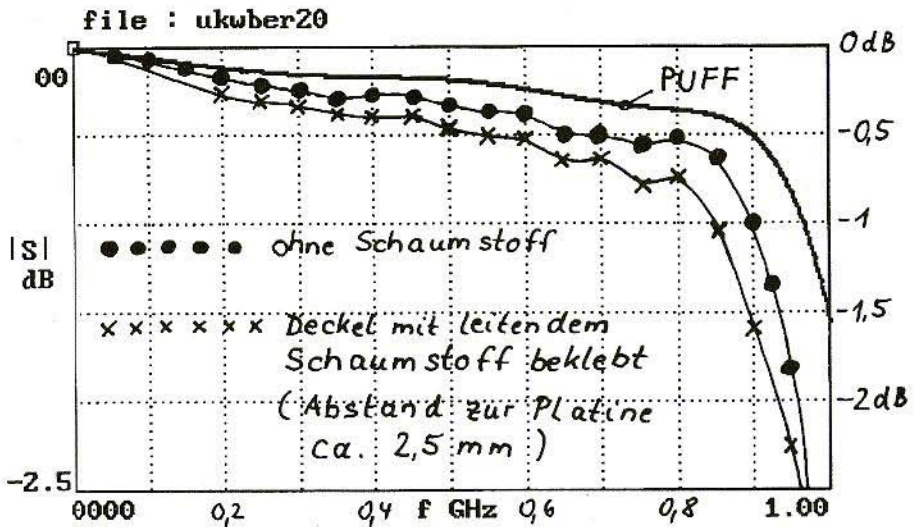


Fig.52: Deep Pass Transmission Attenuation: Range 0...1 GHz; Additional Curves: Influence of Attenuation Material
Ohne Schaumstoff = Without foamed material, *Deckel mit leitendem Schaumstoff beklebt* = Cover glued on with conductive foamed material, *Abstand zur Platine* = Distance from board, *Ca.* = Approximately

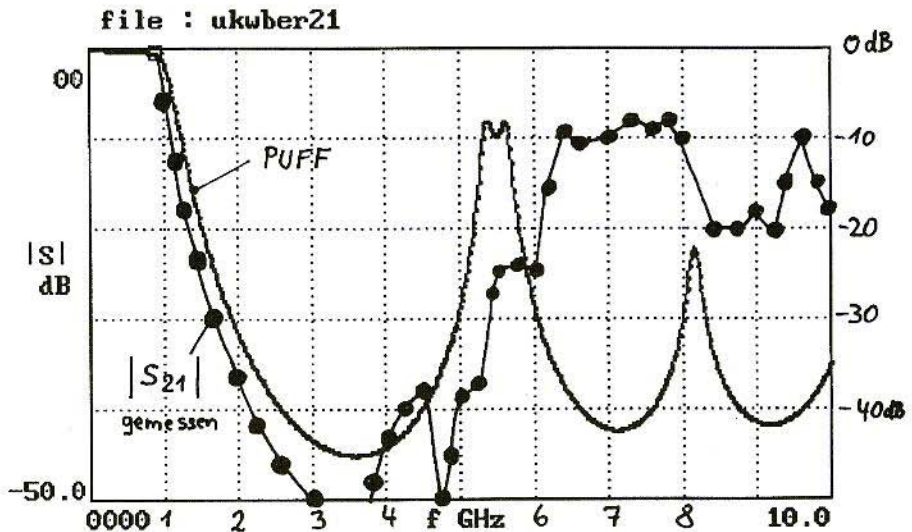


Fig.53: Path of Filter Attenuation of Microstrip Deep Pass, Range 0 to 10 GHz
Gemessen = Measured

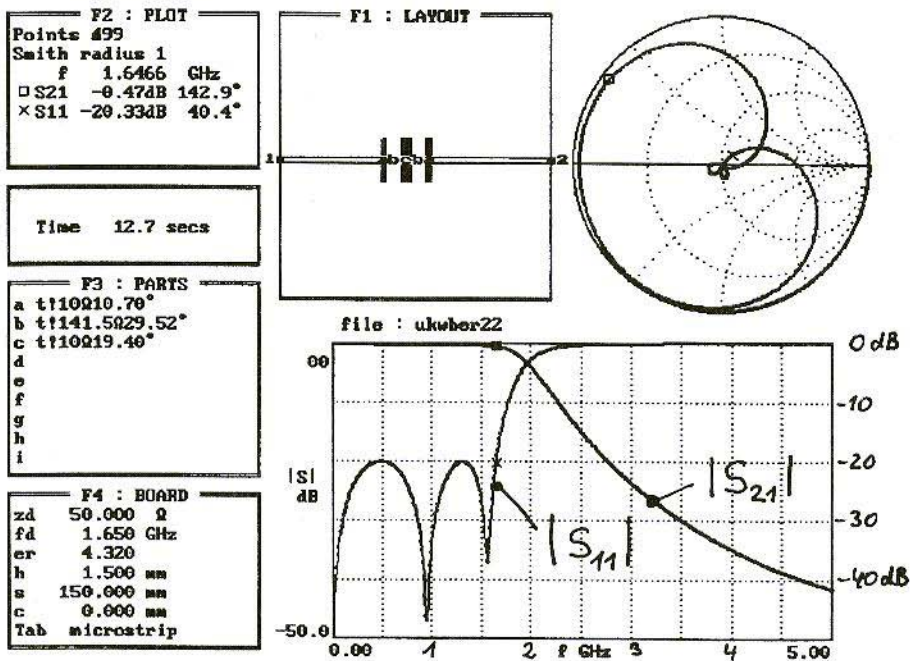


Fig.54: Simulation of 1.7 GHz Deep Pass with Losses following Displacement of Design Frequency

dielectric constants are well matched to the curve path measured, then we come to 4.7 at 900 MHz.

Those who want to work it out for themselves should do as follows. In component list F3, replace all length specifications in degrees by specifications in mm, which have already been entered in the tables above for CAD work. Then change the value for er in field F4, without changing the reference frequency, and continue to plot the new values until the simulated path and the measured result coincide.

b) The conducting foamed material on the underside of the cover certainly attenuates the cavity resonances. Unfortunately, it also increases the transmis-

sion loss, which is approximately 0.3 - 0.4dB at 900 MHz, with a distance of app. 3mm between the foamed material and the board surface. The smaller the distance between the board and the attenuation material and the higher the frequency, the greater the additional attenuation.

Finally, repeat the measurements for |S21| in the frequency range from 0 to 10 GHz (Fig.53). Here again it can clearly be recognised that the component measurements - probably because er has been given as 4.3 instead of 4.7 - are too high, and so the increase in the attenuation from the transmission limit begins too early.

Section	Z	CAD-Width	PUFF-Length	open-end Correction	CAD-Length
a	10 Ω	23.68 mm	2.70 mm	-(0.7 + 0.62) = -1.32 mm	1.38 mm
b	125 Ω	0.22 mm	8.82 mm	+2 x 0.7mm = +1.4 mm	10.22 mm
c	10 Ω	23.68 mm	4.92 mm	-2 x 0.47 x h	3.52 mm

Table 2

9.2. Oscillator Deep Pass

This assembly follows the oscillator circuit, and should still allow the oscillator frequency of 1.6 GHz to be well attenuated. But it should prevent the harmonics (3.2 GHz, 4.8 GHz, etc.) from getting into the subsequent mixer circuit and leading to undesirable mixed products there.

To do this, we call up the Puff file from the previous circuit, alter the reference frequency in field F4 to 1.65 GHz, and obtain Fig.54.

Using component list F3, after briefly deleting the exclamation mark, and entering the equals sign, we return to the electrical dimensions of the stripline sections. In addition, at this frequency - about 1.7 GHz - we can reckon on $\epsilon_r = 4.32$ again in fact.

If we then carry through the open-end correction again, with the values 0.7mm and / or 0.62mm, we already have the data to design the board (Table 2).

The 50 Ω line again requires a width of 2.86mm.

If we examine the list carefully, we see that the CAD outputs become so exotically brief - due to the low impedance

levels of 10 Ω selected and a resulting width of 23.68mm for the "capacitors" - that the open-end shortening is already of the order of the residual length.

This is naturally a very tricky area, due to the dispersion of the board material data, and because of the risk of under-etching the track edges while preparing the board.

Fig.55 shows the board design, which is already displaying a rather unusual look,

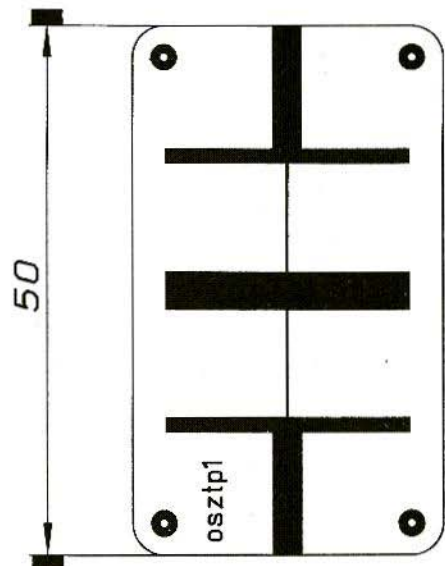


Fig.55: Layout of 1.7 GHz Microstrip Deep Pass, First Version

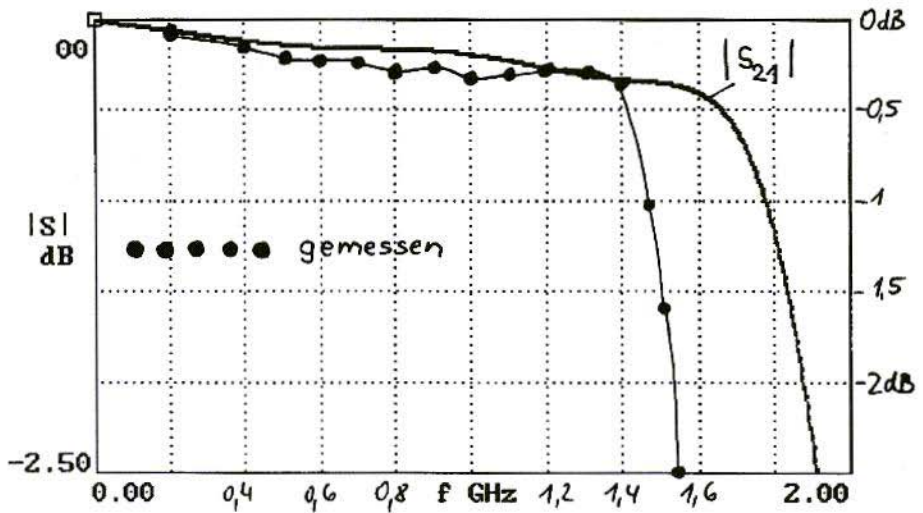


Fig.56: Transmission Loss $|S_{21}|$ in Frequency Range 0 to 2 GHz for First Version of Deep Pass
Gemessen = Measured

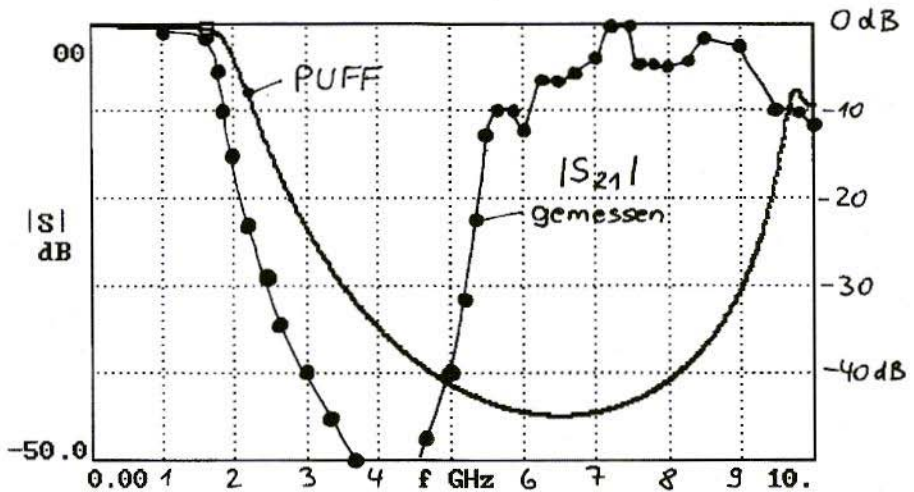


Fig.57: $|S_{21}|$ for Range 0 to 10 GHz in 1.7 GHz Deep Pass, First Version
Gemessen = Measured

for the whole thing almost looks like a Yagi antenna.

The measurement results look just as unpleasant, for whilst the transmission

loss measured corresponds very precisely to the calculated value (see Fig.56), the transition from the transmission band to the filter attenuation band

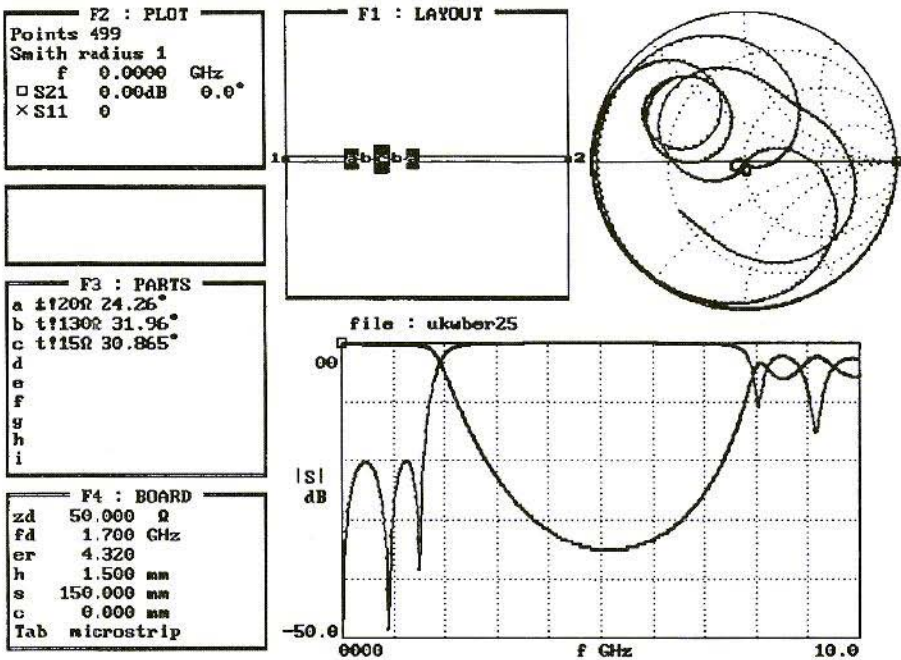


Fig.58: Puff Simulation of 1.7 GHz Microstrip Deep Pass, Second Version

sets in markedly too early. Anyone looking more closely at the diagram can see that there is a deviation of about 250 to 300 MHz, as against the theory - i.e. about 15 to 20% - decidedly too much!

Measurement up to 10 GHz (Fig.57) confirms even more clearly that the transition is far too early, to which are also added, from app. 5 GHz, the usual effects, such as cavity resonance, wave guide behaviour, re-radiation from the track edges, etc., which can not be detected by the simple Puff simulation model being used here.

But the markedly steeper increase in the filter band attenuation from the transmission limit onwards must be evaluated as genuinely positive.

Unfortunately, the attenuation collapses completely from 5 GHz - there is probably not much filter attenuation to speak of here.

One way out which offers itself is an increase in the impedance levels in the "capacitor striplines", for this reduces the track widths and thus the lines must be extended.

The impedance levels at C1 and C2 are experimentally doubled from 10 to 20Ω, while the impedance level at the middle capacitor is increased only from 10 to 15Ω. Its capacity is in fact twice as high as that of C1 and C5, and the conductor length should be kept as short as possible, to keep the "scatter resonances" within bounds.

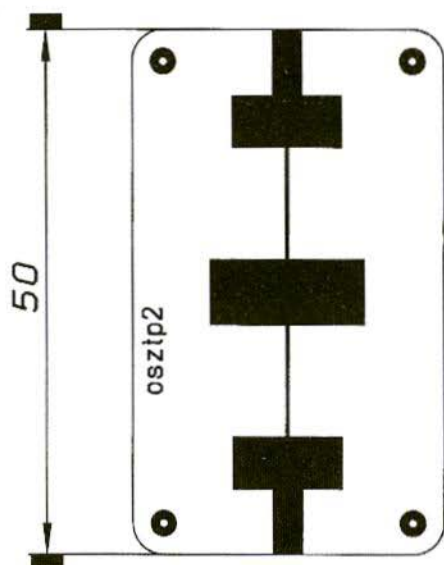


Fig.59: Layout of 1.7 GHz Deep Pass, Second Version

Even the thin lines for the “inductances” are widened, in that the impedance level is lowered to 125Ω . This makes it easier to handle the problems arising

when boards are produced with such long, wafer-thin lines.

The filter calculation and layout, including the fine correction, is once again carried out exactly as in the method described above. The results and / or the component values for the “ideal” deep pass are shown in Fig.58. As expected, the attenuation rapidly falls back above 5 GHz, because the scatter resonances are favoured by the greater conductor lengths. In contrast, the layout shown in Fig.59 works out to be somewhat more balanced. The diagrams in Figs. 60 and 61 include the losses, and the values measured on the board are also entered.

The result now looks a little better, for the deviations from the simulation lie at around 10%, if we deduct the rather higher basic attenuation due to the mounting of the board in the housing, including the junction between each SMA jack and the board.

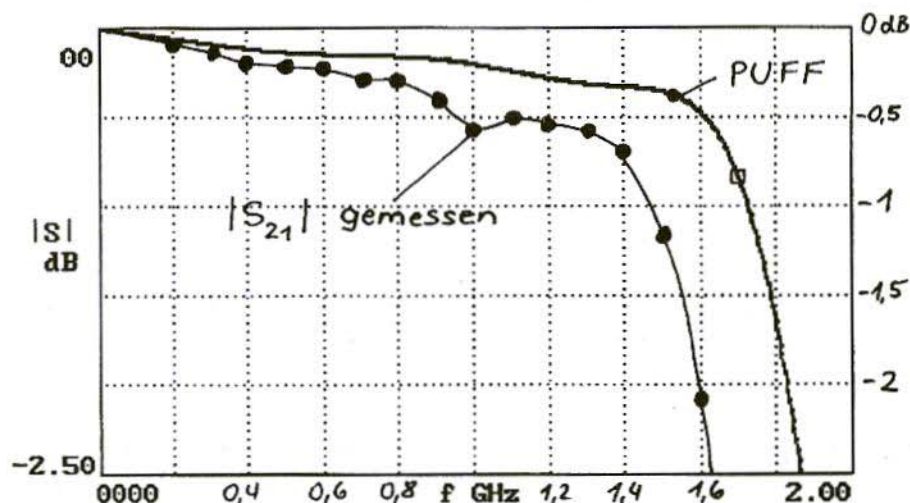


Fig.60: Measured Path of Transmission Loss in Range 0 to 2 GHz of 1.7 GHz Deep Pass, Second Version
Gemessen = Measured

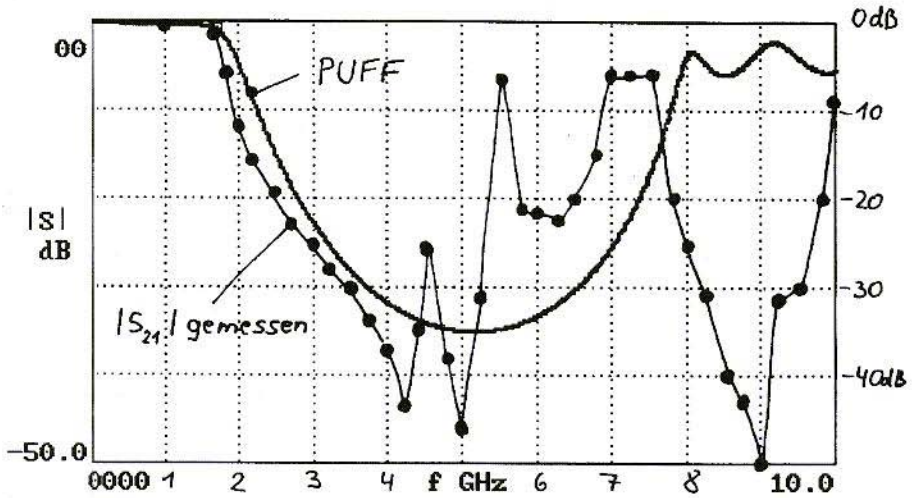


Fig.61: $|S_{21}|$ in Range 0 to 10 GHz of 1.7 GHz Deep Pass, Second Version
Gemessen = Measured

Here too, we can make a quick end of matters, using the following method:

We estimate how many MHz too early the rise to the filter attenuation band begins. In the deep pass board just examined, this is approximately 170 MHz.

We simply increase the limit frequency in field F4 by this amount - so we enter 1,820 MHz there.

Then we use the component list F3 and the equals sign to obtain the new associated mechanical dimensions of the three striplines a to c. After open-end shortening of the wide lines, plus the extension of the "inductances", a new corrected layout is prepared and a board produced.

The measured results obtained in this way are shown in Figs. 62 and 63. There is now a very serviceable match, but of course, once again, only up to about 5 GHz.

9.3. Summary of Experiences and Design Rules for Micro-Strip Deep Passes

1. Start with the standard deep pass, re-dimension it to the desired operational values and calculate the values for the required coils or capacitors.
2. Capacities are realised through very wide striplines (initial value: Z app. 20Ω), inductances by means of very narrow lines (guideline: Z to exceed 100Ω by as much as possible).
3. The line lengths here should not exceed 10% of the wavelengths (corresponding to an electrical length of 36°).
4. In the first Puff simulation, we almost always observe a marked deviation from the ideal path for the S parameter, which is due to the capacitive fraction of the "inductance lines". By carefully reducing the line lengths on the "capacities" under continuous "Puff monitoring", this defect can be eliminated.

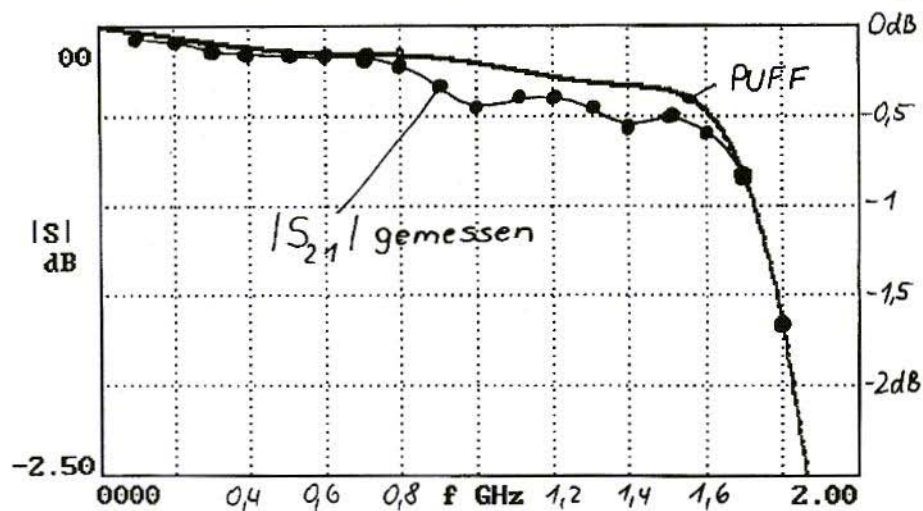


Fig.62: $|S_{21}|$ for 1.7 GHz Deep Pass, Third Version, but with Design Frequency increased by 170 MHz
Gemessen = Measured

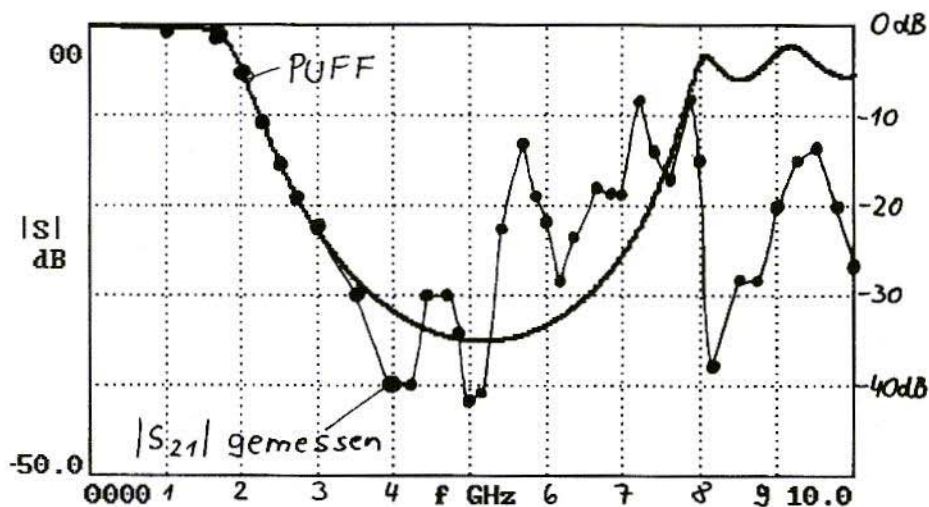


Fig.63: As Fig.62, Range from 0 to 10 GHz
Gemessen = Measured

5. We should try and make the line lengths for the "capacitors" as short as possible, for only in this way can the

unavoidable interruptions in the filter attenuation band (and thus the scatter resonances) be pushed very high up.

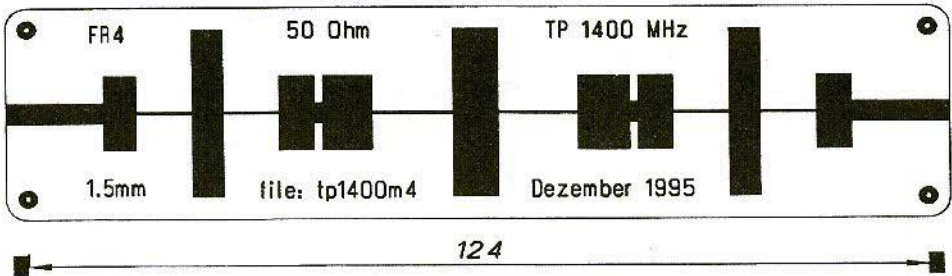


Fig.64: Expanded 23cm Deep Pass with Improved Filter Attenuation
Dezember = December

However, it is not just the board or housing dimensions which set the limits for expanding the line widths of microstrips. As the line lengths increase and the widths grow, structures actually arise which can no longer be correctly represented by simulations using Puff. The difference between theory and practice rises to more than 20%. Moreover, in certain circumstances, the attenuation collapses to zero in the filter attenuation band.

So from experience we can recommend that in the “capacitor lines” the ratio between the track width and the final track length on the board should be selected to be under 8 to 10.

6. When the completed filter is measured, we always detect a displacement of the actual limiting frequency towards lower values. The frequency deviation should then be determined, and the simulation should be repeated, with a design frequency in field F4, increased by the amount of the error.

When the new dimensions have been determined, carry out the correct procedure for an open end correction and re-jig the layout and board, and you will usually be on target.

7. Should there be interference from the interruptions in the filter attenuation band (= the scatter resonances), we can combine the desired deep pass, individually, on the left-hand and right-hand sides, with a further additional deep pass. These additional filters have limiting frequencies which are so high that their filter band attenuation comes in useful where the desired pass is already displaying its interruptions.

Fig.64, a deep pass for the 23cm band, shows how such a solution might look in practice.

8. The following rules generally apply in the selection of boards and materials in the micro-wave band:

- The higher the design and operational frequencies are, the thinner the board material must be selected to be, so as not to favour the re-radiating of energy at the track edges.

- Material with a thickness exceeding 1mm is normally used only in the range under 2 GHz. Even FR4 material should be used only up to this limit. Up to this frequency, it also makes sense to select a copper coating with a thickness of 35µm. For higher frequencies, the skin



effect means the copper layer can be selected to be thinner without any reservations.

– Boards made from pure Teflon material are appallingly impractical to deal with. The advantages of the extremely low losses up to and beyond 10 GHz have to be paid for by inadequate mechanical properties. Not only does it take a lot of skill to drill the holes (because the drill does not meet any kind of resistance). There is also the fact that the board bends even under its own weight.

– Mounting within the housing can be done only by soldering to the continuous-earthed underside, or by using very expensive epoxy conducting adhesive, which contains a lot of silver. There are just as many problems in connection

with the through-plating of the earth connections, since first you have to do a little etching with hydrofluoric acid.

– The later developments (e.g. Teflon or ceramic powder with glass cloth) are rather more good-natured, and mechanically more stable. There are a few recent developments, e.g. from Rogers, the “RT-Duroid inventors”, which point to really tremendous progress.

We can now use strongly cross-linked hydrocarbons as mechanical stabilisers in the carrier material and, with the same or better quality, as against RT-Duroid, we obtain such high strength that such boards can be sawn, drilled, through-plated and screwed exactly like FR4 material (i.e. epoxy glass cloth).

To be continued

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Matjaz Vidmar, S53MV

Wideband & Low-Noise Microwave VCO

1. OSCILLATORS FOR SPECTRUM ANALYSERS

An important piece of radio-frequency or microwave test equipment is certainly the RF spectrum analyser. Spectrum analysers can roughly be divided in two groups: professional and low-cost. Although there are many differences between these two groups of instruments, the most important difference is in the type of (sweep) oscillator used for the first frequency conversion.

Professional spectrum analysers use YIG (Yttrium-Iron Garnet) oscillators. YIG resonators can be tuned over wide frequency ranges (more than an octave in the microwave frequency range) with an external DC magnetic field. YIG resonators also have a high Q allowing a low phase noise when used in an oscillator. Finally, the tuning characteristic of a YIG oscillator is linear, since the frequency is directly proportional to the applied DC magnetic field.

Low-cost spectrum analysers use varactor (varicap) tuned oscillators. The Q of varactor diodes is rather low and is inversely proportional to the operating frequency. Silicon varactors usually have the Q less than 30 at 1 GHz. GaAs varactors are somewhat better, but they are not easily available and are much more expensive. The frequency coverage of low-cost spectrum analysers is therefore limited

below 1GHz and the phase-noise performance is usually 20-30dB worse than that of YIG oscillators.

A spectrum analyser can also be built by a skilled radio amateur. While most circuits of professional spectrum analysers can be reproduced in amateur conditions, the major problem is building a wideband, low-noise VCO for the first (swept) conversion. YIG oscillators probably can not be built in amateur conditions. The price of a new commercially available YIG oscillator is comparable to the price of a surplus professional spectrum analyser.

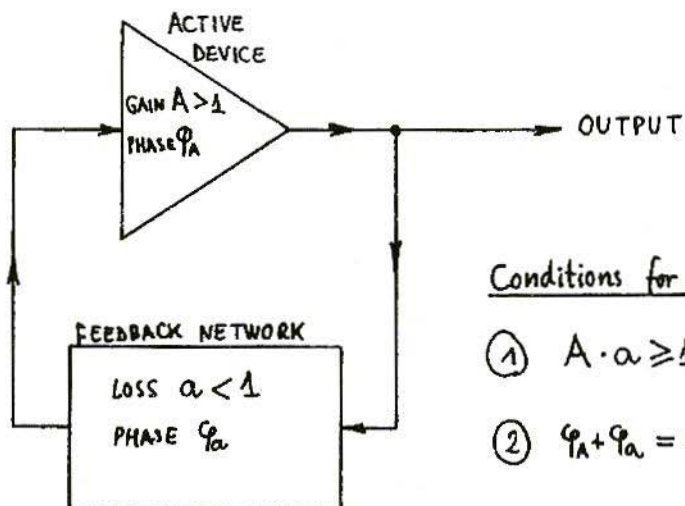


Fig.1: Oscillator Block Diagram

A varactor-tuned VCO covering the frequency band 2-4GHz will be presented in this article. Such a VCO allows the design of a spectrum analyser with the first IF in the 2GHz range, similar to professional instruments. The phase noise of the described VCO is reasonably low, within 20dB of a free-running YIG oscillator. Finally, the VCO design is fully reproducible using standard SMD parts mounted on a conventional FR4 (0.8mm thick) printed-circuit board.

2. SOME OSCILLATOR FUNDAMENTALS

The design of an amateur RF spectrum analyser therefore depends strictly on the type of VCO that is available for the

Conditions for oscillation:

- ① $A \cdot \alpha \geq 1$ (enough gain)
- ② $\phi_A + \phi_a = m \cdot 2\pi$ (phase)

first conversion. In order to explain the design of a wideband varactor-tuned VCO, some oscillator basics have to be discussed first. Any oscillator must contain an active device (gain) and a feedback network, as shown in Fig.1. There are two conditions for oscillation: enough gain (including feedback loss) and correct phase of the feedback.

The frequency of oscillation is determined by both conditions as shown in Fig.2. However, close to the actual oscillation frequency, the gain curve has a broad and flat peak. Therefore, the exact frequency, the stability of the oscillator and the phase noise are all determined by the phase response. The steeper the phase slope, the better the stability of the oscillator and the lower the phase noise.

Low-frequency oscillators are usually designed for operation at a total phase shift of $2 \cdot \pi$ radians. π radians are usually provided by the active device

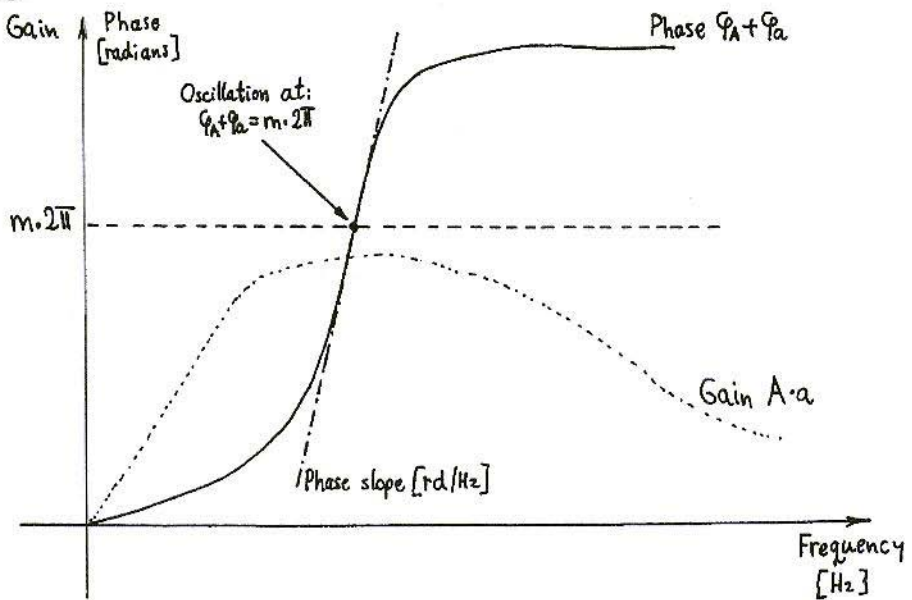


Fig.2: Frequency of Oscillation

(transistor) itself, while the remaining π radians are provided by the feedback network. A steep phase slope is obtained by a high-Q LC tuned circuit or quartz-crystal resonator.

At frequencies above 1 GHz the phase shift of all known active devices is much larger than π radians due to chip and package parasitics. If an oscillator is designed for a total phase shift of 2π radians, then only a small fraction is left to the feedback network. The phase slope of the latter is certainly not very steep resulting in poor stability and high phase noise.

In the case of a variable-frequency oscillator, the frequency coverage is rather restricted since the influence of the feedback network is small compared to the active device itself. Conventional oscillator designs (with an LC circuit or transmission-line equivalent coupled to a

negative-resistance active device) will only provide a restricted frequency coverage and poor stability. Most microwave oscillators are designed in this way, since a negative resistance can easily be obtained from most microwave transistors when considering chip and package parasitics.

Replacing a negative-resistance device with a true two-port, unidirectional amplifier provides the oscillator designer with some more degree of freedom. In particular, the feedback network can be tailored for the desired amplitude and phase response. The feedback network should both match the impedances and compensate the phase shift of the active device as well as introduce its own frequency-dependent amplitude and phase response.

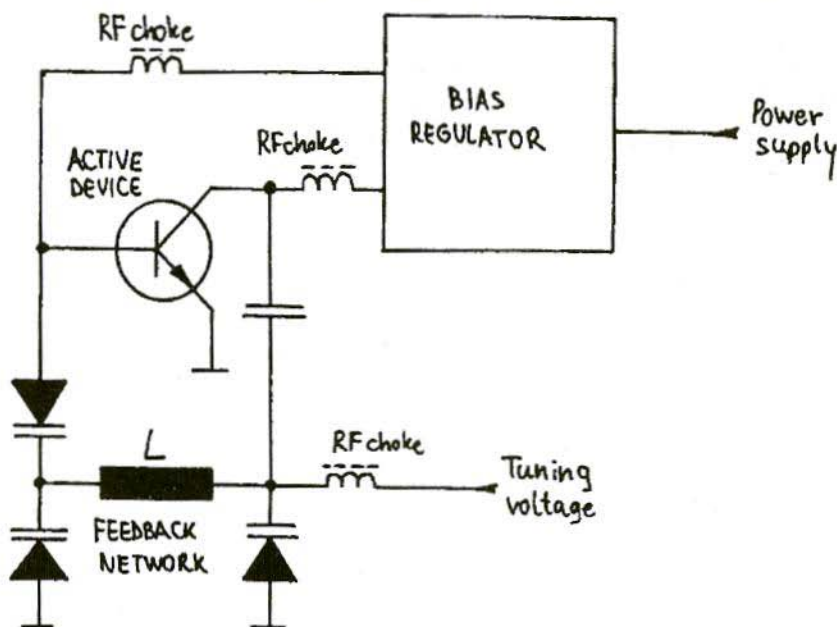


Fig.3: Wideband Microwave VCO

A successful wideband microwave VCO design is shown in Fig.3, covering more than an octave with conventional silicon varactors. Although I developed this circuit for my first spectrum analyser built in 1985, I published the circuit diagram one year later as part of a satellite-TV receiver indoor unit [1], [2]. Many other amateurs used this circuit in their own spectrum analysers and other RF test equipment, but only few acknowledge the original source [3].

The major drawback of the VCO design from Fig.3 is that its operation is still based on lumped components: capacitors (varactors) and inductors. Its upper frequency limit is therefore defined by the parasitic inductance of available (packaged!) varactors to about 2-2.5 GHz. The phase noise can be reduced by a carefully designed bias regulator, to

stabilise the current through the bipolar transistor, so that the impedances and phase shifts do not change.

3. INTERDIGITAL-FILTER MICROSTRIP OSCILLATOR

Active-device phase shifts become much larger at higher frequencies. For example, the phase shift inside a helium-neon laser tube may reach one million radians, although its theory of operation is the same as for other electrical oscillators. The total amplifier plus feedback phase shift still has to be an integer multiple "m" of 2π radians, however

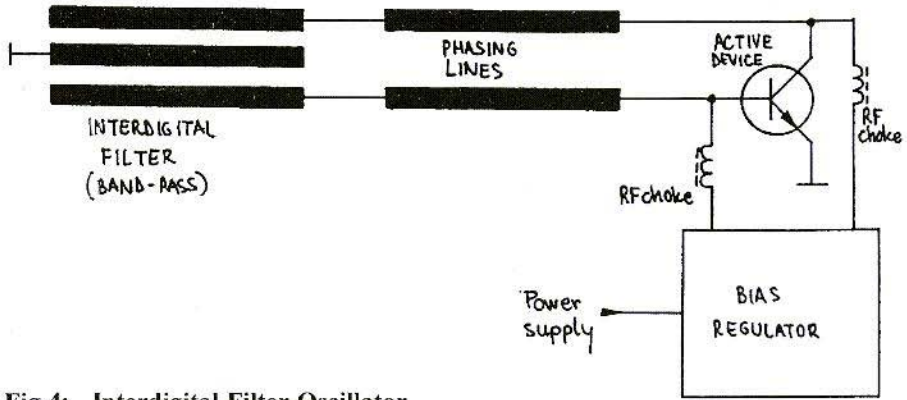


Fig.4: Interdigital Filter Oscillator

“m” is not restricted to unity and may become very large at lightwave frequencies.

Lumped-component oscillator designs become useless at microwave frequencies, since all available components behave as sections of transmissionlines. On the other hand, additional phase shifts can be readily implemented as sections of transmission lines. The total phase shift of a microwave oscillator may be an integer multiple “m” of 2π radians. “m” may be larger than unity, but still relatively small at microwave frequencies.

Most microwave circuits are built in microstrip technology, since the latter is compatible with inexpensive manufacturing techniques like printed-circuit boards and surface-mount components. An example of a microstrip oscillator is shown in Fig.4. An interdigital band-pass filter is used as part of the feedback network. In order to bring the total phase shift to an integer multiple of 2π radians, additional delay lines may be required to obtain the correct feedback phase.

Although the oscillator described here may look complicated, it includes some advantages when compared to conventional low-frequency designs. Although the Q of microstrip resonators is not very high (in the range of 50-100), the phase slope may be made high thanks to the large total phase shift (increasing the multiple of 2π radians). On the other hand, oscillation at unwanted multiples of 2π radians can be suppressed by tailoring the amplitude response of the feedback network.

A fixed-frequency oscillator can be modified into a VCO by tuning the band-pass filter. For narrowband operation it is sufficient to tune one of the quarter-wavelength fingers of the interdigital filter. A varactor is therefore inserted in the central finger, since the latter has the highest loaded Q and provides the highest tuning sensitivity.

The varactor should be inserted either in parallel with a voltage maximum or in series in a current maximum along the length of a resonator. Since the capacitance of most varactors is quite high for microwave frequencies, varactors are

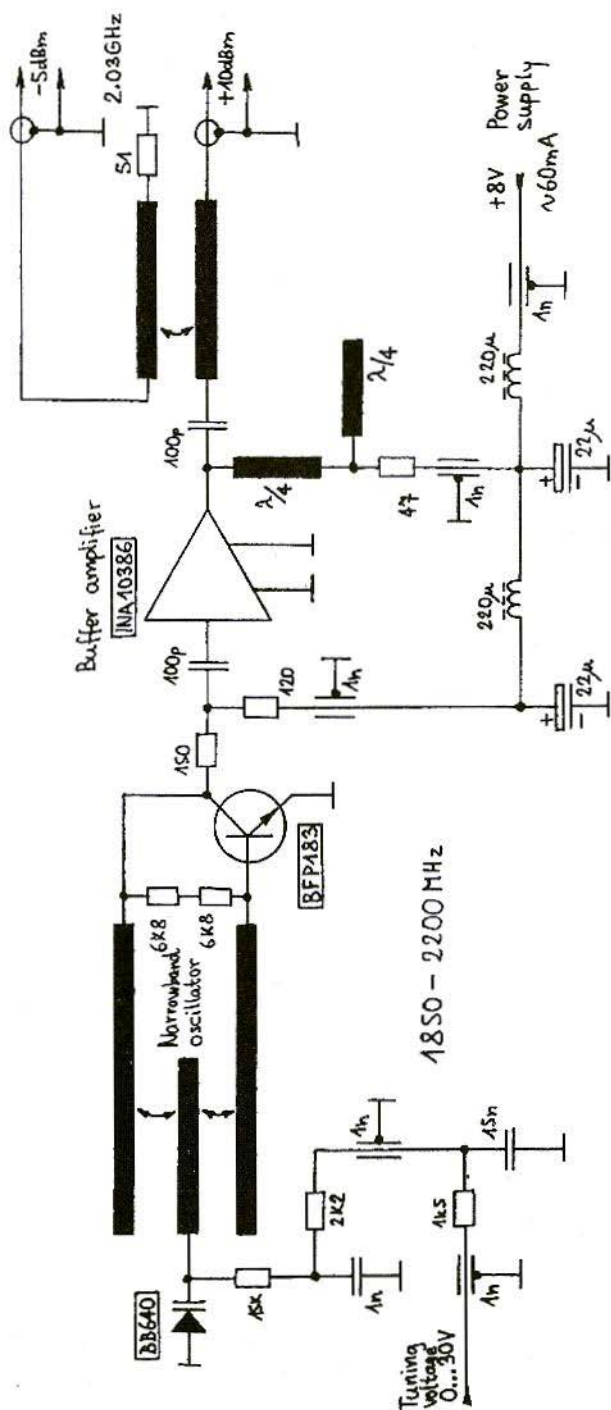


Fig.5: Narrowband Low-Noise VCO

usually installed in series in current maxima. In the case of a quarter-wavelength resonator, the "cold end" of the latter is grounded through a varactor diode.

The circuit diagram of a narrowband low-noise VCO is shown in Fig.5. The circuit diagram includes an output buffer (INA10386) to isolate the oscillator from load variations. Some supply voltage and tuning voltage filtering is included for the same purpose, as well as an output coupler for an auxiliary output.

The tuning range of an interdigital oscillator with one single varactor is limited to about 10-20% of the central frequency. The frequency range of the oscillator in Fig.5 is about 1850-2200 MHz. Outside this frequency range oscillation is not possible, since the gain maximum does not match the correct phase of the feedback.

A narrowband low-noise VCO has many applications in frequency synthesizers. The phase noise is sufficient for both analogue (SSB) communications [4], [5] as well as digital (coherent PSK) communications [8], [9], [10]. Used in a fast PLL it even allows the correct demodulation of complex radionavigation signals [6], [7].

The tuning range can be increased by increasing the coupling in the interdigital filter. This decreases the loaded Q of the resonators, degrading the phase-noise performance. Further, such an oscillator may also oscillate at higher-order resonances of the interdigital filter. This effect can be observed as "kinks" in the voltage/frequency curve, which is no longer monotonic.

4. WIDEBAND LOW-NOISE MICROSTRIP VCO

To obtain a true wideband microstrip VCO, all fingers of the interdigital filter have to be tuned. For example, inserting one BB833 varactor in each of the three fingers of the interdigital band-pass allows a tuning range up to 50% of the central frequency. For example, a VCO with three BB833 and a BFP183 ($f_t=8\text{GHz}$) as the active device operates reliably in the frequency range 2.0-3.2 GHz. Using a better transistor like the BFP420 ($f_t=25\text{GHz}$) allows the frequency range to be shifted up to 2.6-3.8GHz.

A VCO with a contiguous frequency coverage of 1200MHz may look as the upper limit for BB833 varactors (minimum capacitance 0.75pF, series resistance 1.8 Ω). Once again, even better results can only be obtained by changing our way of thinking. Varactors are usually considered as discrete lumped components while our oscillator is built with microwave transmission lines with distributed parameters.

In order to shift the frequency of a microstrip filter, one should preferably change the phase velocity on the microstrip transmission lines. The phase velocity of a microstrip line can be made variable if the line is periodically loaded with discrete variable reactances (varactors). Therefore, several varactors have to be distributed along the transmission lines to obtain the widest possible frequency coverage of a microstrip VCO. Since silicon varactors are inex-

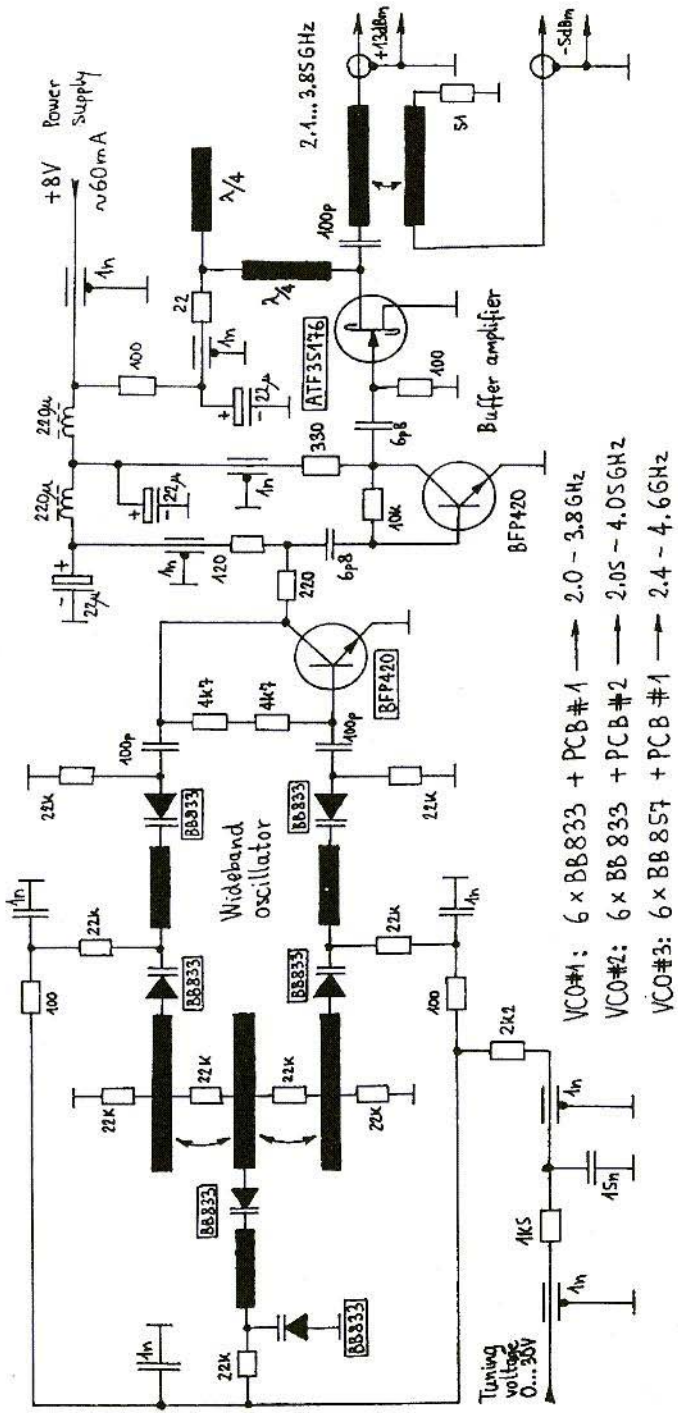


Fig.6: Wideband Low-Noise VCO

TUNING VOLTAGE [V]	f VCO #1 BB833+PCB#1 [MHz]	f VCO #2 BB833+PCB#2 [MHz]	f VCO #3 BB857+PCB#1 [MHz]
0	2013	2041	2388
1	2103	2140	2507
2	2192	2241	2618
3	2283	2356	2732
4	2411	2506	2848
5	2539	2692	2982
6	2660	2895	3124
7	2798	3047	3256
8	2950	3160	3375
9	3076	3250	3485
10	3190	3336	3588
11	3295	3412	3683
12	3349	3478	3772
13	3415	3538	3846
14	3474	3594	3929
15	3526	3645	4011
16	3575	3694	4085
17	3619	3737	4155
18	3657	3778	4219
19	3691	3816	4274
20	3720	3853	4325
21	3742	3887	4367
22	3760	3919	4407
23	3775	3948	4443
24	3786	3973	4476
25	3795	3994	4505
26	3805	4011	4530
27	3812	4024	4551
28	3818	4035	4568
29	3824	4043	4582
30	3829	4050	4593

Fig.7: VCO Tuning Characteristics

pensive, 6 or even 9 diodes can be used in a single VCO circuit.

The circuit diagram of a wideband, low-noise VCO is shown in Fig.6. Each microstrip resonator is tuned by two BB833 varactors, installed in different positions along the same resonator. This circuit configuration allows a very wide tuning range of about 2 GHz around a centre frequency of about 3 GHz. In other words, the tuning range of the described varactor VCO matches the tuning range of YIG VCOs.

On the other hand, a large number of varactors also increases the feedback losses. Further losses are introduced by the complicated varactor bias network (nine 22 k Ω resistors). All these losses must be recovered with the gain of the active device. A high ft transistor like the BFP420 is required for such an oscillator. Maybe the described oscillator could also be modified for other active devices, like GaAs FETs, HEMTs, HBTs or MMICs.

Just like its narrowband counterpart the circuit in Fig.6 includes an output buffer (another BFP420 and an ATF35176 HEMT) to isolate the oscillator from load variations. The buffer frequency response is selected

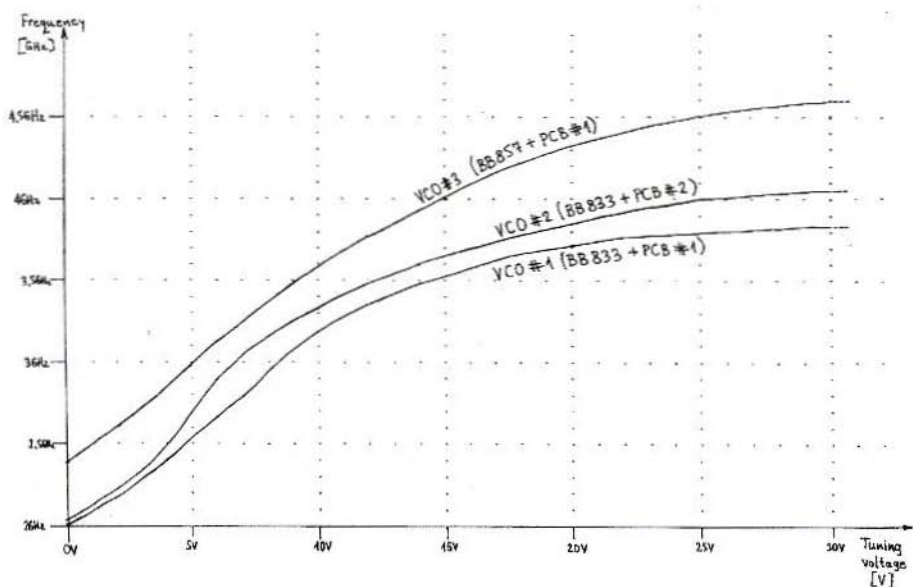


Fig.8: VCO Tuning Characteristics Diagram

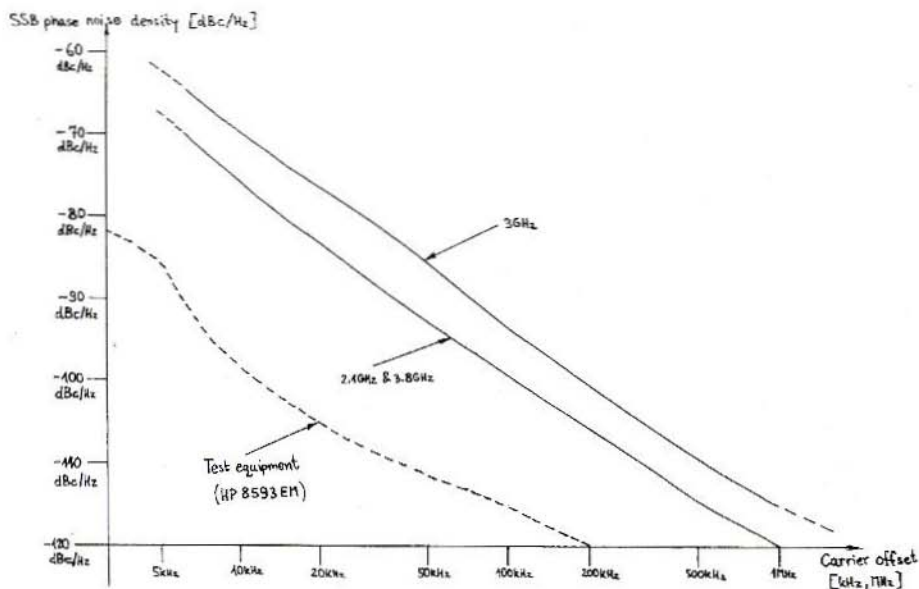


Fig.9: VCO Phase Noise

to partially compensate the output level variation across the octave frequency band. Some supply voltage and tuning voltage filtering is also included. An output coupler provides an auxiliary output, for example to feed a PLL prescaler, a frequency counter or a tracking generator for the spectrum analyser.

The measured tuning characteristics of three wideband VCOs are shown in Fig.7 and Fig.8. The main difference between VCO#1 and VCO#2 is in the printed-circuit board. VCO#2 uses lower impedance lines and the coupling between resonators is weaker. There are also differences in the BB833 varactors used: VCO#1 has rather old BB833s providing a coverage of only 1.8 GHz while VCO#2 has new BB833s providing a coverage of 2 GHz. Finally, VCO#3 is built on the same microstrip board as VCO#1, but uses the new BB857 varactors resulting in a frequency coverage of 2.2 GHz.

The tuning curves of all three VCOs are monotonic without kinks or jumps. The curves are quite non-linear as shown in Fig.8. All three curves are the steepest at band centre around 3 GHz (or at tuning voltages in the 5-10V range). Both below and above the tuning slope decreases, falling at the upper end (30V) down to less than 1/10 of the maximum slope. The frequency coverage of all three VCOs can be extended on the lower end by about 50MHz by applying a small negative voltage (about -0.7V) to the varactors.

The phase noise was accurately measured for VCO#1 and VCO#2 and the averaged results are shown in Fig.9. The

phase noise is about 5dB worse at band centre than at band edges. The phase noise peak coincides with the maximum tuning slope, indicating that at least part of this noise is caused by the varactors used and in particular by the 22 k Ω bias resistors.

In order to decrease the phase noise, the 22 k Ω bias resistors should be replaced by suitable RF chokes. Unfortunately suitable chokes are not easy to find. The resistor values can not be decreased much without introducing additional insertion loss in the feedback network. Finally, a more sophisticated bias regulator for the BFP420 oscillator transistor should also bring some improvement in the phase noise.

5. PRACTICAL MICROSTRIP VCO CONSTRUCTION

Both the narrowband and the wideband VCOs are built as microstrip circuits on conventional, double-sided, 0.8mm thick glassfibre-epoxy FR4 laminate. The upper sides are shown in Fig.10 while the bottom sides are not etched so as to act as groundplanes. Although the FR4 is quite lossy at microwave frequencies, the losses in the BB833 or BB857 varactors are even higher, so using inexpensive and easy-to-handle FR4 laminate is not a drawback.

The dielectric constant and losses of 0.8mm thick FR4 laminate were found quite similar even for materials obtained from different suppliers. One should

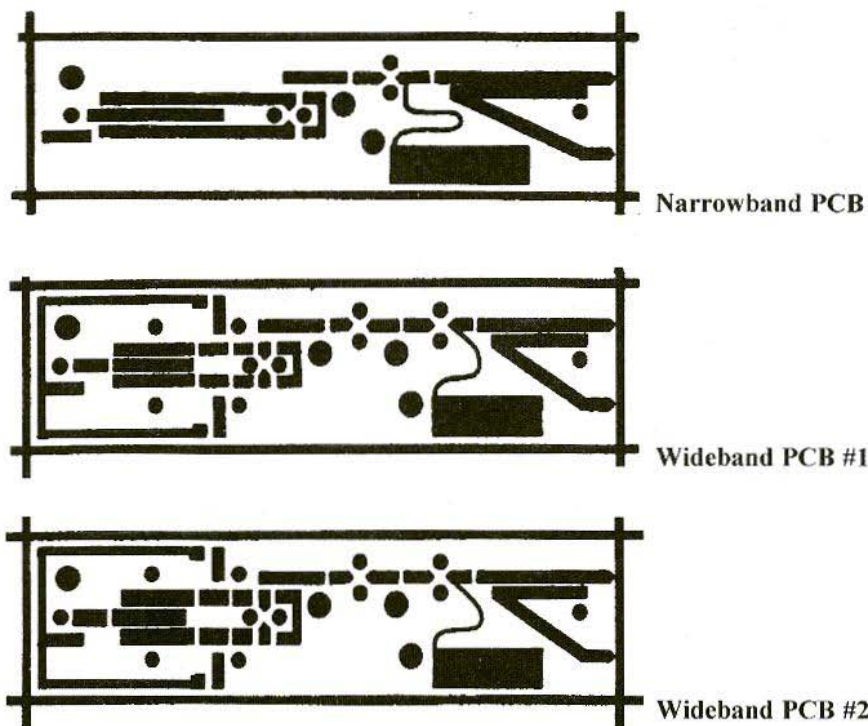


Fig.10: PCB Layouts; Scale = 1:1

only be careful about the thickness, since the same microstrip VCO circuit may not oscillate at all on 0.7mm thick laminate (too weak coupling between microstrip lines) or cover a restricted frequency band on 0.9mm thick laminate (coupling too strong). An additional problem is the large temperature coefficient of the FR4 laminate, shifting the frequency of the VCO downwards with increasing temperature.

All three printed-circuit boards have the same dimensions: 20mm (width) X 80mm (length). The whole surface of the groundplane may be tinned, while one should try to avoid tin-plating the microstrip lines on the upper surface except for the areas where SMD compo-

nents are installed. The boards do not have plated-through holes. All ground connections are made through 2.5mm diameter holes. The latter are first covered on the ground-plane side with thin (0.1mm) tinned copper foil and then filled with solder. The advantages of this grounding method are a low inductance to ground and

easy removal of installed components, without damaging the board nor the component to be removed.

Supply and tuning voltages go through several feedthrough capacitors. Some feedthrough capacitors (3 or 4) are installed in the printed-circuit board itself in 3.2mm diameter holes drilled at the marked positions.

Wire-leaded 1/8W resistors are used to connect the feedthrough capacitors to the microstrip circuit. Electrolytic capacitors, 220 μ H chokes and other filtering components are installed on the groundplane side and are supported directly by the feedthrough capacitors.

Finally, two feedthrough capacitors are also installed in the 0.5mm thick brass walls of the box housing the VCO. The box should be 30mm high, extending 20mm above the board surface and some 9mm below the board bottom.

If the printed-circuit board is well soldered on all four sides to the brass frame, then no additional shielding is required on the bottom side with supply filtering components. On the upper side a shielding cover is required and some microwave absorber is recommended.

In the frequency range of interest only Teflon-insulated coax cables should be used. Both flexible and semirigid Teflon cables can be used. The cable end should be first prepared by tinning both the inner conductor and the shield. Then the inner conductor should reach the microstrip board through a 3.2mm diameter hole in the brass walls, while the shield is soldered all around the perimeter of the hole.

The microstrip boards should be checked before installation in the shielding enclosure. Both narrowband and wideband VCOs should be checked for frequency coverage and output power level. In particular the PCB#2 for the wideband VCO may require some trimming of the central resonator length if the power drops or the oscillator stops at low tuning voltages.

The output power of the buffer amplifier should not drop below +10dBm in any of the oscillators shown.

6. APPLICATIONS OF MICROWAVE VARACTOR-TUNED VCOS

Microwave VCOs have many applications. Narrowband VCOs are suitable for many frequency-synthesiser projects. Wideband VCOs are mainly intended for instrumentation, like CW and sweep generators, spectrum analysers and corresponding tracking generators. In the following paragraph a simple spectrum analyser using both described VCOs will be presented, although for space reasons the description will be limited to the block diagram as shown in Fig.11.

The design of a spectrum analyser is based mainly on the available wideband VCO. Considering the frequency coverage of the described VCOs and the amateur microwave allocations (1.3 GHz and 2.3 GHz) and satellite frequency bands (GPS at 1.575 GHz, weather satellites at 1.7GHz) it makes sense to select the first IF at 2.1 GHz. This is far enough from 2.3 GHz and allows a simple input lowpass filter for the band from 0 to 1750 MHz. On the other hand, the last IF is set down to 10 MHz (10.7 MHz) due to the design restrictions of LC and crystal filters and the logarithmic detector. To simplify image and spurious filtering, the first IF at 2.1 GHz is first converted to a second

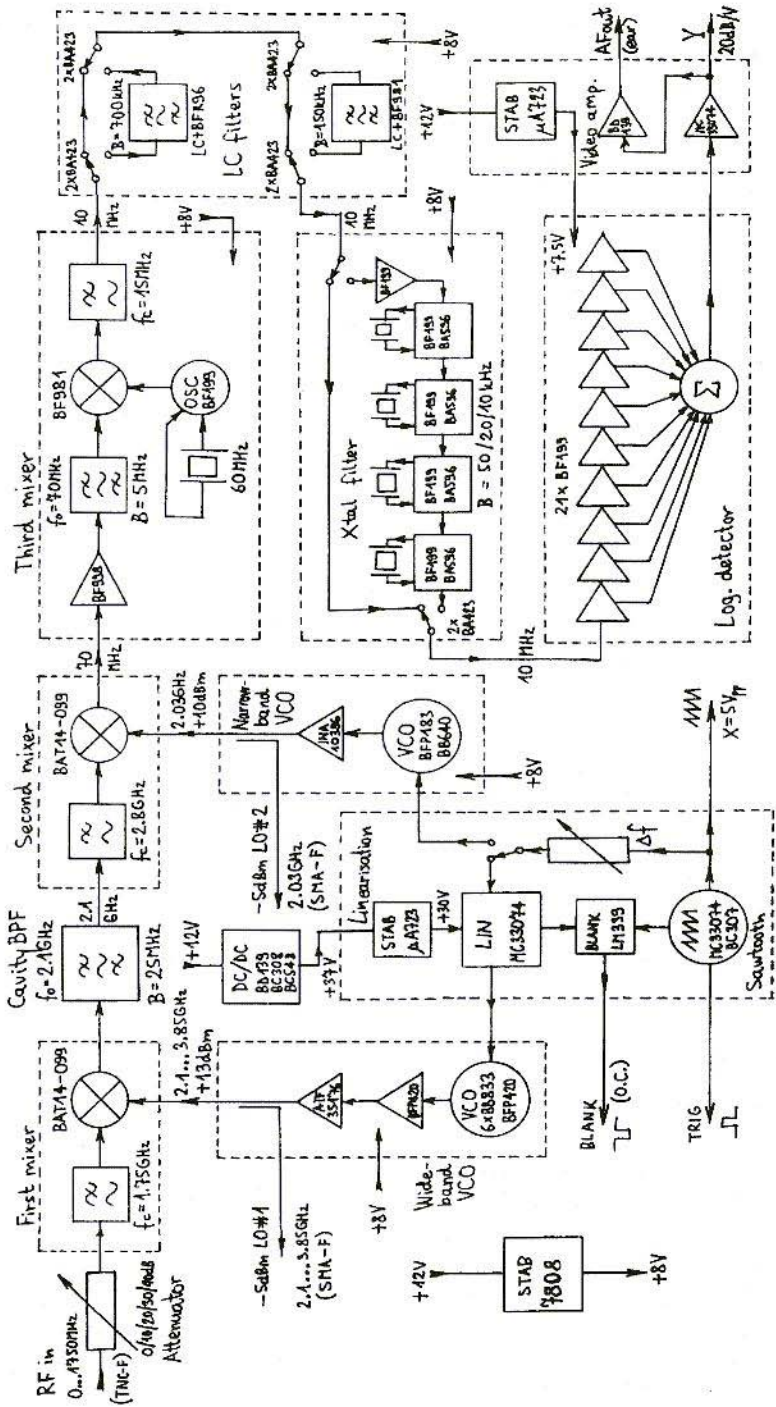


Fig.11: Spectrum Analyser Block Diagram

IF of 70 MHz and the latter is afterwards converted to the final third IF of 10 MHz.

The IF filters offer six different bandwidths: 4 MHz, 700 kHz, 150 kHz, 50 kHz, 20 kHz and 10 kHz. The 4 MHz bandwidth is required for full-band sweeps, considering the limited resolution of the CRT display.

On the other side, the resolution of the spectrum analyser is limited to 10 kHz bandwidth and 50 kHz/div display. Narrower IF filters would require additional stabilisation circuits for all oscillators.

Since narrow IF filters also require very slow scanning, they are practically seldom used and were omitted in this project.

The logarithmic detector is built with discrete components, since available integrated circuits will not handle a dynamic range of over 90dB (achieved already with the 150 kHz IF filter) and many different IF bandwidths at the same time.

The video amplifier provides a 20dB/V output to drive the vertical deflection of the CRT display as well as an AF (earphone) output. Of course the latter is only useful at zero span.

On the other hand, the horizontal deflection signal (5Vpp) is provided by the built-in sawtooth oscillator. The latter may sweep the frequency of the first local oscillator (wideband VCO) or the frequency of the second local oscillator (narrowband VCO for 500 kHz/div or less).

Due to the nonlinear voltage/frequency response, the wideband VCO requires a rather complex linearisation circuit.

The spectrum analyser is designed to use a standard XY oscilloscope display. If an external X deflection is not available, the spectrum-analyser electronics also provides trigger pulses for the time base of the oscilloscope display.

Z-axis or display blanking has two functions in a spectrum analyser: retrace blanking and out-of-band blanking. If a Z-axis (intensity) input is not available on the CRT display, then the blanking output (open collector) may be wired in parallel with the Y output to deflect the trace outside of the visible screen when the blanking is active.

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The ASH Receiver - Principles, Trials and Observations

The letters "A-S-H" stand for a sequence of words which are not particularly easy to pronounce, even in English - "amplifier-sequenced hybrid". If the name of the inventor had not been Darrell L. Ash, "hybrid sequenced amplifier" would have been a better choice. That, at least, represents something that can be imagined. Amplifiers linked in sequence, which are based on two technologies, are nothing new in principle - one example would be a relay behind a tube.

So if US patent 5357206 was awarded for the ASH principle in October, 1994, there must be more to this than meets the eye.

1. THE PRINCIPLE

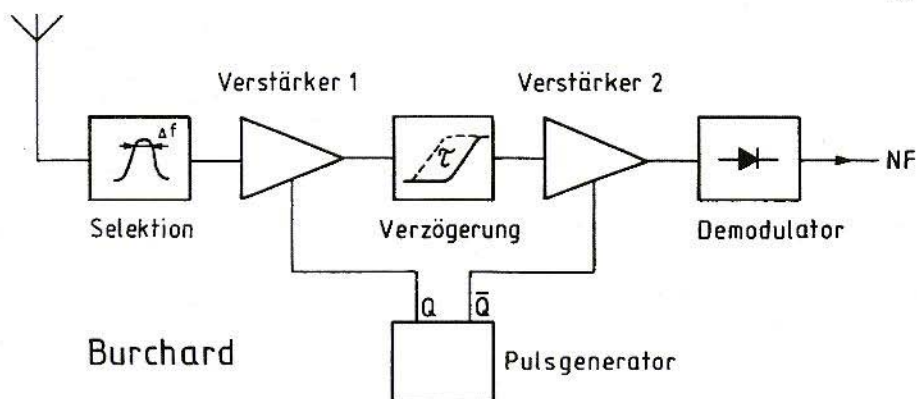
Fig.1, adapted from Ash [1], shows the most important components of the ASH

receiver: a pulse generator, which switches in two HF amplifiers alternately, an input selection, a delay configuration and a rectifier.

The pulse generator switches at a frequency which is matched to the delay. A sample signal, cut out by the first amplifier, is intended to find the second amplifier open just as the delay expires.

The hybrid technology thus consists of the sequential layout of amplifiers and delays wired up. The amplifier wired up first has more the function of a sampling gate. It can be followed by one ASH hybrid stage, or indeed by several such stages.

The idea behind amplifiers wired up in this way is that there are no stability problems, even with a very high total level of amplification. There can be no feedback between the output and the input, because the amplifiers are not all amplifying at the same time. Stability is needed only for each individual amplifier. The stages can be positioned close together, even on the same chip, without reciprocal screening.



Burchard

Fig.1: Basic Wiring Diagram of ASH Receiver

Selektion = Selection, *Verstärker* = Amplifier, *Verzögerung* = Delay, *Pulsengenerator* = Pulse generator, *NF* = Low-frequency

The temporal offset also obliterates crosstalk between the delay input and output. Since any crosstalk occurs with a delay, it finds the next amplifier inhibited.

With the superhet principle, feedback between the output and the input is prevented by frequency conversion, which means mixer stages are required. These create secondary reception points. No frequency conversion takes place with the ASH receiver. It thus forms a tuned fixed-frequency receiver with sampling characteristics, and free from secondary reception. It can process CW, AM and ASK signals, the highest modulation frequency of which is only a fraction of the pulse generator frequency.

As a side effect, the current demand is reduced, since we are certainly always going to arrange things so that an amplifier which is turned off does not receive any current. The reduction is particularly impressive if several ASH stages are used.

As always with tuned radio-frequency receivers, there is a problem, in that the necessary selection has to be made with regard to external signals and the band width matched to the modulation.

Depending on the quality of the means of selection used, the quotient of the mean frequency and the band width lies between 50 and 50,000. The first number can be obtained using filters with concentrated L/C components, and the second using crystal filters. Ceramic, mechanical and acoustic filters lie in between. The rule for all filters is that selection can not be generated without transit time, and delay can not be generated without selection.

So, in the ASH receiver all the filters contribute to the total selection. The input selection must be broad enough for side bands with the highest modulation frequency. However, the delays should have a wider transmission band, so that the pulse is not deformed more and more as it is relayed.

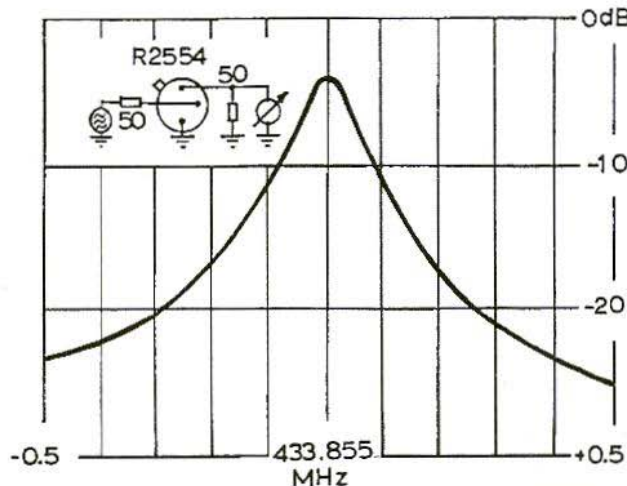


Fig.2:
SAW Resonator as
Quadripole, measured
in a 50Ω System (as
per Burchard [2])

With surface acoustic wave (SAW) filters, the principle allows for formats with an especially long transit time, which can not be attained with other filters with the same band width and the same number of stages. As a manufacturer of surface acoustic wave components, Ash [1] probably had such filters in mind.

Both the selection filter and the delay filter can be created using SAW technology, and can be fitted, together with the amplifiers, the pulse generator and the rectifier, into a single SMD housing with dimensions of 14 x 10 x 2.5 mm.: a real little ash can.

It can be used in the manufacture of remote-control receivers for garage doors and garden gates, car alarms and central locking systems, etc., which receive only about 1 mA from a 3-Volt battery. The frequency used could be, for example, 433.92 MHz. The receiver described in [1] is also designed for this frequency. Using a coded priority output system gives us several channels in a single frequency band.

2.

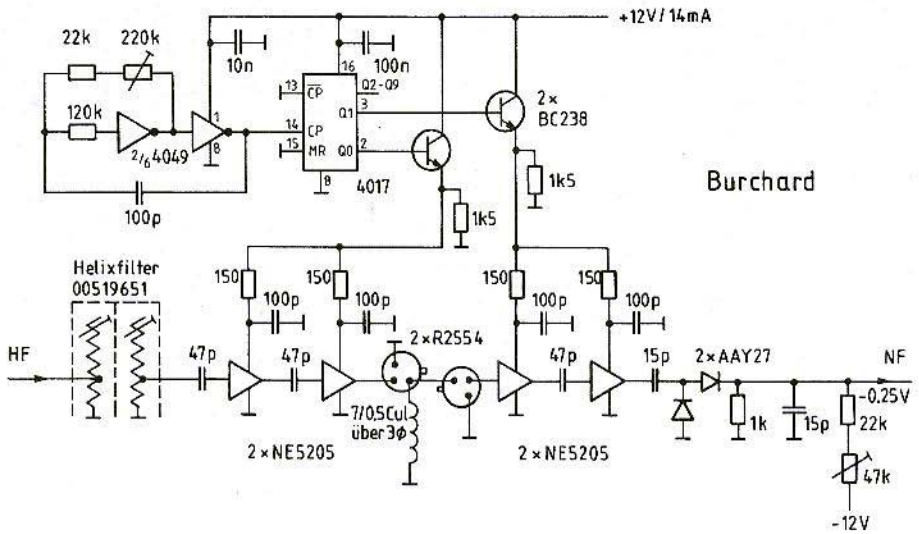
AN EXPERIMENTAL SET-UP

The desire to try out the new principle as quickly as possible inspired a search operation for appropriate components. I sought out spiral filters suitable for 433.92 MHz (Neosid 00519651), SAW resonators (Siemens R2554) and gain-blocks (Signetics NE 5205).

The spiral band filter has a band width of 2 MHz. Measurements have already been carried out using SAW resonators at an earlier date [2], and a band width of 90 kHz was found (Fig.2).

As is shown later in this article, it can be demonstrated that at least a two-stage filter is needed for the delay. So the spiral filter could be used, or one made from two SAW resonators.

In the first case, this leads to switching frequencies of a few MHz, and in the second they are about 10 kHz. The second case is better suited to a 1 kHz modulation of the signal generator, and



Burchard

Fig.3: Wiring Diagram of Experimental Set-Up
Helixfilter = Spiral filter, Über = Through, NF = Low-frequency

so the set-up shown in Fig.3 was selected.

Naturally, it would be desirable - as indeed with any other reception principle - to have the lowest band width immediately at the input. But, as in most cases, this would be impractical here too, for the SAW band filter would have a considerably higher degree of transmission band attenuation. This is another way in which the spiral filter at the input contributes to the distant selection.

The pulse generator was designed to be just 10-phase, so as to be suitable for any possible extension of the amplifier.

The gainblocks also function stably if two are operated in line. This was known from other applications. Thus, amplifiers 1 and 2 each consist of two NE 5205's, and each provide about 40dB of amplification. Since no strobe input is available, they are both simply

switched on by feeding in the operating voltage. Fig.4 shows that this procedure is permissible. This is a mixed-sweep plot, in which the horizontal spot velocity alters after 3.6 screen divisions from 50 to 2µs / division. The top line

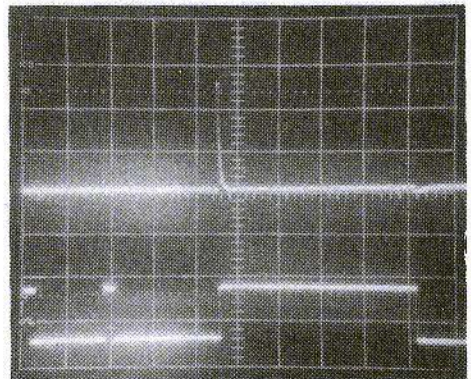


Fig.4: Transient Effect of HF Amp
Y1: NE5202 Output 2V
Y2: CMOS4017 Output 10V/div
X: Mixed Sweep 50µs / div to 3.6 div, then 2µs / div

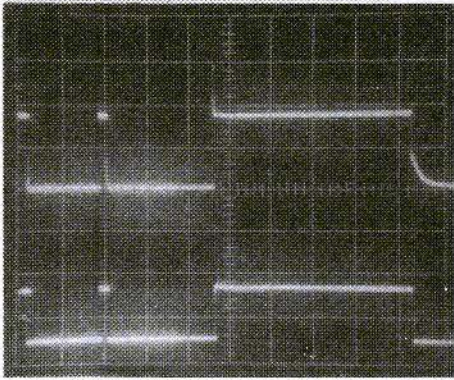


Fig.5: Voltage Surge behind Rectifier
Y1: Low-Frequency 200mV/div
Y2: CMOS4017 Output 10V/div
X: Mixed Sweep 50 μ s/div to 3.6 div, then 2 μ s/div

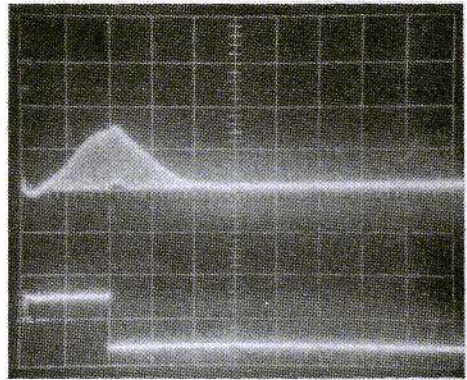


Fig.6: Signal behind HW Filter at -48dBm at HF Input
Y1: Rectified Signal 5mV/div
Y2: CMOS4017 Output 10V/div
X: 5 μ s/div

represents the gainblock's output voltage. It can be seen that the operating point has been set in less than 100ns. When the equipment is switched off, the voltage rapidly falls to less than half. Then another micro-second passes before the output voltage is at zero.

The bottom line is the logic voltage from the 4017 counter component. The pulse width here is 9 μ s, and the repetition rate 90 μ s. For other illustrations appearing later, the pulse width was set to a level 10 μ s.

The rectifier behind the second amplifier receives a surge through the coupling capacitor each time the equipment is switched on or off, as can be seen in Fig.5.

To ensure that this interference decays rapidly, the capacitor has been selected to be small, and the rectifier receives a Townsend current. This gives it a lower impedance, and its discharge time constant is reduced.

The coupling of the two SAW resonators was being tried out to obtain a sufficient delay with a low level of overall attenuation. Here, this optimisation results in a displacement of the mean frequency to 433.88 MHz, around 25 kHz above the resonance of the individual resonators (Fig.2), with a 50 Ω adapted load on each side.

The transit time in the SAW filter can be seen in Fig.6, which was plotted with the help of the rectifier, which here was connected to the output of the second SAW resonator. The signal generator is 100% modulated at 1 kHz, so that an illuminated area is obtained, which comprises all signal amplitudes between 0 and 200% of the mean carrier amplitude.

As expected, when the first amplifier is switched on, no signal is present at first. The signal builds up with the visible rise time, and reaches a final value after 10 μ s. The first amplifier is then switched off and the second switched

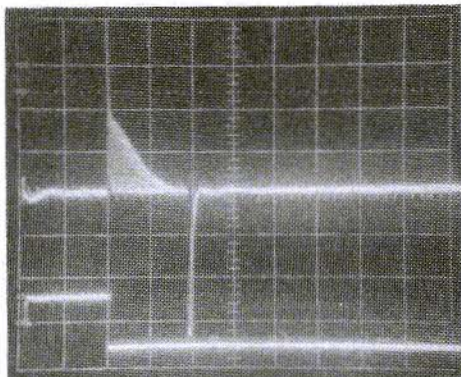


Fig.7: Signal at Low-Frequency
Output at -90dBm at HF Input
Y1: Rectifier Output 5mV/div
Y2: CMOS4017 output 10V/div
X: 5 μ s/div

on. The high frequency now appears at the second amplifier's output, and the low-frequency output looks like Fig.7. The second amplifier clearly does not add a supplementary delay. On the contrary, a rather longer transit time in the SAW filter would be desirable, so that the signal was better positioned in the time window. Reducing the switching-on time of the first amplifier is not a solution, for then the final voltage value would not be attained.

The low-frequency signal as per Fig.7 can now be processed further - e.g. with filters or sample and hold devices. However, the oscilloscope is perfectly capable of demonstrating the operability of the ASH principle. The lowest low-frequency signal which can still be evaluated by means of the oscilloscope is 1 mV. This is associated with an input voltage of approximately 2 μ V (-100dBm) and a total amplification between the input of the spiral filter and the output of the second amplifier of 68dB. It is made up of -3dB in the

spiral filter, 38dB in the first amplifier, -11dB in the SAW filter and - since the system is not operating under load - 44dB in the second amplifier. The experimental set-up selection was in accordance with Fig.8.

The current consumption is only 1mA. If the four gainblocks were permanently switched on, the reading would be over 100mA.

3. IMPROVEMENT OPTIONS

Requirements of a new kind are imposed on the filters here. They must ensure adequate selection, carry out pulse formation, and have an appropriate transit time. All filters contribute to selection.

The pulse formation takes place in the filter behind the first amplifier connected, and all the others should, as far as possible, have a transit time which only alters the envelope curve, but nothing else.

Selection filters are all-pass free chains, made up of concentrated L and C components or their mechanical equivalents. They are usually named after the mathematician who devised the calculation formulae (Cauer, Gauss, etc.). The results are given in the form of tables - e.g. filter catalogues as per Glowatzki [5, 6].

Pulse-forming filters are a mixture, with Chebyshev characteristics in the filter attenuation band and Bessel characteris-

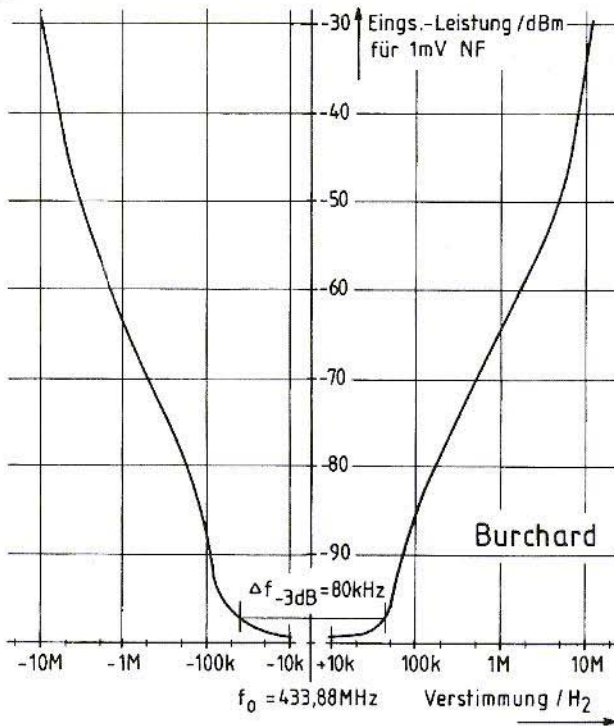


Fig.8:
Experimental Set-Up
Selection
Eings.-Leistung =
Input power,
Für = For,
Verstimmung =
De-tuning

tics in the transmission band, or else they also have all-passes for improving the pulse formation. Such filters can be calculated, for example, as per Feistel and Unbehauen [4], Jess and Schüßler [7] or Jess [8]. Active filters with pulsed properties are also possible, as Fliege [3] has demonstrated.

All filters have a transit time and a final rise time for the envelope. Without going into the reasons or the details, we should just establish here that the approximate transit time (measured at half the height of the pulse) is equal to the reciprocal value of the band width, multiplied by the number of filter stages, whilst the approximate rise time (measured between 10% and 90% of the rise) is simply the reciprocal value of the band width.

For this reason, at least two filter stages are needed for pulse formation.

All-passes are created using concentrated components as Bessel filters. If they have to demonstrate a uniform transit time over the useful band width, their effective band width must be much greater than the area occupied by the signal with its side bands. Thus they must also have more stages than selection filters with the same transit time. This avoids any further distortion of the envelope curve.

Thus those who wish to dimension their filters themselves should take this advice into consideration. For the basic facts on filters - including SAW filters as well - see Meinke and Gundlach [9].

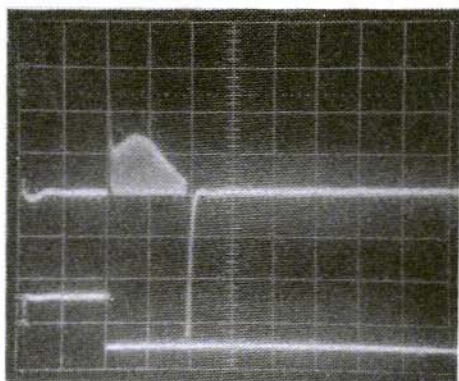


Fig.9: Signal at Low-Frequency Output with De-tuning occurring at +80 kHz and -80Bm at HF Input, otherwise in accordance with Fig.7

Filters with flat tops certainly do not have a flat transit time gradient. This can be seen from Fig.9, which was produced in the same way as Fig.7, except that the signal generator was de-tuned by + 80 kHz. The transit time of the SAW filter has considerably increased, and is even better positioned in the time window. However, this is not an operating condition to be striven for, as the transmission attenuation has increased by 10dB.

SAW filters can not be home-made. Reasonable prices can be obtained only through mass production. Pertinent transit times are possible with them, irrespective of the band width. In addition to Meinke and Gundlach, it is also worth reading Smith [10], Weller and Lewis [11] and the brochures from the manufacturers - e.g. Siemens [12].

The R2554 resonators consist of a piezo-electric crystal, on the surface of which an acoustic wave runs to and fro between reflectors. The electrical input

and output are connected to this through inter-digital converters. The electrical behaviour is very similar to that of a quartz, and an equivalent circuit diagram corresponding to the quartz can also be indicated. The distant selection of such a resonator scarcely exceeds 25dB (Fig.2), and filters assembled using it have markedly higher levels of transmission attenuation than quartz filters.

SAW transit time filters are also based on the propagation of an acoustic wave over the surface of a piezo-electric crystal. However, the latter is not carrying reflectors at its ends but acoustic absorbers. The coupling and de-coupling, once again, are through inter-digital converters. If no further measures are taken, they have a $(\sin x/x)^2$ frequency response - i.e. a round top, a flat transit time gradient, and elimination of the auxiliary maximum as low as 26.5dB. The transit time is proportional to the path between the converters. The power loss in the absorbers and along the transit time path creates a rather high level of transmission attenuation - 20dB is a typical value. SAW filters operating on the same principle, but with shorter lengths, are used for pulse formation.

Ash [1] uses his "ashcan" at the input of an SAW resonator, and uses an SAW transit time line for pulse formation and delay. Thus, without an additional LC filter, he obtains a distant selection of 60dB at ± 6 MHz and 80dB at ± 14 MHz. The ASH receiver described has a band width of 300 kHz with a mean frequency of 434 MHz.

Ash [1] gives no information on the type of amplifier used. If you wish to avoid the inrush current, differential amplifiers

should be fitted. A symmetrical source and load - which an inter-digital converter more or less represents - would make completely crackle-free switching possible.

In emergency, HF transformers could be used for circuit balancing. Integrated differential amplifiers can be found in first and second generation IC's.

4. CONCLUSION OR SUMMARY

Following the presentation of the principle, an experimental set-up was discussed and dimensioned. Reductions in the remaining defects were then discussed. With these specifications, it should be possible to design ASH receivers for other frequencies and tasks.

The principle is naturally not restricted to miniature formats, two-part amplifiers and remote control applications. It provides for high straight amplification without high-quality screens, and this applies, not only to partial amplifiers, but also to transit time filters.

Since all amplifiers are switched on for only part of the time, the current requirement is greatly reduced, as against other concepts.

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Henk Medenblik, B.Sc., PE1JOK

A State-of-the-Art 13cm Amateur Television Transmitter Part-3 (Conclusion)

In this final article of the 13cm ATV transmitter project, a microcontroller unit and 1.5 Watt power amplifier will be described. The microcontroller board takes care of the many features which are possible including a text inserter.

The 1.5 Watt power amplifier described here is different from regular design because of its simplicity and used low cost PCB material.

1. THE MICRO CONTROLLER UNIT

The main part of the micro controller unit is formed by a modern derivate of the 8051 micro family; the AT89C52 from Atmel. This micro is fully compatible with the well known 8051 and carries 8kbytes of flash memory. This has the advantage that it is possible to

reprogram the micro many times, which is very useful during the software development stage. Another advantage is the fact that no external EPROM is needed for the program itself.

The micro has also 256 bytes RAM and needs in fact only an external crystal and 5V power supply. The design of the micro board is therefore rather straight forward. Fig.1 shows a prototype of the micro board, the schematic is given in Fig.2, Fig's.3a and b show the double sided PCB and Fig's.4a and b show the component overlay.

The design consists of a 89C52 and a SAA5254 Teletext decoder chip. The latter is being used as a text inserter instead of a TT decoder. The micro controller just sends the necessary character information downwards to the Teletext chip whereupon this character information is inserted in the incoming video signal. There are four different text lines possible which are being held inside an EEPROM.

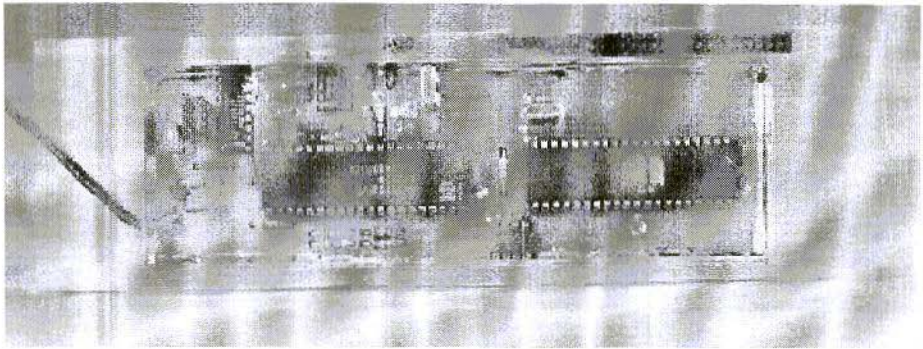


Fig.1: Prototype Micro Board

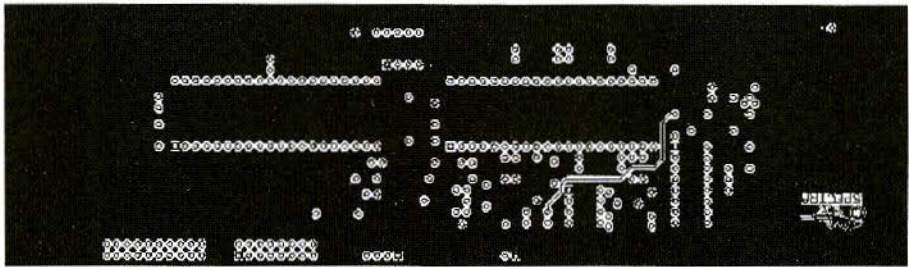


Fig.3a: Micro Board PCB Top Layout; actual size 215mm x 61mm

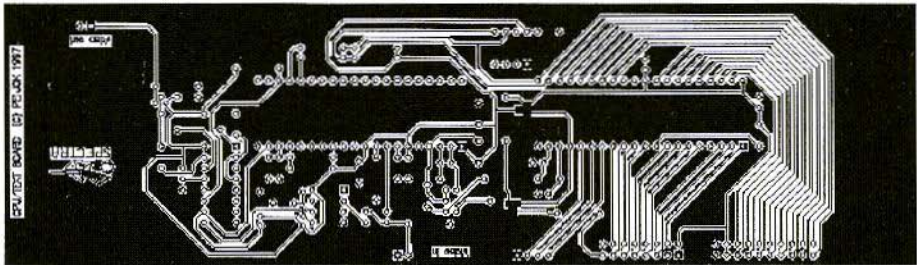


Fig.3b: Micro Board PCB Bottom Layout; actual size 215mm x 61mm

The incoming video signal is forwarded to the SAA5254 and a LM1881 sync regenerator. The generated burst clamp signal coming from the LM1881 is used to maintain the correct black level after the insertion of character information from the SAA5254. The SAA5252 generates the character information on pin-23 and also a special blanking signal

at pin-19 which makes it possible to display a black background behind the white characters. Multiplexing of the signals occurs with a 4053 analogue multiplexer IC. The white level of the characters can be adjusted with potentiometer R14. Finally the composite video is buffered by Q1.

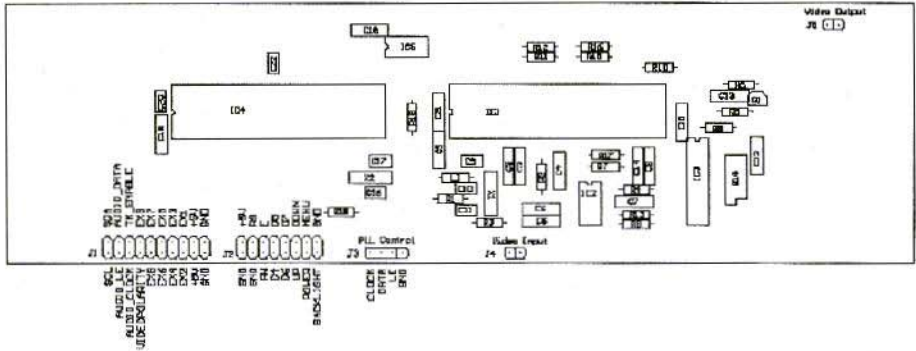


Fig.4a: Micro Board PCB Top Layout; actual size 215mm x 61mm

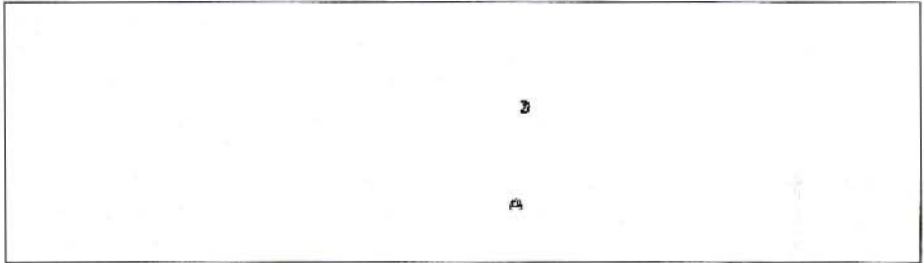


Fig.4b: Micro Board PCB Bottom Layout; actual size 215mm x 61mm

The micro controller controls the several power supply lines. This is accomplished with two switching FET's Q2 and Q3. Q2 switches the back-light of the LCD and Q3 controls the power line to the SAA5254 and the rest of the electronics. Finally a special pin called TX_Enable is available on header J1 to control an external power supply to the transmitter.

A 2x16 character LCD is used for interaction with the user. I used an Optrex LCD type DMC16202NYLYB, but other types could be useful too, because most LCDs use the same chipsets from Hitachi. This LCD is controlled with 4 bit mode to preserve IO pins on the micro. Within 4 bit mode only lines D4-D7 of the LCD are used as

databus. The Anode of the back-light is connected to pin-15 of J2 and the cathode is connected to pin-16 of header J2. Finally 4 simple push-buttons are used for the controlling functions.

A low signal on the corresponding input at the micro activates a function. Therefore the push buttons are connected with one side to ground and the other side to the appropriate input of the micro. They are named as POWER, MENU, UP and DOWN.

A special header is used for the programming interface with the VCO/PLL module. This is header J3. Header J1 is a special connector which carries the controlling lines to the power supply, audio PLL and video polarity selector on the video/audio board. For all inter-

connections there is a major rule; keep lines as short as possible. I used short flat cables. A I2C bus interface is also available on this header together with some other IO pins from the micro. They could be used for future interface extensions.

Finally the power supply is connected to pin-17 (+5V) and pin-19 (GND) on header J1. These lines are re-routed to pins-18 and 20.

1.1 Software

When power is applied for the first time, the micro will scan to see if the SAA5254 is available. If this one doesn't function than an error message is generated. Also, the EEPROM will be checked to see if it holds any data. At that first time 4 predefined text lines will be loaded into the EEPROM. This is only done the first time when power is applied. The text lines can be changed from a special menu function. When power is disconnected, text info is preserved in the EEPROM.

After this initialisation all controlling voltages will be disabled, whereupon the micro will enter idle mode. This means that all functions of the micro go down except the interrupt system. The micro just needs a very low current during idle mode, so +5 V could be applied during the rest of your hobby lifetime. LCD and transmitter are also down during this mode.

The micro can only be awakened by a short pressing of the POWER button. The power supply will be switched on and the unit will awaken except for the transmitter itself. The supply for the transmitter is controlled by the TX_ENABLE signal which will be kept low. The LCD displays the currently adjusted frequency and Transmitter off. The transmitter can be enabled by another short pressing of the POWER button. Repeatedly short pressing the POWER button will toggle the transmitter operation mode between on and off. A long pressing of the POWER button will enable power down mode again. So the POWER button has more functionality than only power on and power off.

During the main menu function (which is automatically entered when powering up) the frequency can also be adjusted with the UP and DOWN buttons.

The MENU button just makes entering other menu modes possible.



Fig.5: An Example of a Menu Display.



The first menu after the main menu is the video polarity control menu. Just pressing UP or DOWN toggles the video polarity. Another pressing of the MENU button enters the first audio carrier menu which makes it possible to adjust the frequency of the first audio carrier between 5.5 and 8 MHz with 10 kHz steps. Another pressing of the MENU button enters the second audio carrier frequency adjustment menu.

The next menu displays the currently selected text line to be inserted through the video. UP and DOWN keys make different selections of this line possible. The next menu controls the text inserter mode; text inserter enabled, text inserter disabled or page switching enabled. The latter repeatedly selects the next text line with an interval of a second.

The next menu enters text edit mode which makes it possible to adjust your favourite text. The current selected text is being displayed on the LCD. The POWER button has another function during this mode; it selects the currently edited character which is underlined. UP and DOWN scrolls the alphabet. The next menu is makes selection of a black background possible with UP and DOWN keys.

The last menu displays the current line position where your favourite text should be positioned. There are 12 vertical different positions possible. During menu functions where the POWER button does not have a special function a long pressing of this button will enter power down mode which will be visible with closing down....

Fig.5 shows an example of a menu display.

1.2 Construction

There are no special considerations to be taken during construction of the micro board. For EMC purposes it is recommended to shield this unit. On the PCB, which can be obtained from SPECTRA BV, a special solder-mask is available, which gives locations of the shielding at upper and lower side.

The height of the shielding at the upper side is 1.5cm and 0.5cm for the bottom side of the PCB. The sizes of the PCBs are specially designed for construction into a BOPLA housing which can be obtained from Conrad Electronics.

2.

THE POWER AMPLIFIER

A special PA has been developed for this project. One main goal during development was the implementation on ordinary FR-4 PCB material and a power output of approximately 1.5 Watt. It resulted in a low cost, straightforward, small design.

Because of component tolerances and the need to get the maximum out of the design, three variable capacitors are needed. These adjustments are accomplished by small SMD trimmers from the TZC03 series from Murata. The well-known SKY trimmers do not work in this design because of the parasitics (they are also a lot bigger than the Murata types).

After adjustment prototype versions showed that typically 1.5 Watt could be

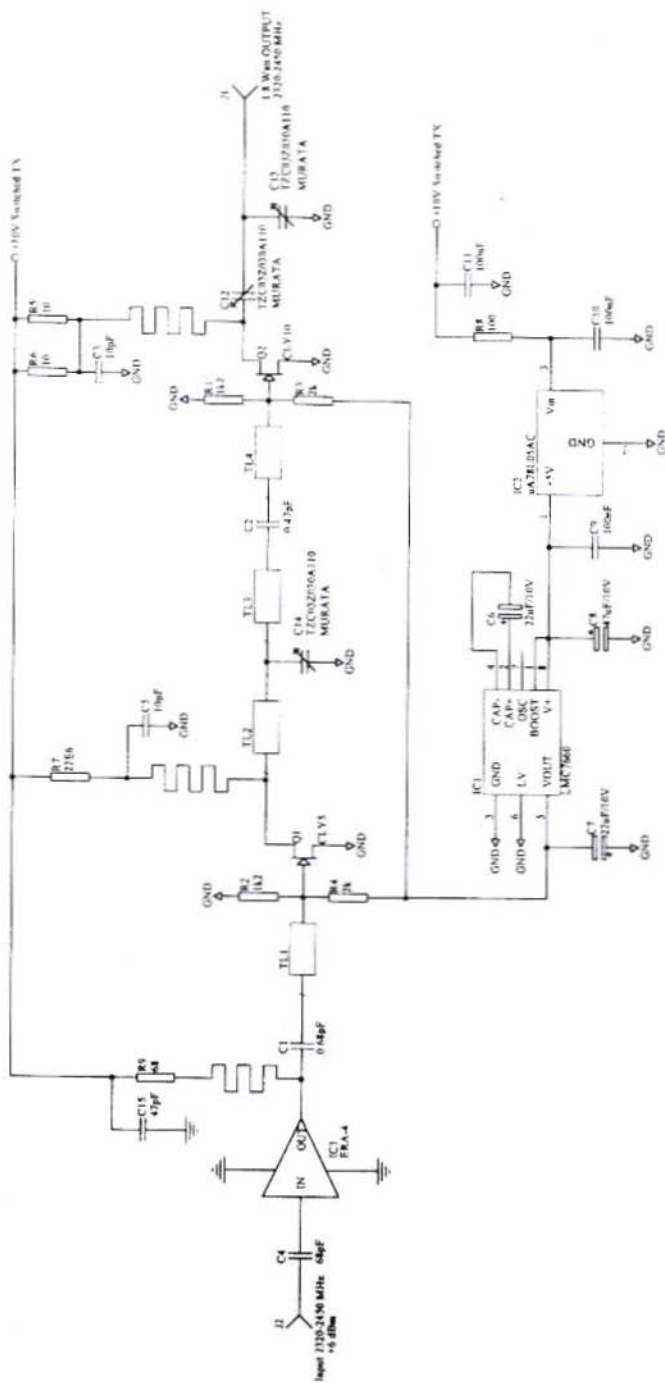


Fig.6: Circuit Diagram of the PA

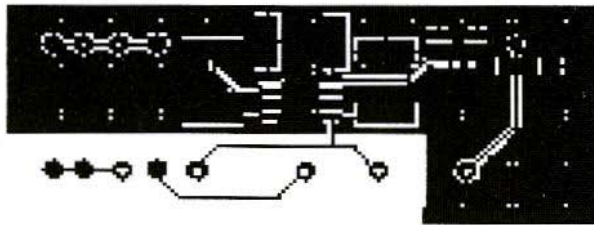
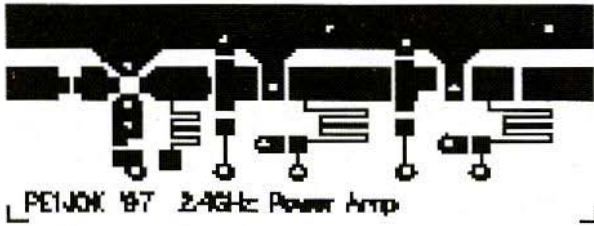


Fig.7a: PA PCB Layout Top Side;
each PCB actual size 54mm x 22mm

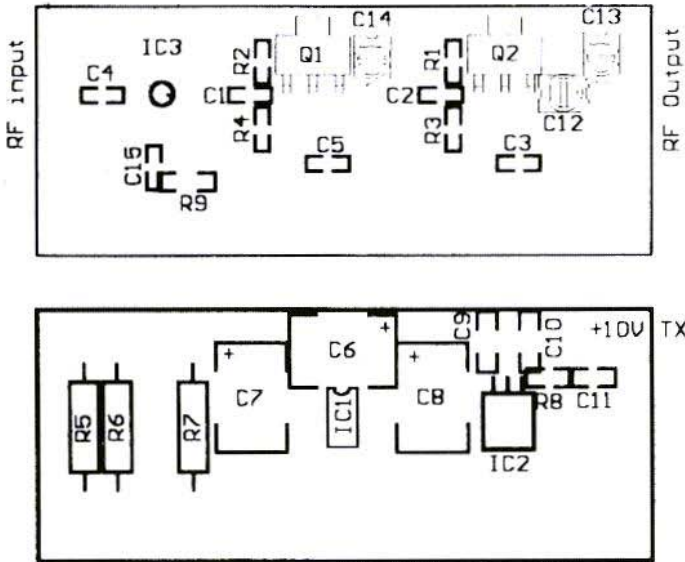


Fig.8: PA PCB's Component Overlay;
each PCB actual size 54mm x 22mm

and needs a drive level of approx. +6dBm.

2.1
Construction

In Fig.6 the schematic of the PA is given. Fig.7 shows both PCB's, with the component overlays in Fig.8.

The first PCB carries the microwave circuit. The rear-side if fully copper. The second PCB carries the biasing components and is also fully coppered on the rear-side. Both PCBs can be mounted back to back inside one shielded housing.

Fig.9 gives a view of an assembled prototype. The ground pins of the ERA amplifier have to be as short as possible; directly to ground on the rear-side. The CLY GaAsFETs from Siemens are

obtained at the output. In one case even 1.8 Watt was measured. The output maintained within 0.6 dB constant over the frequency range of 2320-2450 MHz

very electrostatic sensitive. Use only grounded soldering irons and workplace. Gate voltages exceeding 6 Volt will permanently destroy these devices. The



Fig.9: Prototype of the 1.5 Watt PA

big drain connection is soldered close to the shielding side to get rid of a lot of heat. At the outside of the shielding a heat sink is needed. I used a little heat-sink together with an air blower which is activated with the same controlling signal which enables the transmitter supply (TX_ENABLE).

3.

FINAL NOTES

With this article comes an end with the description of the 13cm ATV transmitter. Although the original unit also uses a special power supply board which is

being controlled by the micro board, I thought that it would not be interesting enough to give a full description of that. The only thing I can say about that is that I used low Ron MOSFETs for switching purposes, whose gates are controlled by the TX_ENABLE signal for the microboard. For the microboard itself I use a simple switcher from National, the LM2594-5, to keep power dissipation low. Supplies to the transmitter itself should be regulated by a linear regulator to avoid EMC problems.

If you have any questions then you can reach me through internet email at: .
 Kits and PCBs are available from Spectra BV, Marga Klompheoie 99, 2743 HW, Waddinxveen, Holland.
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D.Eckart Schmitzer, DJ4BG

Designing Chebyshev Filters for Practical Operation

Part 1

Various processes exist for LC filter calculations. But since the complete calculation process for Chebyshev filters is very detailed, people like to use so-called filter catalogues. The use of data from such a filter catalogue is explained here, with the help of examples.

1. GENERAL

The filter catalogue I have, the book from Pfitzenmeier [1], contains the data for deep-passes, in standardised form, for all possible graph curves, from the simplest right up to very extensive patterns. As a result, the book is over 500 pages thick, and is really impenetrable for someone not used to using such sources. So now we have to learn how to handle the data from this book.

Through a relatively simple adaptation of the computing method, high-passes, band passes and band elimination filters

can also be dimensioned using the same tables.

Using them can be made considerably easier by means of three simplifications clarified below.

Various filters are then examined, once the material has been made rather more accessible, so that anyone should be in a position to lay out filters of this kind for his or her own applications.

2. THE SIMPLIFICATIONS

Basically, there are two different types of filter: symmetrical and asymmetrical.

Symmetrical filters have the same impedance at the input and output (e.g. 50Ω), whereas the asymmetrical formats transform to a greater or smaller degree - i.e., the input and output impedances are different.

Modern equipment design involves a series of stages which all have the same



input and output impedances (e.g. 50Ω). But this applies only to filter circuits which also have the same input and output impedances. For this reason, we shall not go any further into asymmetrical types of filter here. Transformation circuits are a different area. So that already halves the amount of data in question.

The data listed in Pfitzenmeier [1] go right up to fifteenth-grade filters - i.e. those containing a total of 15 structural elements. But firstly, the filter curves which they can produce are hardly ever needed in practical amateur applications, and secondly, they are far too complex to adjust without expensive measuring equipment.

Let us therefore restrict the complexity of filters for practical use to the seventh degree.

The book of tables referred to comprises filter data for transmission curves with a ripple of from (about) 1.25dB, with a reflection factor of 50%, right up to a ripple of 0.00004dB and 0.1% reflection factor. Whilst the one extreme is probably too "coarse-grained" for most applications, the other extreme brings no additional advantages with regard to steeper curves in comparison to a layout without ripple (Butterworth curve).

Thus, for practical purposes, only a certain part of the information available will be required, and so here we are restricting discussion to filters with a ripple of approximately 0.1dB, 0.2dB and 0.5dB, and to Butterworth filters. Thus the entire book can be simplified down to a few easily accessible tables. Since the ripple tends to be rather greater in practice, due to unavoidable

slight tolerances, we always go to the next lowest table when taking the available data from the book of tables. For the applications here, that means 0.0875dB, 0.1674dB and 0.4576dB. Since the tables are laid out for smooth return loss values, correspondingly skewed numbers are obtained for the individual ripple values. For the ripple values referred to above, the associated return loss values are 17dB, 14dB and 10dB, and the reflection factors are 14.12%, 19.95% and 31.62%. For the sake of completeness, the coefficients are also given for the Butterworth cycle - i.e. for filters with a flat-line graph in the transmission range.

3. PRINCIPLES

In contrast to filters with as constant an attenuation course as possible over the entire transmission range, Chebyshev filters allow a certain ripple in the transmission range (usually relatively slight). As a result, the transition into the attenuation range becomes steeper.

Thus in a deep-pass, above the limiting frequency a required minimum attenuation is obtained more rapidly than if the equipment is designed to give the maximum flat transmission range. Another consequence of this is that a simpler circuit with fewer structural elements can already attain the required attenuation value. Fig.1 should make the difference clearer. It shows the curves for a seventh-grade filter, with a Butterworth gradient, and a filter with a Chebyshev gradient with 0.5dB ripple.

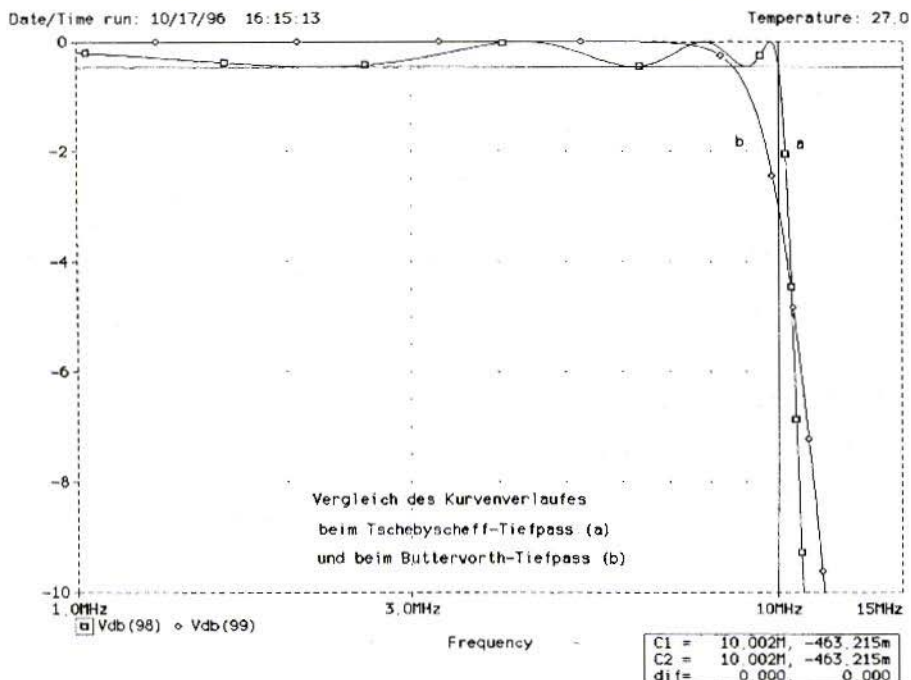


Fig. 1: Gradients of two Seventh-Grade Filters: a) Chebyshev Deep Pass
b) Butterworth Deep Pass
Vergleich des Kurvenverlaufes beim Tschebyscheff-Tiefpass (a) und beim Butterworth-Tiefpass (b) = Comparison of gradient in Chebyshev deep pass and Butterworth deep pass

All these data refer only to deep-pass equipment. The high-pass filters will be considered a little later on.

It will be noted that the Butterworth gradient is completely flat in the transmission range, but that an increase in the attenuation becomes apparent quite some way before the limiting frequency is reached (- 3dB at 10 MHz!). The Chebyshev filter certainly also displays a certain attenuation at individual points (app. 0.5dB), but in contrast continues to maintain this value right up to the limiting frequency, and then the attenuation grows steeper. The two curves intersect at an attenuation level of about

4.8dB. From this point onwards, the Chebyshev filter displays a higher attenuation level as the frequency rises. (The two cursors in the PSPICE simulation were both set to the limiting frequency of 10,000 MHz.)

To provide greater help for the practical users, rather than for theorists, these filters were not standardised to 1 Hz as usual, but were laid out for an application which was practically possible - namely, with a limiting frequency of 10 MHz. From this basis, conversions can be made to other limiting frequencies, using the standard units for the structural elements.

Return Loss 10dB (Riple app. 0.5dB)**Degrees Coefficients**

	a1	a2	a3	a4	a5	a6	a7
3	1,552904	1,105808	1,552904	-	-	-	-
5	1,662454	1,243589	2,495603	1,243589	1,662454	-	-
7	1,694043	1,273338	2,593997	1,362738	2,593997	1,273338	1,694043

Return Loss 14dB (Riple app. 0.2dB)**Degrees Coefficients**

	a1	a2	a3	a4	a5	a6	a7
3	1,187978	1,154234	1,187978	-	-	-	-
5	1,300426	1,345877	2,127107	1,345877	1,300426	-	-
7	1,333498	1,388901	2,238741	1,515988	2,238741	1,388901	1,333498

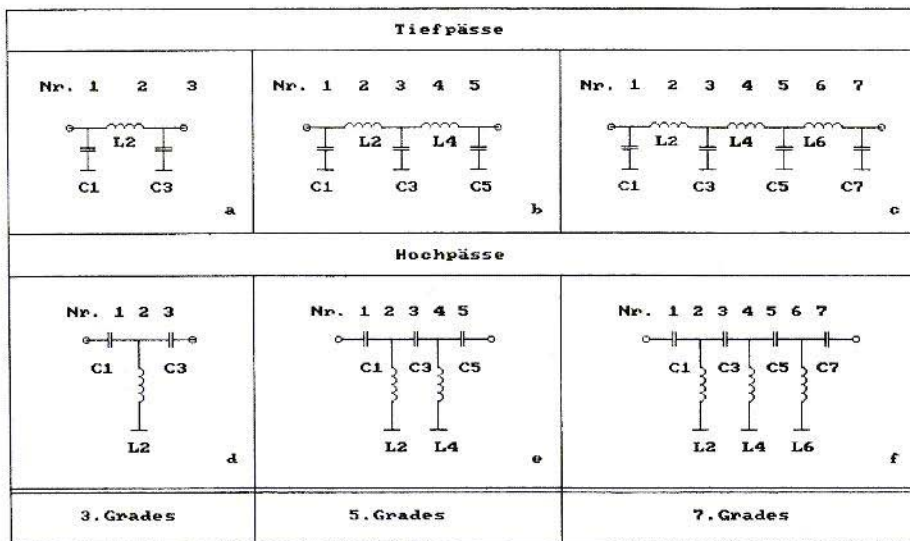
Return Loss 17dB (Riple app. 0.1dB)**Degrees Coefficients**

	a1	a2	a3	a4	a5	a6	a7
3	0,999435	1,142764	0,999435	-	-	-	-
5	1,115397	1,373987	1,944023	1,373987	1,115397	-	-
7	1,150092	1,427636	2,068219	1,583769	2,068219	1,427636	1,150092

Butterworth Gradient (Riple 0dB)**Degrees Coefficients**

	a1	a2	a3	a4	a5	a6	a7
3	1,000000	2,000000	1,000000	-	-	-	-
5	0,618033	1,618033	2,000000	1,618033	0,618033	-	-
7	0,445041	1,246979	1,801937	2,000000	1,801937	1,246979	0,445041

Table 1: Standardised Component Coefficients



Figs.2a-2f: Wiring Diagrams for Low-Coil Deep-Pass and High-Pass Third-Grade to Seventh-Grade Filters (Notice the Component Numbering!)

Tiefpässe = Deep passes, *Hochpässe* = High passes, *Grades* = Stages

It should be pointed out once again that a rise in attenuation to 3dB is usually defined as the limiting frequency for Butterworth filters, whereas the limiting frequency for the Chebychev filters is defined as the point at which the permitted ripple value is exceeded!

4. CALCULATION PROCEDURE

Table 1 shows the selected original values from [1]. They allow us to determine the components for a corresponding Chebychev deep pass.

First we need two simple calculations - namely, the one for "reference induct-

ance" and the one for "reference capacitance". We use the following formulae:

$$L_b = \frac{R_b}{2 \cdot \pi \cdot f_b} \tag{1}$$

$$C_b = \frac{1}{R_b \cdot 2 \cdot \pi \cdot f_b}$$

Here, the impedance (Rb) is given in Ohms and the limiting frequency (fb) in Hertz, the inductance being in Henrys and the capacitance in Farads. These then have to be converted into more practical values, such as microHenrys and picoFarads. If we also fix an operating impedance of 50Ω, we arrive at these simple formulae:

$$L_b = 7.95775 / f_b \tag{2}$$

with Lb in μH, fb in MHz and Z = 50Ω

$$C_b = 3183.1 / f_b$$

with C_b in pF, f_b in MHz and $Z = 50\Omega$

Starting from these two values, we can now calculate the components for the circuit. The wiring diagrams in Fig's.2a to 2f can be used for this.

Three important things can be seen straight away:

- Chebyshev deep-passes are made up only of series inductances and shunt capacitances. The Pi-network is generally preferred, as it needs fewer coils.
- Chebyshev deep passes are made up only of series capacitances and shunt inductances. The T-network is generally preferred, as it needs fewer coils.
- The components are numbered from 1 through to n , these numbers already being identical with the numbers of the associated filter coefficients of Table 1! This is true irrespective of whether we are dealing with a deep-pass or a high-pass filter!

4.1. Calculation Process for a Deep-Pass Filter

For a deep-pass filter, the reference values - reference inductance and reference capacitance - are to be multiplied by the associated coefficients.

In the low-coil Pi-network, component 1 is a capacitance. Its value can be defined as the reference capacitance, multiplied by coefficient a_1 . Likewise, for component 2 we multiply with coefficient a_2 , and so on.

Deep-pass: component $n =$ reference value $\cdot a_n$

4.2. Calculation Process for a High-Pass Filter

For a high-pass filter, the reference values (reference inductance and reference capacitance) are to be divided by the associated coefficients. For the low-coil T-network, component 1 is also a capacitance. Its value can be defined as the reference capacitance divided by the coefficient a_1 . Likewise, for component 2 we divide by coefficient a_2 . High-pass: component $n =$ reference value / a_n

4.3. Calculation for Band-Passes and Band-Elimination Filters

In a similar process to that used to calculate deep-pass and high-pass filters, band-pass and band-elimination equipment can be calculated by conversion from the deep-pass components. Since in the practical application involved these are of less significance, we shall not attempt to clarify them here. These applications are explained in sufficient detail in the literature source referred to, so additional information can be obtained there if required.

5. LITERATURE

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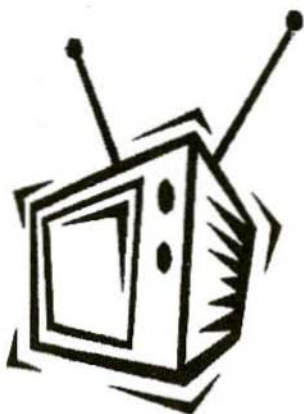
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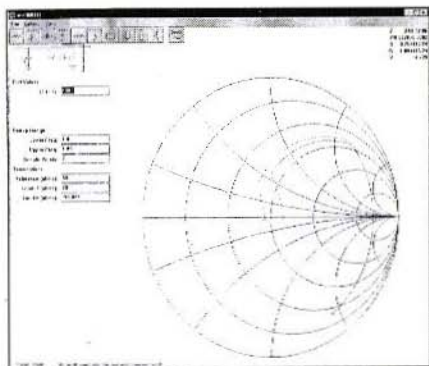
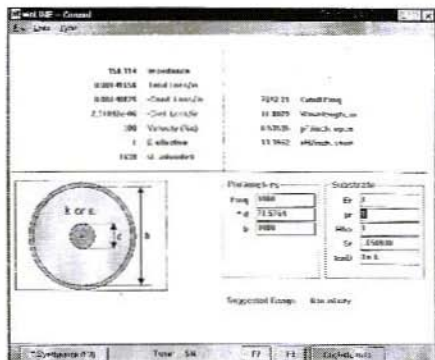
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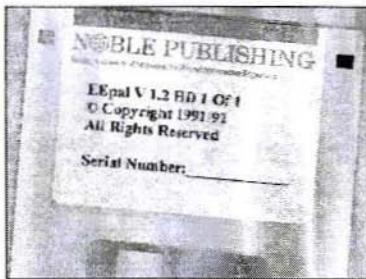
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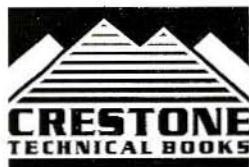
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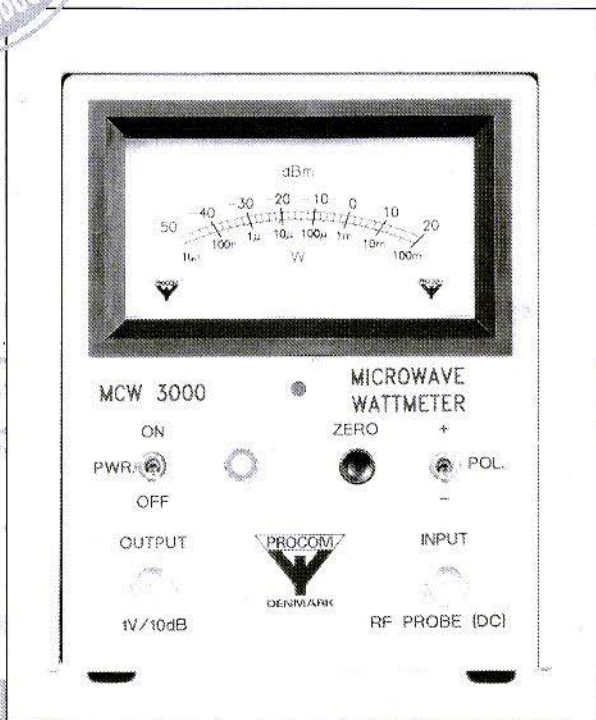
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