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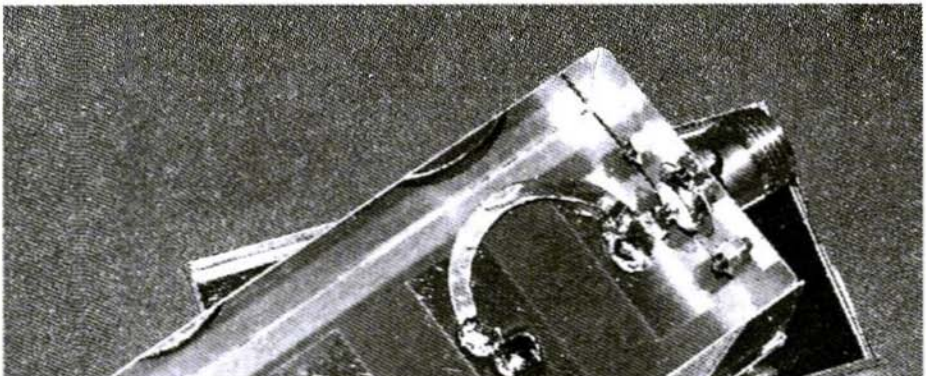

Low Pass Filter for 2 m and 70 cm

Gerhard Schmitt DJ5AP

Contents

The end of my first year as publisher. I would like to wish you all a Happy New Year and thank you all for your support.

I would like to welcome back Christiane Michel as French agent, details can be found on page 256.

73s - Andy

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Wolfgang Schneider, DJ 8 ES

High-Precision Frequency standard for 10 MHz

Since amateur radio began, frequency precision has been one of the most interesting (and simultaneously one of the most important!) subjects for radio enthusiasts. One point has been brought out time and again in various discussions with radio amateurs - frequencies must be measured with extraordinary precision, i.e. much more precisely than currents or component values. In spite of the technical options available today, most of the frequency counters available to us radio amateurs in the shack offer us display "accuracies" which, strictly speaking can not really be described as such. The same is true for the digital frequency displays in radio equipment. But what degree of accuracy will be useful while still being financially affordable?

1.

General

The accuracy of a frequency counter is determined simply and solely through the precision of the time base, if the error of ± 1 digit in the last display place is left out of consideration. The reciprocal value of the time base frequency deviation is equal to the error of measurement. If, as is usually the case, the counter has only a simple quartz oscillator, then its temperature dependence causes a considerable error. The same problem naturally also occurs with quartz oscillators in transverters, test transmitters, beacons, etc.. Nor is a PLL control circuit, which merely uses a simple quartz oscillator as a reference frequency, of any assistance here.

Normal quartzes are specified to have a temperature dependence of ± 30 ppm (parts per million) in the range from 10°C to 60°C. Better, and thus more expensive, quartzes can offer a level of accuracy of ± 10 ppm. Normally such temperature differences seldom occur within rooms, but are decidedly more frequent in portable radio equipment with a transverter, perhaps even up to 10 GHz.

To obtain a clear idea of the possible frequency deviation due to temperature, we should have a clear picture of the following ratios:

- ± 30 ppm corresponds to: ± 30 Hz at 1 MHz or else to a frequency error of:
- ± 300 Hz in a standard frequency counter with a measurement frequency of 10 MHz;

- ± 4,32 kHz at 144 MHz (2-m-Band)
- ± 38,9 kHz at 1296 MHz (23-cm-Band)
- ± 300 kHz at 10 GHz (3-cm-Band)

Against this background, making an sked with a frequency in a higher band (e.g. 10 GHz) is something which is really scarcely possible. Modern and narrow-band digital modes are just out of the question here.

To get a grip on this problem, there are generally two options:

- Stabilising the quartz oscillator temperature
- Re-setting the oscillator with the help of a frequency standard.

Combining both options is naturally better.

2. The temperature-stable oscillator

Special measures are taken in high-quality industrially manufactured frequency counters, and often in DIY equipment as well, to counteract temperature dependence.

For example, we can read in data sheets that the time base is generated with a TCXO. The abbreviation TCXO refers to a temperature-compensated or else temperature-controlled quartz oscillator. Whichever it is, the measures taken have the effect of restricting the quartz frequency to a maximum frequency error of ± 1 to ± 5 ppm in an acceptable temperature range. Better products, such as TCXO4 at 10 MHz from NARDA or a DIY oscillator for GHz applications, e.g. as per DF9LN, attain precision levels of ± 0.1 ppm.

The model HP 10544A quartz oscillator

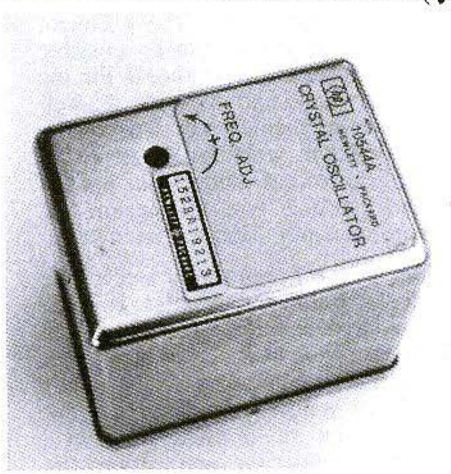


Fig 1: HP 10544A Quatz Oscillator

from Hewlett Packard is almost two powers of ten better, and this at a price which is still acceptable. In the temperature range between -55°C and + 70°C, the oscillator attains a stability of ± 0.015 ppm (1.5 x 10^{-8}). In the range between 0°C and + 70°C, the value improves to 7 x 10^{-9}.

The HP 10544A quartz oscillator is an

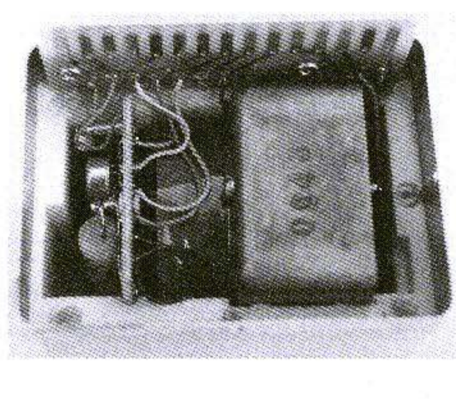


Fig 2 Inside view of HP 10544A

extremely stable reference source for 10 MHz with compact size. The excellent technical specification is made possible by a special production process for the quartz, in combination with a well thought out mechanical construction. A quartz oscillator, a buffer stage and a regulated heater are mounted in a well-insulated housing with dimensions of only approximately 7 cm x 5 cm x 6 cm.

Possible fields of application are, a time base for a frequency counter or as a reference frequency in test transmitters and test receivers or spectrum analysers. The oscillator can likewise be made available as a reference frequency for the production of a GHz frequency. Many other interesting applications are conceivable. All inputs and outputs are fed through a 15-pin connector block on the underside of the housing.

2.1

The essential technical data of the HP 10544A:

- Output frequency: 10.000 MHz
- Output level: 1VRMS at 1 kΩ
- Power supply (DC): +11.0 V to +13.5 V / 15 mA (oscillator, buffer stage)

  +11.0 V to +13.5 V / 5 mA (heating control)

  +20 V to +30 V / 400 mA or 200 mA after 15 min./25°C (heating)

  -5 V to + 5 V (frequency tuning)
- Temperature stability: 1.5×10^{-8} (-55 °C - +70°C), or. 7×10^{-9} (0°C - +70°C)

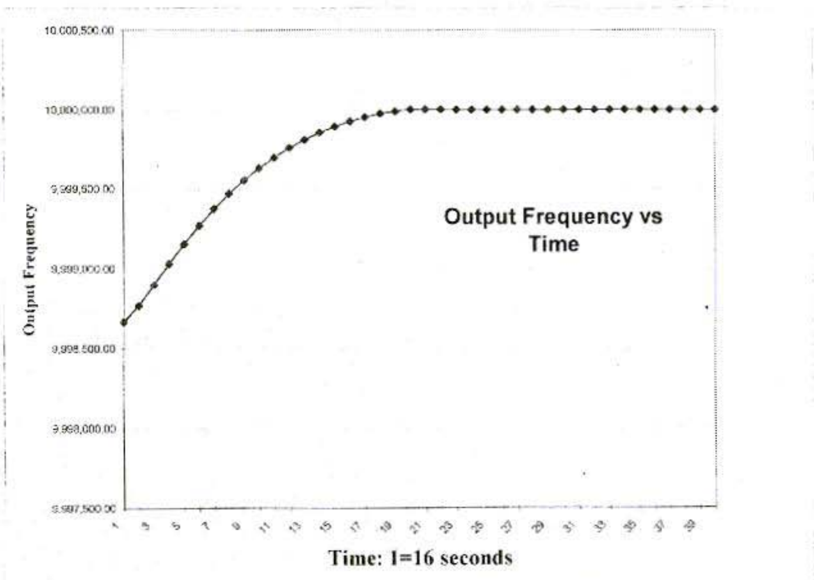


Fig. 2: Switching-on behaviour of HP 10544A

- Short-time stability: 1×10^{-11} / 1s, or 2×10^{-11} / 100 s
- Long-time stability: $< 5 \times 10^{-10}$ / day, or $< 1.5 \times 10^{-7}$ / year
- Side-band noise: -130 dB at 10 kHz
- Coarse tuning: ± 10 Hz (mechanical)
- Fine tuning: ± 0.5 Hz (electrical, -5 V to +5 V)

2.2. HP 10544A pin configuration (15-pin connector block)

Pin Connection

1 Output 10 MHz (1V / 1 kΩ)
2 GND
3 Oscillator (+12 V / 15 mA)
4 GND
5 GND
6 Tuning (-5 V..+5 V)
7 NC
8 Heating monitoring (+12 V 5mA)
9 GND
10 NC
11 Heating monitoring (Low = cold, High = warm)
12 NC
13 NC
14 Heating (+24 V / 200..400 mA)
15 GND

Recommendations:

- The oscillator (pin 3) can be powered from the 12 V source at Pin 8 through a 10 mH choke. Connect a 100µF / 25 V electrolytic capacitor to pin 3 for smoothing to earth.
- The operating voltages for the oscillator, the amplifier and, above all, the frequency tuning should be low-noise. V. Esper, DF 9 PL, has published an interesting circuit proposal for this in [3]

3.

Buffer stage with analogue and digital outputs

The HP oscillator output is comparably high impedance at 1 kΩ. A buffer stage would provide a solution here. In the version proposed, 3 separated analogue 10-MHz outputs, each having 10 mW (-10 dBm) and additional digital outputs with TTL levels of 1 MHz, 5 MHz and, of course, 10 MHz are available.

In this connection it should be mentioned that any analogue outputs which are not in use should be terminated with 50 Ω, otherwise the mismatching will be reflected at the output back into the circuit.

The individual transistor stages are designed in such a way that the input impedance of 1 kΩ is obtained through a parallel circuit and is thus suitable for the HP oscillator. Fifth-order low-pass filters are also provided for the 3 analogue outputs. The harmonic content of the 10-MHz oscillator is thus considerably improved.

The buffer stage assembly (DJ 8 ES 046) is constructed using the SMD components to reduce space the double-sided epoxy printed circuit board measuring 60 x 100 mm.. The fully copper-coated rear side of the printed circuit board acts as an earthing surface. Short pieces of wire are used for feedthrough.

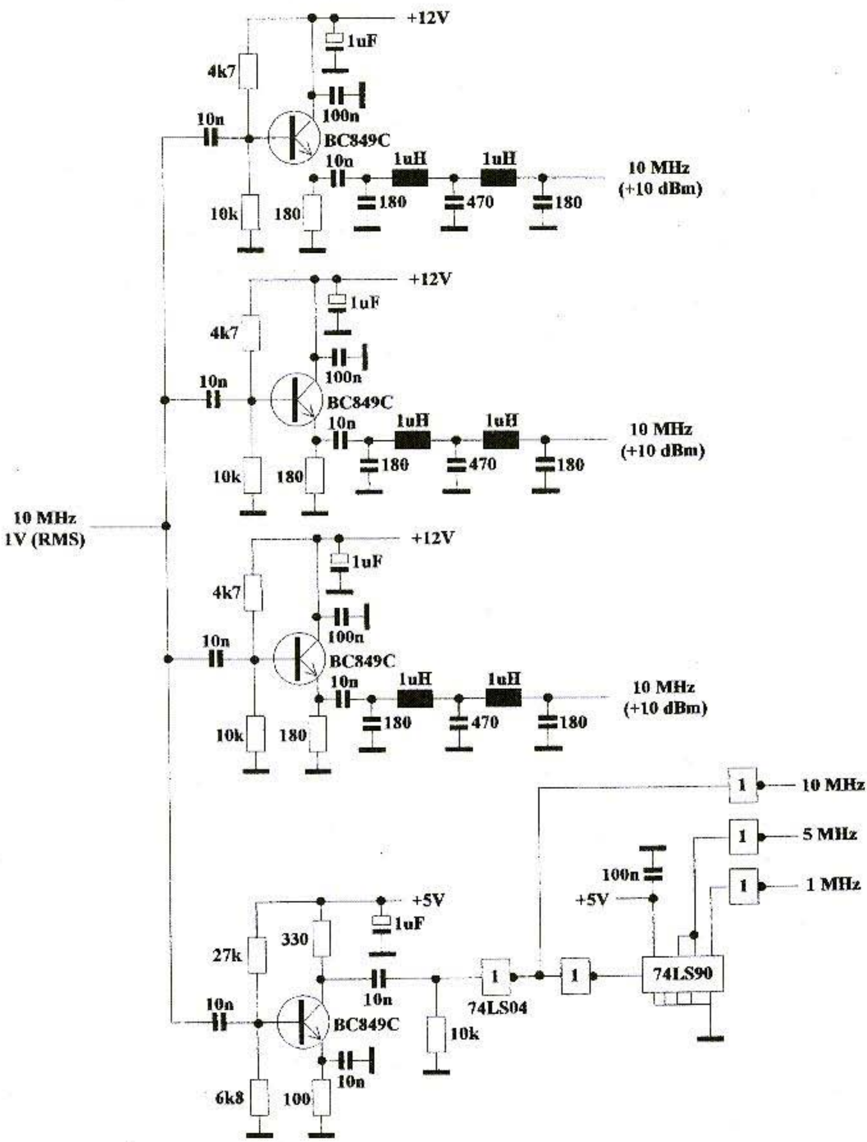


Fig 4: Circuit Diagram of Buffer Stage with Analogue and Digital Outputs

3.1.

Buffer stage DJ 8 ES 046 parts list

4 x BC 849 C, Transistor
1 x 74 LS 04, TTL-IC
1 x 74 LS 90, TTL-IC
6 x 1µH, choke
4 x 1µF / 25 V, tantalum electrolytic capacitor
14 x terminal pin 1,3 mm
1 x DJ 8 ES 046, printed circuit board

Resistances:

1 x 100Ω , SMD
3 x 180Ω , SMD
1 x 330Ω , SMD
3 x 4.7 kΩ, SMD
1 x 6.8 kΩ, SMD
4 x 10 kΩ, SMD
1 x 27 k, SMD

Ceramic capacitors:

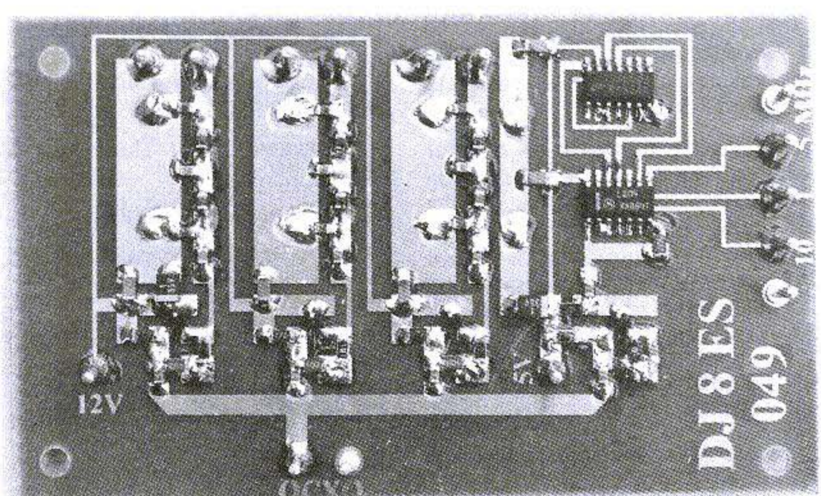
6 x 180 pF, SMD
3 x 470 pF, SMD
8 x 10 nF, SMD
5 x 100 nF, SMD

4.

Outlook

High-precision oscillators of this type provide the user with a frequency standard which corresponds to, and perhaps even exceeds, the objective of attaining additional frequency precision set out at the beginning. This is irrespective of whether the oscillator is used as a reference source or as a time base in the frequency counter.

However, there is one fact which should in no way be overlooked: irrespective of the method selected or the quality of the temperature compensation, quartz ages -



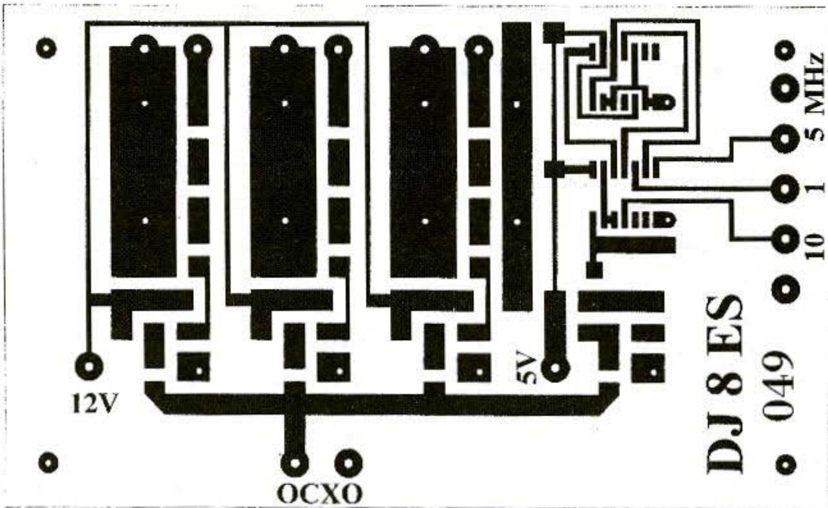


Fig 6: Layout of PCDB DJ8ES 049 for Buffer Stage

nor is this quartz ageing automatically compensated for!

For many high-quality quartzes, according to the manufacturers specifications, ageing can be up to 5 ppm per year, which corresponds to a value of at least 50 Hz at 10 MHz. The HP 10544A, has a value of $< 5 \times 10^{-10}$ / day (0.0005 ppm, 0.0005 Hz at 10 MHz), for operating round the clock, 7 days a week,...$< 1.5 \times 10^{-7}$ / year (0.15 ppm, 1.5 Hz at 10 MHz).

Anyone not able to carry out regular (annual?) calibration still has the option of having this done at one of the amateur radio congresses (e.g. VHF Congress in Weinheim). For it is not simple to carry out such a calibration by ear. An alternative to this is electronic re-adjustment using:

- the line frequency of a television transmitter
- longwave transmitters such as DCF77 or LORAN C, or
- with the help of the satellites of the Global Positioning System (GPS).

But this needs a separate article.

5.

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[1] Frank Sichla, DL 7 VFS:

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Funkamateur 7/95, Theuberger Verlag Berlin

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[3] Volker Esper, DF 9 PL: High-stability, low-noise power supply, VHF Reports 2/1992, Pages 81-93, VHF Communications 1/93, Pages 19-30

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A Quadrifilar Helix Antenna for Orbiting Weather Satellites?

More than ten years ago, I was often able to receive images from NOAA satellites from a distance of almost 3,000 Km., e.g. from locations in the vicinity of Luanda/Angola (2,740 Km.) or even Ndjamena/Chad (2,840 Km.). The turnstile reflector antenna used receives nothing at elevation zero, and thus at 870 Km. flying height, where the radio horizon has a radius of 3,330 Km., could not be expected to cover the full theoretical range. For reception above 15° elevation, it is still 2,200 Km., and down to only 1,500 Km. at 30°. The trees grew taller with time (tropical rain forest!) and reception became worse. So when a friend told me that a quadrifilar helix antenna was better than any other (non-tracking) antenna system, this was just the advice I needed. The following article is based on my experience.

1. The W3KH quadrifilar helix antenna

Ruperto [4] has given a detailed description of the structure of this antenna. Thankfully Fig. 1, was already in existence, I only had to translate the Imperial measurements into Metric. In my version, the PVC carrier tube has a diameter of 110 mm. and the struts are made of conduit with a diameter of 20 mm.. The antenna elements are all made from hydraulic tubing (external ϕ 5.5 mm., internal ϕ 3.0 mm.); a length of RG-188 PTFE coax cable has been inserted into one of them for the feeder.

This tubing is very easy to solder and can be bent into the correct shape by hand. All element lengths specified in [4] turned out to be somewhat too long. That was why the "twists" visible in Fig. 2 were formed, to reduce the length and increase the resonant frequency. Once the setup was correct, the return loss and the band width were measured (Fig. 3) and found to be not too bad. Then the antenna was installed on the roof (Fig. 4) and could be compared with the turnstile reflector antenna.

2. "The song contest"

The turnstile reflector antenna has a reflector with a spacing of 3/8λ. The two dipoles are connected by a 1/4λ phasing line. A further line with

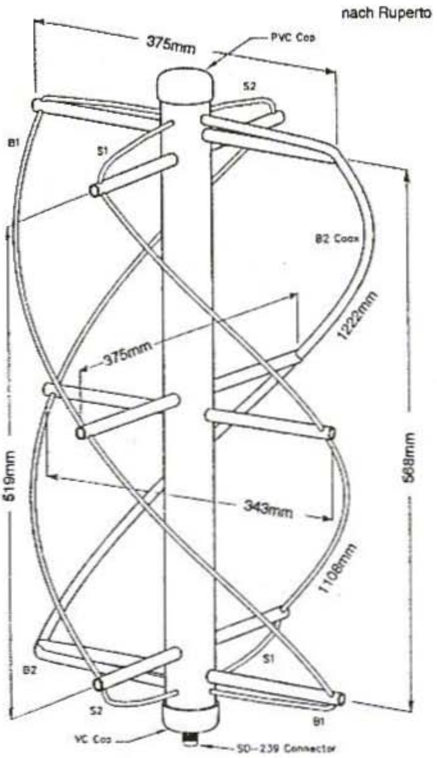


Fig. 1 Diagram of W3KH quadrifilar helix antenna according to Ruperto [4]

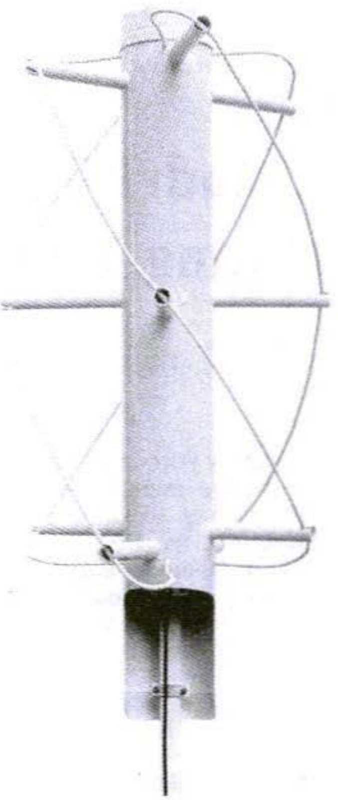


Fig. 2 View of duplicated QHA

$Z=37.5\Omega$ matches the downstream amplifier with G=18 dB. An identical amplifier was also built into the quadrifilar helix antenna and naturally tested to see whether it also had the same amplification and noise performance. There was already a suitable changeover switch with a reed contact available, as it was used for changing over between the VHF and UHF weather images. The reed relay was now switched backwards and forwards, using the feeder, filter and the voltage of a function generator, so that each antenna in turn fed the receiver for 10 seconds. A logarithmic signal was taken from its

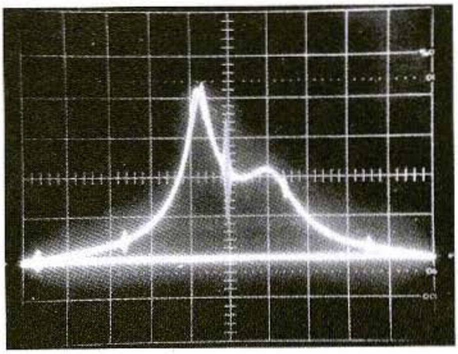


Fig. 3 Return loss of duplicated QHA

X = 5 MHz/div, small marks all 10 MHz, large mark 137.2 MHz

Y = 5 dB/div, reference line 0 dB

Fig. 4 Both antennae ready for reception on roof

intermediate-frequency integrated circuit to record the signal strength. I have already described this earlier in [2]. Fig. 5 shows a schematic diagram of the equipment layout. Finally, several days transmissions were recorded. The surprising result was that the quadrifilar helix antenna was by no means superior. With only one exception, the signals it supplied were about 10 dB lower (Figs. 6 b and c) and it was only very slightly superior in a single direction, to the North-East. As can be seen at the end of recording in Fig. 6a, the quadrifilar helix antennas signal strength in the last few minutes was a few dB above that of the turnstile reflector antenna. In the reception direction referred to were

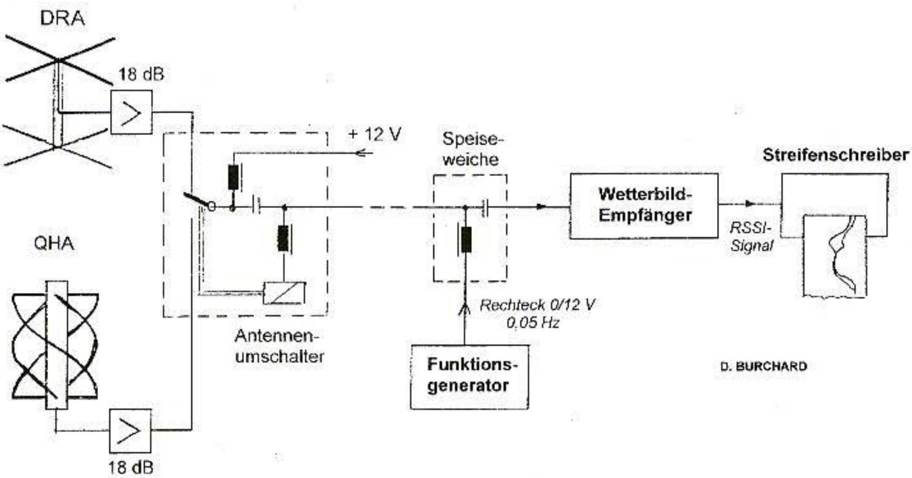


Fig. 5 Equipment layout for recording

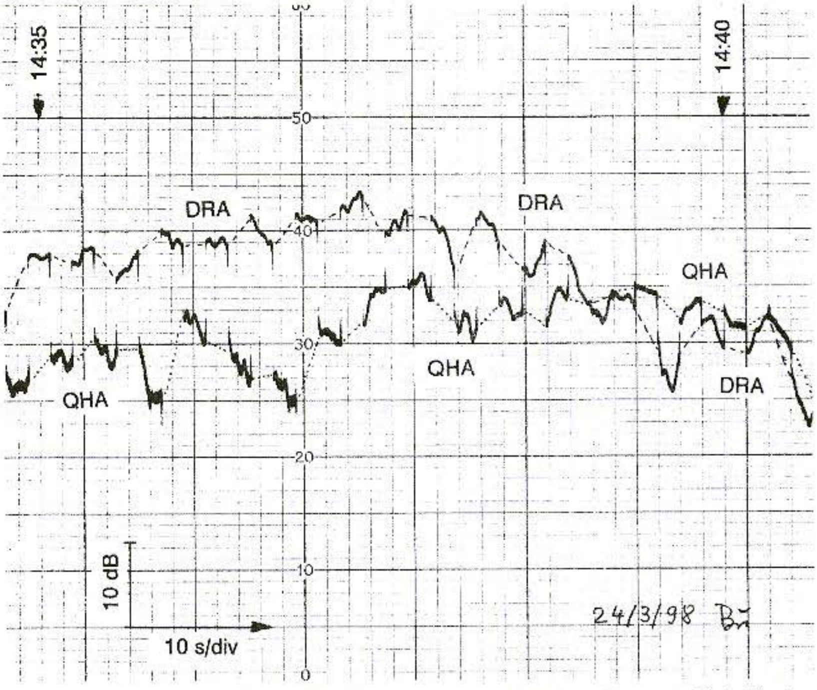


Fig. 6a : NOAA Weather Satellite Signal Strength Recording on 24.3.98. Start 14:35, 49°E, 15°S, Tsaratanana, Madagascar, End 14:41, 47°E, 7°N, Damot,

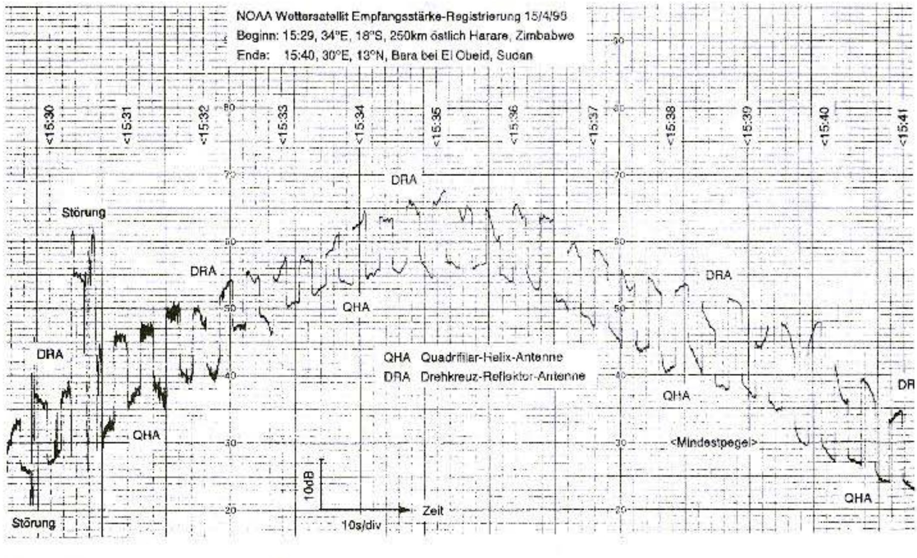


Fig 6b : Recording on 15.4.98.

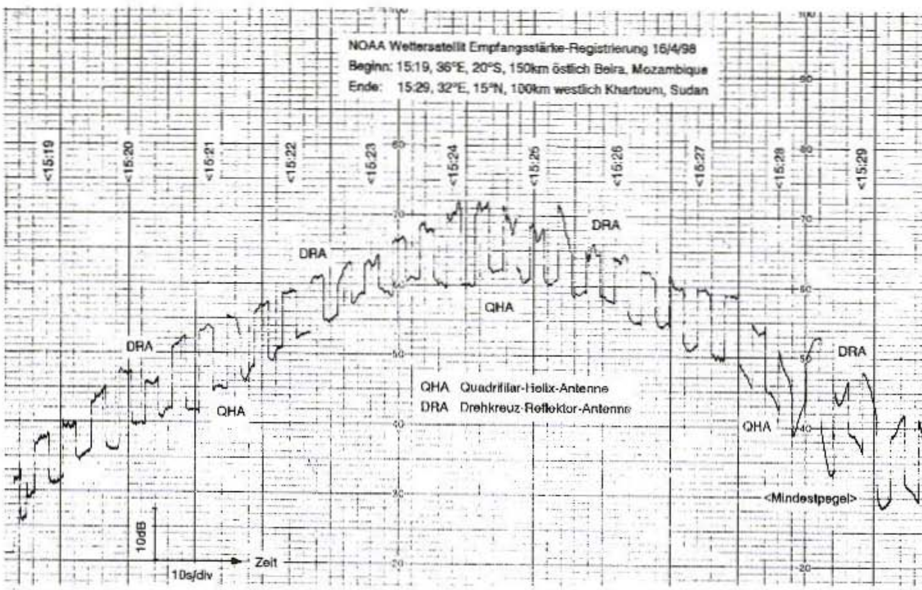


Fig 6c : Recording on 16.4.98.

several jacaranda and croton trees which were 25 m high, a 4 m high gate made of black sheet and, a little distance behind that, a three-storey building. It is certainly a problem direction. Even shifting the reception location by a few meters can produce the effect outlined. But naturally the question now is why the quadrifilar helix antenna performs so badly. If we look at Fig. 4 again, we notice immediately that the quadrifilar helix antenna takes up much less room. It therefore probably also has a smaller absorption area. In order to explain what this obvious fact means in actual figures, some more precise investigations will be required. I shall try to make a contribution.

3.
The 1λ loop antenna

Fig. 7 shows several wire loops. They should all have the mechanical length of a wavelength of the electro-magnetic wave in air. Other forms of loops are conceivable and have been described in the literature - round, triangular, octagonal, etc.. The quadrifilar helix antenna consists of two loops, which are also inter-wound which can very easily be seen in Vidmar [5]. The loops may also be round and flat, which gives an antenna as in Berberich [1]. I have to confess that I dont have enough imagination to be able to understand the functioning of a quadrifilar helix antenna right away. Others may be in a

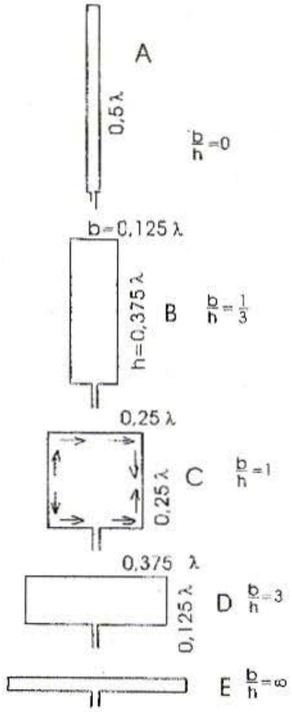


Fig. 7 Rectangular loop antennae

A:b/h=0, B:b/h=1/3, C:b/h= 1, D:b/h=3, E:b/h=∞

similar position. The rectangular loops help you to understand. Then you can twist them and finally inter-wind them.

The first and last of the loops from Fig. 7 are well known to us: A is a 1/2λ parallel-wire line, E a 1/2λ folded dipole. The first emits practically no radiation, the last emits radiation in the form of a balloon tyre, vertically to its axis. Why this is so is explained by the path of the current in the loop in Fig. 7C. Essentially, this form of current path is present, in a very similar manner, in all these loops. Starting from the feed point, the current in both conductors diminishes approximately sinusoidally, becomes zero after a path of ¼λ and then increases again in the inverse direction (phase). Opposite the feed point, it has regained the same magnitude and direction. If energy is radiated off, then the forward-moving wave becomes smaller more quickly than without radiation; the return wave, reflected at the opposite short-circuit, undergoes the same reduction. It is finally absorbed at the power input, because matching takes place there. So it comes about that the currents in the two horizontal conductors are equal and are in phase. In the vertical conductors, the currents, also largely equal, flow towards or away from one another.

The effect of the antenna is given by the vectorial sum of the effects of all the partial elements and their partial currents. The inter-connection can be elegantly represented by linear radiators, but not very clearly for three-dimensional antennae. Thats why I am beginning with the rectangular loop and not with the mathematically simpler round loop.

You must now understand that each of the four sides of the rectangle has a magnetic field arising from the current flow, which, in turn, creates an electrical field vertical to itself and in the direction of the conductor, and that the two of them combine to form an electro-magnetic wave, which moves away from the antenna, vertical to the two field directions. For each of these linear radiators, there is thus an E-plane and an H-plane. If these terms are applied to the entire antenna, they cause confusion. In order to understand the system without the mathematics and without confusion, we should first clarify which conductor can not generate field strengths in which direction. For the two horizontal ones, its the directions to the right and left, whereas for the two vertical conductors the same applies going upwards and downwards. This is always true if the conductor observed has no projection in a direction or, to put it more simply, is invisible from this direction. But for the two

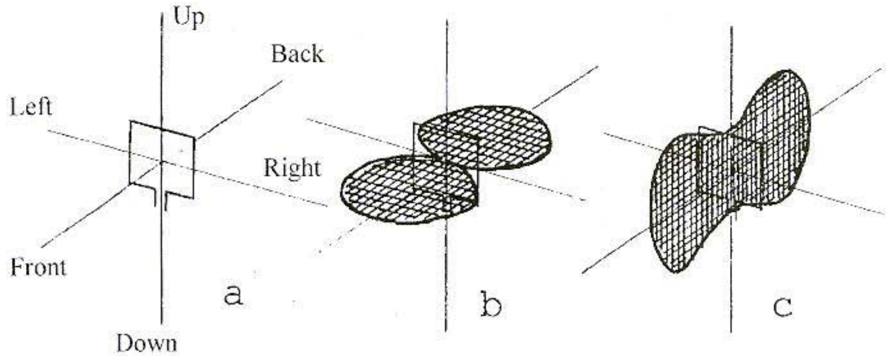


Fig. 8 Radiation diagrams for rectangular loops, horizontal polarisation: a in plane of loop, b vertical to loop and horizontals, c vertical to loop and verticals. There is no radiation in the vertical plane.

vertical conductors it is also true that the vector sum of their partial currents is zero. They therefore also have no radiation in any other direction (except in the immediate near field). The rectangular loops thus behave like two stacked dipoles, which are connected at the ends by means of a non-radiative two-wire circuit, and the lower dipole of which is powered in the middle. Due to the constant length of the loop, the result is either short dipoles with a big distance between them or long dipoles with a small distance between them.

The characteristics can thus be described as - total radiation with horizontal polarisation in the forward and backward directions. This is because here the radiation effects of the horizontal conductors are in phase and are added to each other. In the literature, we read that there is a gain of almost 2 dB if format 7C is used instead of 7E.

In the upward and downward directions, in contrast, the radiation is less, because there is no scalar addition, thanks to the difference in propogation time between the waves of the upper and lower conductors (45...135° from D...B). So, even if the short horizontal conductors of format A still give off some radiation, this is cancelled out in the upward and downward directions. We now know that the radiation is horizontally polarised and in the horizontal plane it takes the form of a figure eight, similar to a dipole. In the vertical plane it can vary between being perfectly circular (E), dented above and below the centre line, and in the form of a figure eight (A), and it is zero in the paper plane. This relationship is shown in Fig. 8. There is no vertical polarised radiation in any of the planes.

4.

The twisted 1λ loop

If the flat loop is wound round a cylinder, then the linear elements above and below are no longer in phase and the linking elements between their ends are no longer linear, even if they almost appear to be so in the overall view in Fig. 9. In spite of the twisting, case A remains a 1/2λ line, which gives off no more radiation than the equivalent

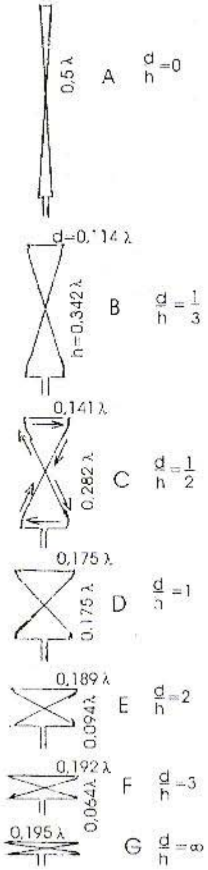


Fig. 9 Helix loop antennae

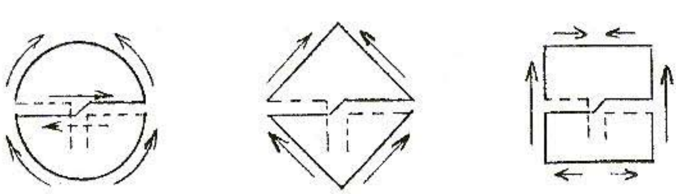
A:d/h=0, B:d/h=1/3, C:d/h=1/2, D:d/h=1, E:d/h=2, F:d/h=3, G:d/h=∞

version A from Fig. 7. But should the elements still give off radiation, due to their not inconsiderable length, then their radiation emissions are added together in the upward and downward directions, whereas they cancel each other out forwards and backwards. We can make a similar observation regarding the other limiting case (version G), in which all the elements lie in the same horizontal plane. In plan view, the antenna then consists of a wheel with intersecting spokes (Fig. 10). Here too, models can be imagined which approximate to the characteristics arising with linear elements and can thus be more easily comprehended (b or c). It can be recognised immediately from either that the radiation effects of the two spokes cancel each other out, as also do those of the front and rear elements running parallel to them. The lateral elements, however, have currents with the same phase and direction, the fields of which are added together in the upward and downward directions and do not quite cancel each other out to right and left. The horizontal ring thus has strong radiation upwards and downwards and weaker radiation to the right and the left, all with horizontal polarisation.

No radiation with vertical polarisation is present at all.

In principle, the twisted loop looks like Fig. 11a. The plan view, b, does not

Fig. 10 Limiting case G and linear approximations



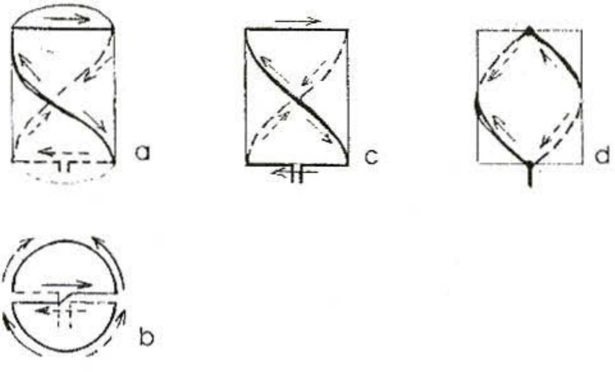


Fig. 11 General representation of twisted loops

look different from Fig. 10a, i.e. all forms have radiation as in case G from Fig. 9 and in addition to this also have that from the twisted lateral elements. With regard to the strength, it can be said that the lateral radiation decreases with the height of the antenna, whereas the radiation upwards and downwards initially increases with the height, but then goes down again to almost zero.

Radiation with vertical polarisation can be emitted only from the lateral elements. In this connection, we can observe Fig. 11c and estimate the levels. In addition, we must now also bear it in mind that the two elements are at a distance from each other. The one drawn solid lies toward the observer and the one drawn in dotted lines is behind it. Currents in opposite directions at the

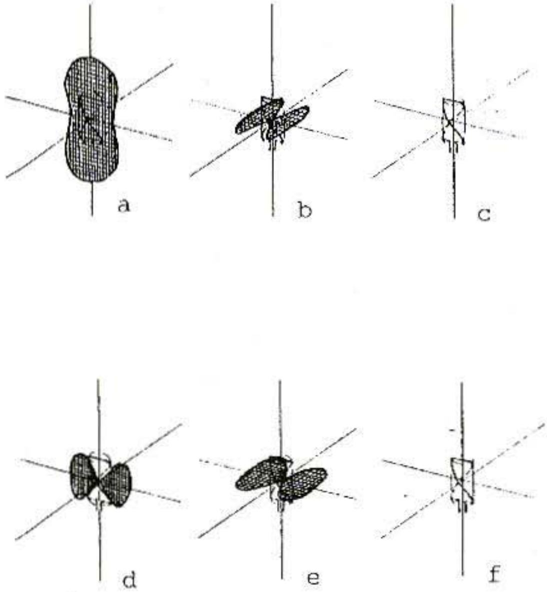


Fig. 12 Radiation diagrams of twisted loops

a - c horizontal polarisation, d - f vertical polarisation. Planes correspond to those from Fig. 8.

There is no radiation in plane vertical to loop.

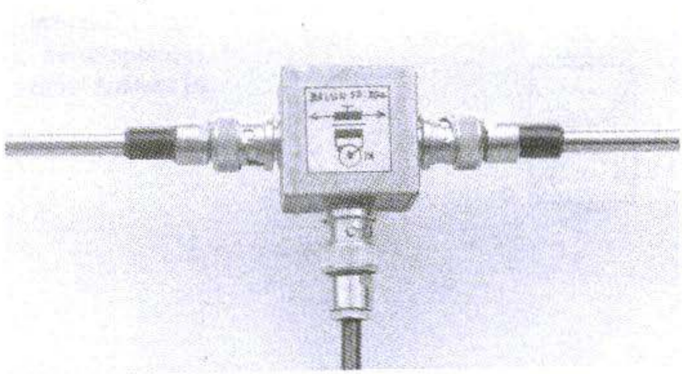


Fig. 13 Hand-held field strength probe and circuit.

Transformer on B62152-A8-U17 core (Siemens)

Has (3+3):6 coils 0.1 CuL wire.

same distance from the observer cancel each other out. This is the case here; there is no radiation forwards or backwards with vertical polarisation. The side view looks different (Fig. 11d). Here there are slanting arrows going in the same direction from the observer. The horizontal components of these currents have already been dealt with in Fig. 10. The vertical components provide for powerful radiation to the right and left, which initially increases with the height of the antenna, but then decreases down to zero, because the conductors continuously move closer together, and the radiation is obliterated. The radiation diagram for a twisted loop thus obtained through observation is shown in Fig. 12. It is largely identical to that which can be created using a hand-held field strength probe (extended by a broomstick if necessary) at a little distance Fig. 13. It also emerged here that there is no preferred half-plane; this loop radiates at the same strength upwards and downwards, right and left and forwards and backwards. Directionality requires elements with $1/4\lambda$ phase displacement and corresponding distances. Neither is present here. If feeding power from the top has any effect, this is less important than the cancellation obtained with the field probe!

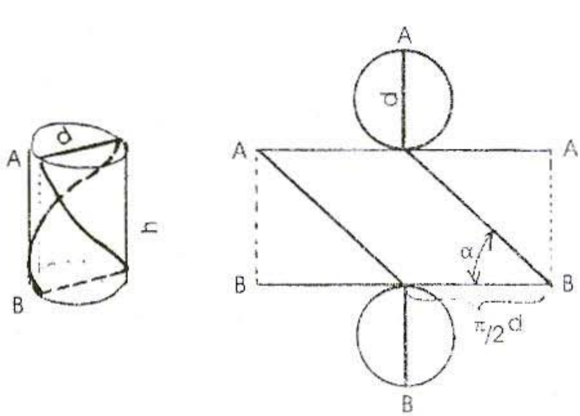


Fig. 14 For calculation of helix loops

The calculation becomes simple if we consider the developed view in Fig. 14. It can be read off immediately:

$$\lambda = 2d + \sqrt{\pi^2 d^2 + 4h^2} \qquad (1)$$

$$\tan\alpha = \frac{2h}{\pi d} \qquad (2)$$

5.

The quadrifilar helix

Two flat loops under 90° do not have a mutual influence on each other; no special evidence is needed to prove this. Two inter-wound loops, forming angles of 90° in the same way in every horizontal plane, can not be assessed so easily. However, what is easy to measure is the change in the impedance of one loop when the other is idling, terminated with 50 ohms or is short-circuited. And some scarcely detectable changes take place here. It is thus not completely wrong to say that the mutual influence is slight. This may be due to the fact that a helix with a half twist still produces no rotary field. This requires the other loop, with the power feed in quadrature.

It can be seen from Fig. 12 that there is no radiation vertical to the loop. But this direction is optimally filled by the other loop, laid out and powered at 90°. Then there is also circular polarisation upwards and downwards. By varying the height and diameter, a balance can finally be achieved for the radiation in all directions in the horizontal and vertical planes. We simply believe that Ruperto [4] suceeded in making the radiation isotropic, that the ratio d/h = 0.62 and the gradient a = 46° are optimal. Compared with the turnstile reflector antenna, radiating only upwards, the spheridian angle of which is approximately $(120°)^2$, whereas the quadrifilar helix antenna approaches $(360°)^2$, the difference in performance actually amounts to 9.5 dB. This explains the differences in Figs. 6 a-c adequately, but not perhaps plainly enough.

A further consideration should thus apply to correct matching. From Fig. 3 we know that the centre frequency is correctly positioned and that approximately 10dB reflection attenuation is present over a band width of 14 MHz (i.e. approximately 10% of the centre frequency); from Fig. 1 we calculate a difference of 10% in the tuning of the two loops. The de-tuning in relation to one another matches the 3-dB band width of a loop of 10%. Then we set a phase angle of 45° (i.e. as against the desired 90°), with matching to the radiation resistance of one loop which is lying at 36Ω. From the readings, no inadmissable departure from these rules can be recognised. So what we are left with is that the poorer reception performance of the quadrifilar helix antenna is a problem in principle, i.e. it is not a good antenna for receiving weak signals.

6 Appendicies

6.1: Loop impedance readings

In the course of the investigations, impedance measurements were carried out on all the loops from Figs. 7 and 9. Since these data could be of general interest, they are reproduced here.

Impedance measurement is relatively simple to carry out; symmetrical and ungrounded sampling heads are not required, because the connection can be

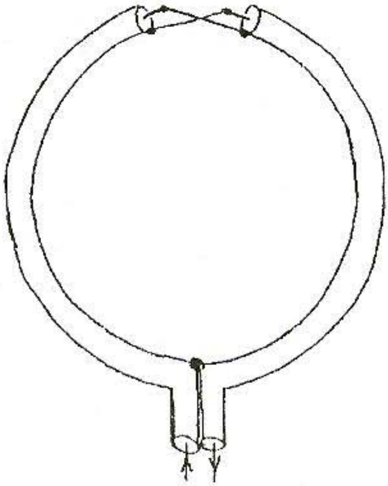


Fig. 15 Z measurement on loop antenna

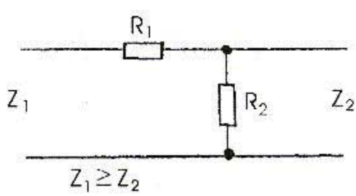
brought about through the interior of the element tubing (Fig. 15). One coax cable feeds in a signal from a generator, the other leads to any rectifier. If standing wave ripple is present, 10 dB attenuators in the coax cables help, and broad-band amplifiers increase the signal if it has become too weak. I have also described this impedance measurement before in [3]. It works most precisely if the mean value remains below 2Zo. It can therefore be necessary to alter Zo using L-networks irrespective of the frequency, as shown in Fig. 16. The resistances should be calculated as shown.

If coax cables are also used for the actual loops from Fig. 15, they can be made into all possible shapes on a suitable wooden frame, namely the shapes required here and many others. Since the loop length remains the same, it will be simultaneously recognised how the resonance frequency also changes along with the shape. For example, a curve like Fig. 17 can be found. This can yield the resonance frequency, the resonance (= radiation) resistance and the resonance band width. They are listed in Tables 1 and 2, together with the values obtained for the angle of inclination and the sharpness of resonance, Q.

6.2:

Quadrature and matching

Two series resonant circuits wired in parallel, the resonant frequencies of which differ by $\pm\Delta f/2$, have reflturn loss curves as in Fig. 18a. The curves were



$$R_1 = \sqrt{Z_1(Z_1 - Z_2)}$$

$$R_2 = \sqrt{\frac{Z_1 \cdot Z_2^2}{Z_1 - Z_2}}$$

Fig. 16 Networks for changing Zo

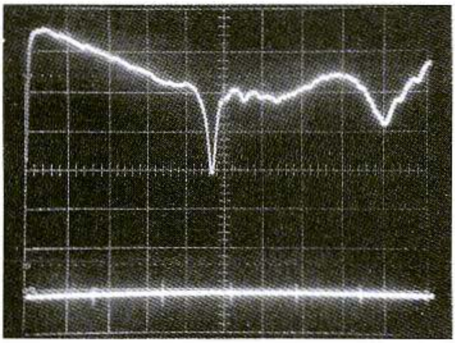


Fig. 17a Example of impedance measurements on flat loop with d/h=0.33

X=30 MHz/div, marks all 50 MHz

Y=impedance 25Ω, zero line superimposed

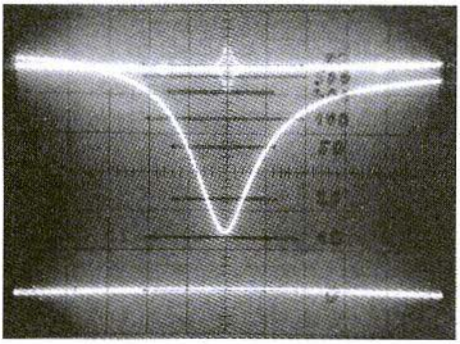


Fig. 17b Example of impedance measurements on twisted loop with d/h=0.03

X=2 MHz/div, large mark 122.9 MHz, small marks all 10 MHz

Y=Impedance corresponding to black scale, O and ∞ superimposed

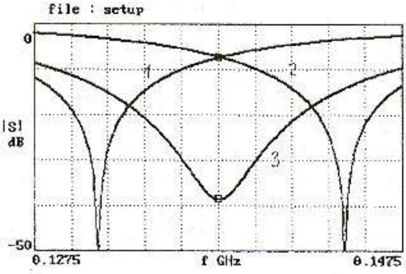
Type	D:h	Resonant Frequency MHz	3 db Bandwidth MHz	Series Resistance Ohms	Quality
A	0.04	125.8	1.8	11	70
B	0.33	137.3	3.4	45	40
C	1.00	139.6	9.8	170	14
D	3.00	132.2	18.5	280	7
E	25.00	123.5	12.3	310	10

Table 1: Flat loops (cf. Fig. 7)

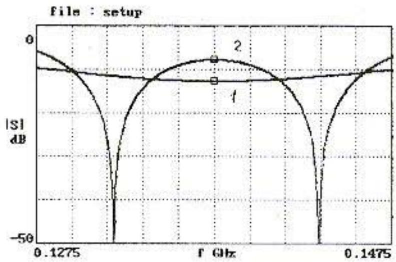
Type	D:h	Angle Deg	Resonant Frequency MHz	3 db Bandwidth MHz	Series Resistance Ohms	Quality
A	0.04	86	125.8	1.8	11	70
B	0.33	62	133.4	3.6	40	37
C	0.50	52	135.5	3.3	38	41
D	1.00	32	137.5	3.2	35	43
E	2.00	18	139.6	3.0	30	47
F	3.00	12	141.9	2.8	25	51
G	25.00	1.5	163.4	6.0	20	27

Table 2: Twisted loops (cf. Fig. 9)

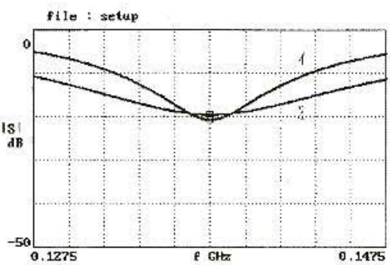
Fig. 18 Matching simulation, S11 = return loss



a. two series circuits (1, 2) with Q = 10, Resonance resistance 50 Ω and tuning difference of ±5%, and parallel-wired (3)



b. as above but with quality factor Q=5 (1) and Q=20 (2) with R_S=50 Ω, parallel circuit



c. as above but with R_S=25Ω (1) or R_S=100Ω (2), which gives Q 20 or 5

generated using PUFF simulation. If the resonance resistances do not coincide with Zo, then the curves from Fig. 18b. If the reciprocal de-tuning is too high or too low for the correct resonance resistance, then the curves look like Fig. 18c. The dimensioning is thus not critical in relation to reflection attenuation, i.e. ripple (a_r = 9.54 dB corresponds to VSWR = 2). However, the correct quadrature displacement is present only in the first case; all other cases cause elliptical deformation of the polarisation. But the same effect is also caused by the amplitude change which accompanies every de-tuning from the resonance curve maximum. Since the frequency range for the orbiting satellites covers only 1 MHz (0.73%), the error this leads to is negligible.

7.

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Carsten Vieland, DJ4GC

Tracking Generator for Microwave Ranges (1.7 to 13 GHz)

This article is part of a series, the other parts of the series will be published in issue 1/2001 of VHF Communications. This part of the equipment description includes the microwave ranges which are otherwise accessible only with commercial kit. Unfortunately, the processing of such a broad signal band can no longer be done entirely with low cost components. But the result provides a high-quality measurement assembly, giving laboratory quality operation from low frequencies to beyond the X band.

1.

Concept of UHF-VHF oscillator

The enhanced HP 8569(A) spectrum analyser forms the receiver section of a set of apparatus for network analysis. It can be used to control the frequency sequences, the range selection and the image display. Although the equipment dates back to as long ago as the first half of the eighties, all its good features can not really be developed in a DIY apparatus. However, other analysers (including those from various manufacturers) follow similar frequency concepts, so that the tracking oscillator presented here should also be adaptable for them. This network analysis process offers very great advantages over the classic wobble measurement.

The spectrum analyser has available an internal, frequency-specific local oscillator, this signal is available for external use on an SMA socket at the rear of the equipment. The frequency range of this YIG oscillator lies between 2 and 4.46 GHz; the level is maintained constant at 8 dBm. To create a generator signal on the analyser reception frequency, a fixed frequency at the level of the first intermediate frequency (321.4 MHz) is mixed in (Fig 1). Above approximately 4 GHz, it is really the various harmonics of this oscillator, arising during the mixing process, which take over the LO function during the mixing. An electronically tuned tracking filter, using YIG technology, cleans the other "signal waste" from the mixed product. A multi-stage broad-band amplifier raises the level of the filtered signal to a maximal 10 mW. The generator signal passes through a broad-band directional coupler which has a broad-band Schottky detector at its measurement output, this enables a constant output is maintained. A PIN attenuator adjusts the level of the intermediate frequency injection at 321.4 MHz. The automatic

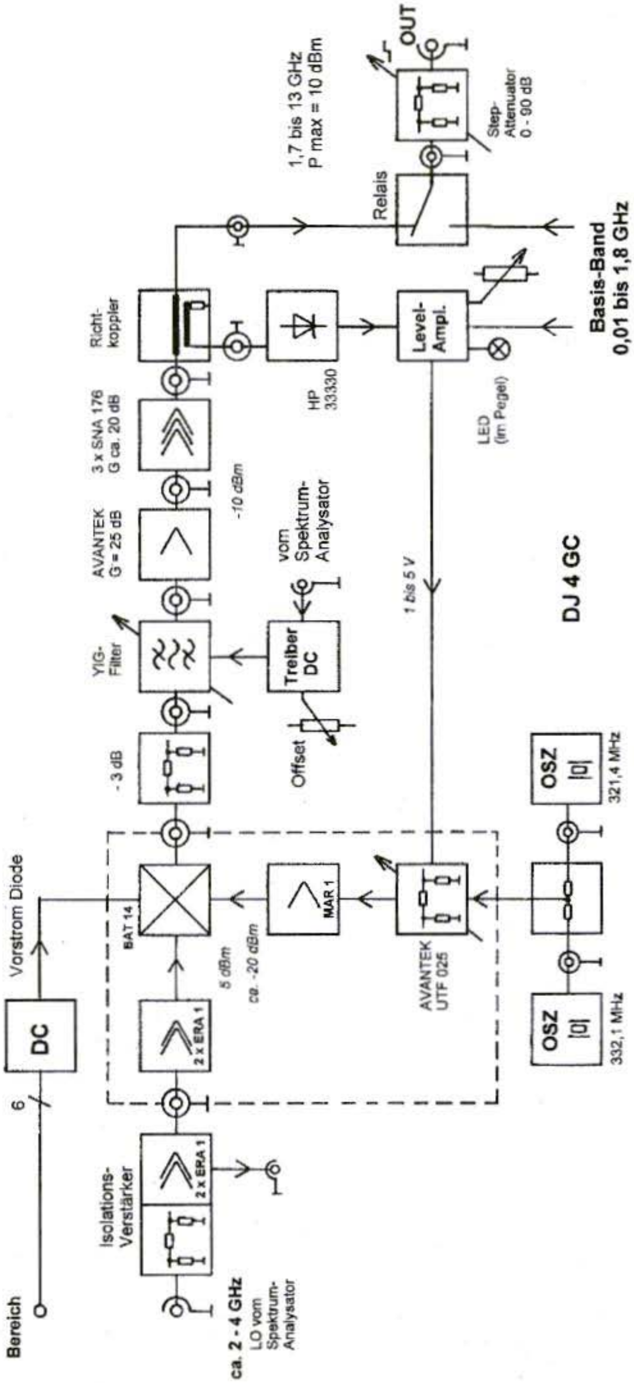

Fig. 1: VHF tracking generator concept

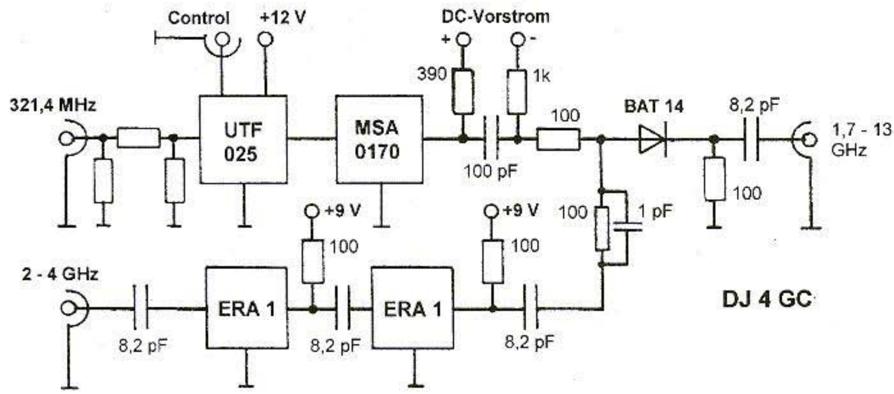

Fig. 2: Harmonic mixer with LO amplifier and level controller

level control, the buffer amplifier [1], and also the step attenuator at the output are used in common with the frequency synthesising of the basic band (10 MHz 1.8 GHz).

2.

Technical data of microwave generator

Frequency range of UHF-VHF section: 1.7 to 13 (15) GHz

Output level of generator: +10 dBm to 5 dBm, directly variable, plus 10 dB steps to 90 dBc

Amplitude ripple of generator: max ±1 dB in range from 1.7 to 13 GHz

Dynamic range of measuring system: min. 70 dB with 1 MHz reception band width, with reduced band width considerably more

Frequency noise: mainly dependent on LO of analyser; here so low that down to 100 Hz band width can be measured over the entire range

Output SWR of generator approximately 1.2 (return loss 20 dB) with 10 dB attenuation

3.

The assemblies

3.1. Harmonic mixer

The HP 8569 spectrum analyser goes up to 22 GHz in 5 frequency ranges which overlap each other to a large extent. In addition to the basic band processing from 10 MHz to 1.8 GHz, described in another part of this series, this home-made tracking generator covers the microwave range from 1.7 to 13 GHz without changing any high-frequency components. The microwave signal is processed separately from the basic band. From 1.7 to 4.1 GHz it is mixed with the fundamental frequency of the

LO signal which the analyser supplies. The second range (3.8 to 8.5 GHz) is produced by the mixing the first harmonic of the LO with the injected first intermediate frequency of 321.4 MHz. The third range (5.8 to 13 GHz) uses the third harmonic of the LO frequency as well as the intermediate-frequency generator at 321.4 MHz. At reduced amplitude, the mixed products reach as far as approximately 15 GHz before the YIG filter fails due to absence of current.

Attempts at signal processing with broad-band ring mixers have failed, since the even-numbered harmonics of the LO are severely suppressed, due to their symmetrical structure. The best mixed results were obtained using a half wave diode mixer with an individual microwave diode, a BAT 14 from Siemens (Fig 2). The required microwave harmonic in the diode current is optimised in each case using a DC biasing current which is adjusted, depending on selected range, to select the correct operating point of the diode. For each frequency range, the HP 8569 supplies a switch signal to a socket strip at the rear. From this signal the optimal diode biasing current can be obtained for the degree of multiplication in question (Fig 6). The adjustment is really critical, but can be carried out and optimised directly on the analyser screen.

The output spectrum after the mixing diode resembles a boiling inferno of harmonics and mixed products which makes it difficult to maintain a constant output level at the desired frequency. The spectrum analyser, with its internal YIG filter, has excellent selection characteristics, this technique could be used to help us get over the marked variations in level, but this requires considerable expenditure. A highly-selective YIG filter, locked into the mixing

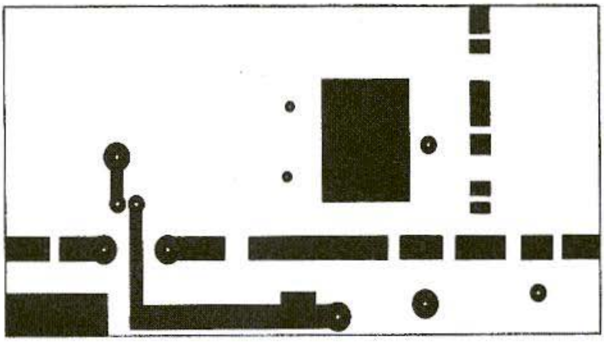

Fig. 3a: Harmonic mixer PCB layout. True size 71 mm x 35 mm

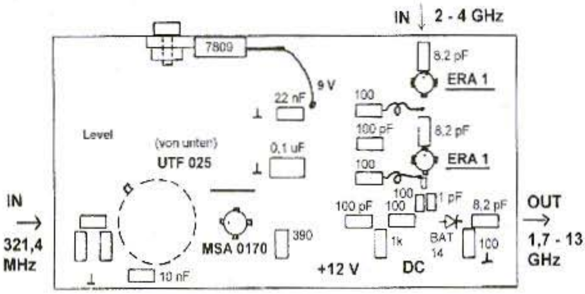

Fig. 3b: Harmonic mixer component layout. The PIN diode controller is on the earth side

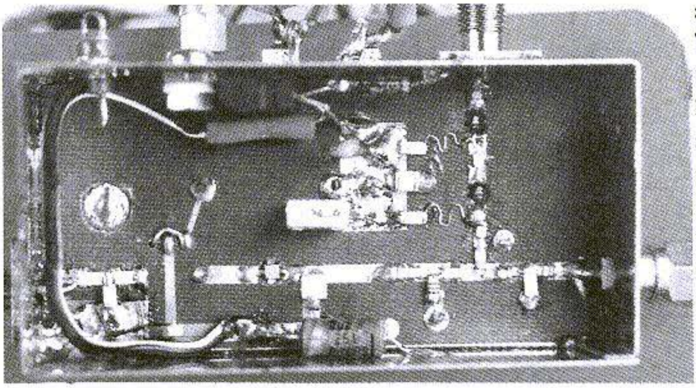

Fig. 4: Practical assembly of mixer. The output impedance of the 2-4 GHz LO amplifiers are supplemented by small additional inductances. These increase the amplification somewhat at the top end of the band.

frequency in question cleans up the signals before broad-band amplification and amplitude levelling take place.

Like a ring mixer, a half wave diode mixer also requires a broad-band termination, as real as possible. A matching amplifier would certainly improve the signal-to-noise ratio in the output spectrum, but is ruled out here because the LO bandwidth (2-4 GHz) is too high. Although further reducing the already limited mixed products after the diode (right down as far as -30 dBm) generates almost physical pain, a compulsory addition, in the form of an attenuator of at least 3 dB, must be inserted between the harmonic mixer and the YIG filter. Otherwise, due to reflections at the filter input, selective level changes, which can no longer be fully compensated, occur outside the transmission frequency.

3.2.YIG filter with driver

Both the main oscillator of the analyser and its pre-selector, as well as the locked-in filter in the tracking generator, are assembled using so-called YIG technology. The actual resonator is formed by tiny little balls (<1 mm, see Fig. 9) made from a mixture of the elements yttrium and iron, plus the semi-precious stone garnet. This mixture, which seems to have emerged from an alchemists

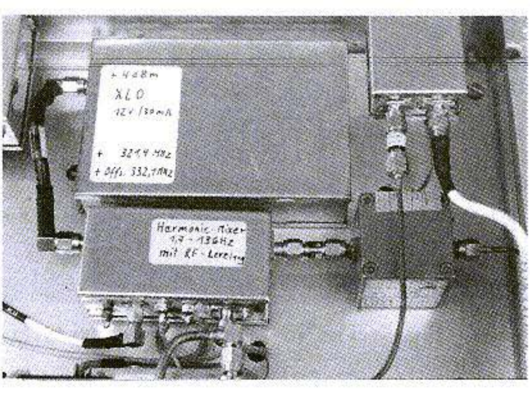

Fig. 5: Section of complete assembly with double XLO, harmonic mixer, 3 dB attenuator and YIG filter. The buffer amplifier for the 2-4 GHz LO is at the right-hand top edge of the illustration. More details of the layout of the two broad-band amplifiers will be in subsequent articles.

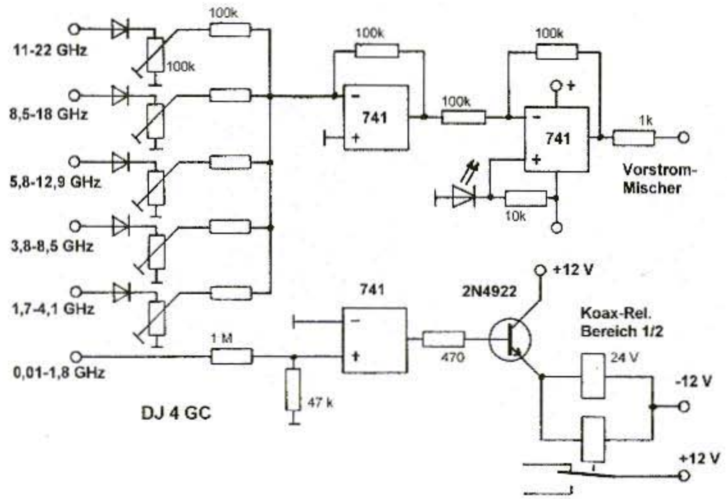

Fig. 6: Automatic biasing current generation for the mixer diode with the help of the turn-on voltages from the spectrum analyser. For range 1 (basic band up to 1.8 GHz), a coax relay switches the genertaor source.

kitchen, displays an inherent resonance in the microwave range when it is immersed in a magnetic field, this can be utilised by using a coupling loop. The level of the tuning current determines the resonance frequency by changing of the strength of the magnetic field. The filters, mainly two-circuit or four-circuit, combine high resonance quality (>1000) with very good isolation (up to 100 dB), a constant insertion loss (here approximately 4 dB) and massive tuning range. Measurements can be made from 1.7 to 15 GHz without any switchover, i.e. merely by altering the tuning current, thus for a first intermediate frequency at only 321.4 MHz, the mirror frequency attenuation will always exceed 80 dB.

The tracking pre-selector in the spectrum analyser generates the required tuning voltage from 1.7 to 22 GHz(!). The tuning voltage for driving the magnetising current is accessible at a socket strip on the rear of the analyser. The control voltage varies between 1.7 and approximately 15 V. Since a markedly linear relationship exists between the tuning voltage and the resonance frequency (1V/GHz), we need only an amplifier circuit which is able to convert the tuning voltage into a proportional current of up to 600 mA (Fig 8). The V-I converter developed for this measures the coil current indirectly as a voltage drop in a low-Ohmic wire

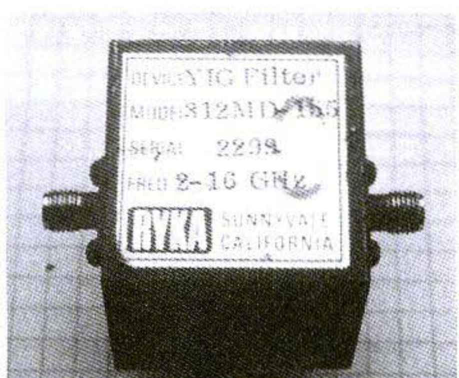

Fig. 7: The YIG filter for 2 to 16 GHz

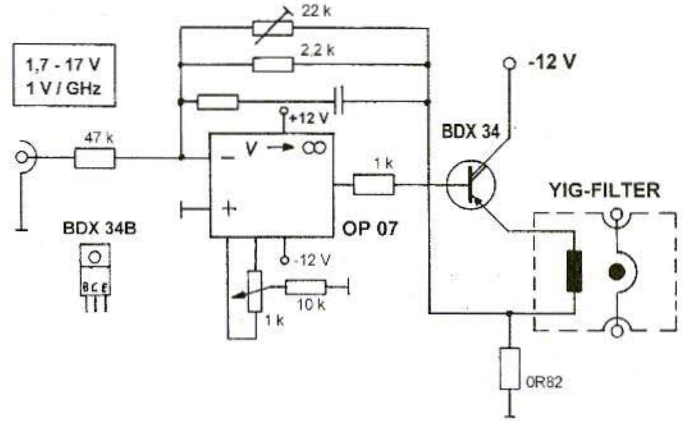

Fig. 8: Driver circuit for YIG filter

resistor, and feeds this voltage to the inverting input of a differential amplifier. Thanks to this little trick, the tuning current is not affected by either the higher temperature correction value, the marked heating-up of the copper coil, or their inductive effects. The V-I converter thus forces a linear tuning current through the YIG coil.

To de-couple the remaining assembly stages, the relatively high tuning current is drawn from the negative supply voltage. A PNP Darlington transistor (here BDX 34) functions as a load controller. Due to its high current amplification, it can be controlled directly by an operational amplifier. Because the YIG filter used here has an extremely narrow band, it is helpful to retain a small, externally adjustable offset variation. Despite the fact that the effect is only marginal, the use of a ten turn potentiometer is advantageous. YIG components should be mounted on a heat sink, or at least on a solid chassis plate, due to their high level of internal heating.

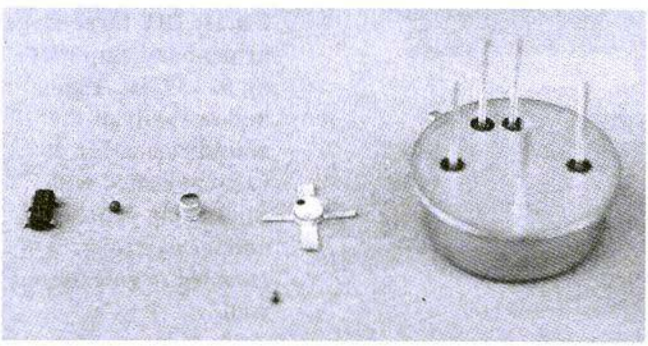

Fig. 9: Small component collection, from left to right: BAR 61 triple PIN diode, dismantled YIG resonator element, BAT 14 microwave diode, MMIC SNA 176 broad-band amplifier, together with UTF 025 PIN diode controller

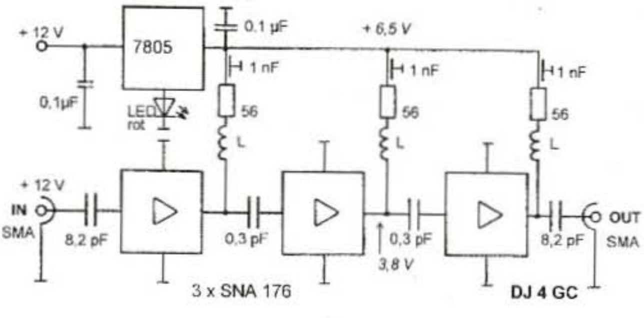

Fig.10: Circuit of second broad-band amplifier with 3 SNA 176s. In order not to overload the voltage reulator (I=50 mA!), the supply voltage was reduced to 6.5 V.

3.3. Broad-band amplification with level controller

The first of two amplifiers available was an Avantek broad-band amplifier (2-18 GHz) with G=25 dB. Mixing products can already be raised to above 10 dBm over the entire range using this commercial high-tech product. This level actually suffices for most measurements. The favourably-priced SNA 176 MMIC amplifier from STANFORD-MICRO-DEVICES, specified for use at up to 10 GHz, turned out to be an outstanding addition to improve the performance further. Three of these transistor-type SMD modules can be cascaded using very small coupling capacitors (0.3 pF) in such a way that an inverse frequency response takes place, giving a maximum gain of over 20 dB between 10 and 12 GHz and only then slowly decreases once more. If optimal matching is used, the amplification is approximately 10db at 2 GHz. This unconventional coupling can be used to offset at least part of the variations in the mixed attenuation between the basic and the harmonic mixing. With slight level compression, the

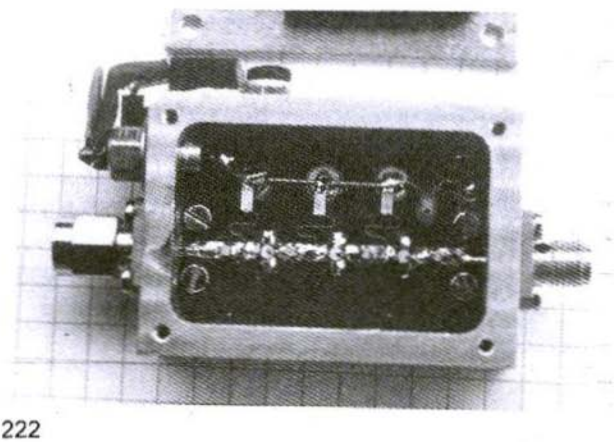

Fig.11: DIY three-stage broad-band amplifier up to 13 GHz. The hollow cavity of the amplifier housing is vacuum coated with conducting foam. The voltage controller is fastened to an external wall.

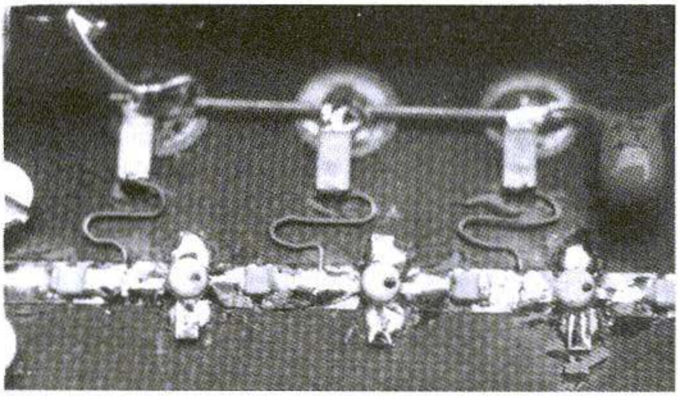

Fig.12: Details of this amplifier. The lateral earth connections of the MMICs are soldered to full tubular rivets; this leads to short earth paths and favourable heat conducting characteristics.

desired 10 mW output can still be attained at up to 13 GHz. Naturally, these excellent low-cost modules could also be used to provide all the amplification required.

The main level regulation is carried out before the harmonic mixer, by means varying the quartz-stabilised intermediate-frequency injection signal at 321.4 MHz. A commercially available monolithic UTF 205 PIN diode controller from AVANTEK adjusts the level prior to the high-frequency mixer in such a way that the pre-selected generator output is maintained at the output socket. The PIN diode controller described in [1], with the BAR 61 triple diode (in emergency, 3 x BA 379) could likewise do the job in a suitably modified printed circuit board design. This solution would be considerably more advantageous in terms of price, but would also take up a larger area than the commercial thick-film circuit in the TO 5 transistor housing. The OP circuit described in the first section is also used, as a control amplifier.

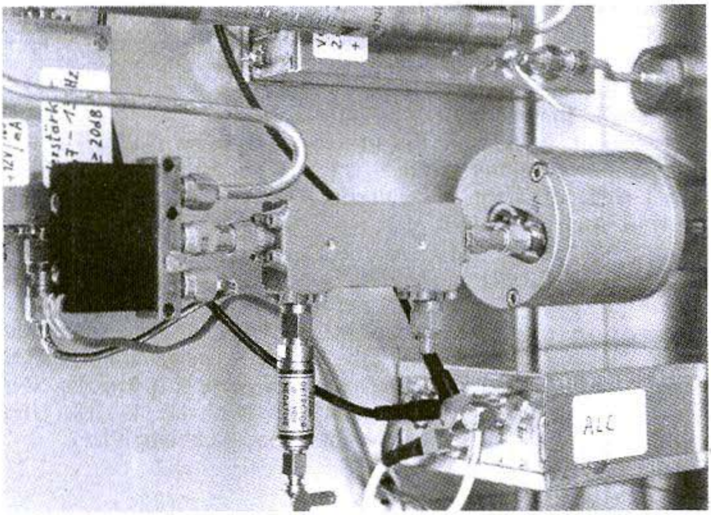

Fig.13: Section with coax relays, broad-band directional coupler, plus detector for level control (ALC), together with switchable attenuator in round metal housing.

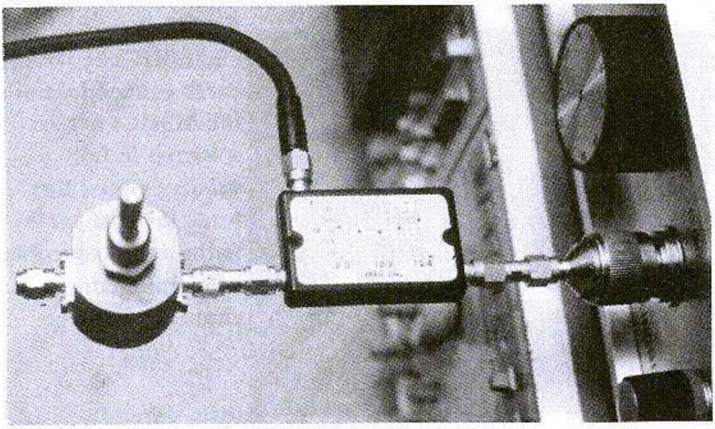

Fig.14: Matching measurement on an X-Band-Resonator brilliantly manufactured by DJ0PQ.

The actual value of the output level is established by a NARDA directional coupler (2-18 GHz, d=10 db, SMA connections), together with an HP 33330 B 50-Ohm detector. The level measured at the output socket (i.e. behind the step attenuator) using a thermal wattmeter are thus less than ± 1 dB from the set value over the entire range (1.7-13 GHz). Such good results can be obtained only if individual high-quality commercial components are used (such as directional couplers, detectors, Yig filters, etc.).

For measuring mixers, there is also a second intermediate-frequency oscillator available for the microwave range (321.4 MHz to 332.1 MHz). Depending on the apparatus, the offset variation of the YIG pre-selector in the spectrum analyser could be too small, so that it has to be expanded when it is integreated with the basic equipment. However, an oscillator offset reduced to, for example, approximately 3 MHz, could also solve the problem.

3.4. Commercial components

Numerous mishaps lie in wait for a generator project with a bandwidth of this order of magnitude. The noise level and, above all, the output control can not be any better than the components determining the level. My own experience suggests that you should not really try and conjure something up out of nothing and build as many components as possible yourself. Many inconspicuous looking little components contain so much unobtainable know-how that the entire project could rapidly turn into a permanent construction site.

The commercial high-frequency components used in the equipment described are listed below are mainly taken from surplus stock. Without them, it would have been considerably more difficult or even impossible to assemble the equipment. At all events, they contribute to the high quality of the finished apparatus.

Broad-band amplifier 2-18 GHz with G = 25 dB in housing with SMA connections AVANTEK AWT 18235

MMIC amplifier - monolithic SMD amplifier module with MAR and ERA range MINI-CIRCUITS wire terminating tabs or SNA range from STANFORD Microdevices, in particular SNA 176 (can be used up to 13 GHz, good-value broad-band amplifier)

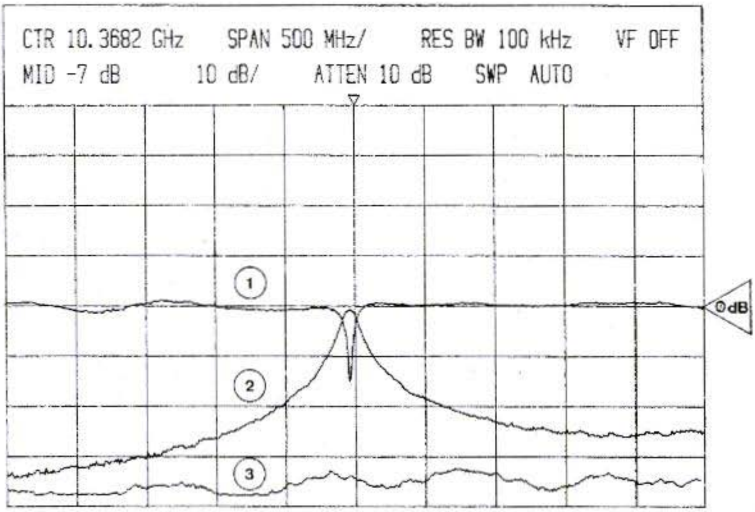

Fig.15: Measurement results for resonator from Fig.14. In the centre of this 5 GHz wide (!) frequency section lies the X-Band frequency of 10368 MHz. All measurement curves have been standardised to the screen mean (0 dBc).

1) Matching measurement using Wiltron coupler from Fig.17. In the non-optimised condition, the return loss amounts to 15 dB.

2) Advance measurement with approximately 1 dB transmission loss, together with remote selection to ±2.5 GHz.

3) Sharpness of directivity of measurement system with precision dummy on Wiltron bridge

Directional coupler 2-18 GHz with 10 dB uncoupling attenuation, NARDA 4203-10 SMA connections (for amplitude leveling)

Broad-band detector with Low-Barrier Schottky diode, 0.01-18 GHz, HP 33330 B APC 3.5 connection (SMA-compatible) (amplitude leveling)

YIG filter (microwave filter with electronically variable frequency) 2-16 GHz, made from old pre-selectors, generator systems, radar apparatus, etc., various manufacturers, for example AVANTEK, WJ, HP, OMNI-YIG, SIVERS, SYSTRON, VARIAN, YIG-TEK etc.

STEP attenuator, switchable attenuator in 10 dB steps, DC-18 GHz , various manufacturers (here WEINSCHEL AF 9010)

PIN controller, electronically controllable variable attenuator, thick film technology within the dimensions of a TO5 transistor housing (had one available, but good-value substitutes also conceivable) AVANTEK UTF 025

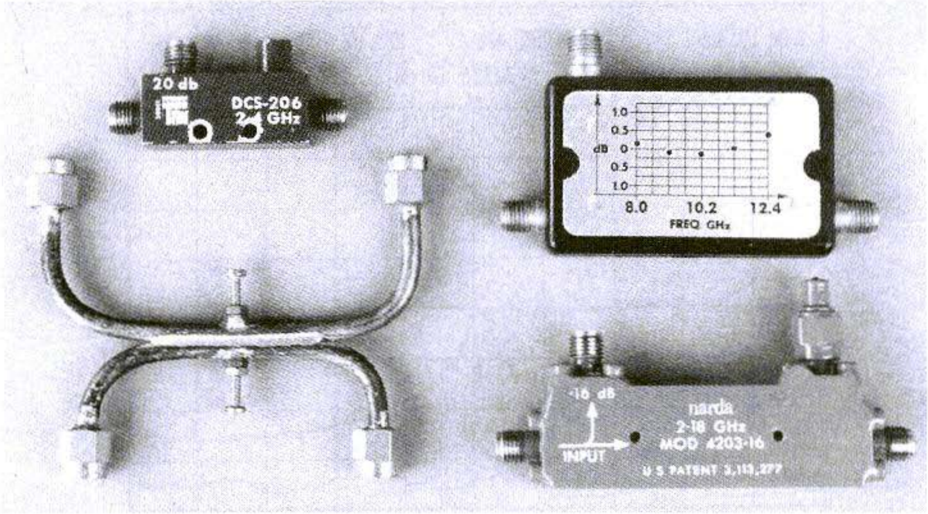

Fig. 16: A small selection from the options of directional couplers or adapter measuring bridges available

Additional small components such as SMA attenuators and a coax relay are the last things to go into the shopping basket. The list is the result of a pretty random search through electronics markets and second-hand dealers. There are undoubtedly many components from other manufacturers of comparable quality. While most individual components could be obtained relatively quickly and at reasonable prices, purchasing a broadband amplifier was a wearisome affair and it cost rather a lot of money. Actually, a high-quality DIY solution has appeared on the scene since then, the advantageously-priced SNA 176 MMICs from STANFORD Microdevices.

3.5. Directional couplers for the microwave ranges

For an efficient network analysis, directional couplers or adapter measuring bridges are required to determine the return loss within a 50-Ohm system. Fig. 16 shows a small selection from the options. Usually, the lower the band width, the greater the sharpness of directivity is. The extremely efficient DIY concept using semi-rigid circuits was described in greater detail in [2]. The SMA broad-band couplers (NARDA, KRYTTAR, OMNI-SPEKTRA etc.) not exactly cheap even in a surplus market - with band widths of 2-18 GHz or even more, are usually optimised towards constant advance decoupling for level controllers, which means the sharpness of directivity is of rather less importance. The coupler used in this tracking generator for power control (NARDA 2-18 GHz) continuously displays a sharpness of directivity exceeding 20 dB at up to 13 GHz.

Matching can take place over a particularly broad band with the help of measuring bridges. In principle, this refers to Wheatstone bridges suitable for high-frequency use, with a symmetry repeater for de-coupling in the bridge branch. The "queen" of measuring bridges, from WILTRON, is also structured according to this principle. The

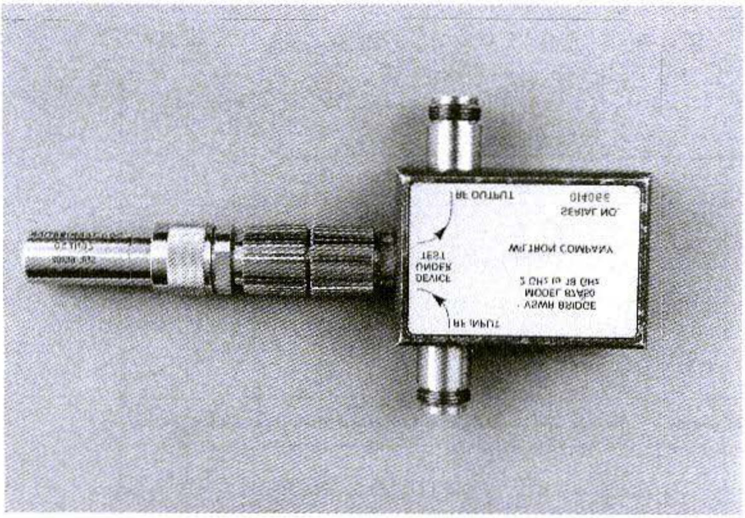

Fig.17: Wiltron measuring bridge with precision dummy as matching reference

measuring port is provided with a high-quality APC 7 connector, and must be correspondingly adapted in each case. Curve 3 in Fig. 15 gives a picture of its sharpness of directivity, which sets a new standard. The basic disadvantage of this (unfortunately scarce) component is its very high price.

3.6. LF transverter

The basic range of the spectrum analyser begins at 10 MHz. Below this frequency, loss of sensitivity and unwanted signals prevent any meaningful measurement. It is of little assistance here that the tracking generator described in another part of this article retains its level down to approximately 1 MHz. Unfortunately, this means that important frequency ranges such as low frequency, electronics applications, me-

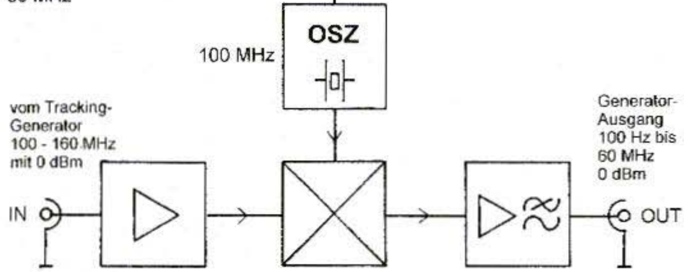

Fig.18: Simplified basic circuit of LF transverter

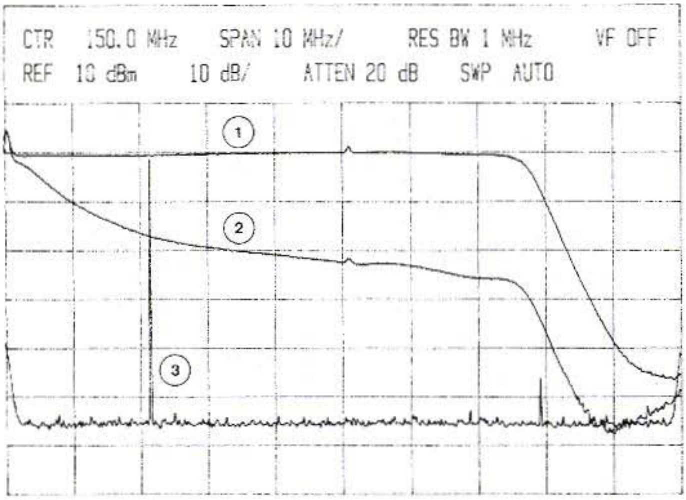

Fig.18: Efficiency of LF transverter from 0-100 MHz transposed to 100-200 MHz. The 100 MHz disruptive breakdowns recognisable at the left-hand and right-hand edges of the illustration are each represented as being as wide as the analyser bandwidth.

1) Frequency response record for entire measurement system. The tracking generator is coupled directly into the receiver input. From app. 75 MHz, strong level attenuation is noticeable as a result of the two low-pass filters. The ripple over the entire useful range from 100 Hz to 70 MHz is only approximately 1 dB.

2) Frequency response over high-ohmic impedance converter at app. half sensitivity. At maximum level setting, the curve is identical with curve1.

3) Noise level of transverter system, using example of a 21.4 MHz quartz filter.

dium wave or the lower short-wave range lie outside the width of the measurement system.

In order to close these unpleasant frequency gaps, a transverter has been developed which can go down as far as 100 Hz. In principle, both the analysis frequency and the generator frequency are displaced by exactly 100 MHz. As a result of this offset. the reception range on the analyser begins at 100 MHz. The tracking generator frequency is simultaneously mixed downwards by this amount. The 100-MHz quartz oscillator used jointly for both mixing processes allows the two frequencies to be equal.

The individual path within the apparatus from this oscillator to the mixer contains both high amplifications and good isolation. There should be at least 80 dB isolation between the two mixers, so that the tracking does not sneak around the test object on this path. The two 3-stage low-pass filters with a 3-dB limiting frequency, each of approximately 60 MHz, prevent the emergence of the 100 MHz oscillator signal, together with signal ambiguities resulting from image frequency problems and harmonics.

For simple level determination, the recieve converter is adjusted to 0dB transmission amplification. Since in the

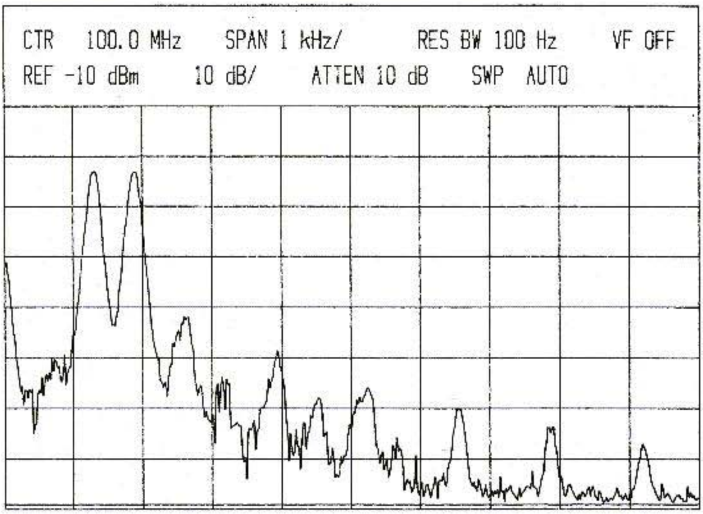

Fig.20: Spectrum analysis of a low-frequency two-tone (1.3 and 1.8 kHz) through transverter.

low-frequency and electronics ranges many signal sources are high impedance, i.e. may not be loaded with 50 Ohms, a high impedance converter with a sensitivity potentiometer is incorporated. The circuit, all in all is really extensive but otherwise conventional, is shown here only in outline, i.e. not going down to details of the solution. There are numerous alternative layouts, only suggestions for the creation of which are given here.

Even if the narrow-band analysis bandwidth of only 100 Hz is not optimal for the investigation of very low-frequency signals, the results are perfectly observable, even in the low-frequency range. Fig. 20 shows the spectrum representation of the low-frequency two-frequency generator, with which the author usually determines the inter-modulation products of transmitters and final stages. Each of the two tones on the left-hand edge of the image stems from a Wien bridge generator and gives a perfect sinusoidal optical display on the oscilloscope. Spectrum analysis shows that the interval between the lower of the two tones (1.3 kHz) and the first harmonic (d2) is a mere 30 dB. The harmonic

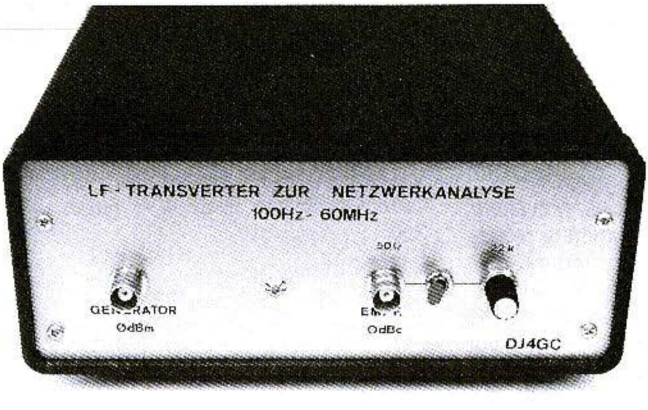

Fig.21: Complete low-frequency transverter.

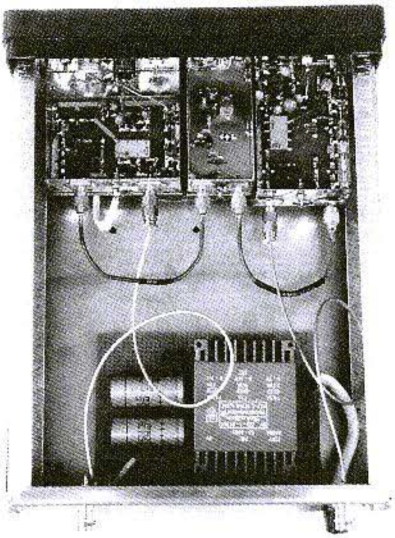

Fig.22: Internal structure with opened tinplate housings.

with the sevenfold frequency can also be recognised on the right-hand edge of the image, and the interval between it and the fundamental oscillation is approximately 55 dB. The rather unexpectedly poor harmonic interval of this generator, however, has only a slight significance in practical measurement, since from 3 kHz at the latest the transmission side filter forces the band to be restricted. The dreaded intermodulation products are mainly outside these filter bandwidths.

4.

Summing-up

By means of the accessories presented here (tracking generator, matching measuring bridge and LF transverter), the spectrum analyser, actually developed for the microwave range, is expanded into an efficient network analyser. The frequency range extends from the relevant range of FFT analysis with the sound card of a computer all the way to the wave guide bands. Including some quite long breaks, the project (excluding the LF transverter) lay on the hobby table for just on a year before it was fully completed. The test rig has now been operating satisfactorily for two years. The not inconsiderable effort (in terms of thought, manual labour and also money) has been rewarded, at the very least, with the availability of a laboratory-grade high-quality measuring system. Its really great when you don't just have to assume that something youve developed yourself works, but you can get the results displayed on the screen through a test rig with, as it were, infinite bandwidth and dynamism, while the plotter is already producing a hard copy.

5.

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Gerhard Schmitt, DJ5AP

Low Pass Filter for 2 m and 70 cm - Part 2

The first part of this article concluded, at the end of Chapter 6, with the construction of a successful 145 MHz filter. The article continues with a 435 MHz filter and filters designed to handle lower power levels.

7. Low-pass filter for 435 MHz

A highly efficient low-pass filter was also developed for the 70-cm band; the wiring diagram again corresponds to Fig. 3b. Since the capacitors have lower values, RT-Duroid 0.79 mm thick was selected as the printed circuit board material. This also produces a somewhat higher level of mechanical stability, since the printed circuit board material with Teflon dielectric, in contrast to the epoxy material, is very soft.

For C1 and C4, the values are 12.5 pF each, for C2 and C3 17 pF each. Using RT-Duroid 0.,79 mm. thick, the required dimensions for C1 and C4 are 32 mm. x 12 mm. each, for C2 and C3 38.5 mm. x 15 mm.. The assembly also takes place in a tinplate housing with dimensions of 111 mm. x 55.5 mm. x 30 mm..

The inductances were produced not by coils or wire but by semi-circular pieces of copper or brass plate 0.1 to 0.2 mm thick. This makes assembly easier, and later adjustment.

The printed circuit board layout for the 70-cm. version on RT-Duroid 0.79 mm. thick can be seen in Fig.10, and the component drawing in Fig. 11. Fig. 12 shows a complete, fully-equipped low-pass filter for 435 MHz. In other respects, what was said in Chapter 6 regarding the assembly of the filter for 145 MHz applies, as does what was said regarding the output compatibility.

The frequency response of the low-pass filter for 435 MHz is seen in Fig. 13. At 870 MHz, the second harmonic, more than 40 dB attenuation is already being achieved. At the third harmonic, in the 23-cm. band, at approximately 1,270 MHz, it is 60 dB, a value which remains constant until approximately 1,800 MHz and then, at a level of 2,100 MHz, falls off to 50 dB. From approximately 3,400 MHz all the way to the limit of measurement at 6 GHz, the attenuation goes back to mean values of 20 dB. The low attenuation values from 3,400 MHz can be traced back to wave-guide effects in the tinplate housing. This effect, moreover, is also present, in a weakened form, in the low-pass filter for 145 MHz. If, as is usual for microwave

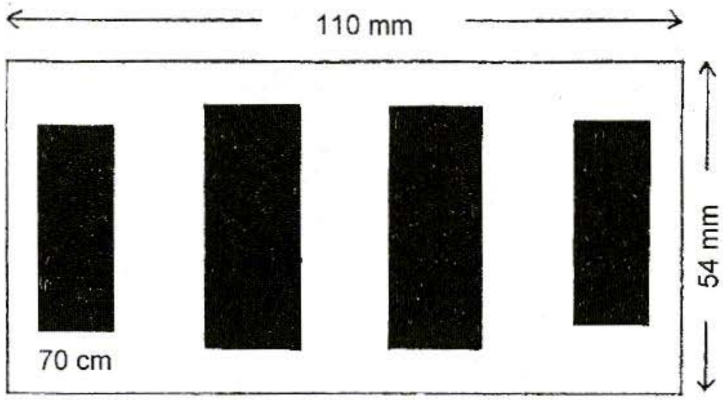

Fig 10: PCB Layout for the 435 MHz Filter using RT Duroid 0.79 mm thick

circuits at 10 GHz, conducting foam is glued into the lid, the undesirable wave-guide effect can be somewhat reduced. The attenuation values in the transmission range at 145 or 435 MHz are not altered by bringing absorber material in!

Attenuation in the 435 MHz transmission range is at a low 0.2 dB.

8. Low-pass filter for lower outputs

High transmission power levels of several hundred Watts, right up to the limit of 750 Watts laid down by the approval authorities, will come up rather seldom for most radio amateurs. For lower

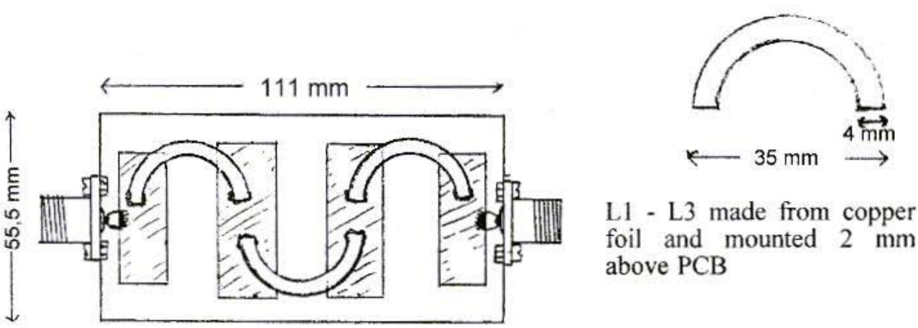

Fig 11: Component drawing for the 435 MHz Filter

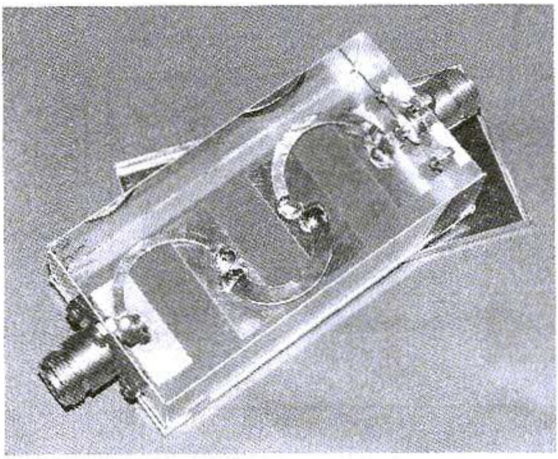

Fig 12: The completed 453 MHz Filter

power levels down to app. 50 Watts, low-pass filters can be made considerably smaller, using small ceramic capacitors. and with similarly good performance with respect to harmonic attenuation. The circuit can again be taken from Fig. 3b. The basic specifications for calculation can be found in [5].

Small ceramic disc capacitors with a diameter of about 5 mm. are used both for 2 m. and for 70 cm. The capacitors usually have an electric strength of 500 Volts. In no case should ceramic tubular capacitors be used, however small or voltage-stable they are!

The connection wires are shortened to a length of approximately 1.5 to 2 mm.. Every additional mm. reduces the attenuation value of the low-pass filter at higher frequencies.

Coil data of small low-pass filter for 145 MHz:

- L1, L3 silver-plated copper wire ∅ 1 mm. , wound on a 6 mm. drill shank, 9 mm. long

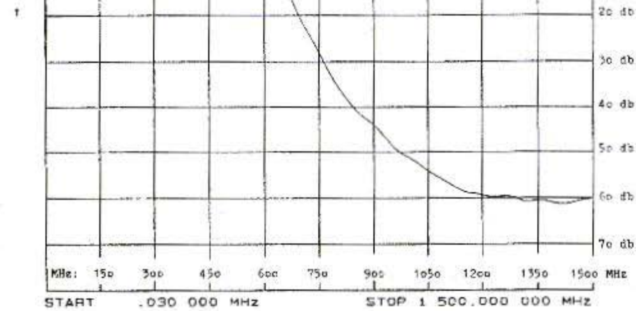

Fig 13: Frequency response of the 435 MHz Filter

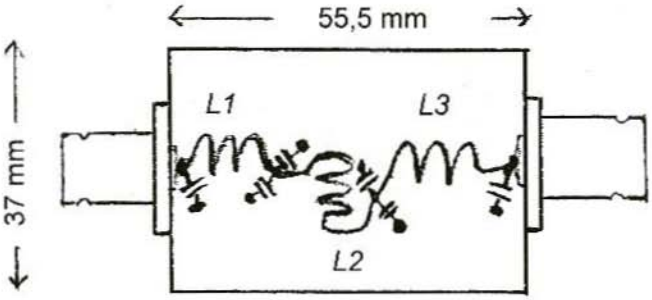

Fig 14: Component layout for the 145 MHz low power filter

- L2 as L1, but 11 mm.long

Small ceramic disc capacitors:

- C1, C4 = 27 pF
- C2, C3 = Two x 27 pF each parallel wired

The parallel wiring, which means that C2 and C3 each have 2 x 27 pF, for an effective capacity of 54 pF, leads to slight undesirable inductances, together with a relatively high current carrying capacity at high frequencies.

For an assembly sketch of the low-pass filter for 145 MHz, see Fig. 14, and for a completely assembled filter, see Fig. 15.

Coil data of filter for 435 MHz:

- L1 to L3 silver-plated copper wire ∅ 1 mm., 32 mm. long, semi-circular, bent as Fig. 17 Balance distance to earth surface to approximately 2 mm.!

Capacitors as described for 2-m. filter:

- C1, C4 = 10 pF
- C2, C3 = Two x 10 pF each parallel wired

For an assembly sketch of the low-pass

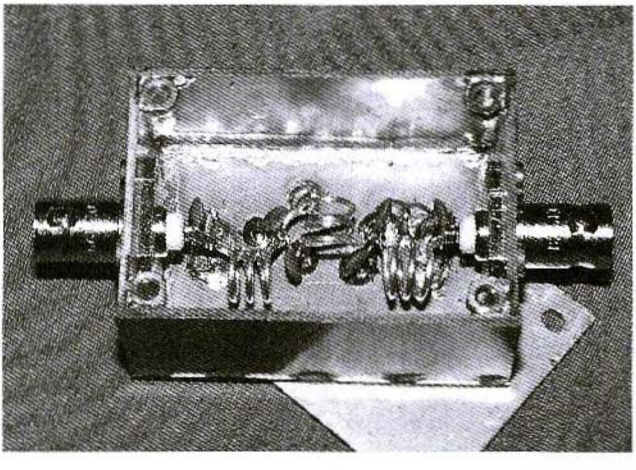

Fig 15: The completed 145 MHz low power filter

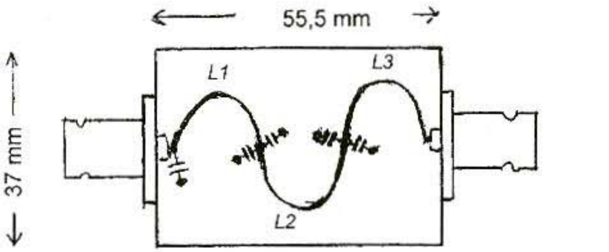

Fig 16: Component layout for the 435 MHz low power filter

filter for 435 MHz, see Fig. 16, and for a completely assembled filter, see Fig. 17.

Since there were actually no suitable tinplate housings available when the filter was assembled, the housings were built out of printed circuit board material or tinplate itself. The internal dimensions correspond to those of tinplate housings with the dimensions 55 mm. x 37 mm. and a height of 30 mm.

The completely assembled and "balanced" low-pass filters for 145 MHz and 435 MHz were each tested at 50 Watts HF. After fairly prolonged operation, the coils were warm to the touch, but the capacitors hardly heated up at all. The upper power limit was not checked out.

As an experiment, an additional low-pass filter was assembled for 145 MHz, using small ceramic high-voltage capacitors for 2-kV DC voltage and with a diameter of 12 mm., though only 22 pF was available. C1 and C4 were thus 22 pF, C2 and C3, after parallel wiring, 44 pF. The increased attenuation in the filter attenuation band was thus not as steep as with the filter described earlier, and the deterioration in attenuation from approximately 1,500 MHz was apparent.

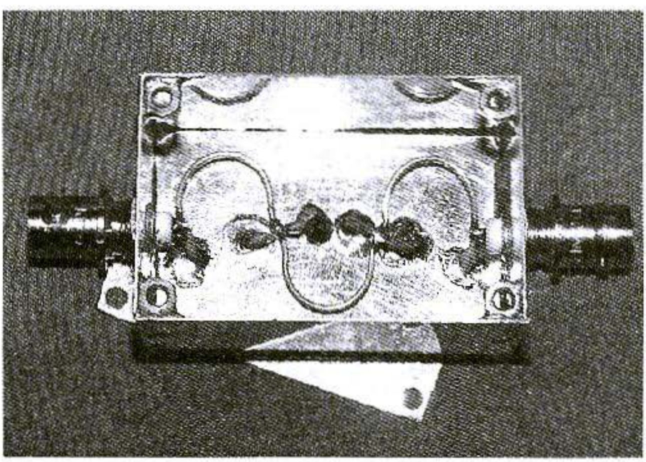

Fig 17: The completed 435 MHz low power filter

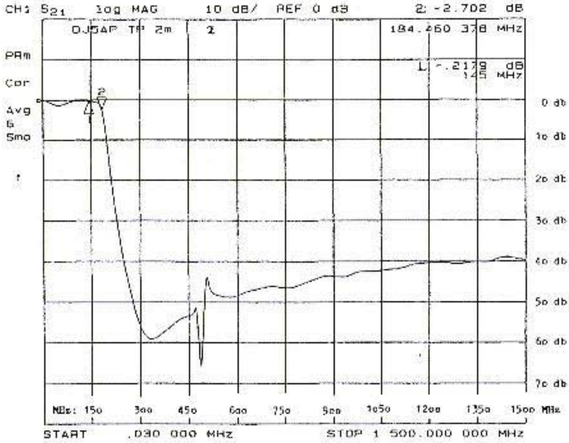

Fig 18 : Frequency response of the 145 MHz low power filter

On the other hand, a transmitter or a high-level stage for 145 MHz which is still producing significant harmonics at 1,500 MHz should just not be put into operation at all! The low-pass filter assembled with these larger capacitors withstood power outputs of 200 Watts without any problems. Here too, the limits were not tested.

The frequency response of the low-pass filter for 145 MHz, as described above, can be seen in Fig. 18, and for 435 MHz in Fig. 19.

In the low-pass filter for 145 MHz, the second and third harmonics, at 290 and 435 MHz respectively, are each attenuated by more than 50 dB, and all other harmonics up to approximately 1,500 MHz by 40 to 45 dB. Not until over 2,400 MHz does the attenuation fall below 30 dB.

In the low-pass filter for 435 MHz, assembled using small ceramic capacitors, the second harmonic, which lies at 870 MHz, is attenuated by approximately 65 dB, and the third harmonic in the 23-cm. band by over 50 dB. The additional attenuation decreases approximately in a straight line. At 2,400 MHz, it still amounts to 30 dB. Then from about 4 GHz, as also happens in small

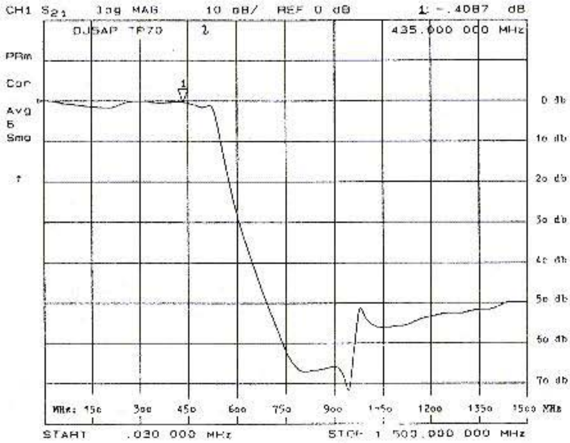

Fig 19 : Frequency response of the 435 MHz low power filter

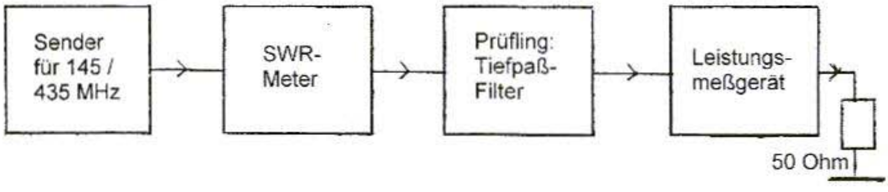

Fig 20 : Measurment set up for balancing filters

filters for 145 MHz, the attenuation goes back to close to 0 dB. The reasons for this may lie in resonance phenomena in the capacitors and coils even the smallest ceramic capacitor has an inductance! As a result of the cramped assembly, even crosstalk phenomena from the filter entrance to the exit, among other things, may play a part.

The attenuation in the transmission range of the small filter for 435 MHz is 0.4 dB.

The reasons for the steeper attenuation increase at the second harmonic lie in the greater capacity values of ceramic capacitors. The transmission loss in the transmission range is certainly again somewhat higher, thanks to the use of greater capacitances.

Since the larger low-pass filters should be usable for up to at least 750 Watts, compromises had to be made in designing the low-pass filters with capacitors made from RT-Duroid; the filters were optimised at the lowest transmission loss level in the transmission range.

It should also be noted that the filters were measured up to 6 GHz and the author does in fact have the readings, but they were of no further use here.

9.
Balancing of low-pass filter

The low-pass filter described can be well balanced using amateur resources.

You will need:

- A transmitter or transceiver for 2 m. / 70 cm., which emits a low-harmonic signal (at least 40 dB attenuation) with a suitable output. For example, modern FM transceivers with outputs of 30 to 50 Watts are very suitable.
- A good SWR meter, e.g. the EME Wattmeter or the older Wattmeter from Götting & Griem.
- An output and standing wave meter, a second EME Wattmeter, is also very suitable for use.
- A dummy load of suitable current-carrying capacity, with a 50-Ohm impedance. Failing this, for example, a long coax cable (e.g. 50-Meter RG 58/U) can be inserted into a smaller and/or poorer dummy load. Thanks to the high cable attenuation, the SWR of even a poor dummy load is very good!

In principle, the measurement rig is

shown in Fig. 20.

When the balancing process begins, the transmission power should be reduced to a few Watts, so as not to put the PA at risk. If the low-pass filter is pre-tuned, the transmission power can be increased, to obtain SWR values as precise as possible.

The tuning is carried out to minimise the standing wave ratio of the low-pass filter inserted between the SWR meter and the dummy load. The minimal SWR coincides with the lowest transmission loss in the transmission range of the low-pass filter.

The balancing procedure is as follows, using the example of the 2 m. filter with capacitors made from RT-Duroid:

First the standing wave ratio is measured at low transmission power. If the SWR is at 1 : 1.1 or better, no balancing is needed!

If the SWR exceeds 1 : 1.1, the low-pass filter should be optimised. First clarify whether the coils must be made bigger or smaller - the values for the capacitors made from printed circuit board material are already fixed. Drawing apart or compressing the coils made from 2 mm. thick copper wire is not recommended in the soldered condition there is a risk that the printed circuit board or the copper coating will be damaged.

A brass core is plunged into the coils (e.g. a brass screw with ∅ 4 mm., 20 mm. long) or an HF iron core with ∅ 4 mm.

When the brass core is plunged into the coils it reduces the inductance; plunging the HF iron core in increases the inductance. If, for example, it is established that the SWR improves when the brass core is plunged into one of the coils, the coils inductance is too high. It should be unsoldered, drawn apart a little and soldered back in. If necessary, the same thing is done to the other coils. If the SWR improves when the HF iron core is plunged in, the coils inductance is too low it should be compressed a little.

With a little patience, the balancing procedure described here can be rapidly completed. An SWR of 1 : 1.1 (return loss approximately 25 dB) can be achieved with all the low-pass filters described here.

For the 2 m. filter, at outputs of up to 50 Watts, balancing can be done somewhat faster. As described above, it is first established whether the coils must be made bigger or smaller. By using copper wire only 1 mm. thick, the coils can be drawn apart or compressed somewhat in the assembled condition; thus the objective can be achieved more rapidly.

With the 70- cm. filters, the balancing takes a very simple form. The coils, made from thin copper foil or 1 mm. thick copper wire, are bent towards or away from the earth surface using a plastic peg. Moving nearer the earth surface reduces the inductance and moving away from it increases it. No further balancing is needed.

10. Summary

The advantages and disadvantages of band-pass and low-pass filters for 145 and 435 MHz are clarified, the development of highly effective low-pass filters for 145 and 435 MHz described, and the assembly is explained in detail.

The values obtained for harmonic suppression, the low attenuation in the transmission range, and the high max. output of 750 Watts of filters with capacitors made from RT-Duroid mean the filters can be used in any application.

Retrofitting is recommended on older existing 2 m./70 cm. transmitters without a built-in harmonic filter. It is achieved by looping the filters described here in between the transmitter output and the antenna relay. When this is not possible, for reasons of space, the filter can be looped in between the antenna jack and cable.

Even with modern high-level stages, which are already equipped with a low-pass filter and obtain good values for harmonic suppression, one of the low-pass filters can be used as well. When transmitting at high power, above all, you can thus always be on the safe side.

To make assembly easier for less experienced radio amateurs, the construction and balancing of the low-pass filter are comprehensively described. Old hands are asked to be patient!

In Section 4, the development of the low-pass filter using printed circuit board material is described. Just so that no-one imagines the author discovered this Holy Grail : René, HB9MPU, remarked, at Ham-Radio 1999 during a visual QSO, that low-pass filters using capacitors made from Teflon printed circuit boards were also manufactured in the commercial sector, with outstanding results.

The author would like to thank various radio amateurs for their theoretical and practical assistance: Helmut, DJ8BF for the first measurements, Horst, DL3QJ for preparing printed circuit boards, Bernd, DJ6DC for the final measurements of the transmission curves, René, HB9MPU for the stimulating discussions, and the OM of the RegTP (callsign unfortunately unknown) at Ham-Radio 1999 for the measurements.

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Dipl.-Ing. J. v. Parpart

Active Directional Coupler

The subject of this article is a directional coupler, which works down to DC. The main signal path is passive. An electronic circuit is used to de-couple the forward or return signal.

1.

Introduction

In high-frequency technology, passive reciprocal directional couplers play a significant role. They are used to split or combine signals; in metering ciruits, they are used to measure matching and output.

In general, a directional coupler has four gates: the signal fed into gate 1 arrives at gate 2 almost unhindered. Gate 2 is the output which can be affected by reflection. At gate 3 a part of the forward signal can be available, at gate 4 part of the return signal is available.

Directional couplers usually consist of coupled circuits which are simultaneously capacitive and inductive, and also of circuits with discrete passive components (coils, capacitors, resistors, transformers, circuits). Broad-band transformer couplers are known from cable television technology. What all these formats have in common is that they have a lower operating frequency limit.

The subject of this article is a directional coupler, which works down to DC with a passive main signal path. A non-reciprocal electronic circuit is used to de-couple the forward or return signal.

This may be an alternative for many applications, as an example a video-coupler, which was developed by the author, is described in this article.

2.

Principle of active directional coupler

We shall first examine the universally matched active directional coupler with three gates. In accordance with their function, gates 1, 2 and 3 are called in, out and coupled. Depending on the direction of the power flow, either the forward or the return signal is at gate 3.

The path between gate 1 and gate 2 is passive and low-attenuation. A path (even more attenuation-free if applicable) should exist from gate 1 to gate 3 to produce a directional effect, whilst at the same time the attenuation from gate 2 to gate 3 with a matched load at gate 1 should be infinite.

How can this be achieved? It is conceivable in principle to connect gate 1 and gate 2 to each other directly, measure the voltage (V) and the current (I) at this point and balance them with the impedance level (Z_w). We thus obtain the source voltage of gate 3 (V_3), which is $V_3 = V + IZ_w$. It can easily be shown that if a matched source (original voltage V_q, internal resistance Z_w) is fed into gate1, V_3 always $= V_q$. If any source is fed into gate 2, and if gate1 is match-terminated, then $V_3 = 0$.

In practise, the current measurement is taken back to a voltage measurement at an ohmic resistance. There is thus a (small) resistance between gate1 and gate 2. In order to guarantee universal matching, this resistance is enlarged into a network (e.g. π structure or double-T structure). Thus, depending on the system, an attenuation arises ap, which is $a\text{p} = 20 \log x$. It will be understood that the attenuation between gate 1 and gate 2 should be independent of frequency and slight, i.e. the parameter x is real and lies very near to "1".

The electronic computing circuit, which provides the signal at gate 3, provides thoroughly welcome isolation in the reverse direction, i.e., there is no transmission from gate 3 to gate 1 or gate 2.

The reflection set out above can be mathematically formulated with the help of the scattering matrix (S). This gives us:

$$S = \begin{vmatrix} 0 & x & 0 \\ x & 0 & 0 \\ 1 & 0 & 0 \end{vmatrix}$$

3.

Specific examples

3.1 Example with precise matching

This example shows a resistance network in π structure (transfer impedance $R1$, transverse resistance on input side $R2$, transverse resistance on output side $R3 + R4$) universally matched.

In order to fulfil the computing rules, the voltage difference between the circuit input and the tap of the transverse resistance on the output side merely has to be amplified. An instrument amplifier serves to do this (high input resistance, low output resistance, defined amplification).

The instrument amplifier supplies the standard level at gate 3; the resistance $R5$ serves to carry out the matching at

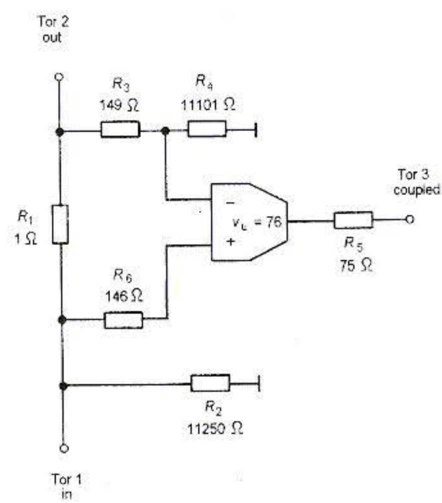

Fig. 1: Specific example with precise matching

gate 3. By adding resistance $R6$ it can be arranged that both inputs of the amplifier see the same impedance; thus the DC operating point will be independent of any (equal) offset currents. Fig. 1 shows the basic circuit, based on the assumption that Zw = 75Ω and x = 0.987.

If parameter x is selected freely, the Calculations are :

$$R_1 = Z_w \frac{1-x^2}{2x} \quad (1)$$

$$R_2 = Z_w \frac{1+x}{1-x} \quad (2)$$

$$R_3 = Z_w(1+x) \quad (3)$$

$$R_4 = Z_w \frac{x+x^2}{1-x} \quad (4)$$

$$R_5 = Z_w \quad (5)$$

$$R_6 = Z_w \frac{3x^2+2x-1}{2} \quad (6)$$

$$v_u = \frac{2}{1-x^2} \quad (7)$$

3.2 Example with slight input mismatch

If a slight mismatch at the input of the coupler can be tolerated, the circuit can be simplified, in that the resistance $R2$ (cf. Fig. 1) is omitted, and the network is re calculated. De-coupling requires:

$$\frac{R_3}{R_4} = \frac{R_1}{Z_W} \quad (8)$$

To be able to cascade several couplers, if applicable, the network output must be well matched. Exact matching can be obtained at gate 2, if we select:

$$R_1 R_4 = Z_w^2 \quad (9)$$

Fig. 2 represents a simplified circuit with Z_W = 75 Ω.

4. Active directional coupler in video technology

4.1 Use in video splitting technology

Video components (sources. sinks. circuits) display a uniform system impedance level (75Ω).

Should one source supply several consumers, the so-called looping-through technology has been used for decades.

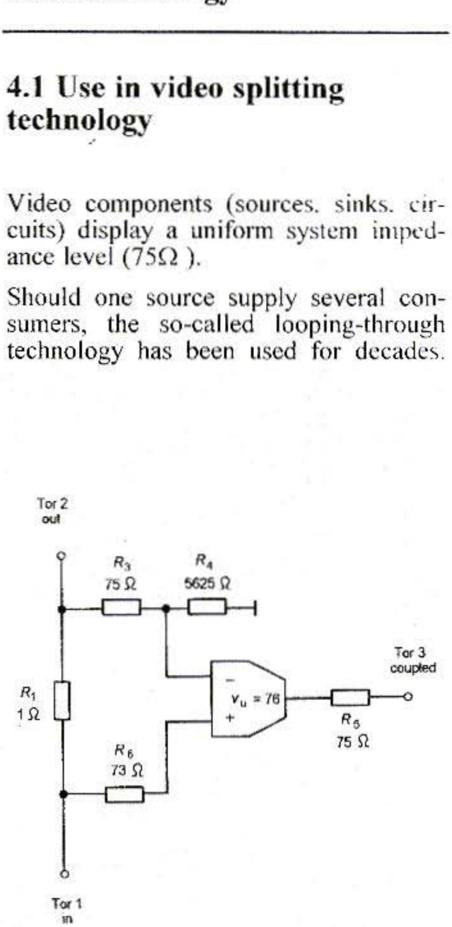

Fig. 2: Simplified circuit

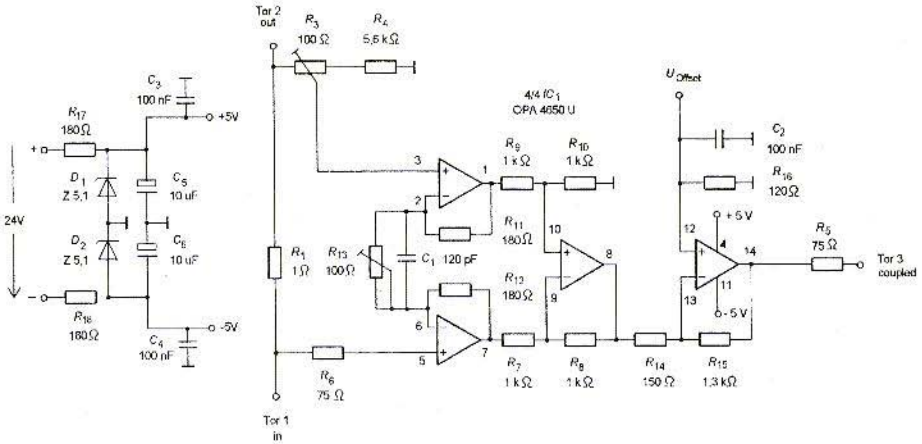

Fig. 3: Video coupler circuit diagram

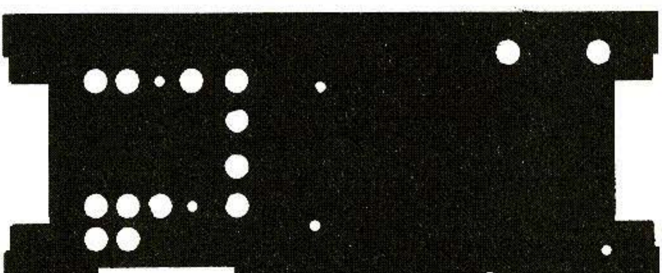

Fig. 4a: Video coupler PCB layout

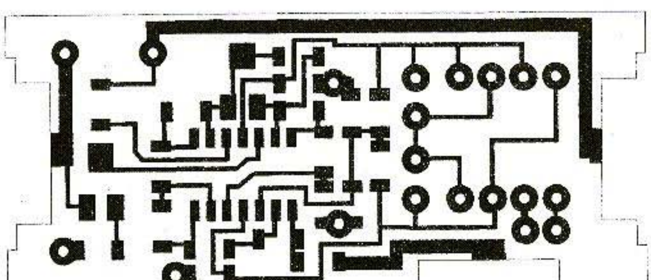

Fig. 4b: Video coupler PCB layout - tack side

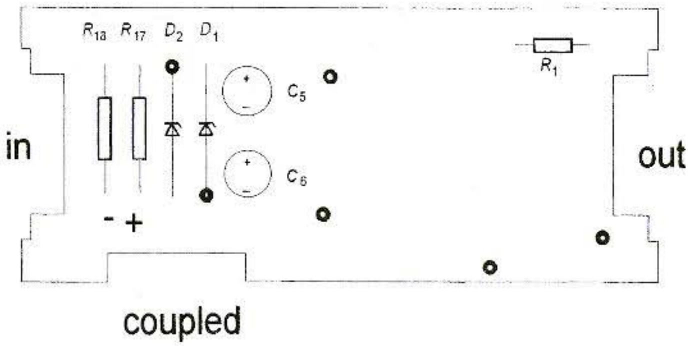

Fig. 5a: Video coupler components layout - wired component side

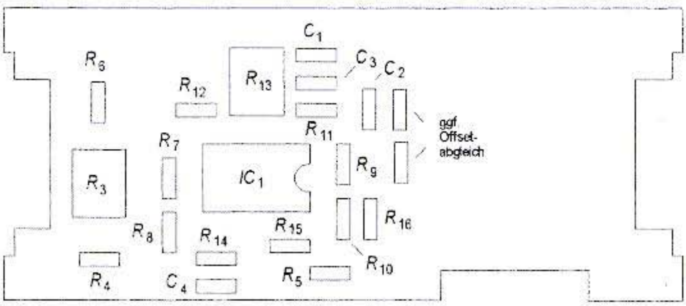

Fig. 5a: Video coupler components layout - SMD component side

The source powers a circuit which is match-terminated. Where a consumer is located, the circuit is tapped, i.e. the signal on the internal conductor is extracted at high resistance as in voltage measurement.

It makes for better operational reliability if the main signal path of the splitting system is passive. If the electronics of a consumer die, this has no consequences for the other participants. This is an advantage which should not be lost unless it has to be.

However, the assumption that every consumer receives the signal allocated to it is justified only if the splitting system is reflection free. An extreme case occurs if the main signal path is interrupted or short-circuited: all consumers now receive the wrong level either double voltage or nothing at all.

It is better if the looping-through technology is replaced by cascaded active directional couplers. An open circuit or a short-circuit on the main signal path will then (in the ideal case) have no influence on the signals de-coupled before the fault location. The video splitting system now functions like the well-known, tried and true passive CATV splitting system.

4.2 Circuit

The circuit (Fig. 3) consists of the resistance network formed by *R*1, *R*3 and *R*4 (s. Section 3.2) together with a discretely structured instrument amplifier (*v*u 8.7) with a subsequent amplification stage (once again *v*u 8.7). Fig. 4 shows the layout, Fig. 5 the components plan.

The signal paths run on the underside of a double-sided, 1.5-mm. epoxy resin board; the top side forms the earth surface. The SMD components are fitted on the bottom side, the wired components come on the top side. Six through platings are necessary (represented by circles in Fig. 5). The printed circuit board is housed in a commercially available in-line module housing.

The quadruple operational amplifier selected (Burr-Brown OPA 4650) is pin-compatible with other known types and behaves in an outstandingly good-natured way and does not tend to oscillate. The OPA 4650 is powered at +/- 5 Volts; it requires approximately 35 mA. A floating DC voltage was available (24 Volts, power supply); with the help of the Zener diodes (*D*1, *D*2), the bipolar supply voltage can be generated from this in a simple manner.

To make the common mode rejection optimal, the resistances *R*7, *R*8, *R*9 and *R*10 must be as identical as possible. A selection must be made; the pairing should be better than 0.1 %, the absolute value is not decisive (note: to improve the high-frequency common mode rejection, it can be advantageous to add a small capacity from pin 9 of the operational amplifier to earth).

The capacitor *C*1serves to linearise the frequency response. The normal video level is available at gate 3 (1 Vss at 75 Ohms); no external driver is needed.

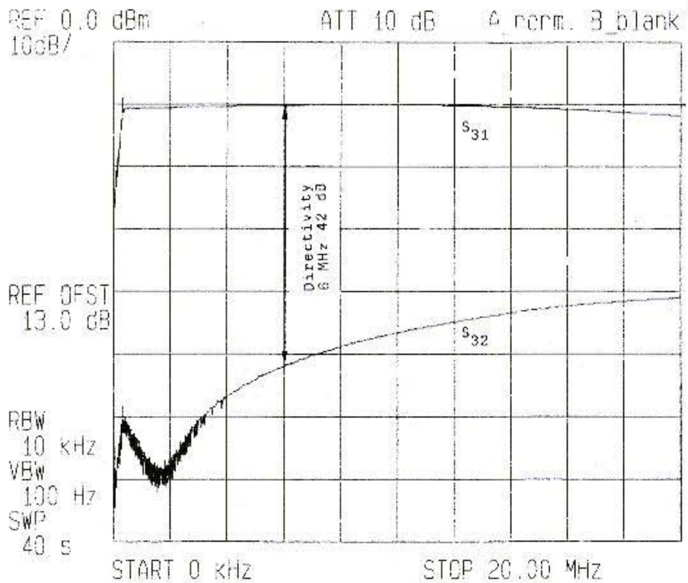

Fig. 6: Video coupler directivity

4.3. Parts list

Wired components

*D*1,*D*2 Zener diode 5.1 V

*R*1 1Ω, 0.1 %, 0.1 Watt

*R*17,*R*18 180Ω , 2 %, 0.3 Watt

*C*5,*C*6 10 uF, 10 V

SMD components

*IC*1 OPA 4650 U, SO14, BURR-BROWN

*R*3,*R*13 Potentiometer 100Ω (3314)

*R*5,*R*6 75Ω (1206)

*R*16 120Ω (1206)

*R*14 150Ω (1206)

*R*11,*R*12 180Ω (1206)

*R*7,*R*8,*R*9,*R*10 1.0 kΩ (1206), paired

*R*15 1.3 kΩ (1206)

*R*4 5.6 kΩ (1206), 0.1 %

*C*1 120 pF (1206)

*C*2,*C*3,*C*4 100 nF (1206)

4.4 Alignment

The transmission from gate 1 to gate 3 is set through the potentiometer *R*13 at 0 dB.

The potentiometer *R*3 makes alignment possible at optimal directivity (gate 2 to gate 3 at 75-Ohms adapted load at gate 1).

If a precise offset alignment is desired (0 V at gate 3 with 0 V at gate 1), then the non-inverting input (pin 12) must be connected with either the positive or the negative operational voltage through a resistance (3 k ...) (fitting option available on printed circuit board).

4.4 Measuring results

The transmission loss of the coupler (gate 1 to gate 2) is approximately 0.1 dB in the 75Ω measurement system.

At 6 MHz, the difference in the signals

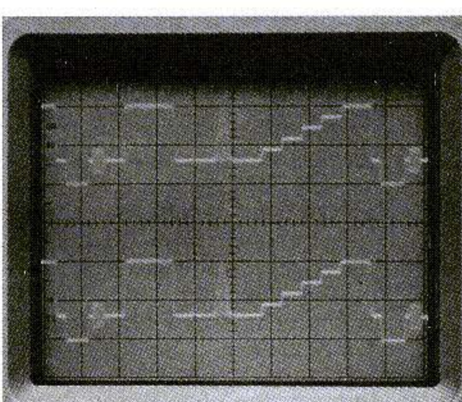

Fig. 7: No recognisable difference between *S*21 and *S*31

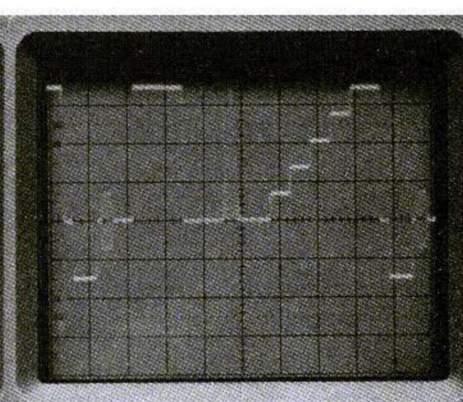

Fig. 8: *S*32 with open circuit at gate1

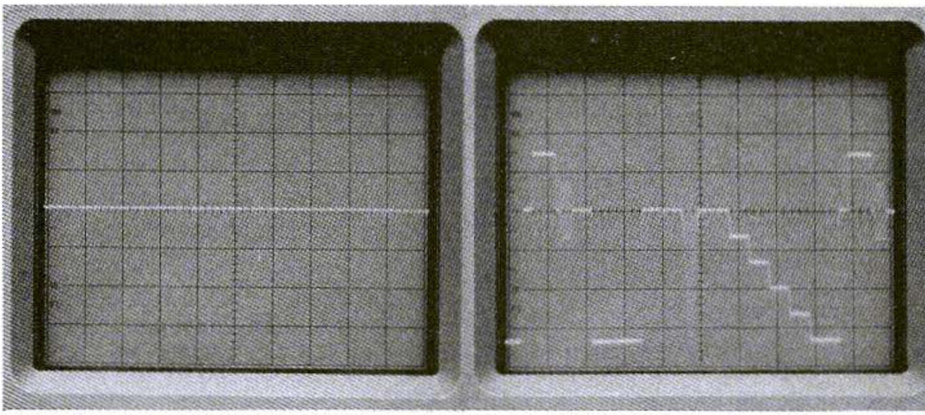

Fig. 9: *S*32 with 75Ω load at gate1   **Fig. 10: *S*32 with short-circuit at gate 1**

in different directions at the coupled output when the passive path is energised still markedly exceeds 40 dB. Because the attenuation of the passive path is slight, this value corresponds to the directivity of the coupler. The measurement is carried out through a network analyser (see Fig. 6).

If a video generator (CCIR 17) is connected to gate1 the outputs for gate 2 and gate 3 can scarcely be distinguished (Fig. 7). More detailed measurements reveal that the signal-to-noise ratio evaluated at gate 3 is better than 60 dB. As a consequence of the high directivity and the good matching of the generator, the signal at gate 3 is almost independent of the load at gate 2.

5.

Reflection measurement

If the coupler is operated in the "reverse direction", i.e. the video generator powers gate 2. The signal at gate 3 gives direct information on the reflection factor ($r1$) at gate 1. The signal at gate 3 when there is an open circuit at gate1 ($r1$ = +1, Fig. 8), 75Ω adapted load at gate 1 ($r1$ = 0, Fig. 9) and a short-circuit at gate 1 ($r1$ = -1, Fig. 10).

6.

Further applications

If we consider Fig. 1 and formulae 2, 3 and 4, it becomes clear that the resistance $R2$ can be cut into two suitable partial resistances, namely $R2 = R3 + R4$.

This makes the network mirror-symmetrical. Using a further instrument amplifier, it is now possible to obtain a signal which is proportional to the signal transmitting power to gate 2; here it is independent of the impedance at gate1. In this way, we obtain a directional coupler which supplies both the forward and the return signal. This has applications in network analysis (e.g. determination of reflection factor, fault location in cables, cable length measurement, power measurement).

With two video couplers, a video duplex installation can be set up: signals can be transmitted and received at the same time, so only one circuit is required.

Last but not least: the transmission range of the video coupler starts at DC; the upper frequency limit is a question of technological progress. Thus the coupler is also suitable for data transmission systems.

7.

Conclusion

The active directional coupler has no lower limiting frequency. The signal path between input and output is passive; only a slight attenuation, independent of frequency, should be expected.

High-frequency old hands will recognise the principle of the transformer-type measuring bridge. The transformer has been replaced by an instrument amplifier and the bridge circuit has been reconstructed in such a way that no power is lost unnecessarily.

Circuits are known which carry out a rectification proportional to the forward or return signals on a resistance network (similar to Fig. 2) in order, for example, to operate suitable display instruments. However, if we obtain the forward or return signal with an active directional coupler, it can be used for secondary purposes (splitting, comparison, measurement). The low-distortion and low-noise amplification of a relatively small voltage difference with a high common mode rejection is required for this and this up to high frequencies can naturally be done only with appropriate modern technologies.

8.

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6 N. Gyula: DIY antenna measuring bridges. CQDL 6/96

Wolfgang Borschel, DK2DO

Circulators and Ring hybrids

For protecting VHF final stages and de-coupling VHF driver transmitters

Readers Ideas on Article in Issue 3/2000

The article [1] in issue 3/2000 took a comprehensive look at this topic. Among the enquiries and comments from readers was a particularly interesting letter from Hans Rohrbacher, DJ2NN.

OM Rohrbacher reports on a research assignment which he was working on in the early eighties, in which he had to interlink temperature-resistant ultrasonic sensors. This required a special coax cable with specific characteristics, which a well-known Swiss firm provided for him. When he read the above-mentioned article, Hans remembered that this very cable, some drums of which he still possesses, was very well suited for applications and assemblies involving the ring coupler described, with asymmetric pitch ratios.

The cable data are as follows:

- Internal conductor: Silver-plated steel-cored 0.18 mm copper conductor
- Screening: Silver-plated 3.70 mm copper braid
- Dieelectric ETFE
- External diameter: 4.30 mm
- Min bending radius: 26 mm
- Capacity: 40 pF/m
- Impedance level (Z): 232Ω
- Shortening factor: app. 0.707
- Attenuation 1.4dB/m (435 MHz)
  3 dB/m (1346 MHz)

Hans has adequate stocks of this cable. He is prepared to give the required λ/4 sections away to interested parties wanting to copy the circuit for as long as stocks last. Hans will send the cable lengths required in return for reimbursement for postage (4 DM In Germany - please enquire for price of overseas deliveries) payable in stamps. His address is:

Hans Rohrbacher, DJ2NN

Talstr 24

D-76689 Karlsdorf

Germany

The output pitch characteristic curve (Fig.1) for ring couplers was expanded using software, so that the pitch ratio (Table 1) and the associated second impedance level (48 , in practise 50) of the ring can be read off for the DJ2NN cable. The transmission loss determined at the useful output was 0.2 dB. i.e.: a ring which was assembled using this cable and powered with approximately

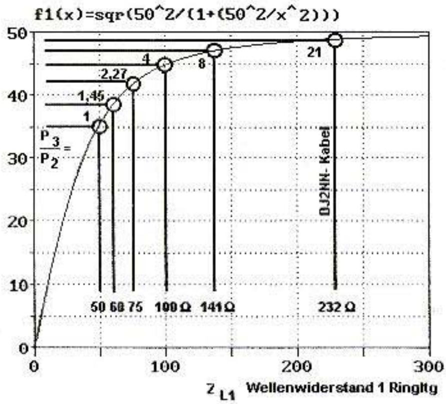

Fig.1: The output pitch characteristic curve of the ring coupler expanded for the DJ2NN cable.

12 W would produce approximately 11. 4 W at the useful output and approximately 570 mW at the auxiliary output. Another important fact to be observed is that as the pitch ratio increases the de-coupled effect of the ring coupler decreases.

Literature references

[1] Wolfgang Borschel, DK2DO: Circulators and ring hybrids, VHF Reports, Issue 1/2000, VHF Communications 3/2 000 pages 179 185.

Table 1: Resistance values for specific pitch ratios, expanded for the DJ2NN cable.

$$\frac{P_2}{P_3} = \left[\frac{Z_L}{Z_{L1}}\right]^2$$

	ZL_1	ZL_2	ZL	a_n
1	50 Ω	35 Ω	50 Ω	3 dB
1,45	60 Ω	38 Ω	50 Ω	2,3 dB
$\frac{P3}{P2}$ = 2,27	75 Ω	41 Ω	50 Ω	1,6 dB
4	100 Ω	44 Ω	50 Ω	0,9 dB
8	141 Ω	47 Ω	50 Ω	0,5 dB
21	232 Ω	48 Ω	50 Ω	0,2 dB

a_n = Durchgangsdämpfung zum Nutzausgang

Gunthard Kraus, DG8GB

Internet Treasure Trove

AGILENT

Anyone who has recently been searching the Hewlett-Packard homepage for data sheets, application notes, S-Parameter files or PSPICE models will have been bewildered. These things have suddenly disappeared without trace and no clues to be found as to where theyve gone!

There is a very simple reason for this, it is currently fashionable to split relatively large companies up into smaller parts and make them into independent entities, naturally with new names!. So we also have to make the corresponding move in this case.

New address: http://www.agilent.com

ANRITSU

A well-known name in the world of gauging equipment manufacturers, especially in the field of microwave, noise and distortion measurement technology. If you search the homepage site carefully, you can find a page with application notes on this subject to be downloaded. Naturally, the companys own equipment is used here, but the basics concerning each individual subject are very well done and can be used both for familiarisation and for personal information on the current state of art.

Address. http://www.anritsu.com

List of German electronic shops on the INTERNET

Have you ever had that problem when you certainly know who might have something, but you just cant remember the address any more? Well, now some friendly person has rectified this by listing the most important shops and their links theyre now just a click away. Moreover, you are cordially invited to help by adding to the list, if you think an important or interesting name is missing.

Address: http://www.schwabmuenchen-.de/~mbpower/elektro.htm

ZRS

Now, do you have any idea what thats about? This is actually a site for Slovenian radio amateurs, and anyone who already knows the name Matjaz Vidmar, S53MV from earlier publications will already have some idea of whats going on. He has now become a professor at the Slovenian Technical University, but his passion for development and DIY projects and his enthusiasm for amateur radio remain undiminished. Thus, if we visit the homepage and go to „Hardware Information Desk, we find a whole lot of projects old and new on every possible subject. For example, he found the No - Tune Transceiver between 1296 MHz and 10 GHz particularly appealing, there are some very comprehensive circuit descriptions, with lots of diagrams and details to be downloaded..

Address: http://www.hamradio.si/hid.htm

Active Filter

Such good operational amplifiers are available today that you can already use them to build active filters for frequencies going up to between 50 and 100 MHz. All a reasonable person needs for a design is the programs already prepared on the INTERNET, and you can even work on line. So heres a small selection.

If you want to do some on-line design work, go to the Onscreen - Design - Program for active filters homepage (Address: http://www.circuitsim.com).

On the Homepage of the Burr-Brown company, you should not only search for the Bbfpro.zip filter program, but also download the application notes:

- AB-017.pdf
- AB-034.pdf
- AB-035.pdf

They contain the theoretical principles of active filters and all the instructions needed to work successfully with the software.

The manufacturer of the PIC micro-controller (= microchip) has something completely new and free of charge, the Windows design program Filterlab for active low-pass equipment. Very well-made and practical specially the option for variations in the component values to lay out the circuit with practical values. You just need to load the file flb1039.zip into your own computer -- but there are obviously still problems on the first version with regard to conversations involving very fast PCs and CD-ROM drives which are just as fast. This will no doubt be put right at some time, but I had to write the file onto my external ZIP drive and install it from there. And then it worked. I found a few more little bugs. But once you start to get used to filter design, you soon recognise things that cant be right...

Address: http://www.microchip.com

You can use the Alta Vista search engine, or look under http://www.rfglobalnet.com to find the following programms and put them onto your own computer:

- afilter.zip

- A-filter.zip
- filter.zip
- filtry10.zip
- filter11.zip

Electronic Repair Service

Anyone looking for a semi-conductor data bank, or for generous helpings of computer repair tips, or for a program which simulates a pattern generator for monitors, or for documentation on TV technology....just take a closer look at this link and download what you need.

Address: http://www.electronic-repair-.com

Tesoft

Anyone who wants to design complete communications or receiving systems and analyse or predict their precise characteristics also needs suitable software nowadays. It is well worth while installing the demo version of such a design program and seeing how it operates, and getting some ideas for your own work. So just take a look at the Tesla System Calculator program!

Address: http://www.tesoft.com

Remcom

The number of EM simulation program manufacturers seems to go up by the hour, and you have to investigate in great detail to see what advantages each new player is offering. So you just take a closer look and install the thing on a trial basis. Not only can you download the demo version of XFDTD 5.0 from the Remcom homepage for this purpose, but you can also obtain the Quickstart tutorial, the complete manual and two very fine simulation examples (= a short vertical antenna and a patch antenna) for familiarisation.

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