A Publication for the Radio Amateur Worldwide

Especially Covering VHF, UHF and Microwaves

VHF COMMUNICATIONS

Volume No.33 . Spring . 2001-Q1 . £5.00

High Precision Frequency Standard Controlled Via GPS

Wolfgang Schneider, DJ8ES and Frank-Peter Richter, DL5HAT

Contents

Wolfgang Schneider DJ8ES, Frank-Peter Richter DL5HAT	High Precision Frequency Standard for 10 MHz Part II: Frequency Control Via GPS	2 - 8
Prof. Gisbert Glasmachers	GMSK, The Modulation Used For Mobile Communications	9 - 19
Dipl. Ing. Hermann L. Aichele	Shielding Technology Using Metallised Non-Wovens	20 - 27
Gunthard Kraus DG8GB	Internet Treasure Trove	28 - 29
	Index of Volume 32 (2000)	31 - 33
Carsten Vieland DJ4GC	Tracking Generator From 1 MHz to 13 GHz for Spectrum Analysers	35 - 48
Gunthard Kraus DG8GB	Modern Patch Antenna Design Part I	49 - 63

Lots of interesting articles again, unfortunately there is no kit available yet for the High Precision Frequency Standard, watch the web site for availablity. Feedback with your 2001 subscriptions was very good, thank you. There were lots of requests for a wider range of articles, if you are a budding author then please contact me. 73s - Andy

KM Publications, 63 Ringwood Road, Luton, LU2 7BG, UK

Telephone: 01582 581051; INT: +44 1582 581051; FAX: +44 1582 581051
Email: andy@vhfcomm.co.uk
WWW: http://www.vhfcomm.co.uk

Wolfgang Schneider, DJ8ES and Frank-Peter Richter, DL5HAT

High Precision Frequency Standard for 10 MHz

Part II: Frequency Control Via GPS

A high-stability frequency standard for 10 MHz can be created using only three system components. Short-term and long-term stability values can be obtained, by simple means, which far exceed the requirements for practical amateur radio operations.

1. Introduction

The most commonly used method for precision time comparisons nowadays makes use of the satellites of the Global Positioning System (GPS). The GPS satellites (there are currently 26 of them) carry atomic clocks of the highest accuracy, the operation of which is carefully monitored by the ground stations. As in all relatively large institutes all over the world, the GPS is also used by the PTB for the international comparison of atomic clocks.

A stable quartz oscillator is regulated so well, with the aid of the GPS, that its maximum frequency deviation always remains better than 1×10^{-11}. This is a precision of 0.0001 Hz in 10 MHz! Or for the GHz amateur: 1 Hz in 100 GHz.

The frequency control via GPS shown in the block diagram (Fig. 1) can offer an accuracy of approximately 4×10^{-10} or, in other words, 4 Hz in 10 GHz. This value results from the imprecision of the counting process built into the system. In frequency counters and this is nothing different the last bit should always be taken with a pinch of salt. Depending on the phase position of the gate time to the counting signal, an error occurs here of ±1 bit (phase error ±100 ns). For a gate time of 1s, that would be 1 Hz for the measuring frequency 10 MHz ($\pm 1 \times 10^{-7}$).

The first practical measurements were based on a gate time of 8 seconds, which corresponds to a resolution of 0.125 Hz. Together with the phase jitter of the GPS signal (1s cycle), there should have been uniform distribution and thus a levelling off of the reading over a relatively long period of time (max. 64 measurements). However, this turned out to be wishful thinking. From the measurement technology point of view, the situation with this gate time was that the oscillator frequency varied very slowly, or perhaps we should say it circled around the rated value of 10 MHz. The absolute frequency here was 10.0 MHz ±0.0305 Hz.

If the gate time is increased to 128

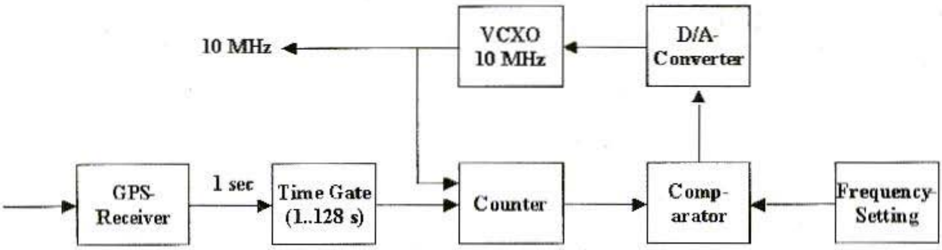

Fig 1: Block Diagram of Frequency Control via GPS

seconds, in theory the reading improves to at least ±0.0078125 Hz (±7.8125 mHz). However, the influence of the GPS phase jitter is now reduced. This results in an effective usable precision for the 10 MHz-Signals of approximately 4 x 10^{-10} or, as already stated above, a frequency counter controlled on such a quartz time basis has a display accuracy of 4 Hz at 10 GHz.

2.

The control assembly circuit

In principle, the control stage (Fig. 2) operates like a frequency counter with an additional numerical comparator. The oscillator frequency - 10 MHz of the HP oscillator HP10544A is counted out here. The gate time of the counter is generated from the 1pps signal of the GPS receiver with a 74LS393. For control operation, it amounts to 128 seconds and 8 seconds in the comparison mode for the OCXO.

The 74HC590 counter module is an 8-bit counter. It can be used, with a gate time of 8 seconds, to measure the input frequency 10 MHz ±16 Hz. The minimum resolution here is 0.125 Hz. In control operation (gate time 128 seconds), this is improved by a factor of 16. This results in the system-determined precision of 0.78 x 10^{-9}, based on the 10 MHz frequency oscillator.

The HP oscillator frequency can be finely adjusted using a tuning voltage of 5 V. This is done by the digital-analogue converter (AD 1851). It offers a resolution of 16 bits for a control voltage range of 3 V. This gives a setting range for the OCXO of approximately 0.5 Hz.

The AT89C52 micro-controller controls all the functions described above within the control assembly. The essential elements controlled which we should mention here are the meter module, the D/A converter and the status in the LC display.

3.

Software description for controller

The software in the micro-controller AT89C52 performs two tasks. Firstly, it should enable a rough comparison operation to be carried out, and secondly it

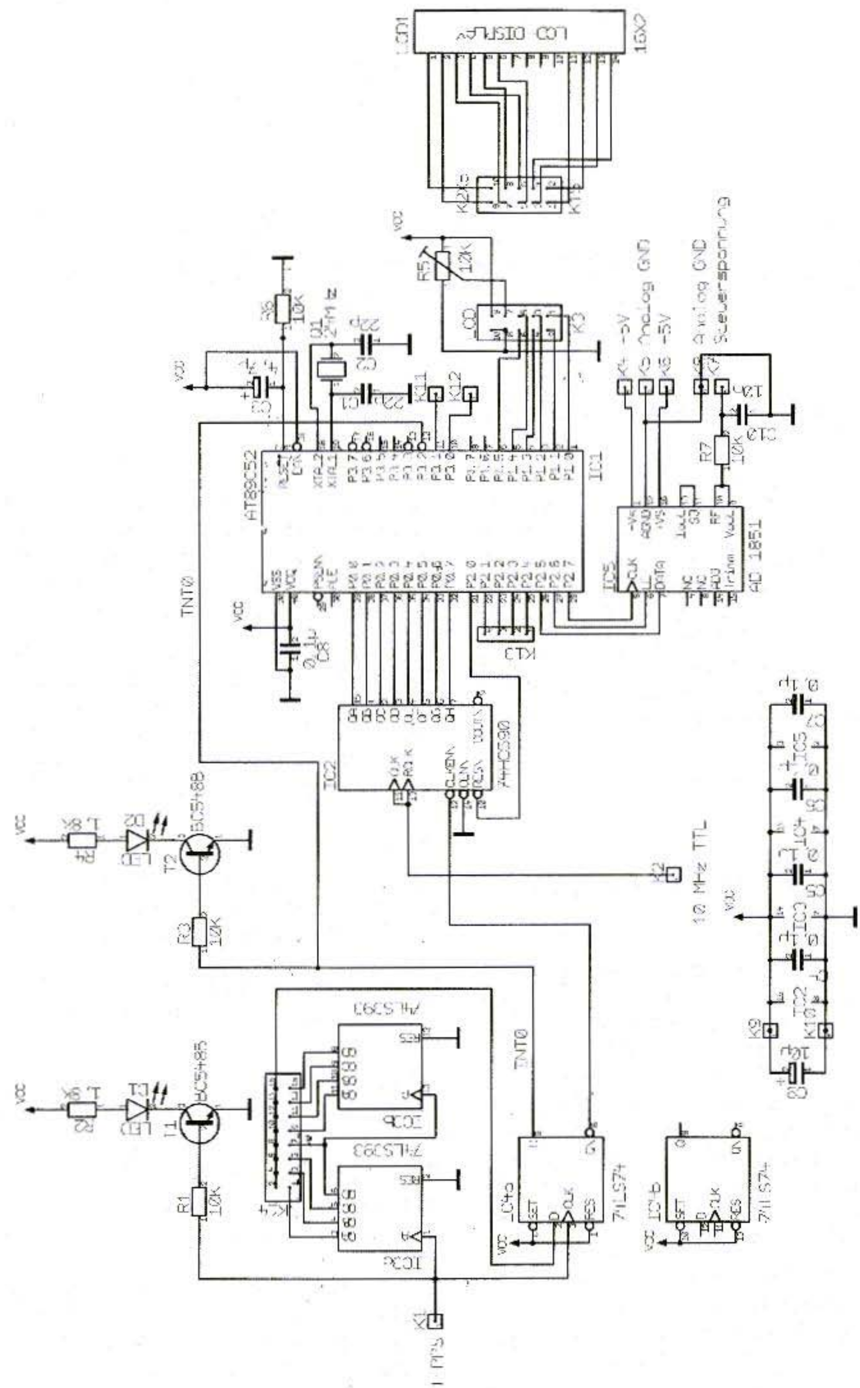

Fig 2: Circuit Diagram of Control Stage

will continuously carry out the final fine adjustment with the GPS signal.

If pin 4 of K13 is open, then when the voltage is applied to the control circuit board the LC display shows Warming Up. If the GPS second cycle is activated, the first value will be displayed after approximately 15 minutes. In order to eliminate any artificial jitter in the GPS second cycle, a mean value is formed and displayed from 64 readings from the 74HC590 meter.

A change in the oscillator frequency on the mechanical potentiometer will thus not display any effect for some time. So after a change on the potentiometer we must just wait for approximately 64 x 8 seconds until the next adjustment takes place. If a value of 0.250 Hz is attained, we can switch over to the basic control.

If the software in the controller is to recognise that the automatic control now has to be carried out, Pin 4 of K13 must be earthed.

The artificial jitter of the GPS signal and the very slight deviation following the rough comparison require there to be a relatively long gate time of 128 seconds. This is obtained by means of a bridge between pins 15 and 16 at K14.

Here too, the first message on the display is Warming Up. If the GPS second cycle is activated, the first value is displayed after approximately 15 minutes; though here it is not the deviation in Hz but the value which is written in the AD 1851digital-analogue converter. This value can reach a maximum of 32767, which means approximately 3V.

The software assesses the condition of the meter and, depending on the polarity sign, calculates the value for the AD 1851 digital-analogue converter. From the present condition of the meter and the mean of the last 64 meter results, the figure is determined which is to be added to or subtracted from the current digital-analogue converter value, which can be seen on the display.

According to pure theory, with the given gate time of 128 seconds and with a mean value formed over 64 readings, the time to reach the finally precision of

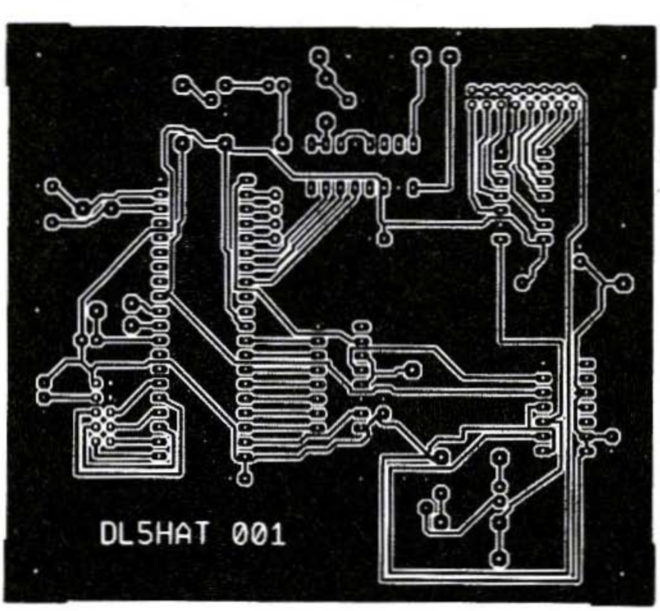

Fig 3: PCB Layout for DL5HAT 001

$4x10^{-10}$ Hz (i.e. 4 Hz in 10 GHz!) is over after 4.5 hours. However, it has been demonstrated in practise that this value has already been reached after approximately two hours. It is important in this connection that the frequency deviation is not determined by the system alone but also depends on the number of GPS satellites which are currently being received, and the signals from which are available for evaluation.

4. Assembly instructions for control assembly

The frequency controller circuit is put together on a single sided epoxy circuit board, with dimensions of 100 mm x 100 mm (Fig. 3). Once the holes have been drilled in the circuit board, all the components can be fitted, in any order. The components drawing (Fig. 4) can be of assistance here.

The micro-controller (IC 1) should have a plug-in socket. Thus if the software is updated later, the processor can be easily replaced.

4.1. Control assembly parts list

1x µC	AT89C52
1x ADC	AD1851
1x TTL-IC	74LS74
1x TTL-IC	74LS393
1x TTL-IC	74HC590
2x Transistor	BC848B
2x LED	green, low current
1x crystal	24 MHz
1x spindle pot	10k
1x socket strip	10-pin
1x plug strip	10-pin
1x stud strip	14-pin
1x jumper	
1x PCB	DL5HAT 001

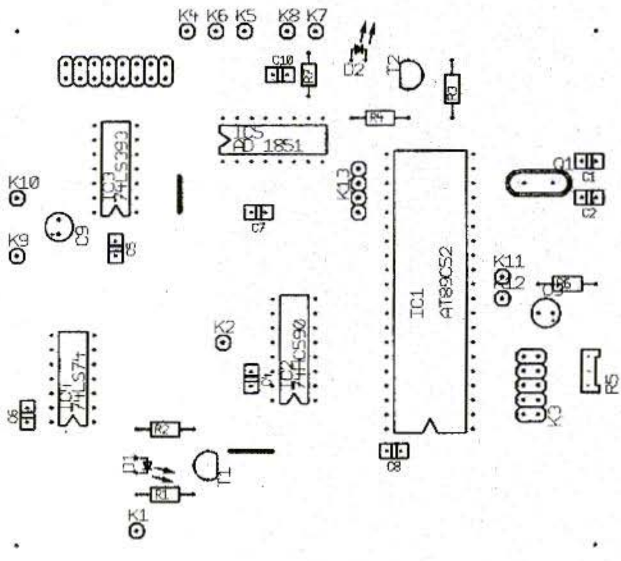

Fig 4: Component Layout for PCB

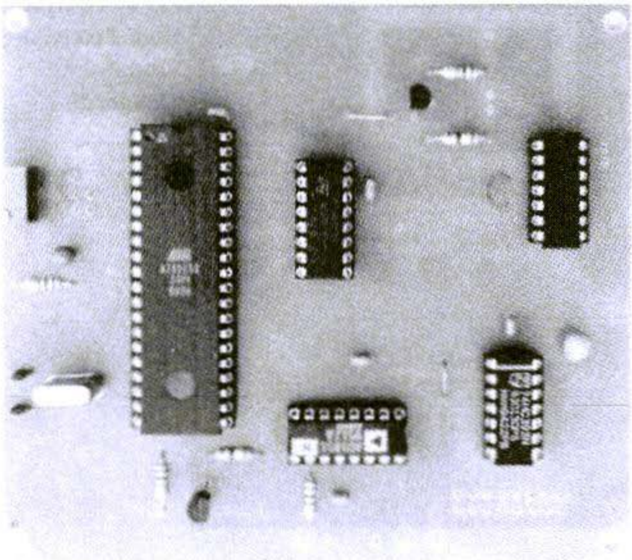

Fig 5: Completed PCB

Resistors

2x 1.8k

4x 10k

Ceramic capacitors

5x 0.1µF

2x 22pF

1x 10nF

Tantalum Capacitors

1x 4.7µF/25V

1x 10µF/25V

5. Inter-connecting all assemblies

The HP 10544A crystal frequency oscillator powers the DJ8ES 049 buffer stage. This makes available outputs for TTL levels (1. 5 and 10 MHz) and 3 separate 10 MHz sine signals on the output side. All connections are made to BNC sockets on the front plate of the apparatus.

The control assembly (DJ8ES 050) initially needs the synchronisation signal (1 pps) to generate the gate time for the frequency counter. This signal is generated in the specimen apparatus by means of a GPS receiver manufactured by GARMIN (GPS 25-LVS receiver board). The external aerial required is coupled through an SMA connection.

The frequency measurement input (10 MHz, TTL level) is connected in parallel to the corresponding BNC socket.

The control assembly output supplies the control voltage for the HP oscillator.

The tuning voltage, 5V, must be separately generated in the frequency controller power supply. The following concept would be ideal for the power supply in the frequency standard for10 MHz:

+24V	HP oscillator
+5V	GPS receiver
+5V	control assembly
5V	control voltage

Fig 6: Completed High Precision Frequency Standard

The three keys on the front panel of the apparatus are intended to make it possible to show the status in the LC display. They are not taken into account in the present software version.

6. Operational experience

Long-term observations of the 10-MHz frequency standard over approximately 4 weeks confirmed the assumptions made in the introduction. Since it is not easy to measure the frequency directly in this way with the necessary precision, the analysis must be carried out through the tuning voltage of the OCXO. The oscillator frequency varies with a time constant of several hours around the value of $4x10^{-10}$ Hz. The control software is being optimised again at the moment. The aim is to improve the deviation by a factor of 10.

To sum up, it must be admitted that a high-quality frequency standard for 10 Hz has been created using what is actually a decidedly simple method, and with minimal expenditure on hardware. This means that a reference frequency is now available in your own home at any time for frequency meters or frequency synchronisation, or for other applications. The accuracy of 4 Hz at 10 GHz (and eventually 0.4 Hz!) is certainly more than adequate for amateur applications.

7. Literature references

[1] Frank Sichla, DL7VFS: Precise frequency measurement but how? Funkamateur 7/95, Theuberger Verlag Berlin

[2] Wolfgang Schneider, DJ8ES: 10 MHz frequency standard using GPS: Congress Proceedings, Weinheim 1999

[3] Hewlett Packard (HP): 10544A 10 MHz crystal oscillator: Technical Data May 74

Prof. Gisbert Glasmachers

GMSK

The Modulation Used For Mobile Communications

Modern communications technology must be matched by modern electronics. So that the precious resource of frequency space is not wasted, new types of modulation are being tested and brought into use; the article below introduces the GMSK type of modulation currently in use for GSM mobile telephones.

1. Introduction

In many branches of technology, a change from analogue to digital techniques is taking place. Current examples of this are video technology and telephony. Modern equipment transmits and stores what were originally analogue signals in digitised form. To this can be added continuously increasing transmission via wireless paths.

Radio technology must provide facilities for the transmission of digital signals. One elementary problem here is the restriction of the bandwidth. Analogue signals, rather like human speech, encounter a natural band width restriction at approximately 10 kHz. Consequently, the medium wave radio transmitters, with analogue amplitude modulation, require precisely this bandwidth. Digital signals, due to their fast rise times, require a much higher bandwidth. If then (to stay with the example of human speech), we digitise at a high resolution, e.g. 12 bits, the bandwidth required for a transmission could easily be increased to several MHz. It is clear that technology is trying everything in order to reduce this unbearable drain on precious frequency space. The main aim of this article is to introduce GMSK, the preferred type of modulation for radio transmission. To this end, the state of the art is described, together with the options for the radio amateur wishing to become familiar with this technology using simple resources.

2. Type of modulation

The type of modulation is usually determined by the efficiency, security and costs of the transmission path. A basic distinction is made between amplitude modulation and phase modulation. Amplitude modulation in the field of long wave and medium wave radio has been known since the nineteen-twenties. Its susceptibility to interference makes it largely unsuitable for modern communi-

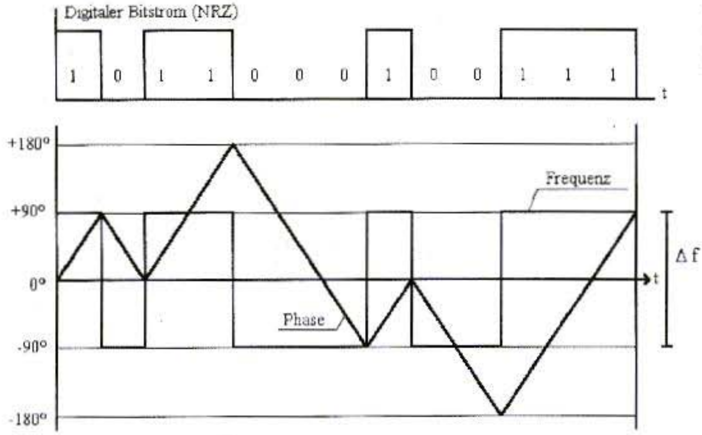

Fig 1: Digital signal Showing the Effect on Frequency and Phase

cations systems.

Phase modulation affects the frequency and phase of a carrier. With digital frequency modulation, the frequency must be switched over in sympathy with the data signal, see Fig. 1.

There is a direct relationship between the frequency and the phase. If the frequency goes up, the phase rises linearly. It is clear that a big frequency change, Δf, is easier to detect in the receiver than a small one. But unfortunately the spectrum width of a radio signal is directly proportional to the size of the frequency change. A reasonable compromise must therefore be found between the frequency shift, Δf, and consideration of the restricted width of the radio channel. At a low received field strength, the channel width plays an important part, yet the noise output increases linearly with the bandwidth. For this reason it is also desirable that the channel width should be as small as possible.

Examples of the effects of the type of modulation on the spectrum are shown in Fig. 2. The frequency axis is divided up into 4 transmission channels. The ideal spectrum would be a uniform distribution of the output over the channel width. In addition, propagation into the adjacent channel would be prevented completely. Such ideal conditions can not be provided in practice. Two real spectra are also shown, GMSK in channel 3 which can still be described as good; MSK in channel 2, by contrast, is not good. The frequency shift keying shown in Fig. 1 corresponds to MSK.

From the spectrum point of view, the bandwidth of an MSK signal is unnecessarily large, we can attempt to reduce it using suitable filtration. Filtration after modulation would be possible, so that we could simply cut off the edges of the

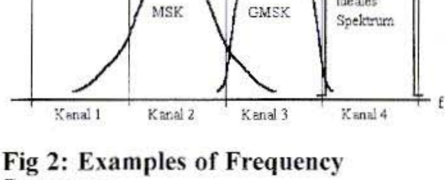

Fig 2: Examples of Frequency Spectrum

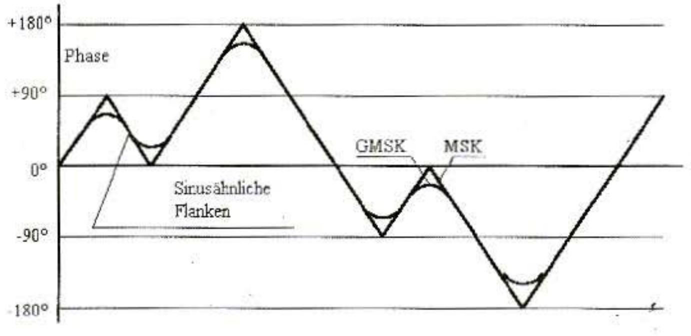

Fig 3: Transition from MSK to GMSK by Rounding Corners

spectrum using a band pass filter. This would also mean that the information itself was also cut off. It seems more sensible to filter the digital signal before modulation, which leads directly to the GMSK spectrum. The concept is known as Gauss Minimum Shift Keying. It includes filtration using a Gauss filter. The results are shown in Fig. 3.

The corners in the phase gradient widen the spectrum. If they are rounded off, the spectrum becomes narrower. We can vary how distinctly the rounding off is made, this is characterised by the so-called BT product, where

B = Bandwidth of Gauss filter

T = TBIT, the bit time.

For MSK, BT approaches infinity. Frequently used values for BT are 1, 0.5 and 0.3. It should be pointed out that the European mobile radio systems (GSM) are operated using GMSK modulation with BT = 0.3.

3. Generation of GMSK signals

Basically, any phase modulation and this includes GMSK can be obtained by changing the phase or frequency. A simple method of generating frequency-modulated signals consists of using a voltage-controlled oscillator (VCO). The frequency is changed using a control voltage fed in from outside the control loop. This produces only an approximate adjustment of the frequency shift.

A precise adjustment, as required for MSK and GMSK, requires the phase to undergo a modulation instead of the frequency. For this, we generally use quadrature modulators, the functional principle is shown in Fig. 4.

The function of this modulator is to generate a carrier with any phase displacement desired at constant amplitude. For this purpose, the carrier is divided into two signals, one with zero degrees phase (sin) and one with 90 degrees phase (cos). These two signals are generated by producing a carrier at twice the frequency, followed by a 2:1 frequency divider and an inverter. The two signals can be set to any amplitude between +1 and 1 by means of a multiplier (mixer). Finally, the two signals are brought back together. The multipliers are controlled by the digital data which is divided into two signals (I and Q) and shaped through digital filters. This produces a phase transitions of 90 degrees, as shown in Fig. 5.

In practice quadrature modulators are on

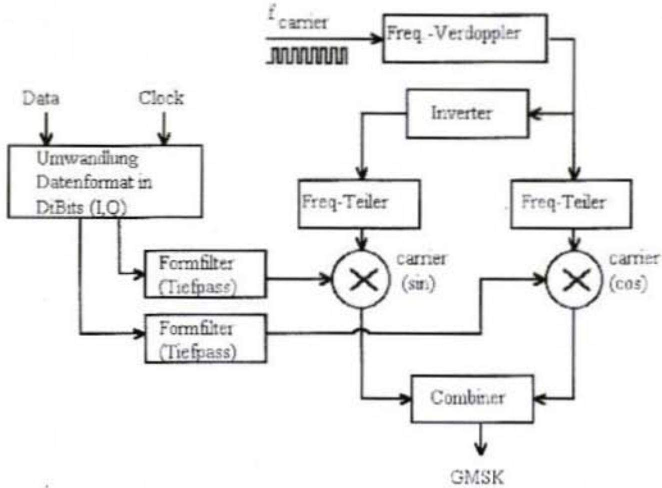

Fig 4:Block Diagram of a Quadrature Modulator

Freq. - Verdoppler = Frequency Doubler

Freq. - Teiler = Frequency Divider

Unwandlung Datenformat in DiBits (IQ) = Convert data to I Q format

sale as complete circuits, e.g. from HARRIS, MAXIM, TEMIC and many others. Splitting the digital signal up into I and Q components and filtering it is usually left up to the electronics manufacturer. This can be achieved by using a micro-controller, for example one from the MICROCHIP PIC range, which reads off the necessary I and Q components from a table. The filtration required for GMSK can also be incorporated into the tables without any problem. For this purpose at least one up to a maximum of three past and future bits

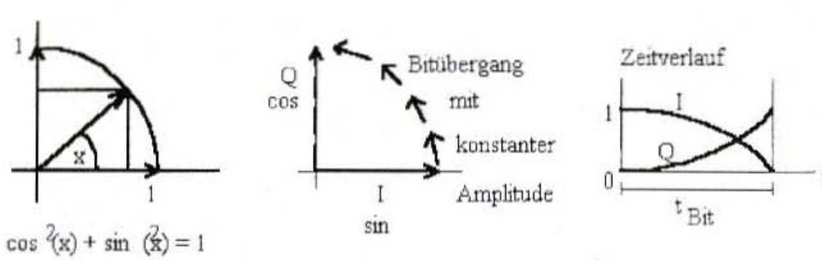

Fig 5 : Combination of sine and cosine components

Bitubergang mit konstanter amplitube = Bit transition with constant amplitude

Zeitvertauf = Time period

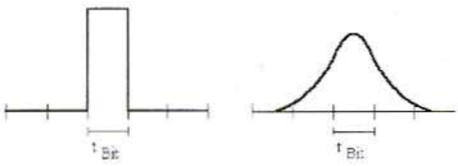

Fig 6: Effect of Gauss Filter on a "One Bit"

around the current bit must be examined; see Fig. 6.

The filtered bit extends into the adjacent bits, this is referred to as symbolic interference if adjacent bits interfere with one another. The lower the limiting frequency of the filter, the wider (and flatter) the bit becomes. This means that the phase rotation, which should be precisely 90 degrees per bit, changes depending on the bits immediately in front and behind. The exact phase rotation can be calculated mathematically. If we takeas an approximation that the change extends to only a single bit, we obtain 8 variants in all, which are shown in Fig. 7.

If the consideration extends over plus/minus two bits, this gives a total of 32 different phase gradients. These must each be calculated for the I and Q components in four possible quadrants (from 0 to 360 degrees). At least 10 measurements must be provided per bit, this would require the preparation of 2560 storage locations. This number can be considerably reduced by skilful use of symmetries. The phase jumps arising from the quantisation can be sufficiently dampened through simple RC filters. Integrated circuits for the conversion from digital signals to the I-Q signals, e.g. the CMX 589 are available from the British manufacturer CML.

4. Demodulation

There are various demodulation techniques. We can distinguish between:

1) Frequency demodulation
2) Quadrature demodulation
   - Coherent
   - Non-coherent

In all cases, an uncorrupted bit flow must be available at the demodulator

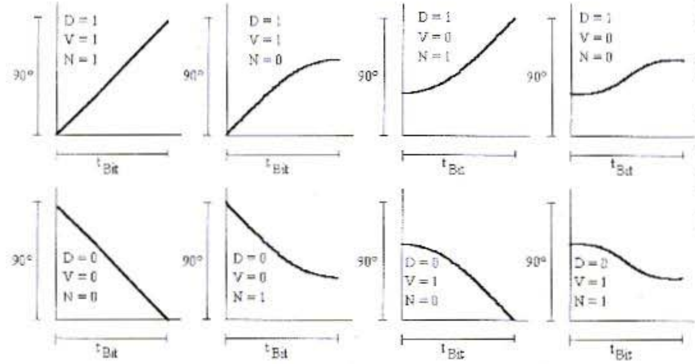

Fig 7: Phase Change by Filter For a Single Bit Approximation Giving 8 Variants.

D = Data Bit

V = Previous Data Bit

N = Next Data Bit

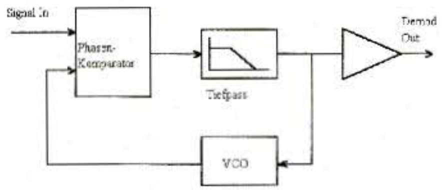

Fig 8: PLL as an FM Demodulator

output. Because the functioning is simpler, frequency demodulation will be discussed first.

4.1. Frequency demodulation

As can be seen from Fig. 1, for MSK the frequency gives a direct reflection of the bit flow. Thus an elementary FM demodulator is sufficient, with either a tuned circuit or a PLL [1]. Technical problems arise from the frequency shift, which in some circumstances is small, this leads to relatively low voltages at the discriminator. Attempting to increase the sensitivity of the discriminator results in a narrower frequency range. This causes the negative effects of drift and component ageing to become noticeable. Similar findings apply for the PLL, with the tuned circuit of the VCO operating as a discriminator. A basic circuit diagram for the functioning of a PLL as an FM demodulator is shown in Fig. 8.

Both circuits basically operate using oscillation circuits, which cause transient effects when the frequency changes suddenly, as happens with MSK. This gives a problem finding out the optimal damping while maintaining a sufficiently high limiting frequency. With the PLL, its structure as a control circuit adds a further difficulty, with the low-pass filter having a considerable influence on the frequency response. GMSK, with its smoother frequency transitions, does not make such high demands on the frequency response of the discriminator, yet it is more difficult to retrieve data securely using this type of modulation.

4.2. Coherent quadrature demodulation

The principle of quadrature mixture can also be used for the reverse transformation of the modulated signal into the base band. The output signal in this case is a rotating voltage vector, the direction of rotation contains the digital information; see Fig. 9.

In order for the demodulation to precisely reproduce the base band, the local oscillator must be twice the frequency of the transmission signal, due to the subsequent halving of the frequency, and coherent with the input signal. In order to achieve this, it must be derived from the input signal in some way. Various processes for achieving this are known, which are known as carrier recovery.

One is taken from television engineering: the unmodulated carrier frequency,

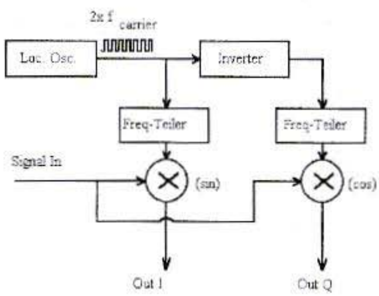

Fig 9: Quadrature Demodulator

the so-called synchronisation burst, is transmitted for a specific time. The local oscillator is synchronised to this frequency and retains it in the subsequent phase, in which modulated signals are transmitted. Thus the carrier and the data signals alternate in the transmission signal (Fig. 10). A disadvantage here is the time lost for the synchronisation; the effective bit rate falls. There are complete circuits available for this demodulation process, such as the above-mentioned CMX 589. The greater the time interval between the synchronisation burts, the more the frequency will be displaced from the coherent condition in the subsequent data phase.

Another option for generating a coherent carrier consists in the obtaining of a phase-locked signal from the phase-modulated signal. For hard phase shift keying e.g. QPSK - the carrier can be obtained by means of double squaring of the phase-modulated signal [2]. This procedure does not work with soft phase shift keying such as MSK and GMSK. Here special control circuits are needed (Costas loop), but because of their time constants these react slowly [3]. Consequently, the effective bit rate is reduced for these systems as well.

4.3. Non-coherent quadrature demodulation

If the local oscillator oscillates freely, a precise conversion will not occur. Even if precise quartz oscillators are used either on the transmitter side or in the local oscillator of the receiver, a residual error remains,

$$\Delta f = f_{sender} - f_{Loc.Osc.}$$

In this case, we speak of non-coherent demodulation. We can only try to keep this error as small as possible. The I and

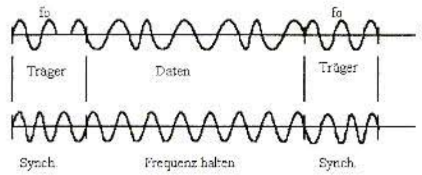

Fig 10 : Synchronisation of the Local Oscillator

Tager = Carrier, Daten = Data, Frequenz halten = Frequency Held

Q outputs of the demodulator produce an output, even when no modulation is present on the input signal. But such a system can be used as a serviceable demodulator. Various cases are possible here:

1) The resulting output (modulation plus frequency error) rotates in varying directions, depending on the value of the data bit. This condition occurs if

$$\Delta f \lhd 4 f_{Bit} = \frac{4}{T_{Bit}} ist$$

At high bit rates, or for high-precision oscillators, this condition is easy to maintain. The value of a bit can then be recognised directly from the direction of rotation of the output.

2) The resulting output rotates in only one direction. The speed of rotation is merely reduced or increased by the value of the bit. It then becomes more difficult to recognise the bit. The processor employed for recognition must measure the actual speeds of rotation and read out the data from their difference. More recent developments take up this idea and attempt demodulation

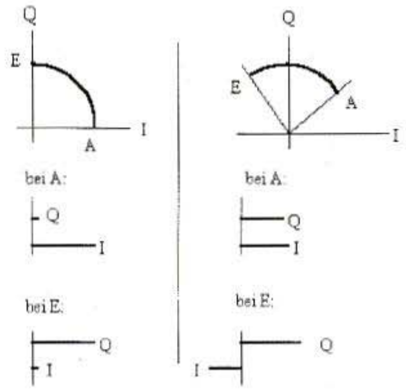

Fig 11: Examples of phase changes
A = Start of Bit, E = End of Bit

directly on the intermediate frequency, using very rapid analogue/digital converters and correspondingly fast processors. This avoids frequency conversion into the base band with the associated additional circuitry.

To sum up, it can be stated that non-coherent demodulation gives the best efficiency with regard to the effective bit rate. A burst is not necessary for synchronisation. Moreover, little circuitry is needed. There are no tuning processes, such as are almost unavoidable with tuned circuits.

5.
Decoding

The I and Q voltages produced by the demodulator are still purely analogue signals. They must be reconverted into digital signals. This task (data reconstruction) is undertaken by the decoder. First, the I and Q signals produced by coherent demodulation must be examined. The phase transition for a logic one corresponds (for MSK) to a rotation through +90 degrees; see Fig. 11 above.

Even in coherent demodulation, the phase of the local oscillator, as against the phase of the reception signal, is not defined, the angular gradient of a bit can lie anywhere on the circle with radius 1. Fig. 11 shows two possible cases for this.

5.1. Analysis of direction of rotation of phase

In order to analyse the direction of rotation of the phase without using extensive technology,a sampling rate which corresponds precisely to the bit frequency is sufficient in principle. This means that the I and Q components are measured at the output of the demodulator at the beginning (A in Fig. 11) and at the end (E) of the bit. The second measurement (E) is then simultaneously the initial measurement for the next bit. The direction of rotation is particularly easy to recognise if the output unambiguously changes the quadrant: see Fig. 11 on right. The associated direction of rotation is determined from the difference between two successive measurements, as can be seen from Table 1 (see Fig. 11 on left and right).

Table 1 can easily be extended to cover the entire phase circuit and both directions of rotation. In systems created in practise, more than two measurements must take place, because it is not so easy to determine exactly when a bit begins. Having several measurement points also creates the advantage that the noise can be suppressed a little by forming a mean value. The comparison operations listed in the table can be carried out using simple micro-controllers. These can also regenerate the cycle; see below.

The analysis of the direction of rotation

Message	I and Q (left of fig 11)	I and Q (right of fig 11)
A	I(A) > 0, Q(A) = 0	I(A) > 0, Q(A) > 0
B	I(B) = 0, Q(B) > 0	I(B) < 0, Q(B) > 0
Quadrant Changes	Not defined	Yes
Check	I(B) < I(A), Q(B) > Q(A)	N/A
Direction	Left	Left
Bit	One	One

Table 1: Determining Direction of Rotation

described above leads to satisfactory results for the coherent demodulation of MSK signals. In order to arrive at the case of the demodulation of GMSK with a non-coherent local oscillator, the procedure can be extended as follows:

- Establishment of differential frequency
- Re-transformation from GMSK to MSK

The differential frequency can be measured accurately only if there is no modulation, or for a known bit sample. Otherwise, it is subjected to drift, which is mainly conditioned by temperature. Additional problems arise from the phase noise of the oscillators. If we look at it from this viewpoint, we can recommend data transmission in blocks using a prefixed synchronising burst.

But this is not essential: provided that the frequency error is small in relation to the bit rate, we can dispense completely with any correction of the frequency error. Taking the frequency error into account requires the calculation of the absolute phase angle for each measurement. This is mathematically possible, with some effort, but it can be done faster using tables (although unfortunately a lot of memory space is needed). But if the determination of the direction of rotation has to be corrected again, with the help of the frequency offset, simple micro-controllers rapidly become overloaded here. All that can be done then is to use programmable logic circuits (FPGAs). This increases the price and the amount of programming considerably.

For GMSK signals, the phase angle change per bit is often considerably less than 90 degrees; see Figs. 3 and 7. It is obvious that this makes satisfactory recognition of the direction of rotation more difficult. Since the GMSK signal on the transmitter side has emerged from an MSK signal via Gauss filtration, the filtration can be reversed again by carrying out the inverse operation on the receiver side The more samples per bit are captured, the better this succeeds. The digital filter required for this can be programmed in a sufficiently rapid processor.

The simple recognition of the direction of rotation referred to above requires only low expenditure. It functions with a low bit error rate right down to BT = 0.3. A pre-condition is that there must always be a stable local oscillator with low phase noise. For the radio amateur interested in practicable solutions, there are thus several variants open:

- FM demodulation at lowest possible intermediate frequency;
- Coherent demodulation with integrated circuits, using a 'synchronous burst;
- DIY non-coherent demodulation.

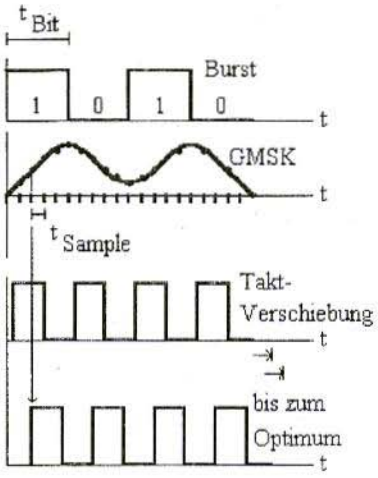

Fig 12: Burst of 4 Bits With 5 Samples.

Takt Verschiebung = Cycle displacement, bis zum optimum = to the optimum

5.2. Obtaining the Clock

Serial bit flows generally require a clock, the edges of which should, if possible, lie in the centre of the bit. It is the task of the decoder to generate this clock. A pre-condition is a bit rate which is maintained quartz-accurately on the transmitter side and which is known on the receiver side. It is also recommended that the signal should be scanned on the receiver side using a whole-number multiple of the bit rate, for example with 5 or 10 samples per bit. In this way, there are always several measurement points in a bit. If, using ten-fold sampling, we now generate the clock edge with the fifth sample, then the edge will be positioned correctly in the centre of the bit time. The only question still to be answered is how to recognise the beginning of a bit. The following alternatives are available here:

- The beginning of a bit is recognised by the change in the direction of rotation
- The beginning of each bit is defined by synchronisation with the help of a synchronous burst.

The change in the direction of rotation takes place each time there is a change from zero to one or vice versa. If we acquire this moment, we can synchronise the cycle to it. But at the actual start of the data transmission no real change is present. The very first cycle is thus not so easy to recognise. In addition, with GMSK specifically, this change is also sharpened: see Fig. 3.

A synchronous burst at the beginning of the block transmission is available as a remedy. This consists of a precisely defined one-zero-one- ...etc. sequence. The smoothing should be deactivated during the burst. The clock is generated by a quartz oscillator, the frequency of which is precisely matched to the data frequency. However, its phase shift relative to the data bits is undefined. With a suitable delay, the phase can be brought into the optimal position, i.e. in the centre of the bit. The decoder must generate this delay. The synchronous burst at the beginning of a transmission block serves, among other things, to pick out the optimal delay time. This is then retained, unchanged by the subsequent data transmission, see Fig 12.

The clock is not used until the first actual data bit is encountered. Since it is generated with quartz accuracy, it will be accurate for the subsequent bits. In this way, several hundred, or several thousand, bits can be transmitted without post synchronisation.

6. Summary

The transmission of digital signals through radio channels using GMSK modulation will achieve greater and greater penetration over wide areas. Even progressively orientated radio amateurs can become involved in this field. The present report is intended to provide some background knowledge, together with a certain amount of orientation to assist in practical implementation.

7. Literature

[1] Best, Roland; Theory and application of phase-locked loop, AT-Verlag, Stuttgart

[2] Glasmachers, Gisbert; Digital phase modulation; VHF Reports 2/1999

[3] Mäusl, Rudolf; Digital modulation procedures; Hüthig-Verlag, Heidelberg

Dipl.-Ing. Hermann L. Aichele

Shielding Technology Using Metallised Non-Wovens

Examples of installation and planning of copper non-woven shielding

Metallised non-wovens and fabrics are being used more and more in the technical world, and by the public, since the shielding sector extols innovation and novelty. Although it was first introduced on the American market, this technology has also been established for over 6 years in Germany for the field of lower-value shield attenuation requirements. The only new point is that this special shielding technology is continually being re-discovered as an alternative to the well-known classic steel plate shielding.

The special advantages of this type of shielding technology are demonstrated, using specimen applications from practice, and specific question, to show that users could find it to be an alternative to the well-known shielding technology using metal plates.

1. Introduction

Due to the rising demand in practice for advantageously priced and easy-to-use shielding, which is also encouraged by the know-how and the production capacities in the fields of metallising plastics and fibres, numerous metallised non-wovens, fabrics or knitteds have now penetrated the market in the most varied areas of use.

In contrast to non-wovens, fabrics and knitteds consist of a weave pattern or knitting pattern of very thin and long polyester fibres or other synthetic base materials. The woven or knitted material naturally has far better strength properties, in comparison to the non-woven. For this reason, this material is used predominantly where high strength is important (e.g. with regard to tearing) or else where elasticity is required and varying continual stresses can arise (e.g. clothing, moving machine components, seals, cable sheaths, etc.). Since fabrics and knitteds have no practical significance in the shielding of rooms, they need not be examined any further in this article.

Only basic non-woven material is certainly relevant for shielding technology primarily for reasons relating to costs and handling. A non-woven is likewise a synthetic material, consisting of an irregular arrangement of millions of short, thin (polyester) threads (see Fig. 1). For shielding, non-wovens are manufactured in pressed form as a kind of wallpaper on reels 1.4 m wide and approximately

Fig. 1: Structure of copper non-woven under the microscope

0.3 - 0.5 mm thick. Non-wovens can be handled by specially trained personnel or by craftspeople such as painters and decorators or floor-tilers, etc., and often using simple tools.

The following article first gives information on the material properties and then on some typical installation procedures, and selected examples from practice are discussed.

2. Metallised non-wovens

2.1. Material properties

The technical and mechanical properties of various non-wovens and fabrics are compared in Table 1.

It is thus particularly surprising that the shielding attenuation values in the table are still sometimes significantly exceeded in practice. Metallised copper non-wovens are the main substances to have won through as shielding materials, on grounds of cost and handling. Because of the possible health hazard, the handling of material mixed with nickel calls for supplementary health and safety at work measures.

2.2. Installation procedure

The installation of the copper non-woven begins with the preparation of the surfaces onto which the non-woven is to be placed. These must be free from dirt and grease, smooth and without sharp projecting components. The corners and edges should be formed first (Fig. 2), followed by the ceiling and wall surfaces. Ducts, outlets and electrical fittings require specialised subject knowledge. The floor is covered last.

In order to attain the specified shielding attenuation in each case, overlapping is required. The non-woven, usually single-ply, can also be installed double-ply in a second stage of the installation. In

Metal		Copper	Nickel/Copper	Copper	Nickel/Copper	Copper	Nickel/Copper
Type of substance		Non-woven	Non-woven	Fabric Ribstop	Fabric Ribstop	Taffeta	Taffeta
Weight in app. g/m²		app. 50 - 75	app. 61 - 102	app. 68 - 92	app. 78 - 112	app. 61 - 102	app. 61 – 102
Thickness in app. mm		app. 0.5	app. 0.5	app. 0.2	app. 0.2	app. 0.5	app. 0.5
Metal fraction in app. g/m²		app. 10 - 24	app. 20 - 47	app. 14 - 24	app. 24 - 44	app. 20 - 47	app. 20 – 47
Specific resistance in Ohms/□ as ASTM F390		< 0.1	< 0.1	< 0.1	< 0.1	< 0.1	< 0.1
Attenuation in app. dB	1 - 100 MHz	60	60	60	70	70	70
as MIL Std. 285	0.5 - 10 GHz	80	90	60	60	70	70
Resistance to tearing in app. kg/cm as ASTM D5035		1.3	1.3	10.7	10.7	8.9	8.9
Linear expansion in app. % as ASTM D5035		9	9	27	27	34	27
Special features		flexible	flexible, corrosion-resistant	flexible, drapable, tear-resistant	flexible, drapable, tear-resistant and corrosion-resistant	flexible, adaptable	flexible, adaptable and corrosion-resistant

Table 1: List of technical and mechanical identifying characteristics of non-wovens and fabrics

this way, the shielding attenuation is increased by an additional approximately 20 dB.

All components coming into contact with the shielding wallpaper must be galvanically compatible, in order to exclude corrosion, as far as possible. When the surfaces have been prepared (ceiling, walls and floor), the RF-shielded connection should be created to the fittings such as doors, mounting plates, filters, etc.. One of the most important shielding elements is the door, since it must be designed for mechanical and shielding technology strength in continuous operation. Fitting the door therefore requires special knowledge and has to be done with care. In addition to this, depending on subsequent use or special requirements such as, for example, the integration of the shielding room within office buildings, hospitals, doctors premises or elsewhere, the doors must also fit into the overall architectural planning, and may not be considered as alien elements. Other important criteria which may be imposed on such a door include supplementary access monitoring, easy opening and closing, and entry into the room without a threshold.

Air feed and air extraction apertures are needed for ventilation or air conditioning, and are built into the shielding as so-called honeycomb chimneys. Provide for connections outside, for example with adapter connecting pieces for the air conditioning equipment provided by the customers, and inside for the connection, e.g. to a no-draught air distribution system. All electrical lines must be fed in through radio interference suppression filters, which are mounted on a filter mounting plate, together with the earthing point, and are let into the shielding. Media or RF ducts are built into an additional, detachable mounting plate provided for the purpose at defined points in an RF-tight manner.

The completion of the interior can now take place. For reasons of shielding technology, no special surface protection

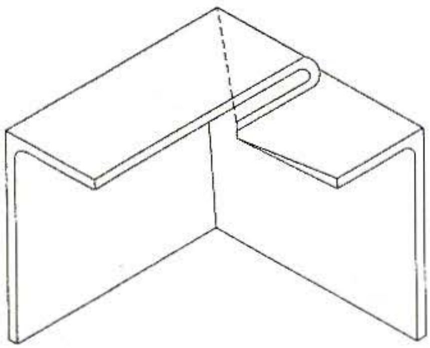

Fig. 2: Expert corner formation using non-wovens

is needed. In areas with above-average atmospheric humidity, or where there is a risk in connection with the formation of heavy condensation or the penetration of moisture, for example in pits - in these rather rare cases it is recommended that nickle-plated copper non-wovens should be fitted as anti-corrosion protection.

Depending on the intended subsequent use or on special requirements, the interior can be completed in the following possible ways, among others:

- Direct painting onto the metallised non-woven using dispersion or latex paints.
- Papering (e. g. fibreglass wallpaper and, if applicable, additional coating) directly onto the shielding
- Incorporating a timber or metal substructure to strengthen the internal structures see Fig. 3 (e.g. sandwich-type plasterboard wall with intermediate layer)
- Lining with absorbers and use as absorber room for electro-magnetic compatibility or aerial application.

There are likewise many formats possible in the floor area:

- Laying down of a wooden floor with an overlay directly onto the non-woven laid out on the floor. For low-stress applications, the overlay alone can even be glued directly onto the non-woven.
- Fitting of an elevated false floor (see Fig. 3)
- For very high loads and stresses (e.g. shearing forces), you are recommended to incorporate an industrially produced coating as per RAL or a concrete floor. If special guidelines are adhered to when the non-woven is installed on the floor, it withstands any kind of structural stress which is found in industry.

2.3. Typical application

Two shielding chambers made from copper non-woven, which have not been in operation for all that long, can be used to show how numerous the uses of this type of shielding technology are, and which points must be taken into account.

In one case, a college wanted to create an aerial measuring room with additional shielding, but had only a limited budget available for this. Instead of burying their heads in the sand because the financial resources were not available for a large 8 x 4 x 4 m chamber made from high-quality steel modules, they took an alternative route. The carcass of the chamber which was subsequently shielded was erected as a structure of wooden bars, planked with plywood panels, and financed from the construction budget of the college. The costs for the pure shielding measures with the copper non-woven and the built-in components could thus be kept comparatively low. Naturally, there were

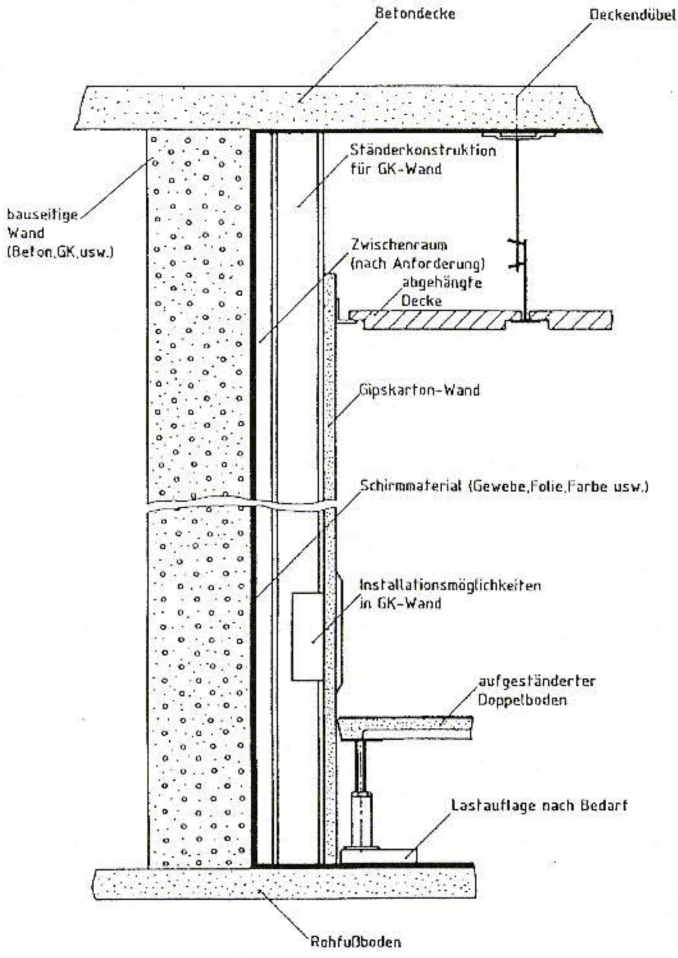

Fig. 3: Ceiling-wall-floor structure of typical copper/ nickel room shielding with internal structure

also reservations in relation to the shielding characteristics. However, the guaranteed values of > 60 dB from 30 MHz to 100 MHz and > 70 dB from 100 MHz to 10 GHz were exceeded by 10 dB on average.

The most important factor at the planning stage of this construction project was close agreement with the user and the construction office. Interfaces had to be defined, tasks and details of execution discussed and agreed, and sometimes clarified with the help of drawings. The acceptance criteria were laid down, in particular the type and extent of the shielding attenuation measurement. And naturally, when questions of detail were examined a little more closely, the intended use after completion namely the application as an aerial measurement chamber from 450 MHz up to far beyond 40 GHz was not to be lost sight of.

A completely different application developed from an enquiry by a customer regarding the possibility of dividing an existing area with complicated geometrical fittings into two smaller shielding

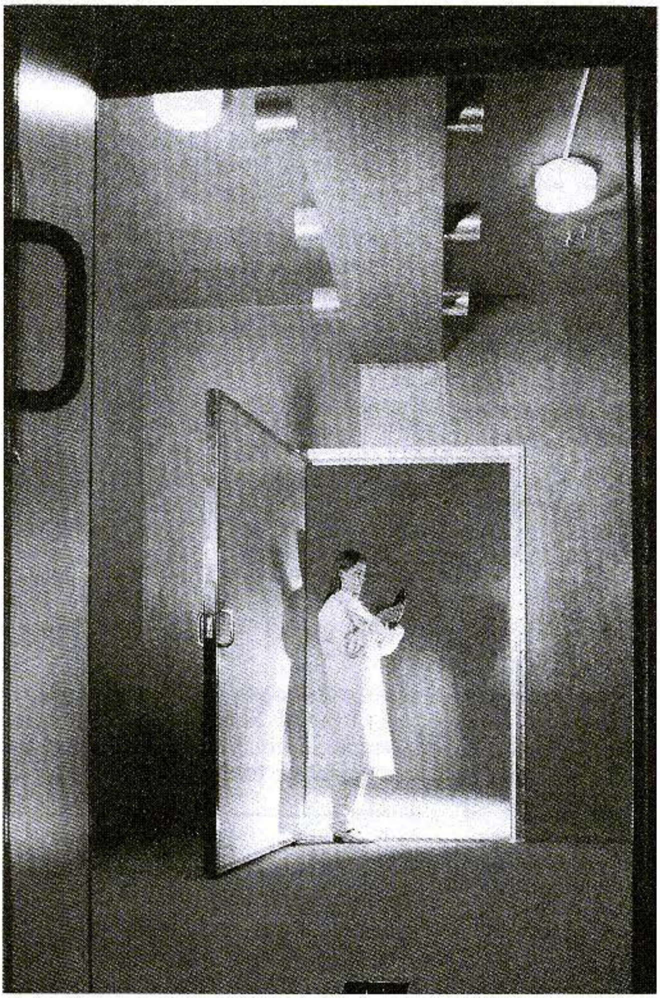

Fig. 4: Copper non-woven area shielding of a double cabin with details of shielding doors and the integration of existing ventilation elements

rooms with a joint dividing wall and doors. Parallel measurements were to be taken here in the mobile frequency range in both parts of the area. Because flexible copper non-woven is outstandingly suitable for the covering and binding-in of complicated geometries, is advantageously priced and in addition has attenuation characteristics of > 70 - 75 dB at 1 GHz, it was particularly attractive for the user. And the photo in Fig. 4 of the room we have just described demonstrates that copper non-woven shielding can look good, even without any additional internal lining.

2.4. Users introduction

The following introduction is intended as a kind of check list to provide support for the increasingly broad spectrum of those whose task it is to find an efficient and price-conscious overall solution for future measurement tasks, or to extend and/or modernise existing equipment. These will predominantly be in the field of aerial, communications and electro-magnetic compatibility metrology, as well as in the data security and medical technology areas a solution involving shielding technology, with the associated (absorber) internal structure and final certification. The following questions should be answered as completely as possible in the preliminary stages:

a. Clarification of technical requirements for shielding, i.e. what shielding attenuation should/must be attained, what upper and lower limiting frequencies are important, where might it be possible to reduce the level of the requirements?

b. What construction measures are necessary for the application of the copper non-woven?

Who is to be responsible for this, and for what subsequent use should/must the planning already be laid down?

c. Is there a compulsory requirement for a flexible structural method such as, for example, execution as movable modular screw-on panels?

d. Use of area and importance of measurement tasks i.e. use of monitoring technology, data security, medical technology, communications technology, aerial technology, electro-magnetic compatibility, etc. What is the precise measurement task. How precisely or in conformity with the standards would I like to / must I measure (only measurements of a certain quality, or keeping up with developments or correct to a standard). What are the primary measurement tasks, with the associated frequency ranges, measurement intervals, options, measurement guidelines for products, test piece dimensions, weights,

e. Doors (size, number, locking, special features as regards fittings and model, etc.)

f. Ventilation (air conditioning equipment, ventilators, power loss, etc.)

g. Filters (network filters, digital/analogue data filters, control filters, telephone filters etc., U, I, f, number of circuits, adaptation etc.)

h. Electrics and Lighting (plugs and sockets, 1ph., 3ph., luminous intensity, emergency cut-out, etc.)

i. Mounting plates and RF ducts (size, number, N, BNC, LWL, air etc.)

j. Floors (wooden floors, false floors, overlays conducting, anti-static, groundplane etc.)

k. Acceptance criteria/measurement (shielding attenuation measurement as per MIL Std. 285, NSA 65-6, EN 50147 Part 1 and absorber room grade as per IEC 1000-4-3 or EN 61000-4-3, ANSI C 63.4 or EN 50147 Part 2 and so on)

l. Absorber design (layout and performance selected by manufacturer in accordance with technical requirement of user)

m. Special equipment and accessories (e.g. internal lining, acoustic covering, intercom, lacquering , electro-magnetic compatibility, turntable, etc.)

3. Summary and outlook

Shielding made from metallised nonwovens and fabrics can be a favourably priced alternative in comparison to the standard shielding techniques using plates. However, the decision to use this technique must be taken after taking the traditional technologies into consideration as well, together with any preliminary work of your own likewise in connection with costs.

It is particularly important to obtain a basic agreement in good time between the future user, the planner and, if applicable, any planners interposed between them. It must be clear for the user what can be achieved from the copper non-woven using the shielding technology, where the limits lie and what alternatives there are. It is just as important for a standardised, homogenous room concept to be developed. This service can be provided only by a company which has shielding technology as one of its core activities, but which is also well versed in the classic technologies of modular steel construction and absorber technology for high-frequency applications.

The trend in relation to the construction of new aerial measurement chambers and security-related rooms (but also to the retroactive shielding of existing premises) is clearly going in the direction of an efficient, and thus favourably-priced, shielding using copper non-wovens.

4. Literature references

[1] Aichele H. L.: Architectural shielding, EMV-ESD, issue 1/97

[2] Aichele H. L.: Room protection concept for information and data security,

EMV Düsseldorf 2000, Proceedings, pp. 731 - 736

[3] Hemming L. H.: Architectural Shielding Handbook, IEEE Press, 1992

Gunthard Kraus, DG8GB

Internet Treasure Trove

Ansoft

A change is becoming apparent in the sales philosophy of producers of HF and microwave CAD software. Up until now, free test versions have been on offer which either operated for only a limited time (usually 10 to 30 days) or restricted performance to such an extent that even testing the product was just no fun at all. But now there has been a sudden increase in the number of student versions, which are certainly still restricted with regard to their power but now make it possible to handle more demanding projects. Ansoft have also changed their views and have put a free student version of Ensemble on the Internet. This name conceals the well-known microwave simulation program Harmonica and the system-planning program Explorer. One excellent feature is that before downloading you are given a precise list of the extent of the restrictions introduced and of their locations and you can still download 75 megabytes ...

Address: http://www.ansoft.com

IrfanView32

What PC user hasnt come across the problem of a graphics file which the systems on his / her computer cant open? This loophole has now been closed by IrfanView32. With this new freeware you can now open and convert any and I really mean any graphics in even the craziest formats. One further detail is that you can also use it to set up a slide show. So we should all have this useful software handy on our PCs.

Address:
http://newlife-win98.server101.com/irfanview_info.htm

Maury

A well-known name, famous as manufacturers of high-quality and expensive microwave test rigs. Its always worth visiting this homepage for, in addition to data sheets for all their equipment, you can also find some very carefully written application notes. They not only describe the precise applications for the equipment in connection with the most frequent measurement tasks, but they also lay out the theoretical principles involved once more and present them clearly something users can always profit from. We can particularly recommend the two introductory brochures covering the measurement of intermodulation and noise.

Address: http://www.maurymw.com

JFW

Anyone looking for a manufacturer of attenuators and switches of all sizes for the microwave range should download the entire catalogue of this company (over 150 pages) from their homepage in pdf format . In addition to basic introductory information, there is an enormous range of products available there, and youll undoubtedly find something to solve your problem.

Address: http://www.jfwindustries.com

Test & Measurement World

The homepage of this technical review has an interesting Article Archive, which you can access through the search function in the usual way. There are also some publications on HF and microwave technology in html or pdf format ready for downloading. This service is free and its always worth going back from time to time to see whats new.

Address:
http://www.tmworld.com/index.htm

dB

Everyone knows this measurement unit and so its no surprise that a company should have chosen it as its name. Here you can find microwave amplifiers and oscillators for sale and, as is so often the case, there is also some good technical information available on the homepage.

Address: http://www.microwavedB.com

EZNEC

The mist is slowly clearing away with regard to short-wave and VHF aerial simulation, and it has now become very clear whats worth buying or testing. Anyone going into aerial simulation programs in more depth will know that all sophisticated software uses the codes from NEC2 and / or NEC3 as the core for the simulation. NEC4 is expensive perfection, which can be used even for simulation in the near field and close to the ground. But it is very expensive and moreover it is still an American military secret.

In the amateur sector, the EZNEC program from Roy Lewallen, W7EL, has recently been earning a very good name for itself. For US $ 80 it certainly offers the best compromise currently available between price and accuracy (which again is dependent on the number of individual elements with which the aerial structure can be modelled). An easily comprehensible operator surface, an update service on the Internet and something completely new! a manual in Word format which can be printed out have put this program out in front for the moment.

Anyone who met Roy and talked to him at HAM-Radio 2000 (he had a little stand of his own there) will know how enthusiastic he is and how committed to this sector. Naturally, you can also download a demo version free for testing from the homepage before you spend any money.

Address: http://www.eznec.com

Index of Volume 32 (2000)

Article	Author	Edition	Pages
ANTENNA TECHNOLOGY			
Microwave Multi Band Feed Second Generation	Freddy de Guchteneire ON6UG	2000/3	130 - 136
A Quadrifilar Helix Antenna for Orbiting Weather Satellites	Dipl. Ing. D Burchard	2000/4	201 - 214
AMATEUR TELEVISION			
5.7 GHz ATV Converter	Helmut Neidel, DL1IN	2000/1	24 - 29
Supplement to Article on 5.7 GHz ATV Converter	Helmut Neidel DL1IN	2000/3	186 - 187
2 m BAND			
SSB / CW Transceiver for 144 MHz	Wolfgang Schneider DJ8ES	2000/1	35 - 58
Control of SSB / CW Tranceiver (VHF 2M) for 144 MHz	Bernd Kaa DG4RBF	2000/1	59 - 63
Low Pass Filter for 2 m and 70 cm	Gerhard Schmitt DJ5AP	2000/3	156 - 166
Low Pass Filter for 2 m and 70 cm - Part 2	Gerhard Schmitt DJ5AP	2000/4	232 - 240
70 cm Band			
Low Pass Filter for 2 m amd 70 cm	Gerhard Schmitt DJ5AP	2000/3	156 - 166
Low Pass Filter for 2 m and 70 cm - Part 2	Gerhard Schmitt DJ5AP	2000/4	232 - 240

23 cm BAND

Microwave Multi Band Feed Second Generation	Freddy de Guchteneire ON6UG	2000/3	130 - 136

6 cm Band

Microwave Multi Band Feed Second Generation	Freddy de Guchteneire ON6UG	2000/3	130 - 136

3 cm BAND

Microwave Multi Band Feed Second Generation	Freddy de Guchteneire ON6UG	2000/3	130 - 136

MEASURING TECHNOLOGY

FC4000 Micro Controlled Double Frequency Counter Up to 4 GHz	Bernd Kaa DG4RBF	2000/1	10 - 23
Frequency Divider Up to 4GHz	Bernd Kaa DG4RBF	2000/1	2 - 9
Correction to Marker Counter for Spectrum Analyser in 4/1999	Matjaz Vidmar S53MV	2000/1	30
A Vectorial Aerial Impedance Meter For The Short Wave Range	Dr. Ing. Jochen Jirmann, DB1NV	2000/2	66 - 88
High Precision Frequency Standard for 10 MHz	Wolfgang Schneider DJ8ES	2000/4	194 - 200
Tracking Generator for Microwave Ranges (1.7 to 13GHz)	Carsten Vieland DJ4GC	2000/4	215 230

FUNDAMENTALS & NEW TECHNOLOGY

The I/Q Modulator Described	Dipl. Ing. J V Parpart	2000/2	106 - 117
Logarithmic Amplifer Up to 500MHz With AD8307	Wolfgang Schneider DJ8ES	2000/2	119 - 124

Teflon or Epoxy Base Material ?	Harald Fleckner DC8UG	2000/2	90 - 104
Teflon, Epoxy or RO 4000 Base Material	Harald Fleckner DC8UG	2000/3	152 - 155
Earthing in HF and Microwave Circuits ... A Case for Puff	Gunthard Kraus DG8GB	2000/3	167 - 178
Circulators and Ring Hybrids	Wolfgang Borschel DK2DO	2000/3	179 - 185
High Precision Frequency Standard For 10MHz	Wolfgang Schneider DJ8ES	2000/4	194 - 200
Active Directional Coupler	Dipl. Ing. J V Parpart	2000/4	241 - 249
Circulators and Ring Hybrids Readers Ideas on Article in 3/2000	Wolfgang Borschel DK2DO	2000/4	250 - 251

WEATHER SATELLITE RECEPTION

A Quadrifilar Helix Antenna for Orbiting Weather Satellites	Dipl. Ing. D Burchard	2000/4	201 - 214

MISCELLANEOUS

A Simple TNC for Megabit Packet Radio Links	Matjaz Vidmar S53MV	2000/3	137 - 151
Internet Treasure Trove	Gunthard Kraus DG8GB	2000/3	188 - 190
Internet Treasure Trove	Gunthard Kraus	2000/4	252 - 254

MICROPROCESSOR TECHNIQUES

A Simple TNC for Megabit Packet Radio Links	Matjaz Vidmar S53MV	2000/3	137 - 151

POWER SUPPLIES

Supplement to Article on 5.7GHz ATV Converter	Helmut Neidel DL1IN	2000/3	186 - 187

Carsten Vieland, DJ4GC

Tracking Generator From 1 MHz to 13 GHz for Spectrum Analysers

Expanding an existing spectrum analyser, many of which have very high-precision amplitude and frequency, into a network analyser using a tracking generator is one of the daydreams of many high-frequency technicians. This article is intended to show how such a commercial object can become a DIY project for a modest outlay.

1. Project Description

The analyser equipment in question consists of four individual components, some of which are more expensive than others.

1. Tracking generator for the basic frequency range of the analyser, which here is 10 MHz to 1.8 GHz;
2. Tracking generator for microwave ranges; here, 1.7 GHz to 13 GHz;
3. Resistive reflectometer measuring bridges for the microwave ranges;
4. Transverter for expanding the network analyser into the low-frequency range, here 100 Hz to 50 MHz

2. Network Analysis with Tracking Generator

RF technicians regard spectrum analysers rather as doctors do X-ray machines. They make it possible to obtain qualitative and quantitative information on frequencies, amplitudes, subsidiary signals, modulation sidebands, noise fractions, harmonic fractions, spurious oscillations, and many other aspects of active signal sources.

A spectrum analyser is really a highly selective receiver with extremely wide frequency tuning. The result of a frequency scan is not made audible in the loudspeaker, but is documented as an amplitude log on the screen. Modern units have digital image processing with the option of printing out a hard copy.

Large amounts of equipment, most of which is fairly old e.g. from Hewlett Packard, Tektronix, Advantest, Alltek, Marconi, etc. have recently come into the hands of committed radio amateurs. Even if you once needed a mortgage to buy some high-end creations, you can now, in certain circumstances, get them on the second-hand market for the price of a good transceiver. However, we

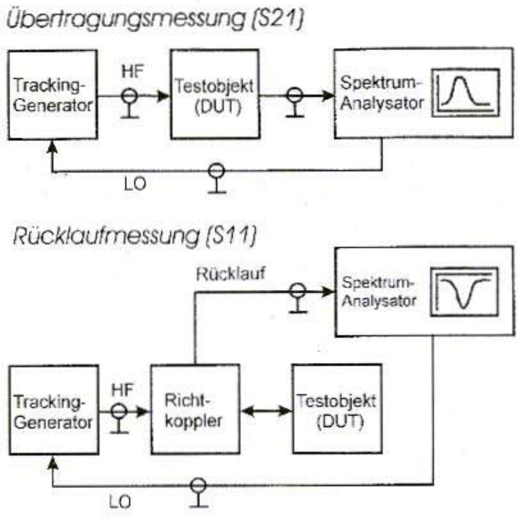

Fig.1: Principle of network analysis for transmission (S21) and reflection (S11)

(Übertragungsmessung = Transmission measurement, Testobjekt = Test object, Spektrum-Analysator = Spectrum analyser, Rücklaufmessung = Reflection measurement, Rücklauf = Reflection, Richtkoppler = Directional coupler)

must qualify this by pointing out that the spectrum analyser itself can still investigate only external signals.

A further dimension in RF measurement technology can be opened up through the tracking generator equipment, namely measurements on components which are not themselves oscillating but which are frequency-dependent. For example, the transmission behaviour of filters, selective matching measures, attenuation values, amplifiers, power splitters, repeaters, line resonators, stubs and much else besides can be shown in dB-linear scale on the screen at the same time as the frequency response log, and can be printed out.

Moreover, an additional directional coupler makes it possible to obtain information on the matching and / or return loss of such components within a 50-Ohm system.

3.

Principle of Tracking Generator

The generator presented here corresponds to the expanded HP8569A or B Hewlett-Packard analyser. The frequency and signal levels are tailored for this equipment but must, in principle, be able to work together with all spectrum analysers which have an LO output.

The frequency coverage of the measurement system covers the amateur radio bands over a range extending from 10 MHz to 13 GHz, from shortwave transmissions all the way to the X band. The four overlapping frequency ranges correspond to those of the spectrum analyser and, at least on the HP 8569, are also switched centrally from there. The entire frequency range is available at only one

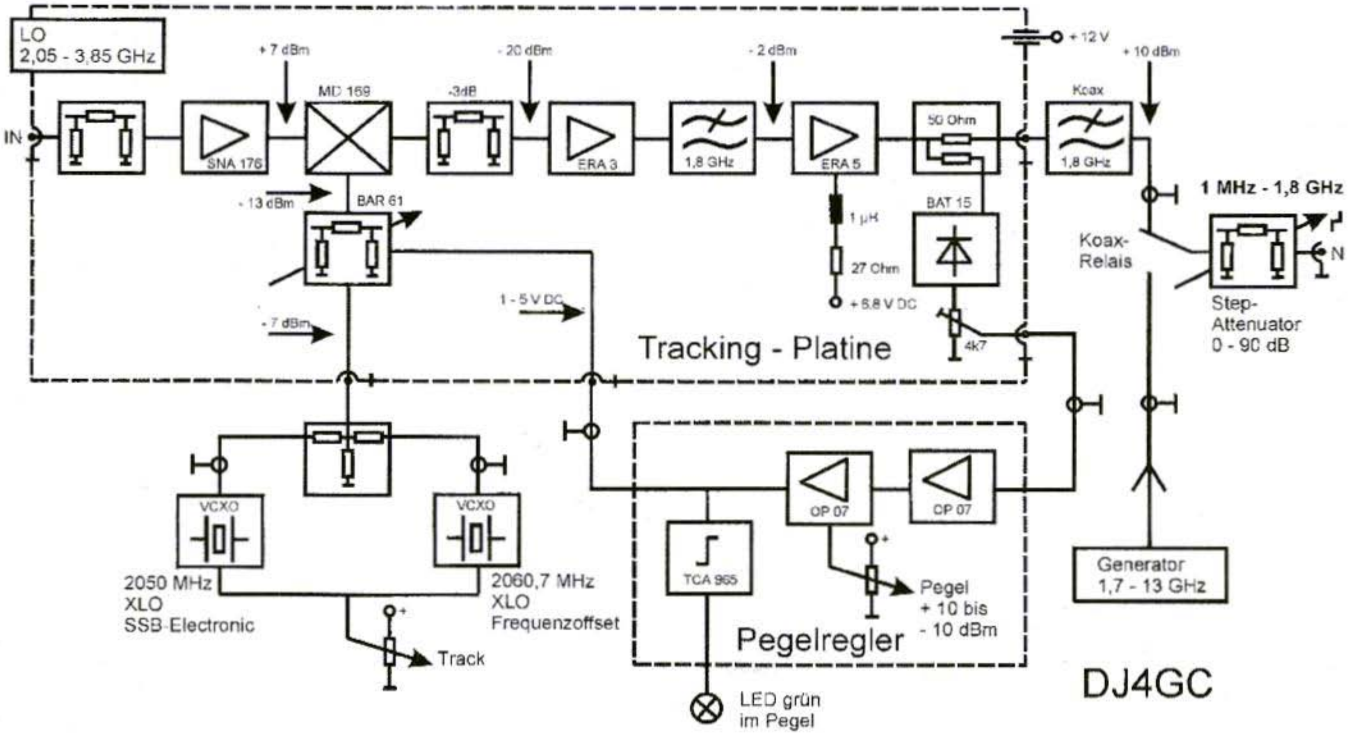

Fig 2: Block Diagram of Tracking Generator for 1 MHz to 1.8 GHz
Tracking Plantine = Tracking PCB, Pegelregler = Level Control, LED grun im Pegel = Green LED within level

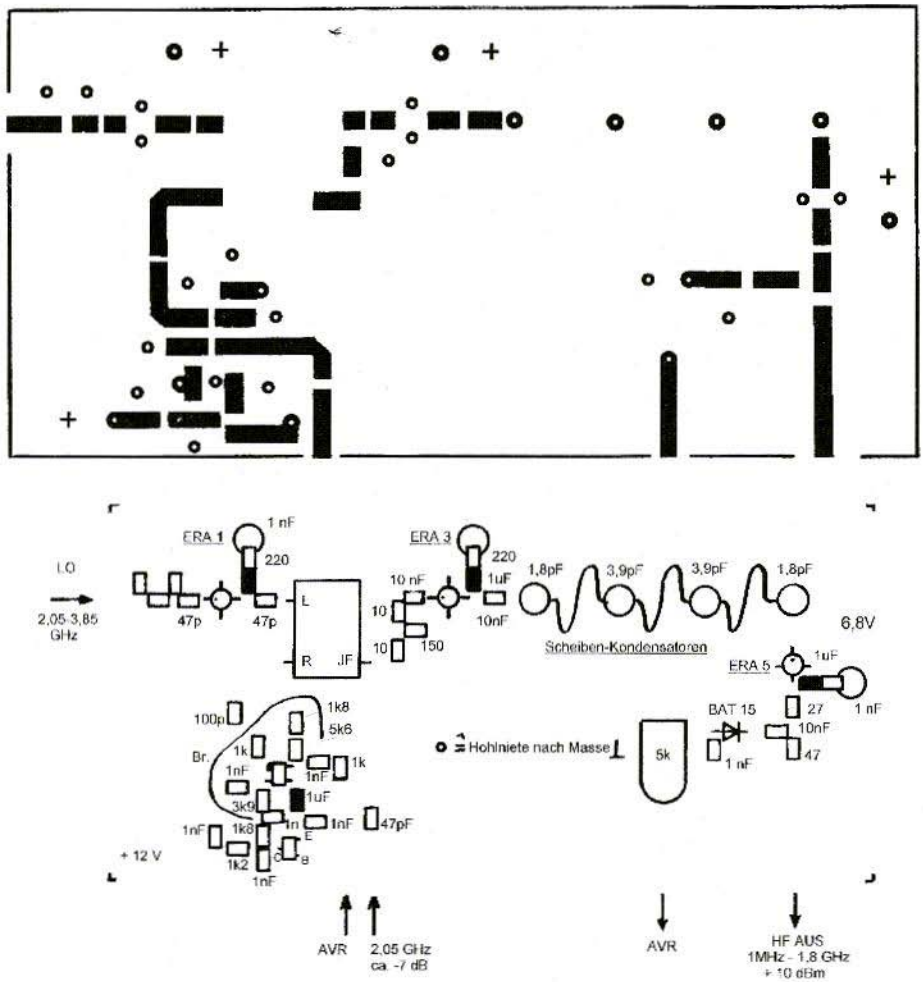

Fig. 3: Layout (100 mm. x 50 mm.) on RT-Duroid, Ultralam or the like (Scheiben-Kondensatoren = Disc capacitors, Hohlniete nach Masse = Full tubular rivets to earth)

N-socket of the generator. In principle, a separate control system for this equipment is therefore not necessary.

In every tracking generator (Fig. 2), a quartz-stabilised oscillator, which corresponds precisely to the first intermediate frequency of the analyser used, is added to the frequency-determining oscillator signal (LO) of the spectrum analyser. The sums or, in general, the differences of the two oscillations then give the analysis and / or tracking frequency precisely.

However, a high degree of isolation,

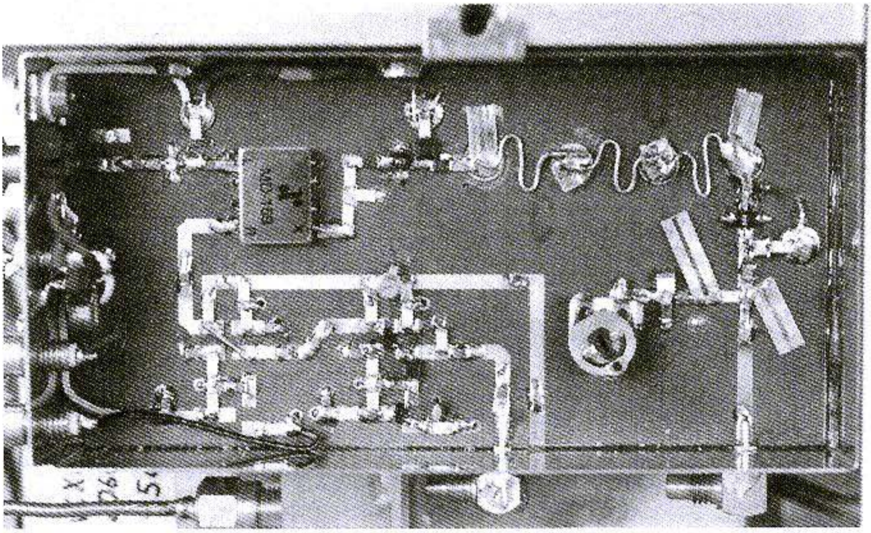

Fig. 4: Generator printed circuit board with PIN controller. The metal lugs serve to optimise the frequency response.

approximately 90 to 100 dB, must be created between this auxiliary frequency oscillator in the tracking generator and the input of the intermediate-frequency amplifier in the spectrum analyser. Such high degrees of de-coupling are difficult to obtain, because the two mixers in the separate units are powered from the same frequency-determining LO source. The degrees of de-coupling of the ring mixers involved (LO intermediate frequency) are nowhere near adequate here. If the isolation values are poor, any residual signals appear on the analyser display, and reduce its dynamic range and sensitivity.

In order to prevent even this intermediate-frequency oscillator signal of the tracking generator from creeping over the Lo circuit into the spectrum analyser, a so-called isolation amplifier is interposed.

Instead of the expensive broad-band circulators and / or one-way circuits which were previously standard, each with approximately 20 dB return loss, here two advantageously-priced modular broad-band amplifiers (ERA 1) are wired up with two attenuators. The forward amplification is sufficiently high at approximately 0 dB over the entire range (approximately 2 to 4 GHz). At the same time, the isolation (attenuation in return direction) still attains 60 dB. Thus, when the tracking generator is connected, the basic noise on the analyser display remains unchanged.

4. Basic Frequency Range From 10 MHz to 1.8 GHz

The measurement system presented consists in practice of two individual units independent of one another, with some individual components (such as, for example, housings, power supply, DC section of level regulator, together with output-side step attenuator) being operated in common.

The part of the equipment presented

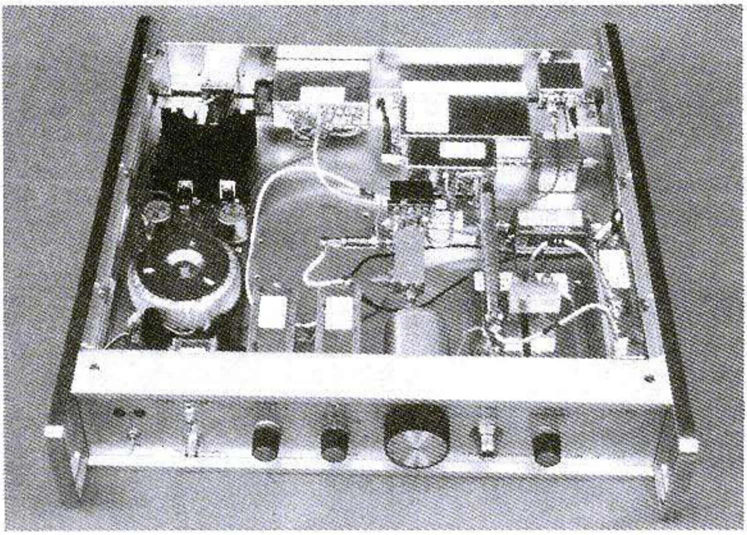

Fig. 5: Internal view of the completed unit

here operates from approximately 1 MHz, but the entire measurement system does not attain high quality until the basic frequency processing range of this spectrum analyser, i.e. from 10 MHz to 1.8 GHz. Thanks to modern, advantageously priced broad band components, this tracking generator can be constructed with amazingly little material and soldering (Figs. 3 and 4). The frequency coverage is suitable for most situations specific to amateur radio.

The frequency determining LO signal of 2.05 GHz to 3.85 GHz supplied to the rear panel of the analyser is mixed with a crystal-stabilised oscillator of 2.05 GHz. This signal injection lies precisely at the first intermediate frequency of the spectrum analyser, and can be adjusted somewhat to balance out a slight frequency drift (VCXO). The difference frequency at the output of the mixer then again lies at the input frequency of the frequency analyser. This combination frequency is amplified again after the ring mixer (here: MD 169 from Anzac), passes through a three-phase low-pass filter with a limiting frequency of 1.8 GHz, and is then amplified once more. At the output of the final stage, a low-barrier Schottky diode (here a BAT 15) detects the level value and feeds it to the level amplifier. This DC voltage amplifier should keep the high-frequency level constant, with the aid of a PIN diode attenuator this corresponds to an internal resistance of ZERO (!). To match the impedance to the circuit system, a chip resistance of approximately 50 Ohms is series wired to the final amplifier. This thus also means the output socket has to have an impedance of 50 Ohms. Even if this resistance uses up a lot of the precious output, it is necessary for matching for correct measurements. Nevertheless, a maximum of + 10 dBm is available for specific measurements with high levels over the entire frequency range.

An electronically controllable variable attenuator, consisting of three PIN diodes in an SMD housing (BAR 61) ensures an automatically levelled output level.

The fixed frequency oscillator at 2.05 GHz is a commercial product. This is a

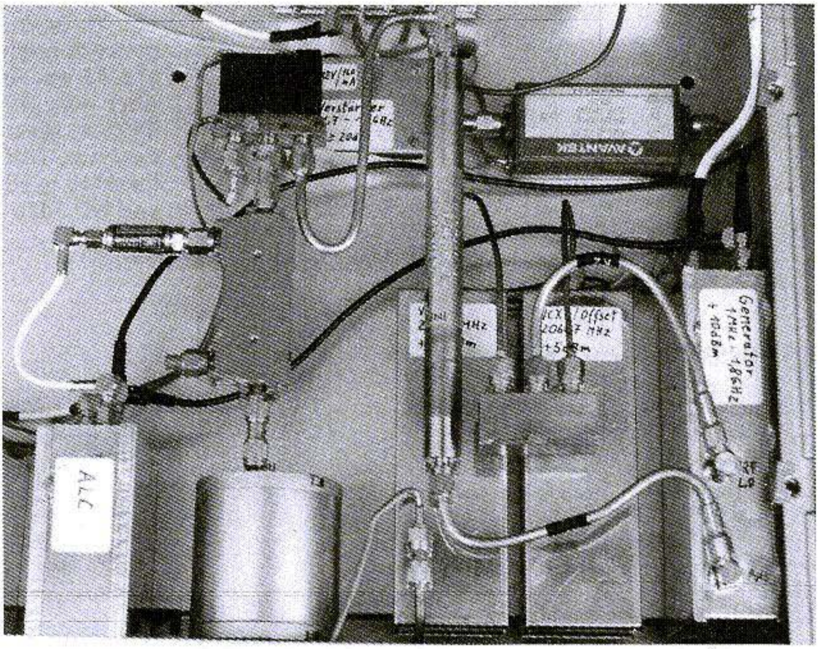

Fig. 6: Section with essential components for basic frequency range up to 1.8 GHz. The long, round, coaxial low-pass filter is in the middle of the picture. The thick, round metal body contains the switchable attenuator. The coax relay, together with the SMA directional coupler and the detector, forms part of the SHF section.

universal oscillator processing (XLO) [1] from SSB-Elektroniks transverter series. This circuit, which can be adjusted for output frequencies between 1.5 and 2.5 GHz, stands out through a high degree of spectral purity. The numerous publications conçerning similar circuits make it superfluous to go over it again at this point. An additional capacity diode in the oscillator circuit of the upper harmonic oscillator creates a pull-in range of several kilohertz. Thus measurements can be better synchronised, with narrow analysis band widths, i.e. low residual noise and high dynamics.

Thanks to modern, advantageously priced broad band modules, the entire high-frequency processing system (without an XLO) fits into a small tinplate housing measuring 53 mm. x 100 mm. (Fig. 4). In principle, the level amplifier and its four ICs could also be fitted in using SMD technology. Some improvement in the measurement dynamic (a few dB) can be obtained by externally fitting a coaxial 1.8-GHz low pass filter with a very high cut off gradient (Fig. 6). Thus the harmful injection signal on the first intermediate frequency (2.05 GHz) at below 90 dBc in the output spectrum is no longer displayable.

Thermally measured, the amplitude of the tracking signal varies by a maximum of only ± 0.5 dB around the set value. This amplitude ripple is thus less than the actual frequency response of the spectrum analyser. However, the HP 8569, like most of todays equipment, has digital image conversion, through

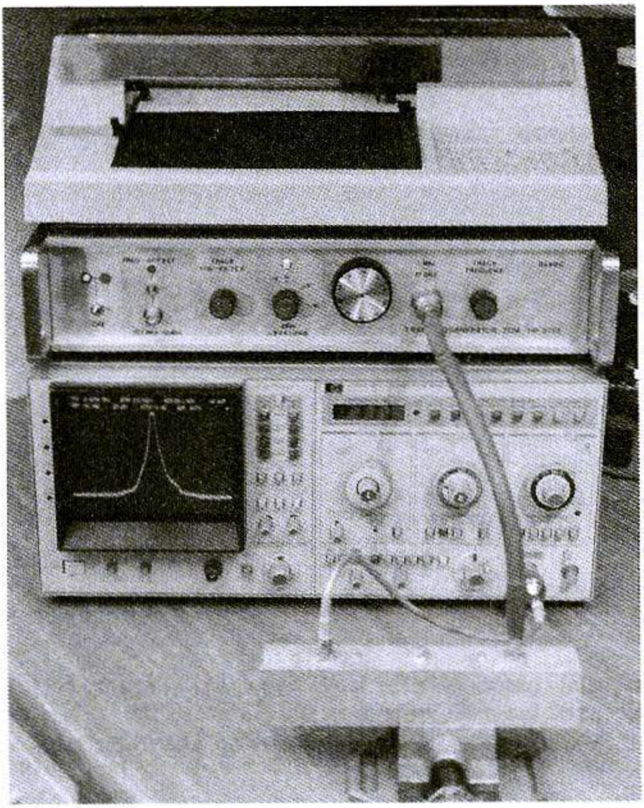

Fig. 7: Scalar network analysis of a 13-cm. finger filter with HP 8569(A) spectrum analyser, DIY tracking generator and HP 7470 A plotter

the useful special function of standardisation, which enables the reference level (0 dBc) to be made into a straight line in the centre of the screen. In this way, all the units inherent ripple is eliminated. Although this is only a digital image conversion, the interpretability of the print-out is recognisably improved, especially in the microwave ranges. Amplifications appear above the screen centre and attenuations below.

The HP 8569 is equipped with an HP-IB interface. Even if this operates in a slightly leisurely fashion by present-day standards, splendid hard copies can be created with the help of an HP 7470A plotter. The documentation appended was put on paper in this way.

5.

Details of Signal Processing

The four MD 169 (Anzac) ring mixers used here are certainly relatively expensive, but they and all their data are within the frequency range of this measurement system. It was obviously developed for similar applications. Other mixers such as, for example, TFM 42 from Mini-Circuits, certainly display a slightly higher amplitude ripple at the output, together with a rather higher LO disruptive breakdown, but apart from this they are also suitable.

The resistive attenuator at the LO input of this printed circuit board is calculated in such a way that the mixer approxi-

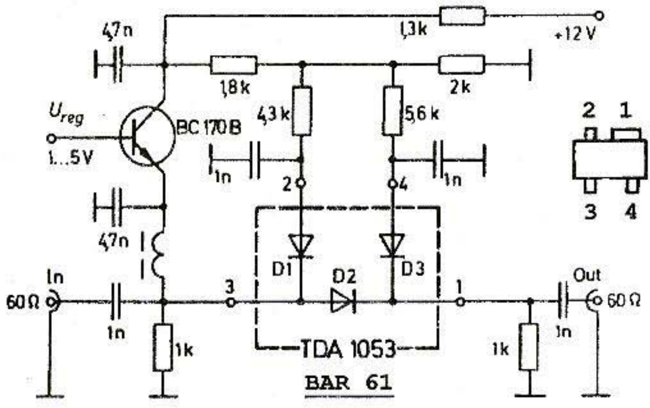

Fig. 8: Circuit of PIN diode controller for adding frequency of 2,050 MHz. The peripheral circuit with a transistor developed by ITT is suitable for all types of PIN diodes

mately attains its rated level of + 7 dBm. The mixer output level at the intermediate frequency port is approximately 30 dBm. This very low amplitude is a favourable compromise between the dynamics of the signal (useful signal to basic noise) and the lack of distortion. It is well known that the distortion spectrum of ring mixers, which grows rapidly with relatively high output signals, generates so-called amplified spurious signals (non-harmonic ghost signals), which can no longer be removed by filtration measures.

The PIN diode regulator used here (BAR 61 from Siemens) provides an infinitely variable attenuation of up to 20 dB. The design parameter of this newly developed SMD module is above 30 dB at 2.05 Ghz. It is thus considerably better than the old TDA 1065, UTF 025, or 3 BA 379 units, which were designed for the UHF television ranges, these still display approximately 10 dB of control coverage at this frequency, but with poor matching. This is naturally not sufficient for exclusive levelling at a constant output level.

In a departure from normal practice, it is not the fixed-frequency, but the variable LO signal from the spectrum analyser which is high-gain and drives the mixer.

Since ring mixers react very tolerantly to level inputs at the LO port, relatively small levels break through to the intermediate-frequency output.

At this point, the useful signal is subjected to a ripple of maximum 2 dB over the entire range. The 3 dB attenuator at the mixer output represents a kind of circuit hygiene. It certainly feeds on the useful signal but, on the other hand, it ensures a certain level of matching between the mixer and the amplifier.

The three-stage low-pass filter is designed for the highest output frequency of the generator. At the limiting frequency of 1.8 GHz, the input and output capacitors have a value of 1.8 pF. The two central capacitors have double this value 3.6 pF. Suitable disc capacitors available also small-format chip capacitors are also suitable. The capacitors can be mounted on the board surface but the disadvantage is that their height has a tendency to cause surface resonances. The three inductances can each be calculated to be 4.4 nH, they consist of 0.5 mm. thick silver-plated wire, which is bent in an S-curve with a length of

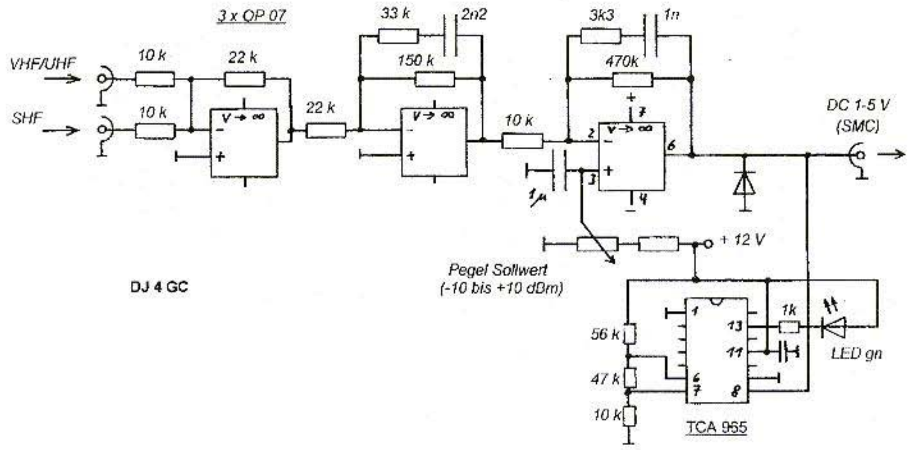

Fig. 9: Control amplifier for controlling PIN diode controller with two inputs and window comparator for level monitoring

(Pegel Sollwert = Rated level value, Bis = To)

approximately 8 mm. (see Fig. 4).

To reduce the ripple levels in the transmission band, the inductances are slightly adjusted by being slightly bent, and the capacitances with the help of metal lugs.

Naturally, a computer-optimised, self contained filter could also be used. A low-pass filter is necessary because the other signals at this point, such as image frequencies or the LO affect the automatic level control and would lead to considerable control errors.

To terminate this low-pass, the final stage must display a continuously good input matching. This demanding task is undertaken by the monolithic ERA 5 broad band amplifier from Mini-Circuits which gives a guaranteed +10 dBm output with a 50 ohm input impedance. Thus even ring mixers can be measured with this equipment.

For practical measurement operation a level of approximately 10 dBm is advantageous. In these circumstances, the tracking generator generates 0 dBm with low distortion. To improve the broad band output matching, 10 dB attenuation is used from the step attenuator. The power then remaining, of 100 micro Watts (- 10 dBm) will admittedly not make any DX connections possible, but is a signal at the limit of the spectrum analyser receiver. Depending on the level set, and on the analysis bandwidth, a measurement dynamic of at least 70 dB is available on the display. With skilled use of the measurement level, with the help of the two stage reducers, dynamic differences of markedly more than 100 dB can be measured.

The wobble measurements which were previously used, with broad-band diode receivers, give quite considerably lower dynamic values. For very broad-band measurements, the detector diodes also generated harmonic distortions, which produced false readings.

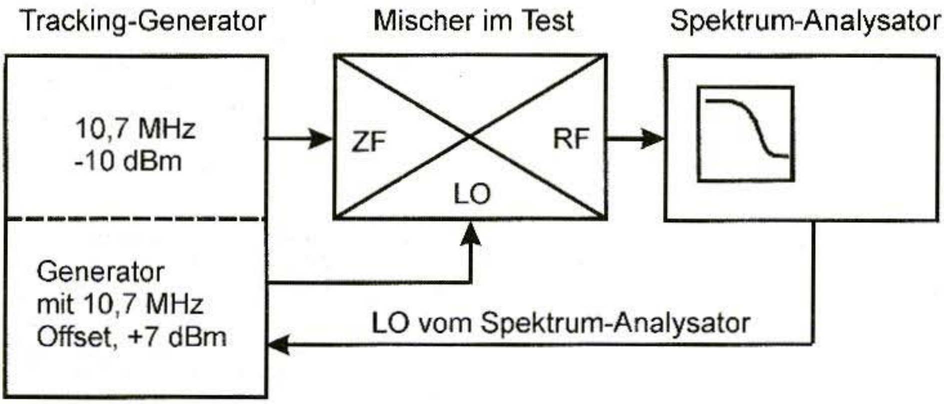

Fig. 10: Measurement of conversion loss readings (gain values) from frequency mixers

(Mischer im Test = Mixer in test, Spektrum-Analysator = Spectrum analyser, Mit = With, Vom Spektrum-Analysator = From spectrum analyser, Messung

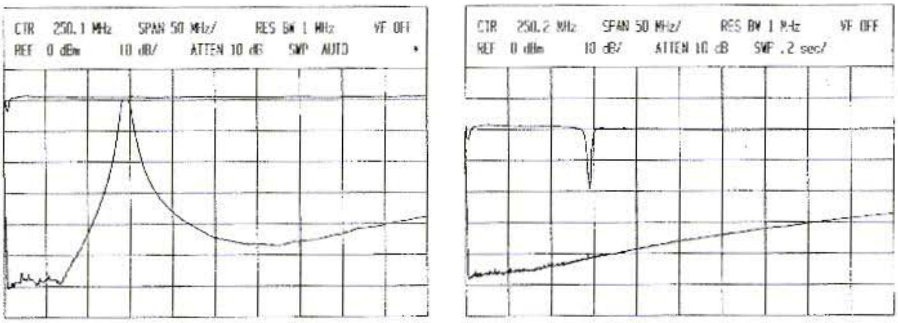

Fig. 11: Measurement on an LC two-circuit filter for 145 MHz:

a) S21 transmission behaviour from 0 500 MHz

Top curve: reference level 0 dBc

Bottom curve: filter curve (attenuation in app. 1 dB transmission band)

b) S11 reflection behaviour

Top curve: reflection loss of two-circuit filter (app. 20 dB)

Bottom curve: sharpness of directivity of measurement system

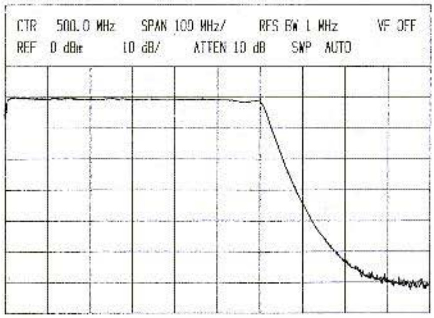

Fig. 12: Coaxial low-pass filter with a limiting frequency of 600 MHz. Up to 800 MHz, the attenuation increases by 50 dB.

6.

Control Amplifier

The circuit of the PIN diode controller, developed back in the seventies for television purposes, requires DC control of between approximately 1 and 4 Volts (Fig. 8). Its design is equally suitable for all low-signal PIN diodes. The rectified voltage at the diode port of the generator printed circuit board is measured and kept constant, as far as possible, due to the automatic control. Approximately 100 mV rectified voltage can be expected for an output level of 1 mW (0 dBm).

A three-stage OP amplifier processes the control voltage. Although in principle this is a DC voltage, level differences during rapid wobbling must be controlled within a few micro-seconds. On the one hand, such control paths must be rapid, but on the other hand they must not bring dynamic errors due to over shoots. The RC frequency response compensation on both OPs prevents both sorts of anomalies.

A summing amplifier at the input processes the detector level of both the VHF / UHF section and the SHF generator, without switching. The 4.7 k-Ohm trimming potentiometer at the detector output of the generator printed circuit board described here makes it possible for the level to be matched to the SHF section.

An externally operated potentiometer

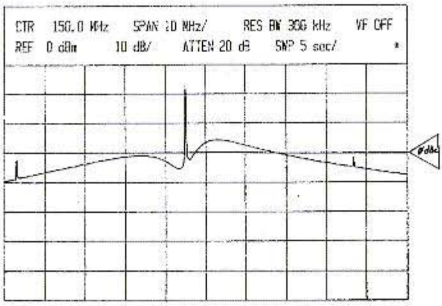

Fig. 13a: Highly-selective 2-m. pre-amplifier with quartz resonator wide selection between 100 and 200 MHz

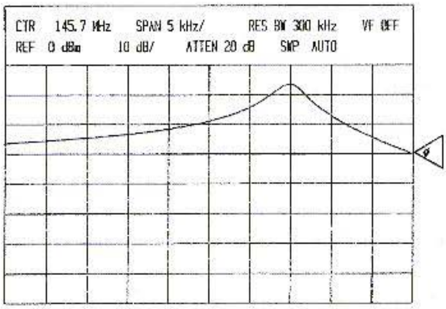

Fig. 13b: Same amplifier at high frequency resolution. The main resonance of the quartz is only 3 kHz wide.

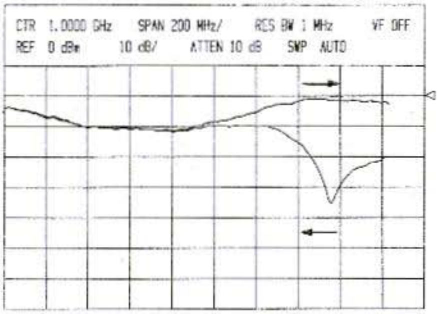

Fig. 14: Transmission behaviour of a commercial directional coupler in transmission and non-conducting directions from 1 to 1.8 GHz. The optimal operating frequency is 1.6 GHz.

pre-sets a freely adjustable DC reference value with which the rectified voltage, already amplified, can be balanced. Since no frequency dependency of any kind has to be taken into account at this point, the potentiometer can be directly calibrated in RF output levels (- 10 dBm to + 10 dBm). With only very slight deviations, this scaling also applies to the microwave ranges for which this DC controller is also used.

A window comparator (TCA 965) monitors that the pre-selected output is being maintained. All that this requires is for the DC voltage fed to the PIN diode controller to be monitored to ensure that the limiting values are maintained. Voltage values of between 1 V and 4 V mean that the PIN diode controller is being controlled within its rated range; this is indicated by a green LED. Above approximately 4 Volts, the LED goes out and alerts the user to a malfunction. The same thing happens if the voltage fall below approximately 1 Volt.

7.

Measurements on Frequency Mixers

This tracking generator has an interesting additional option. The fixed-frequency oscillator required for mixing at the first intermediate frequency of the spectrum analyser used can be switched up to 10.7 MHz upwards, i.e. from 2,050 MHz to 2,060.7 MHz. This task is undertaken by a second XLO [1]. This makes it possible to investigate the attenuation and frequency response of frequency converting four pole networks such as ring mixers. The two oscillators are brought together into a small coaxial housing through two 50-Ohm resistances onto a summation line (Fig. 6). A crystal-stabilised 10.7 MHz oscillator with a calibrated level of 10 dBm is incorporated into the equipment as an auxiliary signal. It is automatically activated when the frequency shift is switched on.

Naturally, only mixers, or else complete converter systems, with intermediate

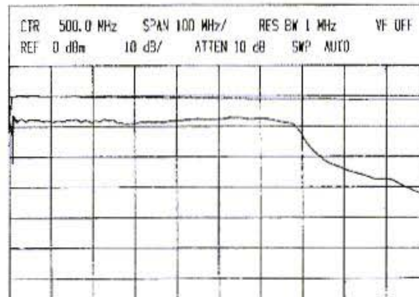

Fig. 15: Conversion loss of a 500-MHz ring mixer in range between 0 and 1,000 MHz. The top curve indicates the reference level of 0 dBc.

ports which can take 10.7 MHz can be measured. The maximum tracking output level of + 10 dBm is sufficient for most ring mixers as a direct oscillator signal. The cost of a second XLO is considerable, of course; moreover, it is needed only for mixed systems. However, this measurement option, which is met with nowhere else except in high-end network analysers, provides valuable insights into the field of RF. technology when required.

In Fig. 15, the mixed attenuation curve for the M1B ring mixer (Watkins & Johnson) is documented, as an example, over a range of 0 to 1,000 MHz. The measurement procedure is as follows:

First, a reference line is stored in the screen memory of the spectrum analyser with the tracking generator over the entire frequency range for a level of 10 dBm

Using the second quartz oscillator mixed together with the variable LO signal in the tracking generator, the output signal is shifted upwards by 10.7 MHz. The analyser no longer reacts to this new tracking frequency, since it always displays 10.7 MHz frequency offset to the reception frequency in question. However, a crystal-stabilised 10 dBm signal source is available for precisely this difference frequency, which is fed into the intermediate-frequency port of the mixer to be investigated.

The difference frequency of both the mixing products generated in this way falls precisely on the individual analysis frequency and is represented as a continuous line on the screen.

The conversion loss can easily be read off on the display or on the hard copy, with high precision, as the differential between the current and reference lines from the memory. The high selectivity of the analyser prevents faulty measurement products involving other mixed products.

8.

Conclusion

Only the mixer printed circuit boards, an XLO and the control amplifier are required for a minimal version of the tracking generator described here. At a constant output level (for example, 0 dBm), the television PIN diodes which can still be found in many DIYers toolboxes can be called up for active service.

However, the expensive SHF stage reducer, the rare coaxial low-pass and the second XLO add a touch of luxury to the project.

9.

Literature References

[1] Jürgen Dahms, DC0DA: The Microline 3 transverter system the breakthrough in 10-GHz experimental radio technology, Part 1:

VHF Reports, 1/1989, Pp. 35-42

[2] Carsten Vieland, DJ4GC: Tracking generator from 1 MHz to 13 GHz;

Proceedings of 42nd Weinheim VHF Congress, 1977

[3] Dr.-Ing. J.Jimann, DB1NV: A tracking generator for the spectrum analyser;

VHF Communications, 1/1992, Pp. 35 - 46

Gunthard Kraus, DG8GB

Modern patch antenna design Part 1

1. Introduction

The development of modern communications technology is leading to smaller and smaller equipment just look at the newest generations of mobile phones and GPS receivers! The antennas used are also correspondingly inconspicuous; from 1 GHz upwards, patch antennas have thus markedly improved their market position in recent times. Since, due to modern design programs and present-day circuit board technology, they can be made relatively easily and accurately, even by the professional RF DIY enthusiast. This article is intended to point out the road to success, the intention is that the developers imagination should be stimulated.

Because the tracks of Patch Antennas are sometimes very delicate they are not well suited to high-powered transmission, but it can easily stand a transmission power of a few Watts. And complete antenna groups can be assembled with low-noise amplifiers into very interesting active reception antennas we can certainly expect some developments here in future.

2. Principles

A new project always begins with the acquisition of the necessary information. The obligatory reading for the first step is the article on Microstrip antennas by Friedrich Krug from [1] (Unfortunately, some of the fraction bars in the formulae were missed out in the original article please make the necessary changes!)

The standard work, Antennas by John D. Kraus [2] is also very helpful, as is the book Millimeter-Wave Microstrip and Printed Circuit Antennas [3].

First, the theoretical principles behind the patch antennas for the microstrip model described in this article will be investigated. We should carry out an intellectual investigation of the first resonance, which is important for our purposes. Here we shall be working with a few simplifications, but they make it much easier to understand and carry out the design for this operational case. However, anyone planning an application at frequencies exceeding 10 GHz will have to spend rather more money and go for the modern, more expensive 3D-EM simulations programs (Sonnet, Microwave-Office, XFDTD....). Such

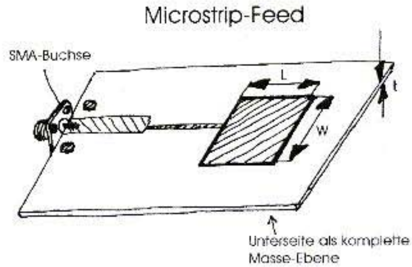

Fig 1a: Microstrip Feed, one patch with matching stripline

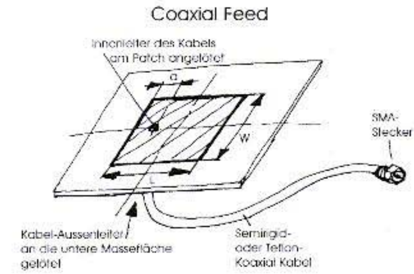

Fig 1b: Coaxial Feed one patch with feed point defining impedance

programs (usually free in the LIGHT version!) are used as a check simulation of the design worked out and for comparisons with the measurements. You can expect various solutions to be worked out during the project using different software, it will certainly be interesting!

In the simplest form, the patch antenna is a circuit board, coated on both sides, with the underside forming a continuous earthing surface. Only a small rectangle (patch) is left on the top side during etching, and there is usually a microstrip feed connecting with the middle of one edge of this patch.

If the connection point is moved from the edge of the patch in the direction of its centre, we can only use a coaxial cable for the power feed. This would emerges vertically from below with its braid soldered to the lower earthing surface, whilst the internal conductor goes through a suitable hole, and is connected to the patch (coaxial feed). The input resistance can be altered by moving the connection point along the centre line of the patch. In Fig. 1,both versions are sketched, and the following additional details are given:

a) With microstrip feed, the length, L, of the patch is always to be thought of and measured as being parallel to the feeder. It determines the frequency of resonance of the antenna and the relationship must be $L = \lambda/2$. For coaxial feed, it is that edge which runs parallel to the centre line from the middle of the external edge to the centre of the patch.

b) The polarisation of the antenna (equivalent to the direction of the electrical field lines for the radiated wave) is parallel to this length, L, of the patch, and thus also to the feed line for microstrip feed and / or the centre line for coaxial feed.

c) The width, W, determines the radiation resistance of the antenna, but in contrast has only a slight influence on the frequency of resonance.

But now to what is probably the most interesting question: how does a little metal plate like this suddenly become an antenna? If we look at Fig. 2 and view the patch as a microstrip which has an electrical length, L = /2,then the phase difference between the voltages at the left-hand and right-hand end is precisely 180 degrees. Now we must also remember that the associated electrical field lines do not end precisely at the end of the line, but are, to some extent, like the

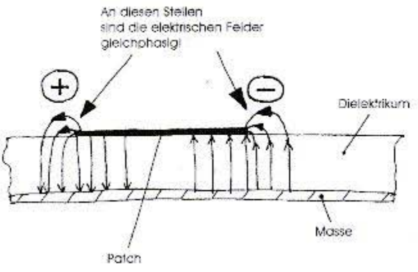

Fig 2: Sketch of a patch antenna, at resonance it acts like two slot antennas

hairs of a shaving brush, standing up vertically over and above it, and reaching into the air above the patch edge. This effect has already been described as open-end-extension or Fringing in [1]. We must therefore design the mechanical line lengths to be shorter than required in the layout design for the strips.

Now, astonishingly enough, we must consider these bristles on the field lines! If we examine the direction of these field lines for a specific instant, using arrows, we suddenly realise that these field lines, and thus the antenna currents are in phase, although the associated voltages on the two patch edges are in antiphase!

If we also know that the substrate thickness (t), is always selected to be very small with respect to the wavelength (the usual maximum is 1 to 2%), then both the left-hand and the right-hand patch edges suddenly behave like two slot antennas powered in phase. So that there shall not be any unpleasant surprises and / or deviations, the dimensions of the earthing surface should be at least one wavelength or more.

One important characteristic of this layout can be easily guessed: as soon as we deviate from the frequency of resonance, the condition of 180 degree phase difference between the voltages on the two edges will very quickly change and the radiation falls. A patch antenna is thus a very narrow-band, like all circuits operated in resonance.

When design is done with a modern program, the desired band width can be entered directly, or it can be calculated. However, it should be taken into account that the circuit board data, such as thickness, dielectric constant, etc., have an influence.

For a slot antenna, things are very different from a dipole. The E-field and the H-field are transposed as regards their direction in relation to the dipole, the E-polarisation forms a circle (or in our case a semi-circle, due to the rear face earthing surface), and therefore in the H-field we suddenly obtain the well-known figure eight (and with the rear face earthing surface only a full circle as half this figure eight). The current distribution is no longer sinusoidal, as for a dipole, but is constant. Literature reference [2], in particular, gives slot antennas a very comprehensive chapter of their own, and refers explicitly, in the patch antennas chapter, to these characteristics of the slots.

The equivalent electrical circuit diagram for the area around the frequency of resonance (this is needed later for investigations using PUFF) is relatively simple: with the microstrip feed it is a microstrip with an electrical length of 180 degrees and with the radiation resistance of the slot at each end. With the coaxial feed, there are two circuit sections, which are connected in the middle and at the two outer ends, terminated by the radiation resistance of the slot antenna. Their electrical lengths corresponding to the selected feed point are unequal, but must once again add up to 180 degrees (see Fig. 3). The width, w, of the circuit is in practice often

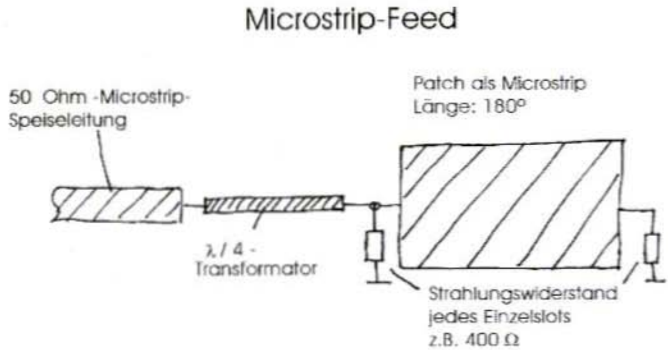

Fig 3a: Equivalent circuit for the microstrip feed to be used by PUFF

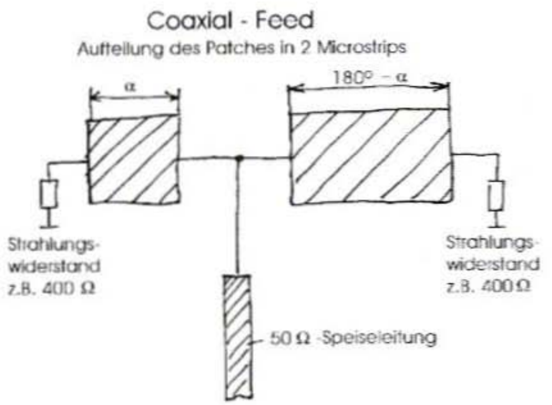

Fig 3b : Equivalent circuit for the coaxial feed to be used by PUFF

greater than the length, L, which reduces its wave resistance at frequencies of up to 10 GHz, to values below 10 to 15Ω.

It should also be recognised that the radiation resistance of each individual slot in a square patch usually lies somewhere between 400Ω and 600Ω . If the patch is made wider than it is long, we can go down below 400Ω . Since both resistances appear connected in parallel at the feed edge, this becomes a /2 circuit with a dummy load at its input. If the feed goes into the middle of an external edge, we always get a total input resistance of at least 200Ω . This value must then be adjusted again to 50Ω .

The choice of circuit board material should be made taking into account performance rather than price. The tried and trusted FR 4 fibreglass epoxy material can be used up to 2 GHz and more. However its dielectric constant falls off from 1 GHz upwards (e.g. from 4.8 at 1 GHz to 4.3 at 1.7 GHz) and the losses increase; the loss factor at 1.7 GHz is approximately 0.02, as against 0.015 at 1 GHz. Because of the decidedly high radiation resistance levels of the slots, these circuit board losses soon become noticeable. FR 4 can be used for a economy antennas, but it might not necessarily be a good idea to use it in conjunction with an extremely low-noise amplifier. Every tenth of a dB of additional attenuation at the input of a receiver is known to increase the noise number of the system by exactly the same amount.

The new RO range of materials from the well-known RT-Duroid manufacturers Rogers are ideal here. They display insignificantly poorer electrical data than Teflon boards up to over 10 GHz, but have the good mechanical properties of fibreglass material, and can be drilled, screwed, riveted, stamped, sawn and simply etched or through-hole plated. Teflon material, in contrast, is as flexible as chewing gum and is less suitable for such an antenna. Moreover, RO materials are better value than Teflon also comprehensive information can be obtained from the German importer [4].

The information above should be sufficient as an introduction. Further information can be obtained in [1], where the various applications and structural forms (e.g. horizontal, vertical and circular polarisation, circular patches, antenna groups, etc.) are really beautifully described.

3. First project

3.1. Required characteristics

In designing a new construction as complex as this, we should not immediately go for a very expensive end product, but rather initially create a feeling for ourselves of the accuracy of the design tools and the manufacturing options. For this reason, an individual patch was planned first, with the following characteristics:

- Resonance frequency: fr = 1691 MHz (Meteosat reception).
- Microstrip feed.
- Input resistance on patch edge = 0.5 x radiation resistance of an individual slot: 200Ω.
- Adjustment to 50Ω through a λ/4 microstrip transformation circuit.
- Feed through an SMA socket and a 50Ω microstrip to the transformation circuit.
- Use of circuit board material RO4003 with a relative permitivity of ε=3.38, board thickness = 1.52 mm, circuit board size 100 x 160 mm (Europa card).

To make things clearer, Fig. 4 shows the layout of the prototype for a printed circuit board coated on both sides, manufactured in accordance with these specifications.

Now, before the actual antenna design begins, a summary of the most important rules for successful patch antenna design can be cited from literature reference [3], page 40:

1. With coaxial feed, a displacement of the feed point on the centre line from the edge of the patch influences only the input impedance.
2. Higher values for the dielectric constant of the circuit board material give larger band widths.
3. Higher values for the dielectric constant unfortunately also give a poorer degree of effectiveness for radiation, and thus a poorer degree of antenna efficiency.

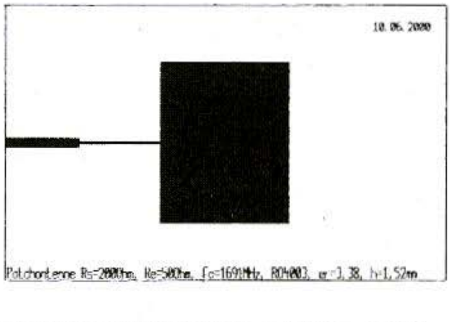

Fig 4: Layout of the prototype patch antenna on RO4003 material for use with an SMA connector.

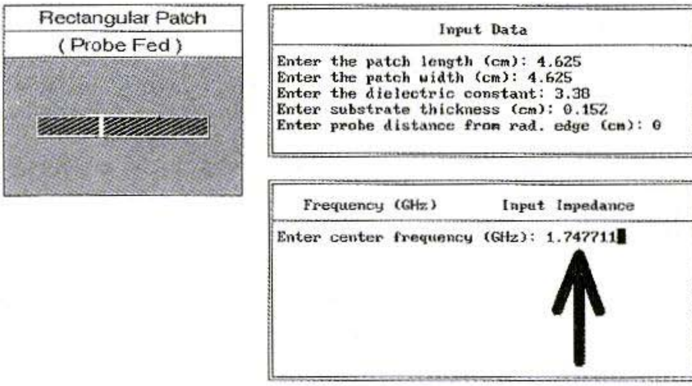

Fig 5: First result from pcaad21 showing a frequency of resonance at 1.747711 GHz

4. The width, w, of the patch is directly related to the radiation resistance and thus the input impedance, but in contrast has scarcely any influence on the frequency of resonance.

5. A single patch antenna only has the simple directional diagrams referred to above (semicircle / half eight) if the circuit board material selected is very thin.

   As a guide, the thickness of the circuit board should be no more than 1 to 2% of wavelength.

   This is not a good directional antenna, because of the large apex angle associated with it. However, as soon as the circuit board is selected to be much thicker than recommended, zero points suddenly appear in the radiation diagram which are caused by additional surface-wave modes.

   The following rule also applies to the circuit board thickness: a thinner circuit board gives a better antenna quality and thus a smaller antenna band width.

6. The antenna gain of an individual patch is very small the typical value attainable is 6 to 6.5 dBi.

7. The design and calculation methods in this article (and thus also the programs initially used) are based on an empirical model and are thus not universally valid. For higher frequencies and for analysing the antenna behaviour over a broad frequency range, the use of modern EM simulators is indispensable.

3.2. Determining the patch dimensions required (microstrip feed)

Two programs are well suited to this: pcaad21 and patch16, they can be downloaded direct from the Internet [5].

A) Design using pcaad21

The operation is very user friendly, and when the appropriate menu has been called up you are immediately asked about the patch length. You could enter any length here as an experiment and see what happens. However, it is considerably more promising to work the approximate length value out first (half electrical wavelength) from the following equation using a pocket calculator and to work with this value:

$$L = \frac{c}{2 \bullet f \bullet \sqrt{\varepsilon}} = \frac{3 \bullet 10^{11} mm}{sec \bullet 2 \bullet 1691 \bullet 10^{6} Hz \bullet \sqrt{3{,}38}} = 46{,}25 mm$$

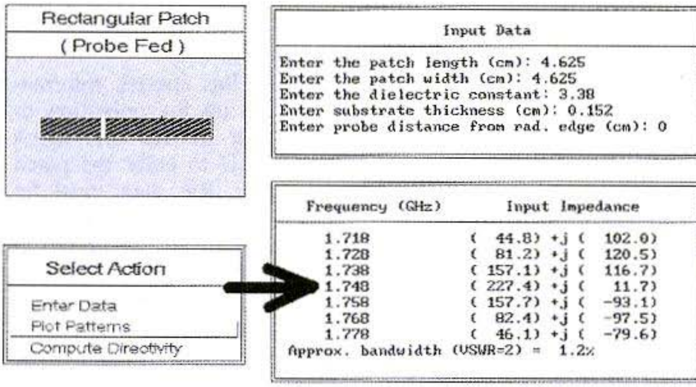

Fig 6: Pressing ENTER shows the impedance gradient. At the frequency of resonance the radiation resistance is too high at 227Ω

For the patch width, we simply take the same value of 46.25 mm and subsequently examine the result. When the dielectric constant (ε=3.38) and the circuit board thickness (0.152 cm) have been entered, you are asked about the distance between the feed point and the edge, w. Since we wish to use microstrip feed, select 0 as the answer and press <ENTER>. Thus we obtain a screen structure like Fig. 5, from which we can read off the value 1.74771 GHz as the frequency of resonance. We press <ENTER> again and the impedance gradient will also be shown in the area around the frequency of resonance, in accordance with Fig. 6.

It can be seen here that the radiation resistance, at 227Ω for the calculated frequency of resonance, is still too high: the small reactive component of j11 can be ignored with confidence. So the procedure is repeated when ENTER DATA has been activated in the left-hand menu, with a greater width, w, and another length, L , until a radiation resistance of approximately 200Ω has been achieved for a frequency of resonance of 1691 MHz.

Fig. 7 shows the end result of our efforts; the menu item plot data gives us the directional diagrams. The directivity can be displayed, if desired, at approximately 6.5 dB; it represents the ratio of the maximum power radiated in a specific direction to the mean value.

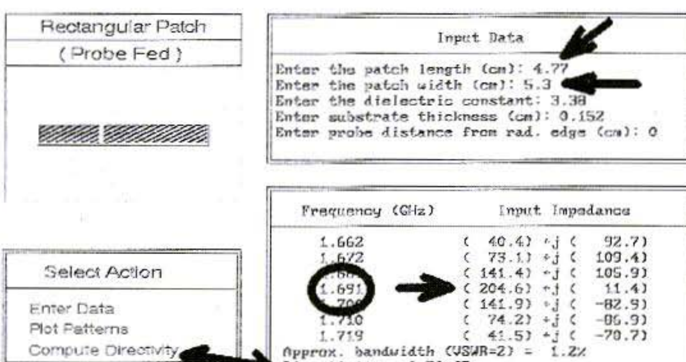

Fig 7: Results from pcaad21 after optimising for a frequency of resonance at 1691 MHz and a radiation resistance of 200Ω

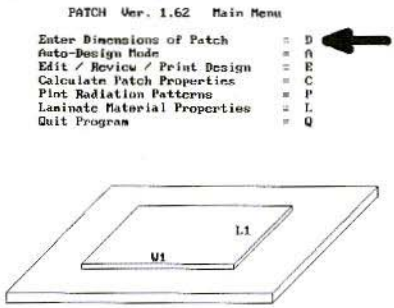

Fig 8: The first menu in patch16. Choose option D to enter data

Correctly expressed in scientific terms, it is the ratio of the maximum value to the mean value using the Poynting vector. It tallies with the normal dBi gain specification, for with the spherical radiator as reference there is no preferred direction and thus it displays a directivity or a gain of zero dB.

It should be noted that the radiator length of 47.7mm shown in Fig 7 is the electrical length, this should be reduced by the open end extension for microstrips. With such wide lines, this is about half the circuit board thickness on each side, which means that in total approximately, 1.5 mm must be taken off.

B) Design with patch16

Once the program has started, information can be called up for operation or we can go straight to the first menu (Fig. 8) and press D to enter the patch data. Unfortunately, the data must be entered converted into inches. It must also be noted that with this program the radiator length will already be in the open-end corrected form which is how it will also be used in the layout!

The following values can be taken over from the previous design:

- Corrected patch length, L = (47.7 - 1.5)mm = 46.2 mm, corresponding to 1.819 inches.
- Patch width, w = 53 mm, corresponding to 2.0866 inches.
- Circuit board thickness, t = 1.52 mm, corresponding to 0.06 inches.
- Dielectric constant ε = 3.38.
- Probe distance from radiating edge = 0.
- Loss tangent of RO4003 at 2 GHz = 0.001.

Once you have worked through the entry procedure, you come back to the same screen (Fig. 9) with the sketch of the patch as at the beginning but this time with data entered as a check; these are

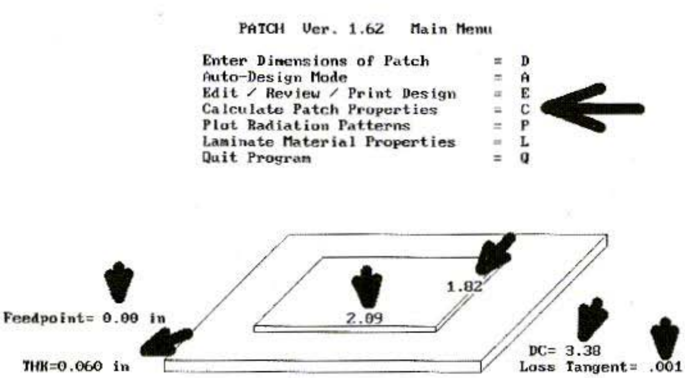

Fig 9: Display after data has been entered. If there are any mistakes thaey can be corrected by pressing E

```
The Resonant Frequency is   1.706 GHz
Qo is  60.5

The Edge Radiation Resistance is  234.78 ohms
Zc of Quarter-wave transformer is 108.3 ohms
Approx. width of the Quarter-wave transformer is 0.029 inches
Length of Quarter-wave transformer is 1.109 inches at the Resonant Freq.

Input Resistance at probe location is 234.78 ohms

The 2:1 VSWR Bandwidth is  1.3%
Upper Frequency Limit =  1.717 GHz
Lower Frequency Limit =  1.694 GHz

Press 'ENTER' to continue:
```

Fig 10: Results after calculation.

rounded off in the display, but for calculation they are stored in the correct form, with all the digits after the decimal point. If we now press C (calculate patch properties), we can see all interesting data (Fig. 10).

So let us look at the results in somewhat greater detail:

a) With the radiator length specified by pcaad21 and open end corrected, we obtain a frequency of resonance of 1,706 MHz in patch16 instead of 1,691 MHz. Which program is actually right?

A comparison with the prototype created shows that a radiator length specified using patch16 is closer to the measured value, because not only is the open end correction automatically carried out here but the modelling is also rather more precise.

b) The radiation resistance amounts to 235Ω and is thus approximately 15% higher than the value from pcaad21-200Ω . Here too, patch16 gives a more reliable prediction.

c) The data for the λ/4-matching transformer required for 50Ω is shown.

d)We also obtain a quality of Q = 60.5, plus the 2:1-SWR-band width with the value 1.3%, outside of the relevant lower and upper limiting frequencies (1707 and 1694 MHz).

Due to the frequency deviation, we now repeat the design again, varying the length, L, until it gives exactly 1691 MHz as a frequency of resonance. Then we change the patch width again until we obtain the required radiation resistance of 200Ω. The exact dimensions and design data, with all the digits after the decimal point, can be obtained as a table by pressing E (edit) (Fig. 11); the electrical characteristics of the finished patch are shown in Fig. 12.

Thus we go to the next step with a length of L = 1.835 inches = 46.6 mm and a width of w = 2.31 inches = 58.67 mm. We call up the microwave CAD program PUFF, to design the necessary matching circuit.

3.3. Matching antenna at 50Ω

Before PUFF is started up, the file

```
These are the design parameters:

Length (L) =  1.835  inches
Width (W) =  2.31  inches
Height (H) =  5.999999E-02  inches
Dielectric Constant (D) =  3.38
Loss Tangent (T) =  .001
Feedpoint Distance (F) =  0  inches

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

Fig 11: Exact dimensions for the patch at 1691 MHz and 200Ω

```
The Resonant Frequency is    1.692 GHz
Qo is  57.2

The Edge Radiation Resistance is  200.38 ohms
Zc of Quarter-wave transformer is 100.1 ohms
Approx. width of the Quarter-wave transformer is 0.036 inches
Length of Quarter-wave transformer is 1.113 inches at the Resonant Freq.

Input Resistance at probe location is 200.38 ohms

The 2:1 VSWR Bandwidth is  1.4%
Upper Frequency Limit =  1.704 GHz
Lower Frequency Limit =  1.680 GHz

Press 'ENTER' to continue:
```

Fig 12: Electrical characteristics that correspond to the data shown in Fig 11

setup.puf is created using a text editor and the following data are entered:

- zd 50.000 Ohms (normalizing impedance. 0<zd).
- fd 1.691 GHz (design frequency. 0<fd).
- er 3.380 (dielectric constant. er>0).
- h 1.520 mm (dielectric thickness. h>0).
- s 200.400 mm (circuit-board side length. s>0).
- c 19.000 mm (connector separation. c>=0).
- r 0.010 mm (circuit resolution, r>0, use Um for micrometers).
- a 0.000 mm (artwork width correction.).
- mt 0.035 mm (metal thickness, use Um for micrometers.).
- sr 2.000 Um (metal surface roughness. use Um for micrometers.).
- lt 1.0E-0003 (dielectric loss tangent.) cd 5.7E+0007 (conductivity of metal in mhos/meter.).

Then PUFF is started up, and first it specifies the microstrip equivalent data for the patch. For this purpose, we place a microstrip in field F3 with Z = 10Ω and an electrical length of 180 degrees. With the cursor is still in this entry line of F3 and the equal sign is being keyed, the associated mechanical dimensions appear in the dialogue window, in the form of the length, L, and the width, w. We now vary the wave resistance, Z, until precisely the required width of w = 58.67 mm is obtained. Fig. 13 shows this condition and the Z value current for it, of 4.957 . For the correct length, the process becomes rather more difficult and more extensive. We must actually work with exact modelling and here PUFF has a characteristic which is mentioned only briefly in the manual -- and if we don't take it into account, we are creating noticeable errors for ourselves.

As soon as we add a exclamation mark

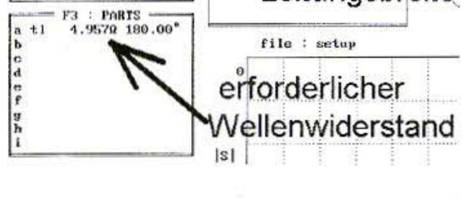

Fig 13 PUFF showing the conditions with width 58.76mm

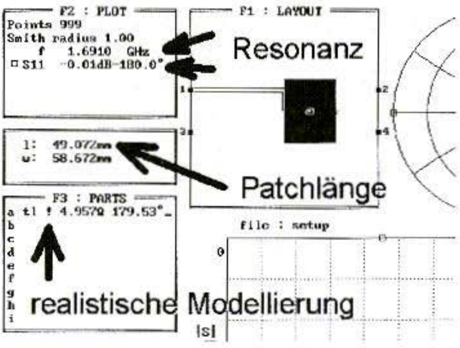

Fig 14: Puff display after modelling for L = 180 degrees at 1691 MHz

behind the tl entry in field F3, it means that a precise analysis of the circuit is carried out, and also that the dependence of the data on the frequency is very precisely modelled. If we now enter the equal sign in F3 again, then the actual wave resistance and the actual electrical length appear in the dialogue field.

But:

These two details are valid only for DC, and there is initially no command (or even a function key) which can be used to make this value directly visible at the required design frequency (here: 1691 MHz) -- which is important due to its dependence on the frequency!!!

So we must help ourselves and quickly draw up a layout in which this line is short-circuited at one end and is connected to port 1 at the other end. If we now plot S11 in a narrow range around the design frequency 1691 MHz -- and with the highest possible resolution of 500 or 1000 dots -- , then we need only to move the curve using the keys <Page UP> and <Page DOWN> and observe the phase of S11 precisely: as soon as it jumps from - 180 degrees to + 180 degrees, we have found the series resonance frequency, at which the electrical length amounts to precisely half a wavelength .Thus we need only experiment a little with the length input in the corresponding F3 line and re-plot until this jump occurs at precisely 1691 MHz. Thus we finally have a microstrip before us, correctly modelled with all losses and dependence on frequency, with L = 180 degrees at 1691 MHz (Fig. 14).

Anyone who now deletes the exclamation mark in the entry (tl ! 4.957 179.530) and enters the equal sign will

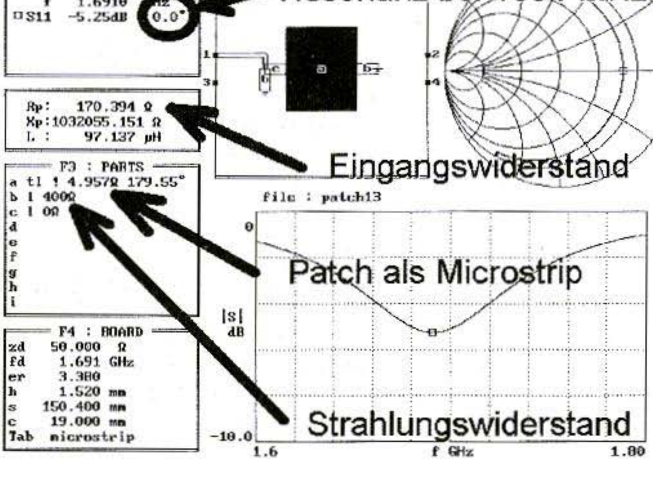

Fig 15: PUFF simulation of a lossy model with the radiation resistance of the slot at both ends

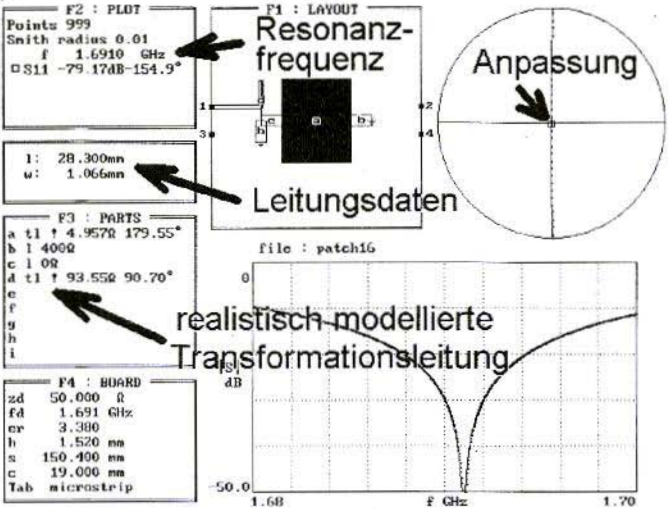

Fig 16 : Final conditions of simulation showing dimensions of the transformation circuit

obtain the associated mechanical dimensions with w = 58.67 mm and a length of L = 49.07 mm.

If we remove the open end extension of about one board thickness from L, then there will still be a remaining patch length of approximately 49.1 - 1.5 = 47.6 mm over to be realised in the layout. But that is a considerable deviation as against the patch16 simulation of 46.6 mm (see above) and one tends at first to look round in bewilderment for the source of the error.

Only when we look at literature reference [6] is there any light in this darkness. The divergence between the microstrip model and the real antenna is physically determined and is the price we pay for making life so very easy with such a simple model. So, just this once, we are not concerned here with the PUFF length specification, but we simply transfer the patch16 value into the layout.

Now we provide this realistic and lossy model of our patch with the planned radiation resistance of the individual slot at both ends (400Ω) and start a new plot. It turns out that a small length correction is required first for correct resonance at 1691 MHz. With the help of the tab key, the Smith chart should first be switched to Electric conductance display, so that the input impedance can be displayed as a parallel circuit. This happens in that, in accordance with the plot in field F2, the cursor is set to S11 and the equal sign is keyed in again. In the dialogue field, we obtain the total resistance of the patch, including the line and transformation losses, at 170.4 Ω and we must then use our transformation circuit to bring this value to 50 Ω (Fig. 15). The wave resistance required for this is :

$$Z = \sqrt{50\Omega \cdot 170.4\Omega} = 92.3\Omega$$

There we make life somewhat easier for ourselves again and we initially provide in F3 just such a circuit through the lines

tl ! 92.3Ω 90°

If we now switch this element between the patch and port 1, then we should in theory have perfect matching at 1691 MHz and consequently land precisely in the centre of the Smith diagram. Since initially this is not precisely the case (for

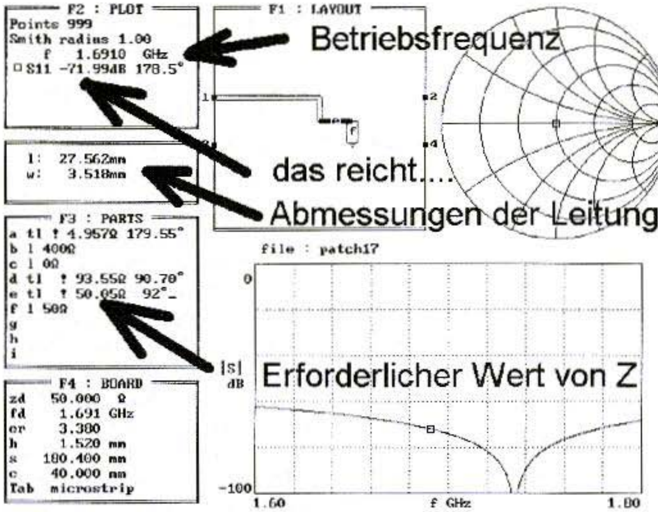

Fig 17: With the line terminated at 50Ω adjustments are made to give an S11 value at 1691 MHz

the data for this circuit are themselves also dependent on frequency...), we simply change the transformation circuit length and its wave resistance round until we have found the correct value. Naturally, we also zoom in on the area around the centrepoint in the Smith chart at high magnification and enter a correspondingly small radius for it (here: r = 0,01).

Fig. 16 then shows the final condition, together with the mechanical dimensions of the transformation circuit required for it.

Now all we need is a 50Ω microstrip with, for example,. L = 90° as a connection to the SMA socket. Because of the above-mentioned dependency of the circuit data on the frequency, their wave resistance diverges from 50Ω again at 1,691 MHz, even if we have created a correct model (using the

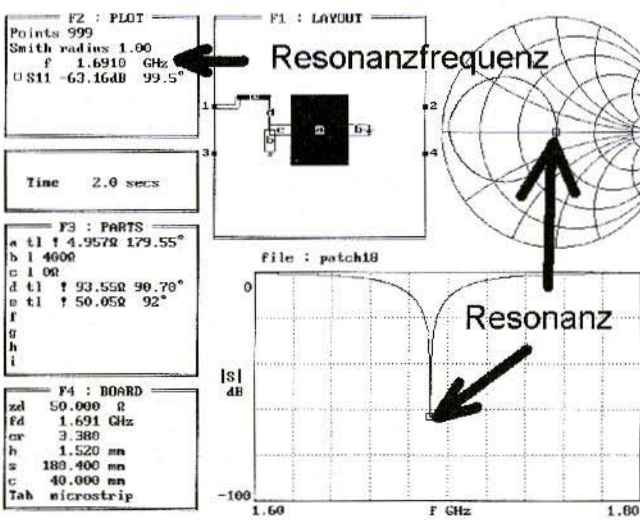

Fig 18: The final PUFF simulation showing that no further adjustmets are needed

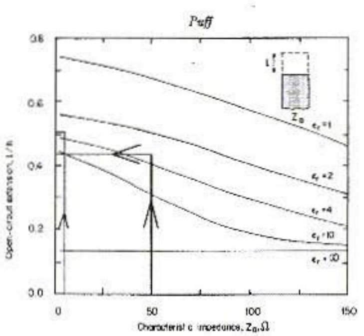

Figure 7.2 The open-circuit end correction in microstrip, plotted from (7.2). The artwork length correction in a parts list should be negative.

Fig 19: Determining the open end correction from the diagram in the PUFF manual

exclamatiom mark in the corresponding entry line) and worked out this value in DC. So we help ourselves by means of a further additional simulation:

We simply terminate such a 50Ω microstrip section with a 50Ω resistance and plot S11 for ourselves in the range 1600 - 1800 MHz. In the loss-free ideal case, there would always be a match, but things don't stay that way, because Z varies with the frequency. Consequently, we now very carefully alter the input value for the wave value in F3 until we have obtained at least one value for S11 for 50Ω at 1691 MHz.

In Fig. 17, the result of our efforts can be seen. And now its full speed ahead with the creation of our first antenna. We add this 50Ω feed to port 1 to our previous circuit formed by the patch and the transformation line and we then start what is hopefully the last simulation. Fig. 18 shows that no further corrections are needed and that everything fits together.

As soon as we remove the call-up sign again in the corresponding lines of F3 the good old equal sign finally supplies us with all lengths and / or widths for the individual microstrips. So first we just print off this table for the draft layout:

Patch: width w = 58.67 mm / length L = 46.6 mm (= patch16 value!)
Transformation line: width w = 1.066 mm / length L = 28.3 mm (uncorrected)
50Ω feed: width w = 3.518 mm / length as required in layout.

Our last actions now concern the required open-end corrections on the centre circuit, and for this we shall once again need the well-known diagram from the PUFF manual (Fig. 19). For a patch with Z = 4.955Ω and ε = 3.38, we can read out an extension here of about 0.51 (51 % of board thickness = 0.78 mm). Since the width of the transformation line, at approximately 1 mm., is extremely small, as against the patch width of 58.67 mm, we must take this 0.78 mm fully into account and select our λ/4 transformer to be correspondingly longer.

Things look rather different on the 50Ω connection side. For 50Ω and ε = 3.38, Fig. 18 supplies us with an extension of approximately 0.44 = 44% of board thickness = 0.67 mm. However, since the conductor widths coming up against each other no longer vary as much as previously, the actual extension turns out to be shorter and can easily be calculated in accordance with formula (7.3), page 36 of the PUFF manual:

$$L_{corr} = L \cdot \left(1 - \frac{w1}{w2}\right) = 0.067mm \cdot \left(1 - \frac{1.066mm}{3.518mm}\right) = 0.47mm$$

Thus this side of the transformation line requires only an extension of 0.47 mm and now the total length can be determined::

$$L_{ges} = 28.3mm + 0.78mm = 0.47mm = 29.55mm$$

Before we start the layout CAD program, we should always put the final dimensions in a simple sketch as per Fig. 20. Errors can creep in only too easily during printed circuit board de-

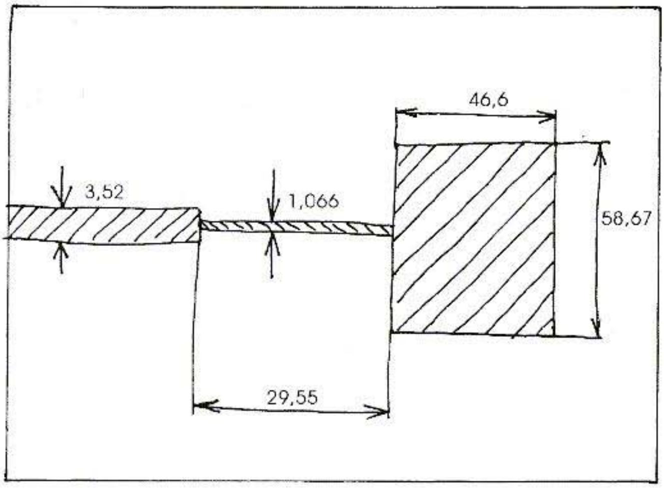

Fig 20: A sketch of the artwork to be produced

sign and bad circuit boards can unfortunately only be thrown away (my own experience....).

This article will continue in issue 2/2001 with simulations using Sonnet Lite and Coaxial feeds

Literature references for Part 1

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

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

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

[4]: Fa Mauritz GmbH & Co., P.O. Box 10 43 06, D-20030 Hamburg, Tel. 040-24 16 54

[5]: http://www.rfglobalnet.com or via search engine http://www.fireball.de under all languages. Then please search for pcaad21.zip bzw.patch16.zip.

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

VHF COMMUNICATIONS

A Publication for the Radio Amateur Worldwide

Especially Covering VHF, UHF and Microwaves

Volume No.33 Spring Edition 2001-Q1

Publishers

KM PUBLICATIONS,
63 Ringwood Road, Luton,
LU2 7BG, United Kingdom

Tel: +44 1582 581051
Fax: +44 1582 581051

Email:
andy.@vhfcomm.co.uk

Editor

Andy Barter G8ATD

VHF *COMMUNICATIONS*

The international edition of the German publication UKW-Berichte is a quarterly amateur radio magazine, especially catering for the VHF/UHF/SHF technology. It is owned and published in the United Kingdom in Spring, Summer, Autumn and Winter by KM PUBLICATIONS.

The 2001 subscription price is £20.00, or national equivalent. Individual copies are available at £5.00, or national equivalent each. Subscriptions should be addressed to the national representative shown in the next column. Orders for individual copies of the magazine, back issues, kits, binders, or any other enquiries should be addressed directly to the publishers.

NOTICE: No guarantee is given that the circuits, plans and PCB designs published are free of intellectual property rights. Commercial supply of these designs without the agreement of the Author and Publisher is not allowed. Users should also take notice of all relevant laws and regulations when designing, constructing and operating radio devices.

Translated by: Inter-Ling Services, 62 Caldecott Street, Rugby, CV21 3TH, UK

Printed in the United Kingdom by: Cramphorn Colour Printers Ltd., 15a Boughton Road Industrial Estate, Rugby CV21 1BQ, UK.

AUSTRALIA - W.I.A. South Australia Division, GPO Box 1234, Adelaide, SA 5001, Australia. Tel/Fax: 8 8522 4859

BELGIUM - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133-77980. Fax: 09133-779833
Postgiro Nbg. 30445-858.

DENMARK - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

FRANCE - Christiane Michel F5SM, Les Pillets, 89240 PARLY, France
Fax: (33) 03 86 44 08 82 Tel: (33) 03 86 44 06 91

FINLAND - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

GERMANY - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133 7798-0. Fax: 09133 779833.

GREECE - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

HOLLAND - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

ITALY - ADB ELETTRONICA di Luchesi Fabrizio IW5ADB, Via Vechia Romana 266, 55100 ANTRACCOLI, (LUCCA), Italy.
Tel: 0583 952612. Fax: 0583 91109

NEW ZEALAND - Judith Holland, AUCKLAND VHF GROUP Inc., P.O. Box 10 138, AUCKLAND 1030, New Zealand.
Tel: 0 9 412 7659

NORWAY - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: +46 40 16 42 66. Fax: +46 40 15 05 07.
GSM: 0705 16 42 66
e-mail: waveline@algonet.se http://www.algonet.se/~waveline

SOUTH AFRICA - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

SPAIN & PORTUGAL - JULIO A. PRIETO ALONSO EA4CJ, Donoso Cortes 58 5º -B, MADRID 15, Spain. Tel: 543 83 84

SWEDEN - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: 040 16 42 66. Fax: 040 15 05 07. GSM: 0705 16 42 66 e-mail: waveline@algonet.se http://www.algonet.se/~waveline

SWITZERLAND - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

UNITED KINGDOM - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051.
Fax: +44 1582 581051. Email: vhfsubs@vhfcomm.co.uk

U.S.A. - GENE HARLAN, ATVQ Magazine, 5931 Alma Drive, Rockford, IL 61108, USA. Tel: Toll free USA only: 800 557 9469; Elsewhere: +1 815 398 2683; Fax: +1 815 398 2688
Email: atvq@aol.com

ELSEWHERE - KM PUBLICATIONS, address as for the U.K.

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

ISSN 0177-7505