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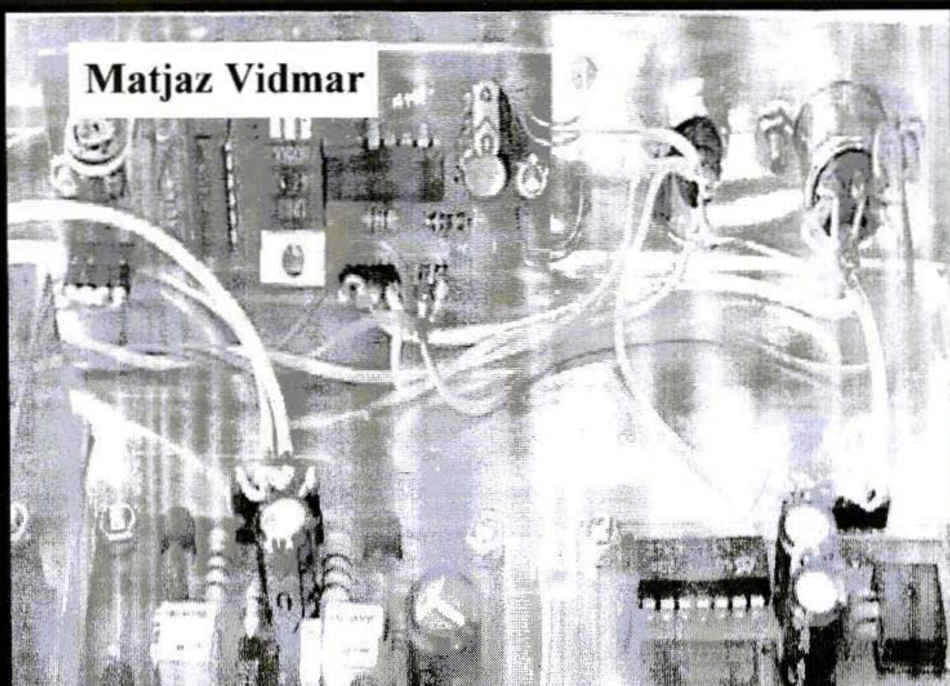

# Contents



Owing to space problems and the fact that some of the articles in this issue are particularly of current interest, part-4 of the Wolfgang Schneider (DJ8ES) series on *Measuring Methods Using a PC* and part-3 of the Gunthard Kraus (DG8GB) series on the *Design and Realisation of Microwave Circuits* have both been held over until issue 3/97. Apologies ... Mike

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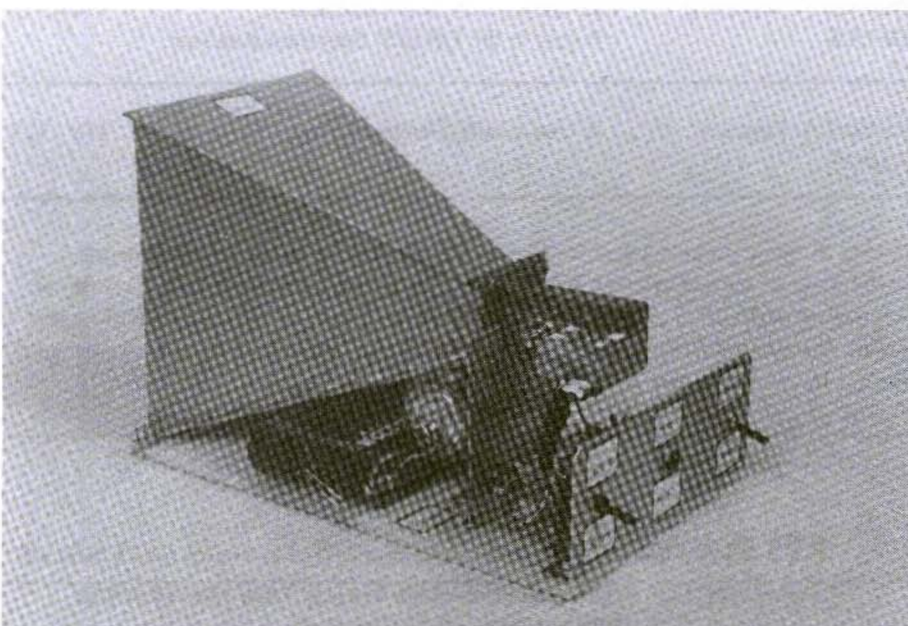
*André Jamet, F9HX*

# Using a DRO as a Transmitter

**To receive TV satellites a low noise block (LNB) is used first to amplify the 10 to 12 GHz signal collected by a parabolic antenna. Then a local oscillator (OL) and a mixer change the signal into a lower intermediate frequency, usually between 950 and 2150 MHz. That frequency comes via a coaxial cable to a demodulator which delivers video and audio signals to the TV set.**

**The local oscillator is a Dielectric resonator Oscillator (DRO). As already described [1], we can make a 3 centimetres super-regenerative receiver using a DRO. We will see how to make a transmitter and try not to destroy transistors!**



## 1.
## THE DRO

If recent LNBs have an integrated circuit which provides both LO and mixer, the older ones have separate circuits. A transistor is used as oscillator and diodes as mixer. That kind of LNB is the one useful for us to make RX and TX. We can obtain them from TV antenna installers [1].

The OL comprises a 10 GHz transistor, a DRO, strip-lines, resistors and capacitors on a glass-PTFE PCB. Inside the LNB, the OL is contained in a sealed can, in order to avoid any radiation from, or to, other circuits. The frequency adjustment is provided by adjustment of a screw located on the cover.

Several forms of oscillators are used, the feedback can be made by the DRO between gate and drain, gate and source, or by reflection. The output power can be drawn from the source or the drain. Figures 1 and 2 give the diagrams of common printed circuits. To make a transmitter all are usable, but that is not always the case for a super-regenerative receiver.

A 5 or 8 volts regulator supplies the transistor as well the other LNB circuits, but, a dropping resistor reduces the voltage to about 3 volts at app. 15 to 30 mA. The RF power is around 20 mW (+13 dBm).

As the OL frequency is not suitable for amateur band use, we have to modify it. As shown in [1], it is better to have a lower frequency DR as required which is about 10.368 GHz.

The 9.75 and 10 GHz OL (Astra satellite) are preferred instead of the 11.475 GHz ones, as it is easier to increase the DR frequency by height reduction rather than to lower it by adding chips of ceramic.

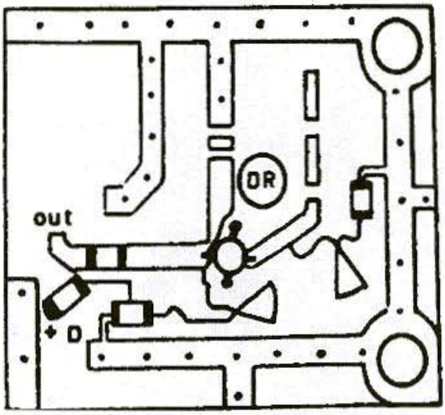



**Fig.1: PCB of a DRO Gate/Drain Feedback Source O/P Dimensions: 25 x 30mm**

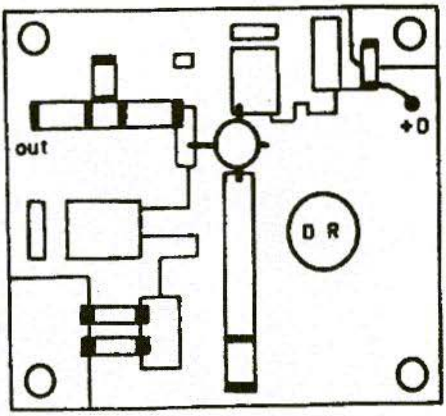



**Fig.2: PCB of a DRO Earthed Drain, Gate Reflection, Source O/P Dimensions: 25 x 30mm**

## 2.
## DRO EXTRACTION FROM AN LNB

First of all open the LNB by removing screws or rivets which fasten the external enclosure. Then find the DRO, that is easy thanks to the frequency adjustment screw and remove the small cover which holds that screw.

Then feed the LNB at its nominal voltage, around 12 volts. Check the DRO drain voltage, nearly 3 volts. Calculate the current by measuring the voltage drop on the drain or source resistor according to the case. Note those values as they will be used in our transmitter.

To be sure that the circuit is oscillating put back the cover, because some DRO do not oscillate, or oscillate badly without it. The oscillation is present if a drain current variation is shown when you finger the DRO output driving the mixer. When we are sure of the operation we need to extract it from the LNB enclosure. A saw will be used to cut the aluminium and scissors for the PCB, in order to extract the useful section, which is limited to the OL output. Beware of mechanical shocks to avoid cracking, breaking or ungluing the DR. If it is cracked it will work but less satisfactory. Unglued it would be difficult to bring back to the right place.

We have now a small closed block with only one aperture clearing the PCB where the line was going to the mixer.

## 3.
## TEST OF THE DRO ALONE

That small block no longer has connections to feed the drain, it is now time to discover a real way to kill transistors!

As the drain-source voltage required is around 3 volts, we can think that two 1.5 Volt batteries will do the job, very convenient for a portable TX, or for indoor tests with a regulated power supply of that voltage: beware "danger".

Everyone knows or must knows that the more efficient high frequency active components as MOS, GaAsFET, HEMT are very sensitive to electrostatic discharges so there need a lot of care for storage, handling and mounting. So grounded conductive wrist strap, soldering iron uncoupled from the mains, no synthetic material clothes, antistatic cover on the working table, etc.

Another kind of breakdown caused the destruction of a dozen of 10 to 24 GHz transistors because no information was available to me: one must not allow fast variations (high dv//dt) in supply voltages, for these kind of components.

If 3 volts or more is applied abruptly to the drain, you run a risk of breaking the transistor between drain and source. If you vary the drain resistor by a decade resistor, at the time of transit between contacts: beware. Likewise if you vary the decoupling capacitor between drain and ground by a decade capacitor. A short-circuit between drain and ground: breakdown, not the power supply , if it is protected, but the transistor. If a modulation signal is applied to the gate, the source or the drain, beware also if the voltage is too high or the reverse and if a high dv/dt occurs.

So respect these rules:

- maintain a resistor in the drain supply that will damp the dv/dt owing to the decoupling capacitor. So a higher starting voltage must be delivered from a 5 or 8 Volt regulator

- not to make the dangerous manoeuvres as above mentioned

if you do not want to see the current drain going to up to 50 mA or more that surely means the transistor is destroyed.

If the price of this kind of transistor is not very high, soldering and unsoldering especially is so delicate you would be better to avoid them.

## 4. DRO FREQUENCY MODIFICATION

First we need a means to measure the frequency, so we have a 10 GHz frequency meter. The easiest and cheapest way is to use a new LNB, or a working second hand, having a 9.75 GHz OL and a 1 GHz frequency meter, as one made from these available inexpensive kits. The DRO radiation will be received by the LNB and we will have:

frequency DRO =
9.75 GHz + frequency meter reading.

So we will read 618 MHz for a 10.368 GHz DRO.

It is obvious than the frequency meter will indicate zero if the tested DRO frequency is 9.75 GHz also. A reading will only appear when we increase the frequency enough.

To move the DRO frequency from 9.75 to 10.368 GHz, needs to reduce the DR height. Articles [1] and [2] explain that operation can be done by sand-paper. That means to unglue the DR from the PCB with the risk to crack or to break it. Furthermore it requires to bring it back to the right place. It is better to keep it in position and to abrade it with a small millstone powered by a miniature electric motor.

The abrasion is more or less fast according to hardness of the DR ceramic. It seems that light yellow DR having a hole inside are easier to abrade than those made with white ceramic ones. As the ceramic dust could be toxic, do not breath it when abrading.

It is essential to check the height reduction by measuring the frequency frequently in order not to exceed the requisite value.

Frequency measurements must be done with the cover in position and the screw at half turn to obtain a sufficient adjustment range. If the screw is too close to the DR to obtain the required frequency, a power loss is caused by an oscillating circuit overdamping. On the other hand, if the screw is to far from the DR the adjustment range can be lower to compensate for the DRO-waveguide coupling effect.

If LNB's having a 9.75 or 10 GHz DRO are not available, but only those very common types having a 11.475 GHz one, you need to lower the frequency by increasing the DRO height. That can be done by fixing with an ultra-fast glue a small chip taken from an another DR which is sacrificed for that. The use of ceramic coming from capacitors or other kind of component is more uncertain. If the ceramic has a high permittivity it has a high temperature coefficient generally that will be disastrous for our transmitter stability. If the permittivity is low, the temperature coefficient will be low generally, but it will be necessary to add a large piece of ceramic and that will create other difficulties, for instance, the screw could be in contact with the DR as soon as it is moved.

As pointed out in [1], it is advisable to make an ageing cycle on the DRO modified in order to remove any mechanical stress due to handling.

## 5. DRO MODULATION

If it is difficult to get the bandwidth for TV [2], but telephony, even wide FM, only requires 50 to 200 kHz. That is easily obtained by applying the modulation to the drain supply. That is done here by audio signal injection into the normally grounded pin of the voltage regulator by means of a resistor. So the DC voltage is modulated with a variation of 100 to 300 mV peak to peak.

In order to make easier low signal finding, a 1000 Hz square-wave modulation is provided by a 555. For voice, an electret mike and a 741 amplifier give an almost equal voltage to the 1000 Hz one, therefore getting the same modulation swing.

The circuit diagram of the transmitter is shown in figure 3.

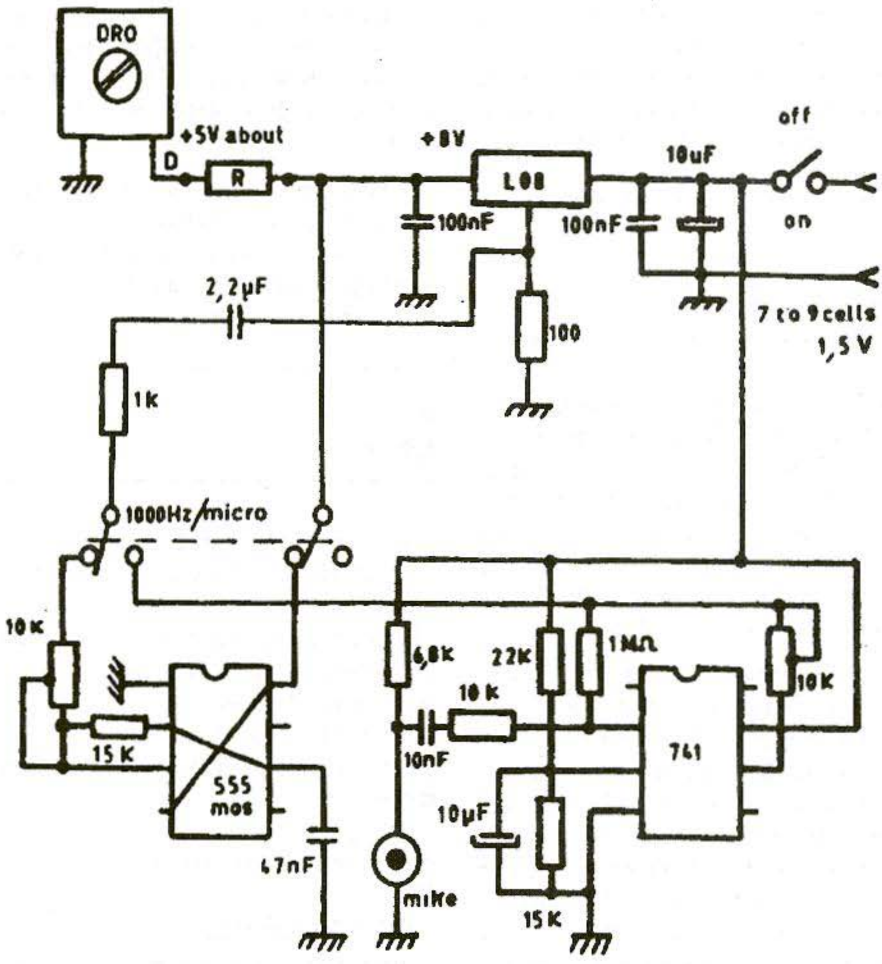



**Fig.3: Circuit Diagram of the DRO Transmitter**

## 6. MECHANICAL ASSEMBLY

The DRO must be joined to an antenna and it is very simple to make a 20 dB horn as described in the article [4] with gives dimensions and method of construction. Using copper-clad glass-epoxy for PCB makes it quite easy.

A WR90/R100 waveguide brings the HF to the horn. A slot into the waveguide lets in a probe to inject the 10 GHz. The DRO is placed against the waveguide with slides in order to fix the probe at the optimum place giving the highest radiated power. Two setting screws are provided for impedance matching between DRO and waveguide so as to improve the transfer efficiency.

The complete horn, waveguide, DRO, supply and modulation circuits are mounted on a plate as shown in the photograph. Figure 5 gives the mechanical assembly of the 10 GHz parts.

## 7. TESTING AND ADJUSTMENT

Once the assembly is completed, we need to check the output voltage regulator, DRO drain current, 1000 Hz voltage and frequency, voltage and waveforms from the microphone amplifier. Resistor R in figure 4 will be chosen in order to obtain the same DRO voltage and current as the original values.

Then, we need to obtain the highest radiated power. To do that, we have to make a field-strength-meter with a horn, waveguide, SHF diode and a micro-ammeter, as described in [5] and [6]. The distance between the TX and that f.s.m. will be around one meter, in order to obtain a meaningful reading, but not to have direct coupling between them. We have to play with the probe fitted in the slot and the two impedance matching screws. Settings are interdependent and we have to act minutely and methodically to get the optimum result. We also have to retune using the frequency screw, as the frequency must remain at the required value because the above adjustments are affecting it. If a frequency analyser is available, we can check the unmodulated wave purity and the 1000 Hz and microphone swings. Otherwise, we can listen the TX to check its quality. The receiver could be a super-regenerative one such as described in [1] or an LNB as used for frequency measurement, followed in this case, by a 600 MHz receiver as a scanner.

Warning: even the radiated power (erp) is modest (1 to 2 watts), so DO NOT look inside the horn when the TX is working, because the human retina is very sensitive to SHF.

## 8. QSO

As already quoted in [1], this kind of TX has already allowed (the time this article was written), QSOs up to 48 kilometres line-of-sight. It is foreseeable to get much longer distances and also by reflection, refraction or scattering as obtained by other TX. We can use a parabolic dish to increase the radiated power.

## 9. CONCLUSIONS

It is very instructive to work at 10 GHz with very simple and cheap means, owing to the experience and experiments we can do during realisation and adjustment, as well as making QSOs: difficulties due to components miniaturisation, settings interdependence on the one hand, and propagation irregularities even in sight due to the clouds, rain and fog, on the other hand.

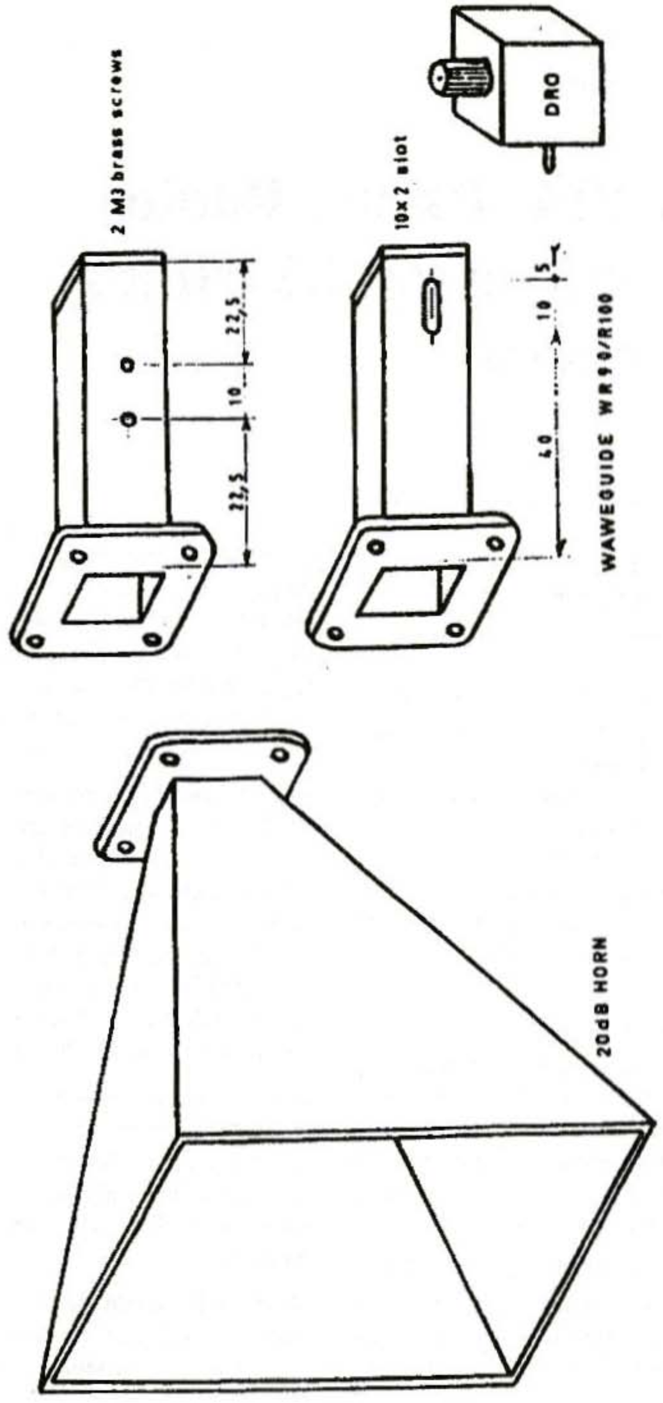



**Fig.4: Mechanical Design including Waveguide Front and Rear Views**

*Matjaz Vidmar, S53MV*

# 23cm PSK Packet Radio Transceiver for 1.2 Mbit/s User Access

## 1. WHY BI-PHASE PSK MODULATION?

Upgrading the packet radio network to higher data rates also requires using more efficient modulation and demodulation techniques both to reduce the signal bandwidth and to increase the radio range of the system. In particular, inefficient modems coupled to standard FM transceivers have to be replaced with custom-designed radios for data transmission.

Considering the bandwidth and TX power available to radio amateurs, it is necessary to switch to coherent demodulation techniques at data rates around 100 kbit/s in terrestrial packet radio and at even lower data rates in satellite communications.

One of the simplest forms of digital modulation that can be demodulated in a coherent way is biphase PSK. The usual amateur approach to implement biphase PSK, is to use already existing equipment like linear transverters or SSB transceivers coupled to custom-designed modems operating at an intermediate frequency. While this approach may be acceptable for satellite work, it is rather complex and inconvenient for conventional terrestrial packet radio.

On the other hand, professionals developed very simple and efficient digital radios like GSM cellular telephones. Professionals also found out that they cannot use the frequency spectrum efficiently with narrowband FM radios; all new cellular phone system use high-speed TDMA techniques or even spread-spectrum modulation.

If we radio amateurs want to improve our digital communication, it is therefore necessary to develop and build new equipment. The only place for obsolete narrowband FM equipment is in a museum!

Maybe PSK modulation is not considered very efficient by many amateurs, since it is used on satellites at data rates of only 400 bit/s or 1200 bit/s. On the other hand, in Slovenia (S5) we installed our first 1.2 Mbit/s PSK links in 1995, operating in the 13cm amateur band at 2360 MHz. This equipment proved very reliable and the PSK links never failed, even when the 70cm and 23cm 38.4 kbit/s links were out due to heavy snowfall in the 1995/96 winter.

The 13cm PSK 1.2 Mbit/s link transceiver used in these links (shown in Weinheim in September 1995) was only the first attempt towards a dedicated PSK radio. The 13cm transmitter was simplified by using direct PSK modulation on the output frequency, but the 13cm receiver is still using a double downconversion followed by a conventional, if squaring-loop, PSK demodulator.

The construction of this transceiver is not simple, there are several shielded modules and especially the double-conversion receiver requires lots of tuning.

## 2. DIRECT-CONVERSION PSK DATA TRANSCEIVER

Similarly to an SSB transceiver, a PSK transceiver can also be built as a direct-conversion radio as shown in Fig.1. The Costas-loop demodulator can be extended to include most of the amplification in the receiving chain. Since such a receiver does not require narrow bandpass filters, the construction and alignment can be much simplified. In addition, some receiver stages can also be used in the transmitter (like the local oscillator chain) to further simplify the overall transceiver.

A direct-conversion PSK receiver also has some problems. Limiting is generally not harmful in the signal amplifier, however it increases the noise in the error amplifier chain. In practice, the loop bandwidth has to be decreased, if no AGC is used and both amplifiers operate in the limiting regime. It is also

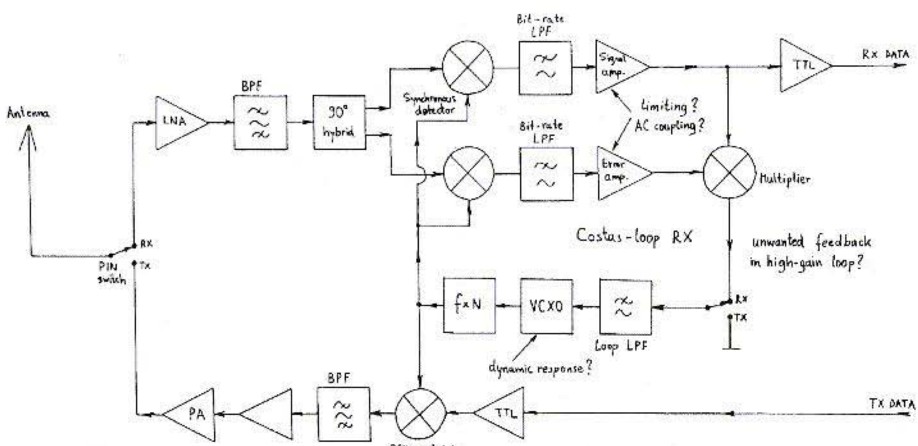



**Fig.1: Direct-Conversion PSK Data Transceiver**

very difficult to have both amplifiers DC coupled as required by the theory. If AC coupled amplifiers are used, randomisation (scrambling) has to be applied to the data stream and some additional noise is generated. However, in a well-designed direct-conversion PSK receiver, the signal-to-noise ratio degradation due to AC coupling can be kept sufficiently small.

Building a real-world, direct-conversion PSK receiver one should also consider other unwanted effects. For example, the Costas-loop demodulator includes very high-gain stages. Unwanted effects like AM modulation on the VCO or FM-to-AM conversion in the multiplier stages can lead to unwanted feedback loops. However, the most critical component seems to be the VCO. In a practical microwave PSK transceiver, the VCO is built as a VCXO followed by a multiplier chain. Although the static frequency-pulling range of fundamental-resonance and third-overtone crystals is sufficient for this application, their dynamic response is totally unpredictable above 1 kHz. The latter may be enough for full-duplex, continuous-carrier microwave links, but it is insufficient for CSMA packet radio, where a very fast signal acquisition is required.

## 3. ZERO-IF PSK DATA TRANSCEIVER

Most of the problems of a direct-conversion PSK receiver can be overcome in a so called "zero-IF" PSK receiver, as shown in Fig.2. Incidentally, a zero-IF PSK transceiver requires very similar hardware to a direct-conversion PSK transceiver. The main difference is in the local oscillator. A zero-IF PSK receiver has a fixed-frequency, free-running local oscillator, while the demodulation is only performed after the main receiver gain stages.

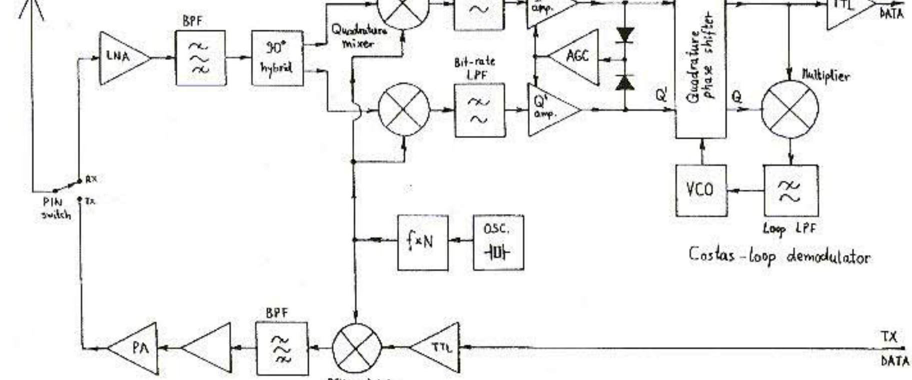



Fig.2: Zero-IF PSK Data Transceiver

A zero-IF PSK receiver includes a quadrature mixer that provides two output signals I' and Q' with the same bandwidth as in a direct-conversion RX. The signals I' and Q' contain all of the information of the input RF signal, but they do not represent the demodulated signal yet.

Since the zero-IF RX contains a free-running LO, its phase is certainly not matched to the transmitter. Further, if there is a difference between the frequencies of the transmitter and of the receiver, the phasor represented by the I' and Q' signals will rotate at a rate corresponding to the difference of the two frequencies.

To demodulate the information, the I' and Q' signals have to be fed to a phase shifter to counter-rotate the phasor. The phase shifter is kept synchronised to the correct phase and rate by a Costas-loop feedback. Since the whole Costas-loop demodulator operates at high signal levels and at relatively low frequencies, it can be built with inexpensive 74HCxxx logic circuits that require no tuning at all!

A zero-IF PSK receiver requires linear amplification of the I' and Q' signals. Limiting of the I' and Q' signals is very harmful to the overall signal-to-noise ratio. If the zero-IF amplifiers are AC coupled, data randomisation (scrambling) is required. On the other hand, a zero-IF PSK transceiver does not include any critical stages or unstable feedback loops and is therefore easily reproducible.

Searching for a simple PSK transceiver design, I attempted to build both a direct-conversion and a zero-IF PSK transceiver for 23cm. The 23cm band offers sufficient bandwidth for 1.2 Mbit/s operation. Further, the whole transceiver can be built on conventional, inexpensive glassfibre-epoxy laminate FR4. Finally, the propagation losses without optical visibility are smaller in the 23cm band than at higher microwave frequencies.

A direct-conversion PSK transceiver for 23cm proved very simple. The signal and error amplifiers used just one LM311 voltage comparator each, operating as a limiting amplifier. The only limitation of this transceiver was the VCXO.

Due to the undefined dynamic response of the VCXO, the capturing range of the Costas-loop RX was only about +1/-5 kHz. Further, even this Fig.ure was hardly reproducible, since even two crystals from the same manufacturing batch had a quite different dynamic response in the VCXO.

A zero-IF 23cm PSK transceiver resulted slightly more complex, due to the linear IF amplification with AGC and the additional Costas-loop demodulator. On the other hand, the zero-IF 23cm PSK transceiver resulted fully reproducible, since there are no critical parts or unstable circuits built in.

Since the additional complexity of the zero-IF transceiver is in the IF part, using only cheap components and no tuning points, it does not add much to the overall complexity of the transceiver.

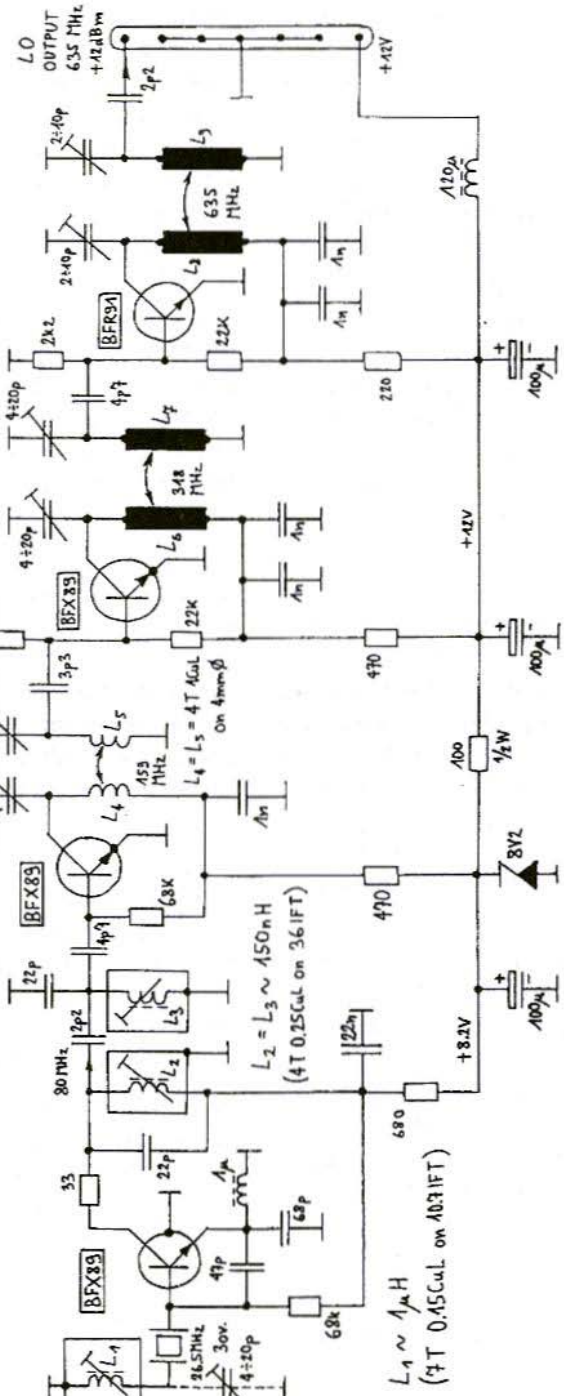



Fig.3: 435 MHz Local Oscillator

## 4. DESIGN OF THE ZERO-IF 23CM PSK TRANSCEIVER

In this article I am therefore going to describe the above mentioned successful design of a zero-IF PSK data transceiver. The transceiver is built on seven printed circuit boards, four of which (the RF part) are installed in metal shielded enclosures. The RF part is built mainly as microstrip circuits on O.8mm thick glassfibre-epoxy laminate FR4.

Sub-harmonic mixers are used both in the transmitter modulator and in the receiver quadrature mixer. Sub-harmonic mixers with two antiparallel diodes are simple to build. Since the LO signal is at half of the RF frequency, RF signals are easier to decouple and less shielding is required. Finally, it is very easy to build two identical sub-harmonic mixers for the receiver quadrature mixer.

The whole transceiver therefore requires a single local oscillator operating at half of the RF frequency, or at about 635 MHz for operation in the 23cm amateur band. The local oscillator including a crystal oscillator and multiplier stages is shown in Fig.3. The LO module is built on a single-sided PCB, as shown in Fig.4 and Fig.5.

To speed up the TX/RX switching, the receiving mixers are powered on and are receiving the LO signal all of the time. On the other hand, the LO signal feeding the modulator has to be turned

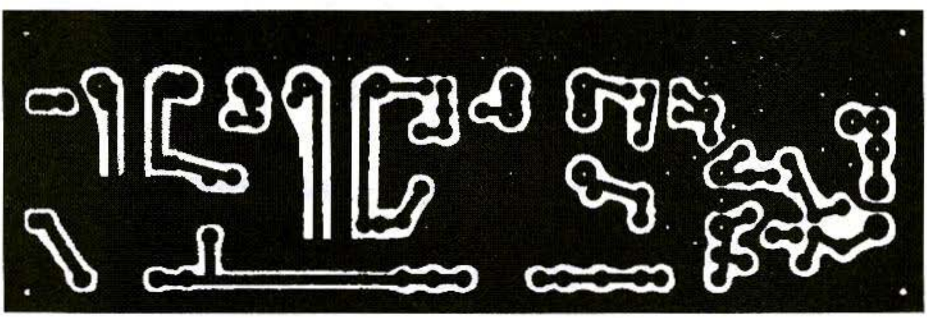


**Fig.4: 635 MHz Local Oscillator PCB - actual size 120 x 40mm 0.8mm single-sided FR4**

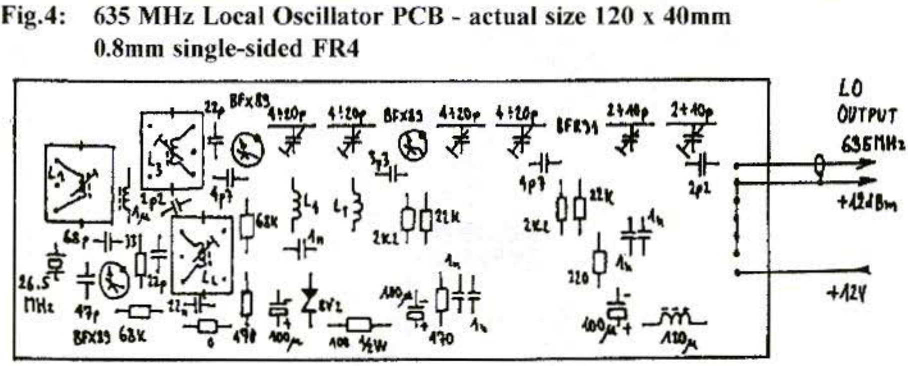



**Fig.5: 635 MHz Local Oscillator Component Overlay**

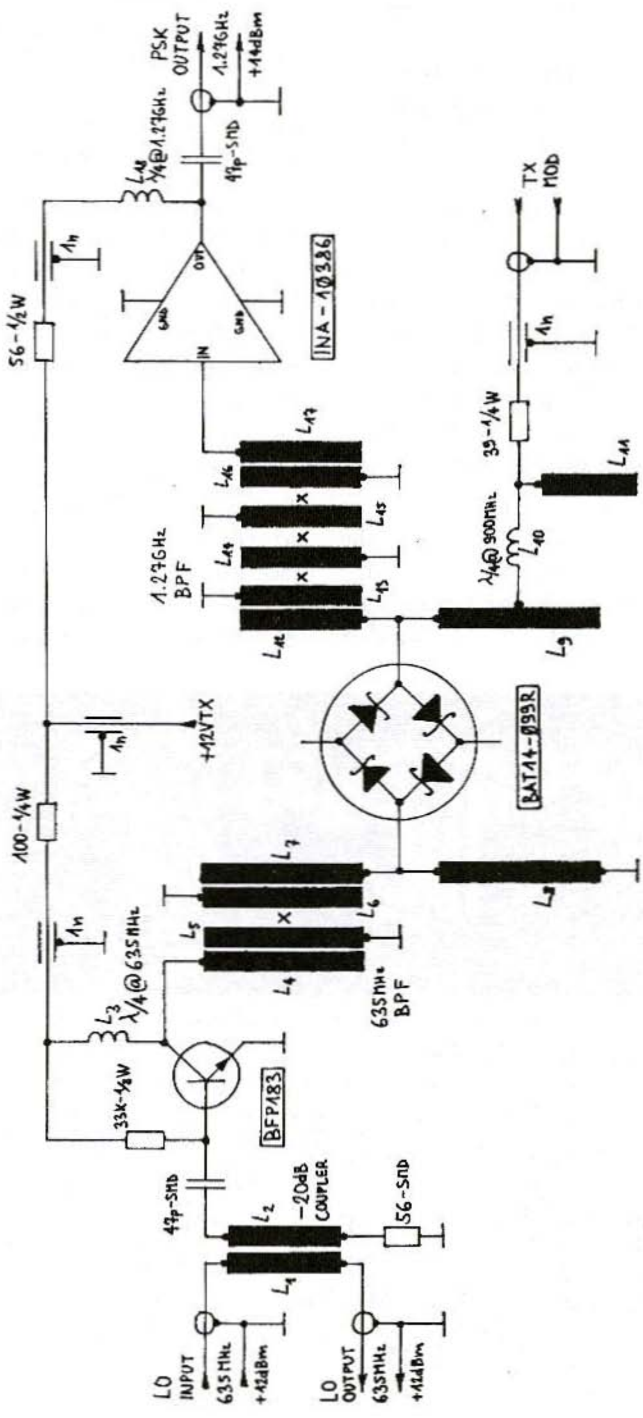



Fig.6: 1270 MHz PSK Modulator

off to avoid any interference during reception. Therefore, the LO signal is fed to the receiving mixers through a directional coupler located in the 1270 MHz PSK modulator module as shown in Fig..6.

Only a small fraction of the LO power (-20dB) is fed to a separation amplifier stage (BFP183). The 635 MHz BPF ensures a good residual carrier suppression (>30dB) in the PSK modulator. The 1.27 GHz BPF is used to suppress the 635 MHz LO signal and its unwanted harmonics. Finally, a two-stage MMIC amplifier (INA-10386) is used to boost the signal level to +14dBm.

The 1270 MHz PSK modulator is a microstrip circuit built on a double-sided PCB as shown in Fig.7 and Fig.8. The bottom side of the PCB is not etched to

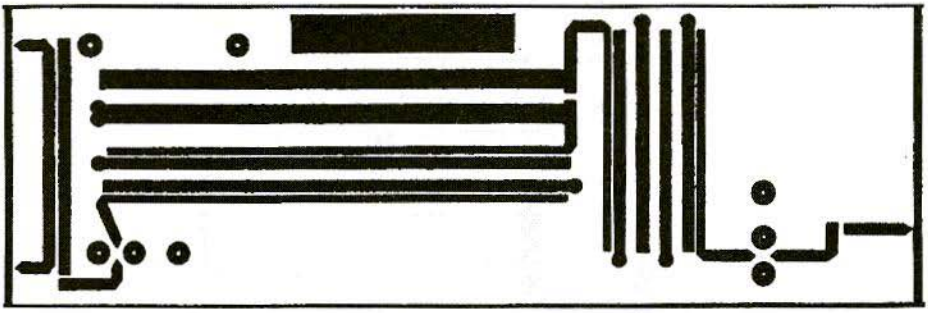


**Fig.7: 1270 MHz PSK Modulator PCB - actual size 120 x 40mm 0.8mm double-sided FR4**

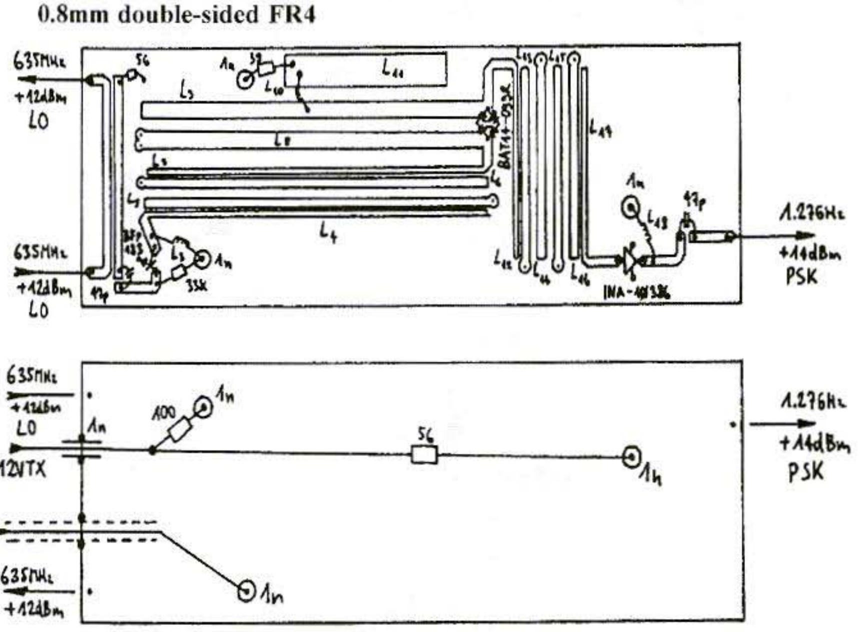



**Fig.8: 1270 MHz PSK Modulator Component Overlay**

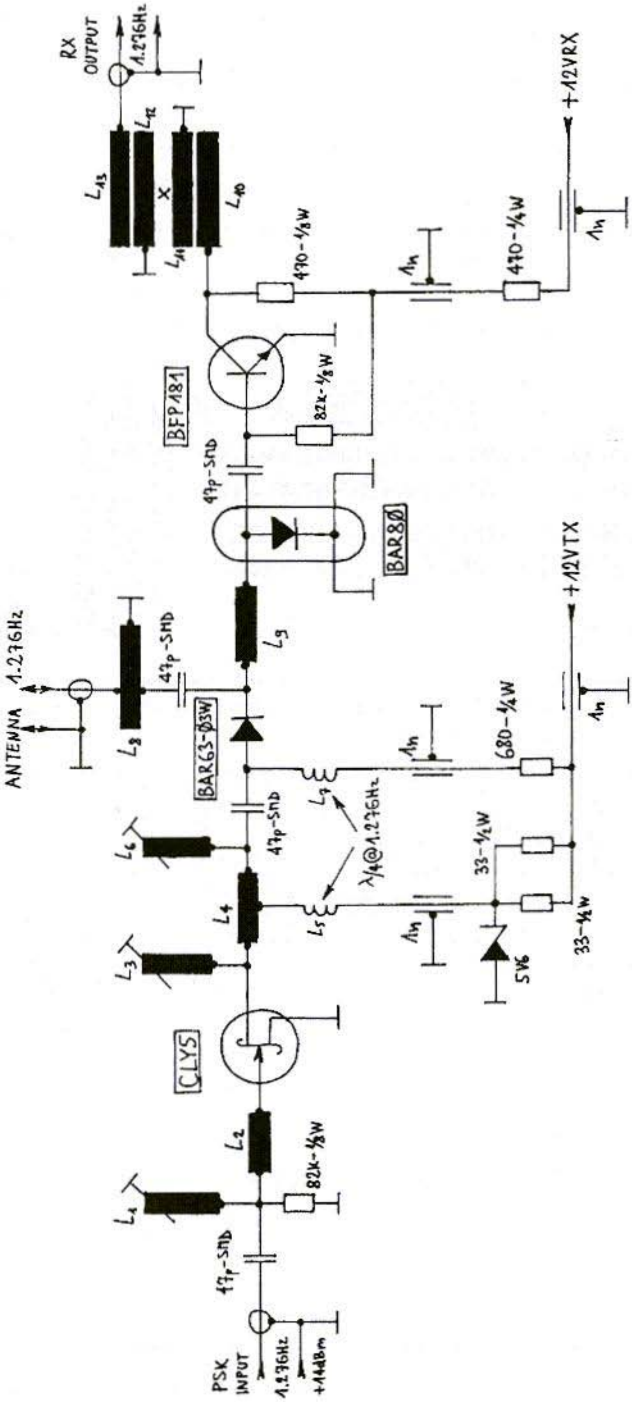



Fig.9: 23cm PSK Transceiver Front End

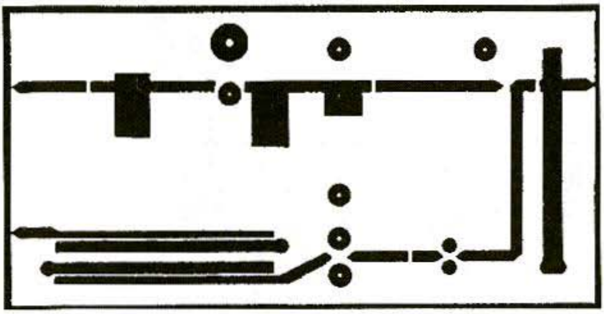


**Fig.10: RF Front End PCB - actual size 80 x 40mm 0.8mm double-sided FR4**

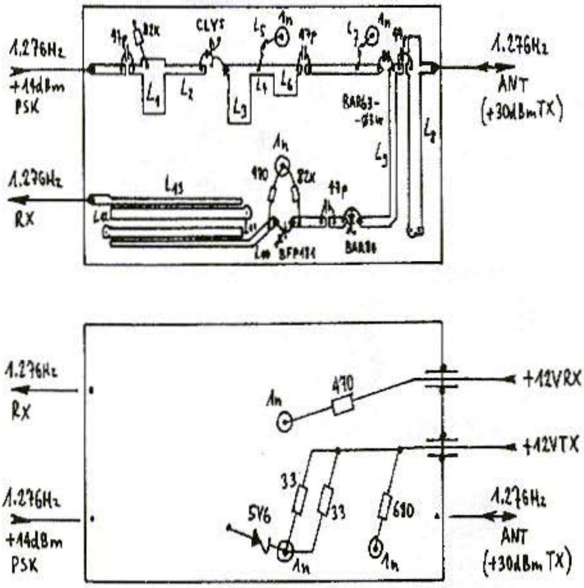



**Fig.11: RF Front End Component Overlay**

serve as a groundplane for the microstrip circuit. The RF signal losses in the FR4 laminate are rather high at 1.27 GHz. For example, the 1.27 GHz BPF has a passband insertion loss of about 5dB. On the other hand, all of the microstrip bandpass filters are designed for a bandwidth of more than 10% of the centre frequency and therefore require no tuning considering the laminate and etching tolerances.The RF front end of the 23cm PSK transceiver, shown in Fig.9, includes a TX power amplifier with a CLYS power GaAsFET to boost the TX output power to about 1W (+30dBm), a PIN diode antenna switch (BAR63-03W and BAR8O) and a receive RF amplifier with a BFP181. The latter has about 15dB gain, but the following 1.27 GHz BPF has about 3dB passband loss. The RF front end is also built as a microstrip circuit on a double-sided PCB as shown in Fig.10 and Fig.11.

The quadrature I/Q mixer for 1270 MHz, shown in Fig.12, includes an additional gain stage at 1.27 GHz (26dB MMIC INA-03184), two bandpass filters at 1.27 GHz (3dB insertion loss each), a quadrature hybrid for the RF signal at 1.27 GHz, an in-phase power splitter for the LO signal at 635 MHz, two identical sub-harmonic mixers (two BAT14-099R Schottky quads) and two identical IF preamplifiers (two BF199).

Since the termination impedances of the sub-harmonic mixers depend on the LO signal powerl the difference ports of both the quadrature (RF) and in-phase (LO) power splitters have to be terminated to ensure the correct phase and

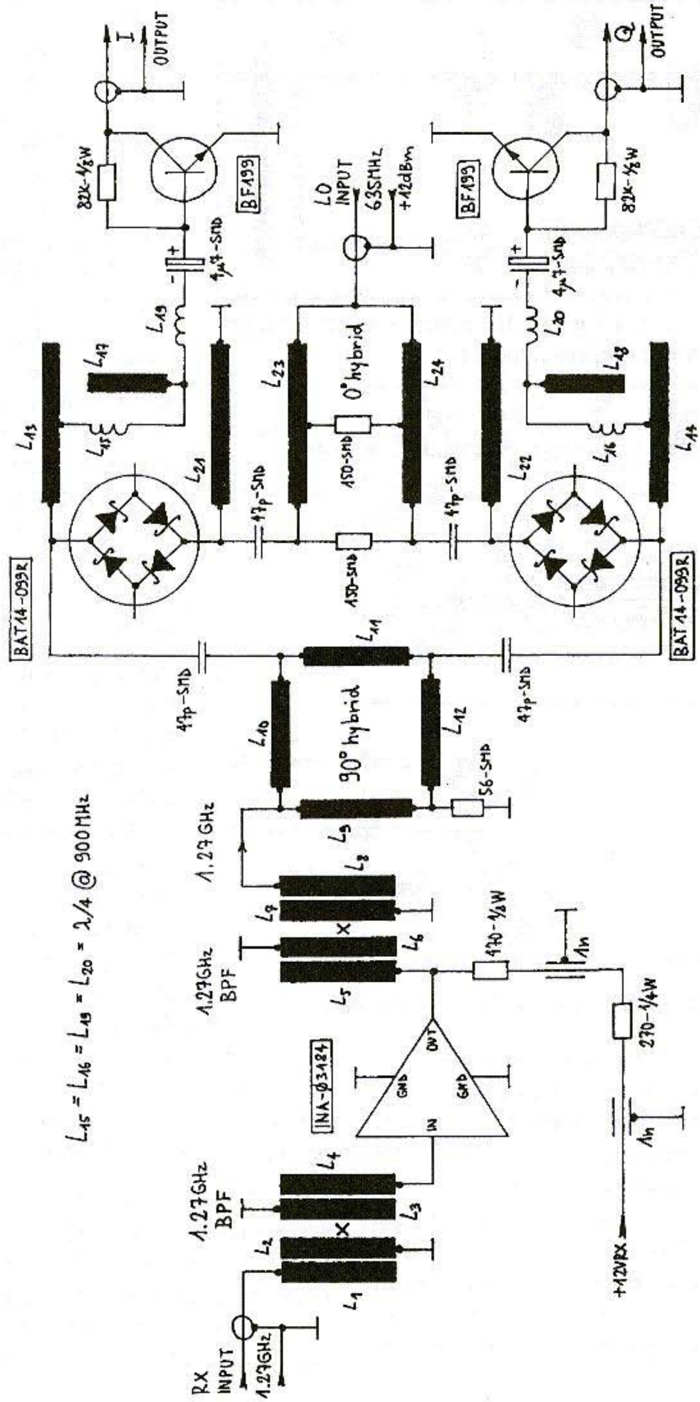



Fig.12: Quadrature I/Q Mixer for 1270 MHz

amplitude relationships. Considering the manufacturing tolerances of the microstrip PCB shown in Fig.13 and Fig.l4, the amplitude matching is usually within 5% and the phase shift is within +/- 5degrees from the nominal 90 degrees.

A zero-IF receiver requires a dual IF amplifier with two identical amplification channels1 but a single, common AGC. Since DC-coupled amplifiers can not be built, the lower frequency limit of AC-coupled stages has to be set sufficiently low. At a data rate of 1.2 Mbit/s, a convenient choice is a lower frequency limit of 1kHz. The latter allows all of the time constants in the range of 1ms (TX/RX switching time!) and causes a distortion of about 4% of the amplitude of the IF signal.

Of course, the AGC time constant should also be in the same range around 1ms. Such a fast AGC can only be

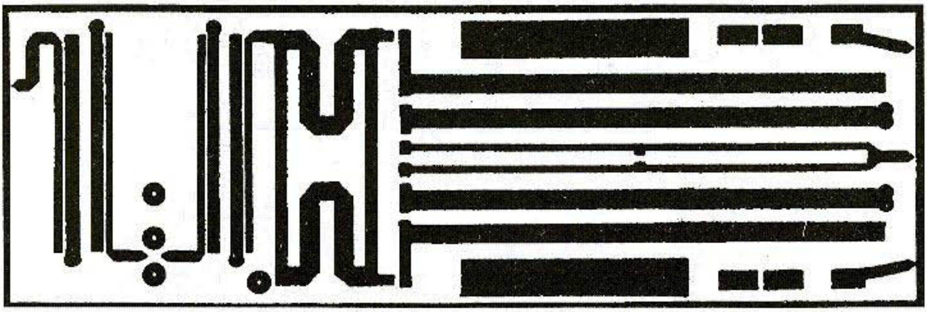


**Fig.13: Quadrature Mixer PCB - actual sixe 120 x 40mm 0.8mm double-sided FR4**

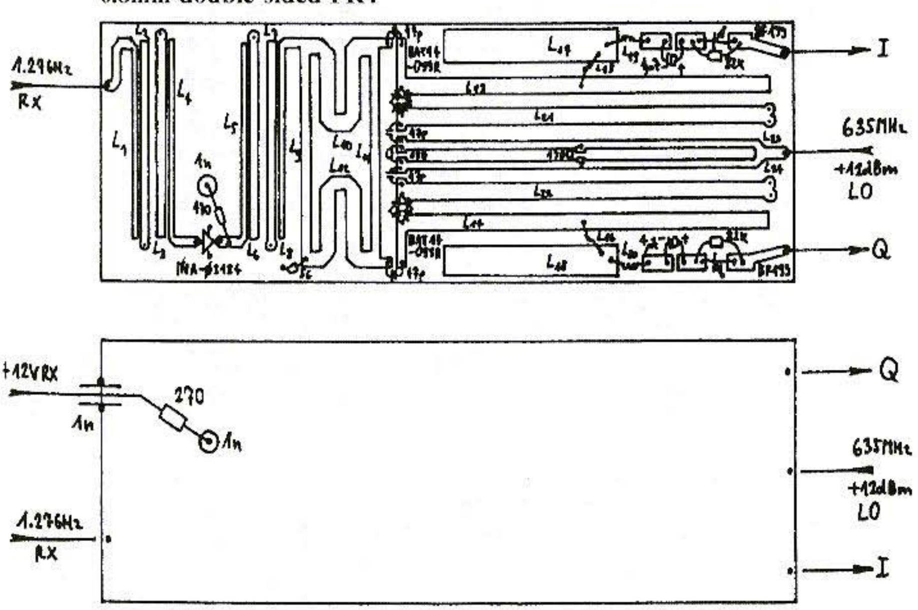



**Fig.14: Quadrature Mixer Component Overlay**

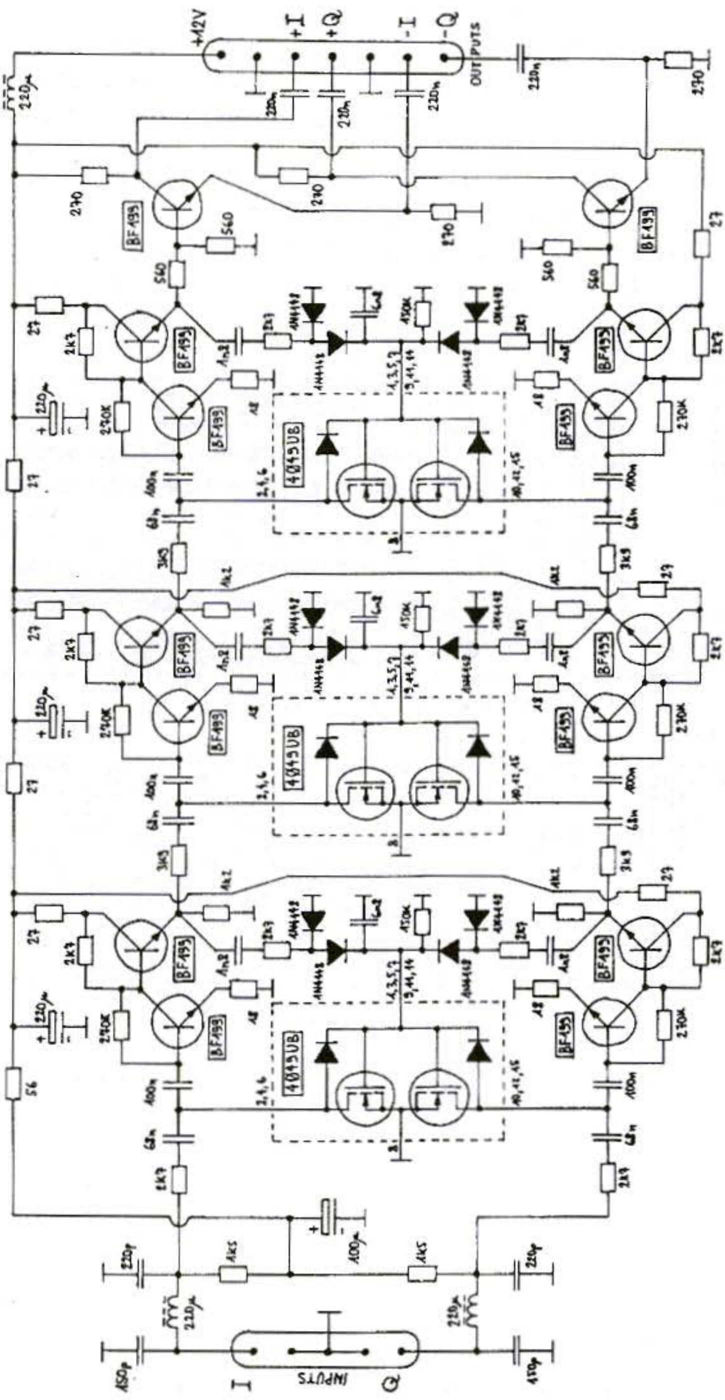



Fig.15: I/Q Dual Amplifier with Common AGC Stages

applied to low gain stages to avoid unwanted feedback. A simple technical solution is to use more than one AGC in the IF amplifier chain. The I/Q dual amplifier shown in Fig.15 has three identical dual amplifier stages and each of these dual stages has its own AGC circuit using MOS transistors (4049UB) as variable resistors.

The I/Q dual amplifier module also includes two identical lowpass filters on the input (that define the receiver bandwidth) and two phase inversion stages on the output to obtain a four-phase output signal (+I, +Q, -I and -Q) to drive the following phase shifter. The I/Q dual amplifier is built on a single-sided PCB as shown in Fig.16 and Fig.17.

The Costas-loop I/Q PSK demodulator is built entirely using cheap 74HCxxx logic as shown in Fig.18. The four-

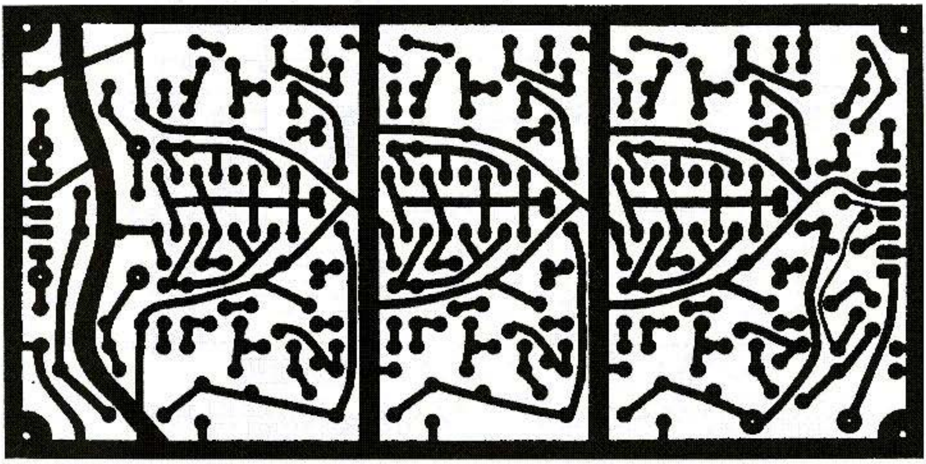


**Fig.16: I/Q Dual Amplifier PCB - actual size 120 x 60mm 1.6mm single-sided FR4**

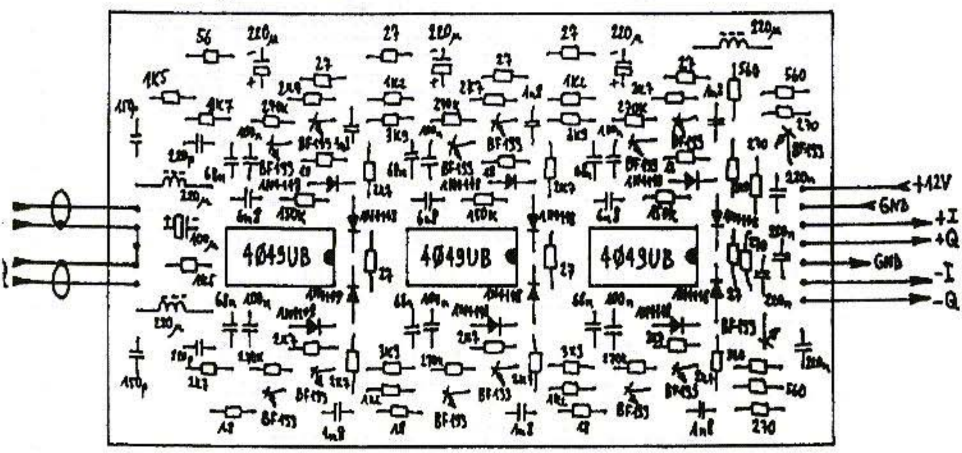



**Fig.17: I/Q Dual Amplifier Component Overlay**

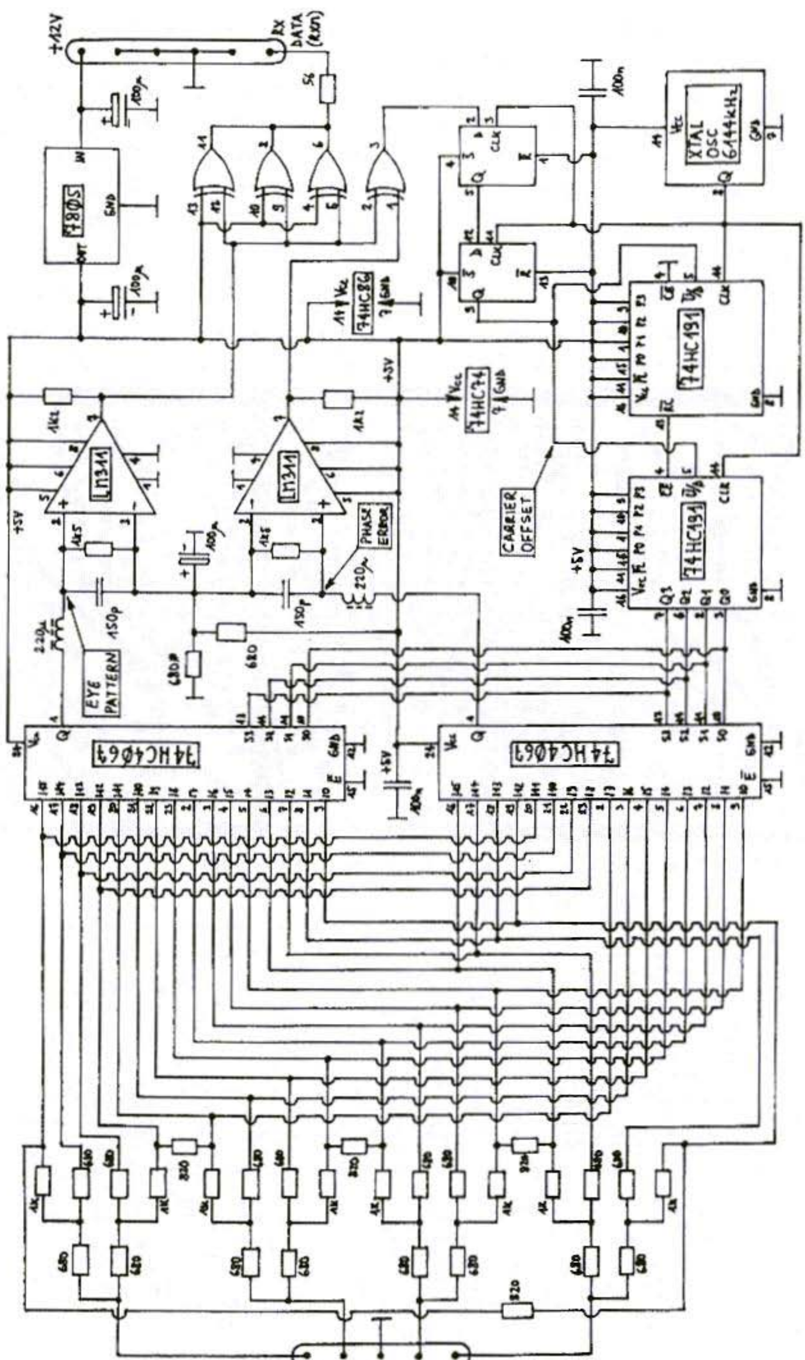



Fig.18: Costas-Loop I/Q PSK Demodulator

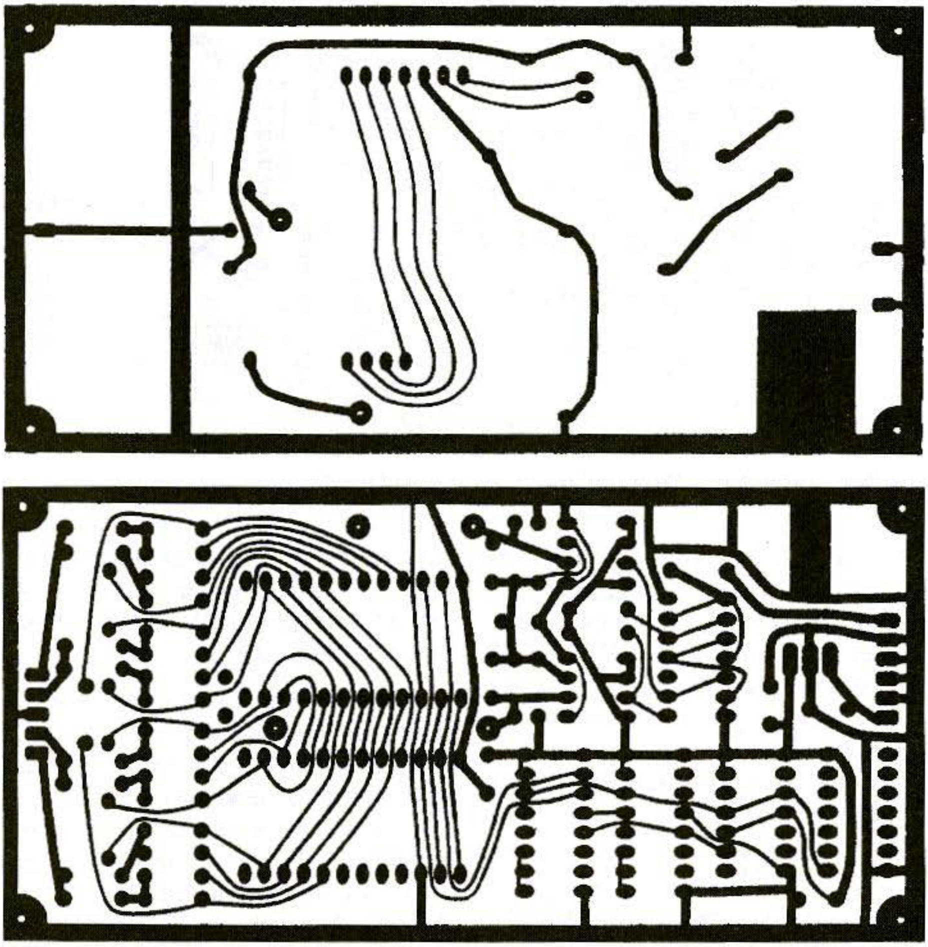


**Fig.19: Costas-Loop Demodulator PCB - actual sixe 120 x 60mm 1.6mm double-sided FR4**

phase input signal (+I, +Q, -I and -Q) feeds a resistor network that generates a multiphase system with a large number (16) of phases. Two 74HC4067 analogue switches are then used to select the desired signal phase. The inputs of the two analogue selectors are offset by 4 to provide the required 90 degree phase shift between the signal and error outputs.

Both the signal and error are first fed through two lowpass filters (to suppress the 74HC4067 switching transients) and finally to two LM311 voltage comparators to obtain TTL-level signals. The signal and error are then multiplied in an EXOR gate and feed a digital VCO. The digital VCO includes a 6.144 MHz clock oscillator and two 74HC191 up/down counters.

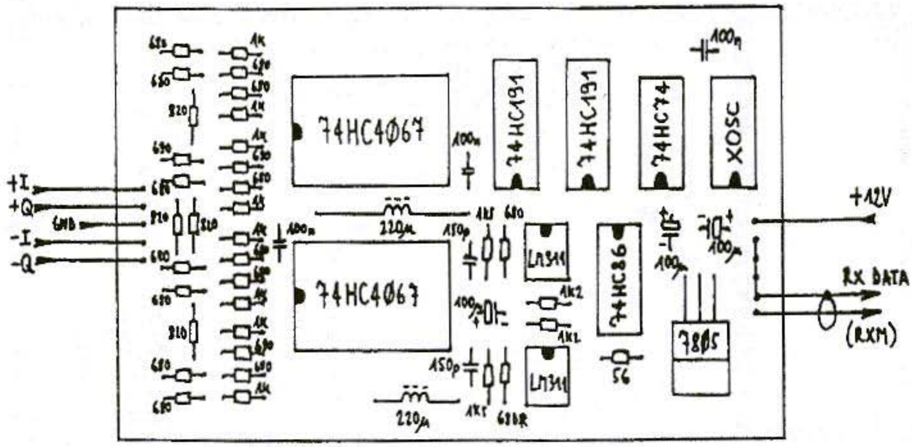



**Fig.20: Costas-Loop Demodulator Component Overlay**

The up/down control is used as the VCO control input. If the latter is at a logical ZERO, the up/down counter rotates the two 74HC4067 switches FORWARD with a frequency of 24kHz. If the input is at a logic ONE, the up/down counter rotates the two 74HC4067 switches BACKWARD with a frequency of 24 kHz. Finally, if the control input toggles, the result depends on the ON/OFF ratio of the control signal. At 50% duty the 74HC4067 switches stay in the same position.

The overall circuit therefore operates as a first-order, Costas phase-locked loop that is able to correct carrier-frequency errors between -24 kHz and +24 kHz. The loop gain is defined by the dividing ratio of the 74HC191 up/down counters and the clock frequency. If a wider capturing range is desired, the clock frequency can be increased up to 20 MHz, but the resulting higher loop gain also increases the phase noise!

The Costas-loop demodulator is built on a double-sided PCB as shown in Fig.19 and Fig.20. The circuit includes its own +5V regulator and an output stage capable of feeding a 75Ω cable with the emodulated RX data.

The overall PSK transceiver requires a few additional interface circuits (shown in Fig.21) including a supply voltage switch and a modulator driver. The modulator driver includes a lowpass filter to decrease the high-order sidelobes of the modulation spectrum. The supply switch interface is built on a single-sided PCB as shown in Fig.22 and Fig.23.

The overall PSK transceiver is enclosed in an aluminium box with the dimensions of 320mm (width) X 175mm (depth) X 32mm (height). The location of the single modules is shown in Fig.24. The four RF modules are additionally shielded in small boxes made of 0.5mm thick brass sheet as shown in Fig.25. The groundplane of the PCBs is soldered along all four sides to the brass frame to ensure a good electrical contact.

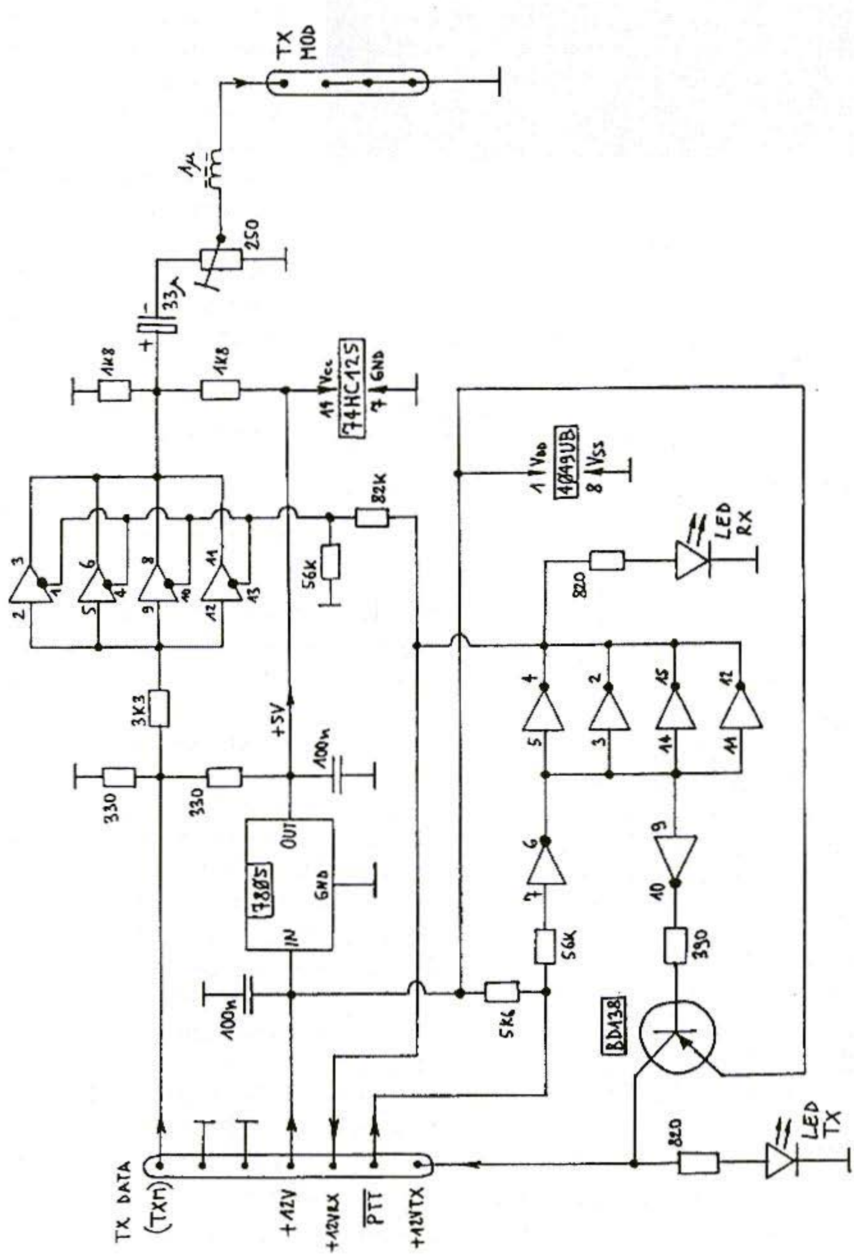
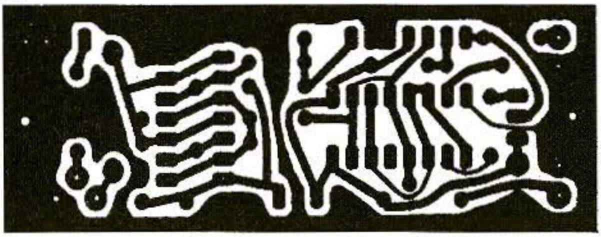



Fig.21: Supply Switch Interface



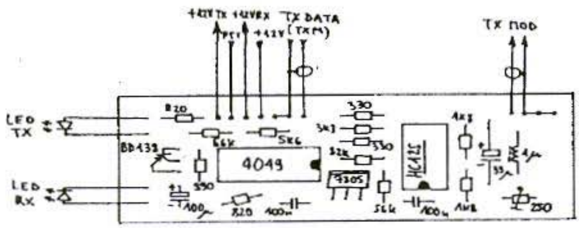



**Fig.23: Supply Switch Interface Componet Overlay**

Special care should be devoted to the assembly of the microstrip circuits. The microstrip resonators are grounded at the marked positions using 0.6mm thick CuAg wire. The SMD components are grounded through 2.5mm, 3.2mm or 5mm diameter holes at the marked positions. The holes are first covered with a piece of thin copper sheet on the groundplane side, then they are filled with solder and finally the SMD part is soldered in place.

The assembled PSK transceiver requires little tuning. The only module that needs to be tuned in any case is the local oscillator module. Since most of the stages are just frequency doublers, it is very difficult to tune this module to the wrong harmonic.

The TX power amplifier may need some tuning to get the maximum output power. As printed on the circuit board, L1 in the RF power amplifier should not require any tuning if the interconnecting 50Ω Teflon cable from the modulator is exactly 12cm long. Tuning L3 and L6 the output power can only be increased by less than 100mW. All of the other microstrip resonators should not be tuned. Finally, the 250Ω trimmer in the supply switch interface is adjusted for the maximum TX output power (usually 2/3 of the full scale).

## 5. INTERFACING THE 1.2 MBIT/S PSK TRANSCEIVER

Amateur packet radio interfaces for data rates above lOO kbit/s are not very popular. One of the most popular serial interfaces, the Zilog Z8530 SCC, only includes a DPLL for RX clock recovery that can operate up to about 25O kbit/s. Other integrated circuits, like the old Z8OSIO, the MC68302 used in the TNC3 or the new MC68360 do not include any clock recovery circuits at all. In addition to the RX clock recovery, data scrambling/descrambling and sometimes even NRZ/NRZI differential encoding/decoding have to be provided by external circuits.

The circuit shown in Fig.26 was specially designed to interface the described PSK transceiver to a Z8530 SCC,

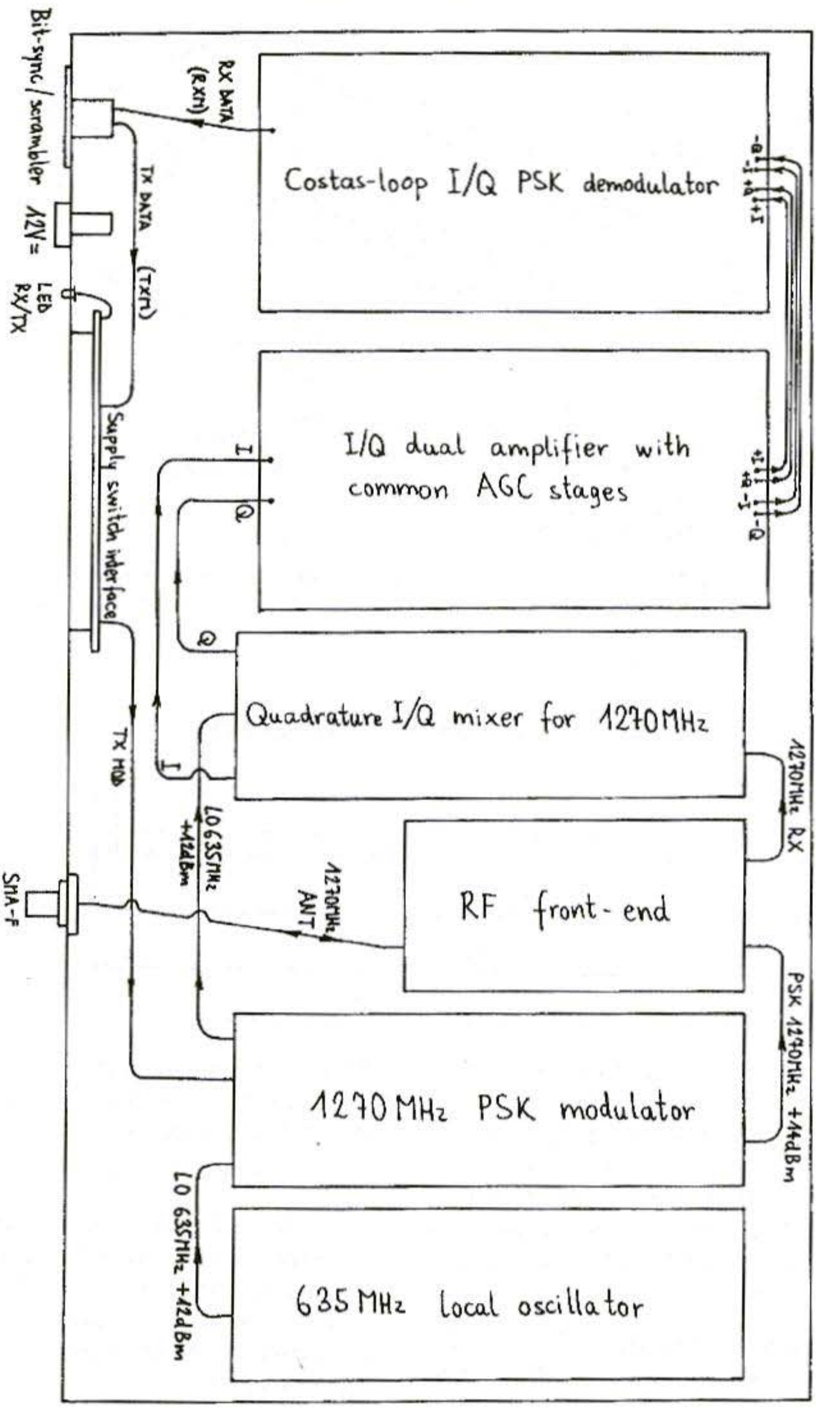



Fig.24: 23cm PSK Transceiver Module Location

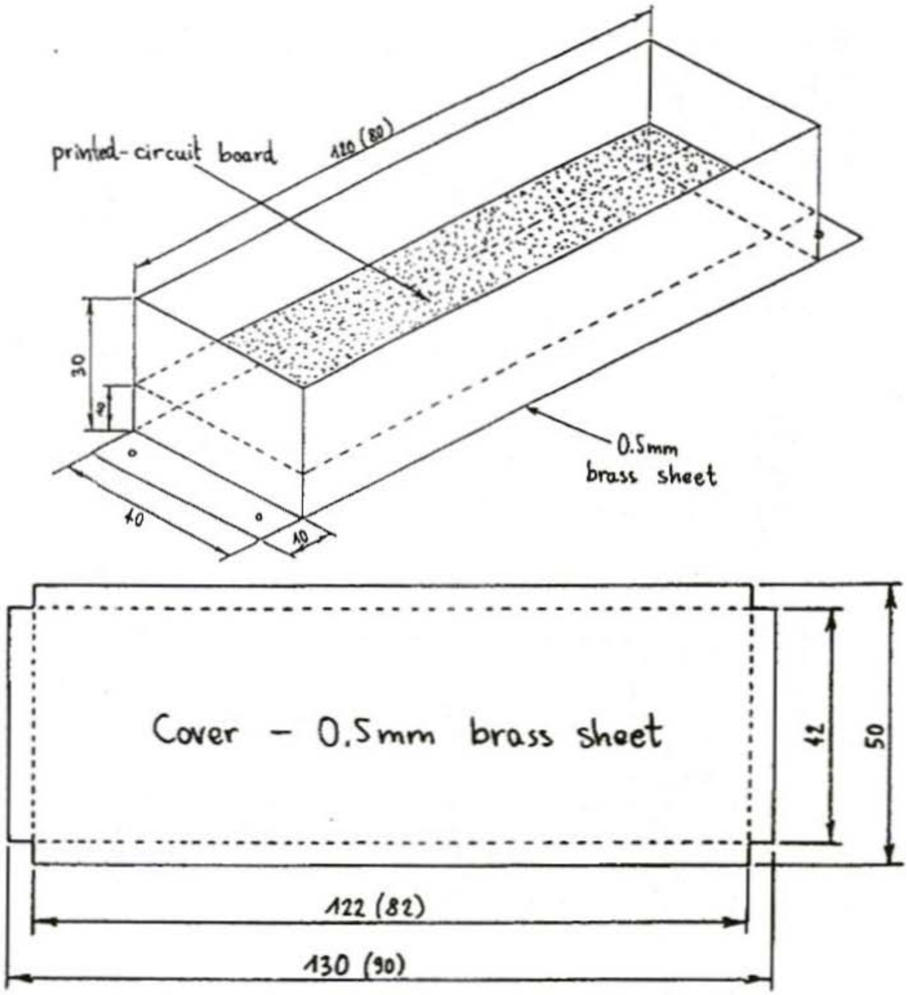



**Fig.25: 23cm PSK Transceiver Shielded Module Enclosure**

although it will probably work with other serial HDLC controllers as well. The circuit includes an interpolation DPLL that only requires an 8-times higher clock frequency (9.830 4MHz), although provides the resolution of a /256 conventional DPLL with a 315 MHz clock.

The scrambler/descrambler uses a shift register with a linear feedback with EXOR gates. The scrambling polynomial is the same as the one used in K9NG/G3RUH modems:

1+X**12+X**17

Due to the redundancy in the AX.25 data stream (zero insertion and deletion), a simple polynomial scrambler is completely sufficient to overcome the AC coupling limitation of the described PSK transceivers.

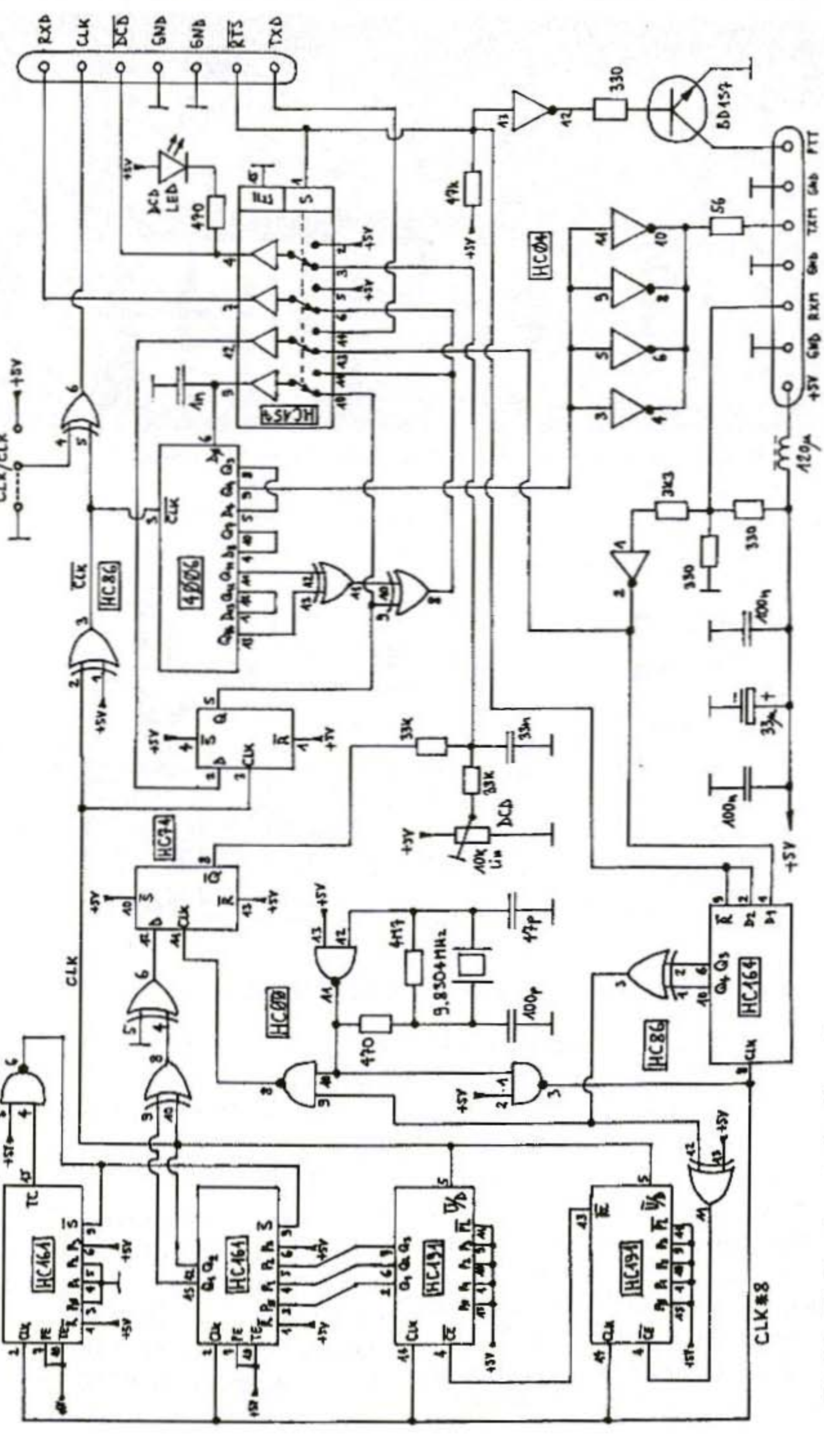



Fig.26: Bit-Synchronisation/Scramber

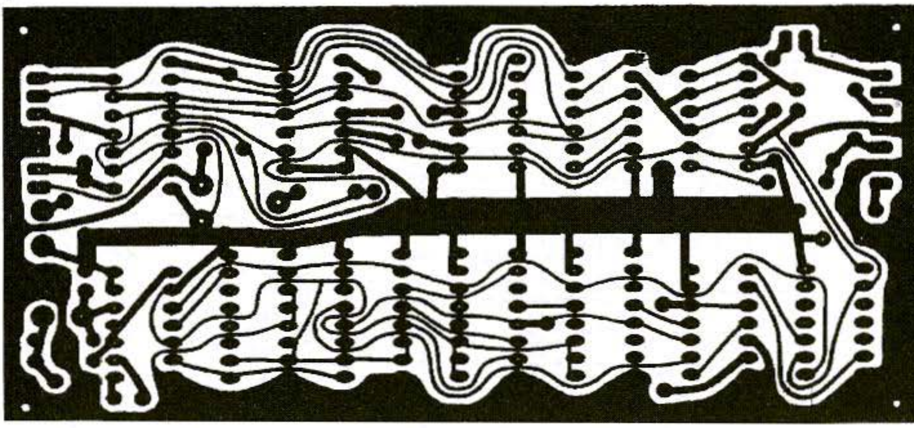


**Fig.27: Bit-Synchroniser/Scrambler PCB - actual size 120 x 60mm 1.6mm single-sided FR4**

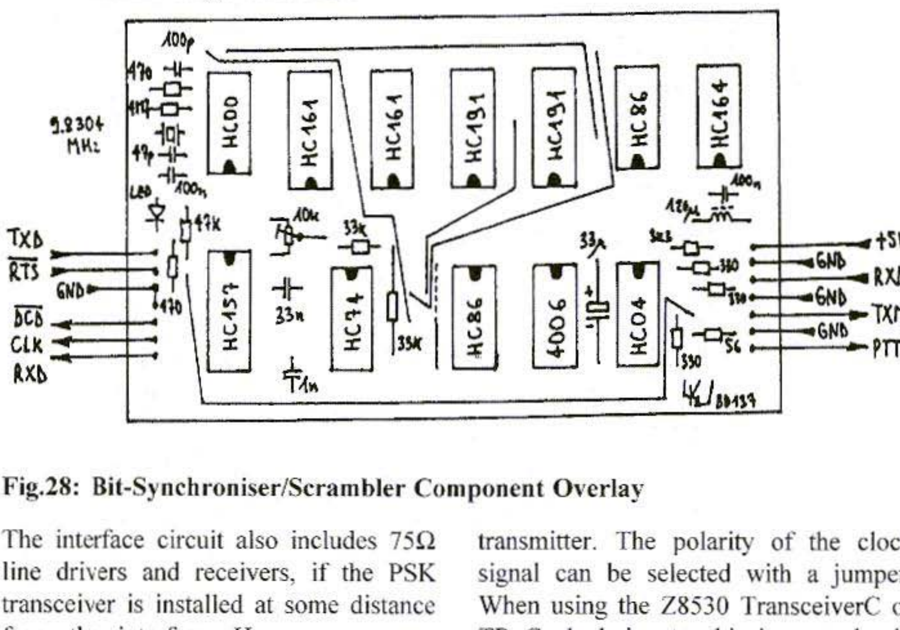



**Fig.28: Bit-Synchroniser/Scrambler Component Overlay**

The interface circuit also includes 75Ω line drivers and receivers, if the PSK transceiver is installed at some distance from the interface. However, connections have to be kept short on the side towards the computer serial port. The described interface only provides one clock signal, since it is intended for simplex operation with the described PSK transceiver. Of course the DPLL is disabled during transmission, so that the circuit supplies a stable clock to the transmitter. The polarity of the clock signal can be selected with a jumper. When using the Z8530 TransceiverC or TRxC clock inputs, this jumper should be connected to ground.

The bit-synchronisation/scrambler circuit is built on a single-sided PCB as shown in Fig.27 and Fig.28. It only requires one adjustment, the DCD threshold, and the latter can only be performed when noise is present on the RXM input.

*D.Eckart Schmitzer, DJ4BG*

# Danger - Parasites!

**The design of circuits is generally based on what is still a very idealised way of representing the circuit diagram. No-one worries about so-called dirty effects until something really inexplicable happens. Even so, real components have characteristics quite different from those desired.**

## 1. INTRODUCTION WITH EXAMPLE

The following example shows how much the "same" circuit can differ in theory and practice. A very simple low-pass is assembled, in the form of a Pi-filter, which is intended to filter out the harmonics of a short-wave transmitter.

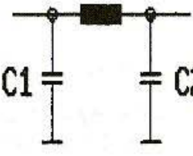



**Fig.1: Ideal PI Filter Circuit**

This low-pass has a rated limiting frequency of app. 32 MHz. The circuit is shown in Fig.1. Fig.3a shows the attenuation curve to be expected theoretically, determined purely by calculation.

If this circuit is measured over a sufficiently large frequency range, some surprising discrepancies are noted at higher frequencies, as against the purely theoretical filter curve. Additional attenuation peaks and even gaps appear, which were completely unexpected. The question is - why?

Any real component has additional parasitic characteristics which must be taken

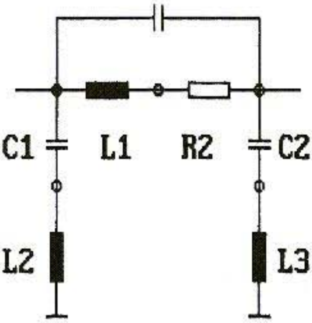



**Fig.2: PI Filter Circuit with Parasitic Elements**

into account. Wired-up capacitors have feeds, which display the characteristic of having very low levels of inductance. Coils have an additional capacity, admittedly very small, from one end to the other, as well as losses due to radiation and Ohmic losses, which can be included (much simplified) in a series resistance.

If these unwanted but unavoidable additional components are entered in the original circuit diagram, the low-pass suddenly looks quite different (Fig.2).

The precise values of the parasitic elements can be estimated in all cases. They are very dependent on the mechanical structure and on the components themselves. A coil which is relatively thin but stretched lengthwise has a somewhat lower parasitic capacity from one end to another than a short coil with a larger diameter. A short capacitor with broad, strip-form feeds displays a markedly lower serial inductance than a long tubular capacitor with thin feed wires. The rule of thumb applying here is:

1cm wire corresponds to app. 10nH.

A coil made of thick, silver-plated wire, which is also suitably screened, will have a lower loss resistance than a freely structured coil made of thin lacquered wire without silver.

The parasitic elements depicted in Fig.2 are pure estimated values, which can turn out considerably differently in practice. But it is the principle which is important here, not their precise values.

The coil was assumed to have a quality of 100 so that, in accordance with the definition of coil quality, the loss resistance is:

$$Rv = X_l / Q$$

This can be more expediently specified at the limiting frequency - here in the example, at 32 MHz.

The different feed inductance values at C1 and C2 are intentional. This was dealt with in the clarification to Fig.2. It was also assumed here that only one component was incorporated in each

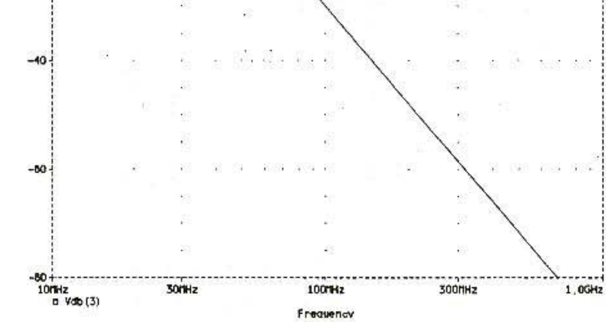



**Fig.3a: Ideal Filter Curve**

***LC-Tiefpass ohne paras. Elemente*** **= LC Low-Pass without parasitic elements**

case - i.e., for example, the capacitors did not consist of several individual capacitors wired in parallel. Otherwise, these would have had to be shown separately with their individual feed inductance values.

## 2. FILTER SIMULATION

In order to see the effect of these parasitic elements, you can assemble the circuit and measure it thoroughly. An alternative, for which no actual measuring equipment is required, is to simulate the circuit on a computer. A suitable program for this, by way of example, is PSPICE. The full version of this program is a costly item in itself. But there is a free demo version, which you are explicitly allowed to copy (see literature reference [1] at end of article).

This demo version of PSPICE is fully operative, but is restricted as regards the number of circuit nodes. "Smaller" circuits of up to about 100 nodes can be processed without restrictions. This is sufficient for most of our applications.

For this purpose, all "nodes" (connection points between the components) are to be consecutively numbered. Thus every component can be described by a name, its nodes in the circuit, and its value.

The syntax for entering the circuit in PSPICE runs: component name, node numbers, component value.

The star, "*", indicates a comments line. For examples, see C1 and L1 in the command list below.

*Symbolic wiring diagram with node numbers. (Not required for entry! Given here only to show nodes.)

```
         .-------C3--------.
         |                 |
1---R1---2---L1---4---R2---6--------.
|        |                 |        |
|        C1                C2       |
|        |                 |        |
Vi       3                 5        R3
|        |                 |        |
|        L2                L3       |
|        |                 |        |
-   0    -                 -   0    -
```

LC-Low Pass with parasistic elements

| | | | |
|---|---|---|---|
| Vi | 1 | 0 | AC 2V |
| R1 | 1 | 2 | 50 |
| C1 | 2 | 3 | 100p |
| L2 | 3 | 0 | 10n |
| L1 | 2 | 4 | 0.5µ |
| R2 | 6 | 4 | 1 |
| C3 | 2 | 6 | 6p |
| C2 | 5 | 6 | 100p |
| L3 | 5 | 0 | 8n |
| R3 | 6 | 0 | 50 |
| .ac | dec 100 | 10 meg | 1000 meg |
| .probe | | | |
| .end | | | |

Clarifications which are not entered:

Heading (There must be one!)

- Vi = EMK 2V generator
- R1 = 50Ω generator source resistance
- C1 Filter capacitor C1 (100pF) lies between nodes 2 and 3
- L2 Parasitic series inductance L2 lies between nodes 3 and 0 (earth) with 10nH
- L1 L1 filter coil, 0.5µH
- R2 Loss resistance for Q = 100
- C3 Parasitic coil capacity
- L3 Parasitic series inductance
- R3 Moving load from knot 6 to 0
- .ac Command for: AC simulation, decadic, with 100 measuring points per decade from 10 MHz to 1000 MHz

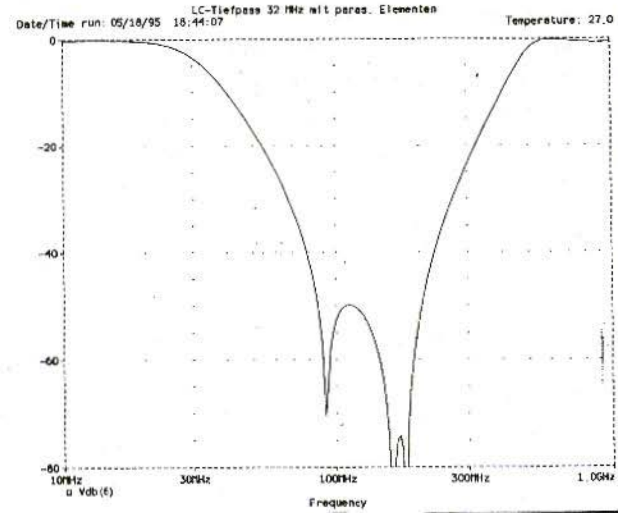



**Fig.3b:**
**Filter Curve with Parasitic Elements (Estimated Values)**

***LC-Tiefpass 32 MHz mit paras. Elementen =***
**32 MHz LC low-pass with parasitic elements**

.probe Calls up graphic output program

Only the data in the box are to be entered!

Fig.3b shows the result of the simulation. It can be seen that it is blatantly different from the theoretical curve of Fig.3a! Apart from unique resonance points, it is clear that at very high frequencies the low-pass filter becomes fully conducting again.

The attenuation peak at 91 MHz arises from the parallel resonance of L1 and its stray capacitance (C3), whilst the attenuation peaks at 158 MHz and 177 MHz are series resonances of C1 and C2 with the series inductances, which are assumed to be different.

At frequencies of app. 500 MHz and above, the low-pass filter becomes fully conducting again. C1 and C2 are now

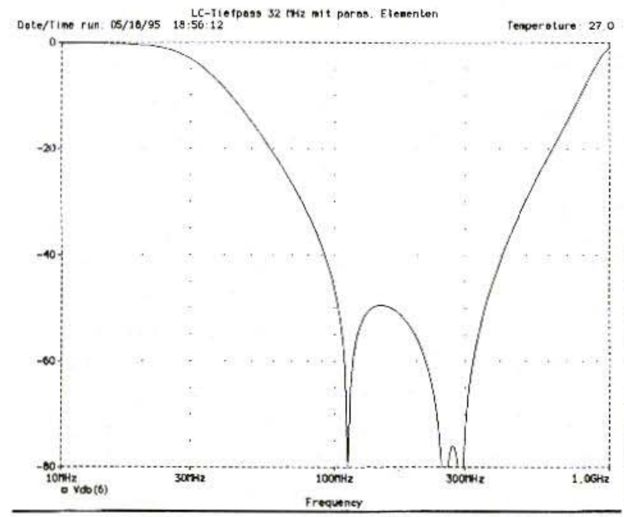



**Fig.3c:**
**Filter Curve with Reduced Parasitic Elements**

***LC-Tiefpass 32 MHz mit paras. Elementen =***
**32 MHz low-pass with parasitic elements**

negligible, as against their feed inductances, whilst L1 is negligible, as against its parasitic parallel capacity. Thus in the UHF television band all harmonics of a transmitter would be radiated to their full extent if it was fitted with this low-pass filter.

If the series inductances of the capacitors can be reduced to app. 3 or 4nH by shortening their feeds still further, and if the filter can be assembled with a slightly slimmer coil with a self-capacitance of only 4pF, we obtain the curve in Fig.3c.

The parallel resonance of the coil has risen to around 112 MHz and the series resonance of the capacitors to 252 or 289 MHz. The filter does not now become fully conducting again until slightly above 1 GHz, whilst at the top end of the television band 5 we still have about 6 dB attenuation at 860 MHz.

With more extensive filter circuits, the number of resonance points caused by parasitic elements becomes correspondingly greater and less definable.

## 3. CONSEQUENCES

It may appear illusory to be considering harmonics right up into the region of 1 GHz for a short-wave transmitter. But if we consider that, for example, T-MOS-FET's of medium power display limiting frequencies of 1 to 2 GHz, it can be seen that it is still perfectly possible for something to happen here with a CW or FM transmitter in C mode.

In SSB mode, where the transistors are usually operated linearly, the risk that such harmonics will be created is certainly markedly lower, but can not be completely excluded. For this reason, even in the short-wave range, "VHF-type" structures should generally be preferred. All the more so at higher frequencies, at which the component values tend to be correspondingly nearer to the values of the parasitic elements. Other forms of component must be used here under certain circumstances, depending on the frequency range - e.g. strip lines - for the processing and simulation of which PC programs are also available - for example, [2].

## 4. NOTE ON SECONDARY EFFECT

Finally, we should also mention another effect of parasitic elements, which is usually overlooked in practice. If we compare the theoretical curve (Fig.3a) with the curves in which the parasitic elements were considered (Fig.3b or Fig.3c), we can establish that the 3dB limiting frequency has also been somewhat displaced - from the predicted value of 32 MHz (more precisely, 31.83 MHz with the level component values selected) to 29.08 MHz or 29.9 MHz, depending on the estimated values for the parasitic elements.

Transverse capacitors have the effect of bringing the Ohmic values down somewhat, due to their parasitic series inductances, as if their capacity were rather greater. Even far away from the resonance frequency, a series resonance circuit has lower Ohmic values than its individual components. The same is true of the series inductance, which becomes rather more high-Ohmic due to its parasitic parallel capacity, as if its inductance were somewhat higher than the actual value. Even far away from the resonance frequency, a parallel resonance circuit has higher Ohmic values than its individual components. The limiting frequency is thus displaced to frequencies which are lower, or not as low.

In practice, this means that the theoretical limiting frequency must be set slightly higher, so that a quite specific limiting frequency can be obtained in reality for the low-pass in question here.

## 5. LITERATURE


[1] Simulation using PSPICE Dietmar Ehrhardt and Jörgen Schulte Vieweg-Verlag (Book includes voucher for free demo version of PSPICE)

[2] Puff CAD Software VHF Communications 2/91, pp. 66-68

*Dr. Ing. Jochen Jirmann, DB1NV*

# Noise Behaviour of Zener Diodes

**While assembling a broad-band amplifier for a magnetic active aerial, the author observed a number of remarkable effects when he used Zener diodes for phase coupling and potential displacement. This led him to study the phenomenon of noise in Zener diodes in greater detail.**

## 1. INTRODUCTION

Noise from power supplies has already been discussed frequently in "VHF Communications" in relation to low-noise oscillator circuits. Previous articles have been more concerned with practical aspects, namely the assembly of "cleaner" power supply systems, whereas no attention has been paid to the main cause of the phenomenon - noise in the voltage reference.

As manufacturers do not give information about the noise behaviour of their products, the only alternative was to carry out my own measurements. No attempt was made to investigate the "flicker noise" at frequencies of a few Hertz, as the measurement technique is not exactly simple. Due to the small quantity of equipment involved and to the fact that testing was limited to three companies' products (ITT, Motorola and Philips), my results are certainly not a hundred per cent representative, but comparison measurements on a few ancient Z diodes from the sixties and seventies showed that the trend is correct.

## 2. THE PHYSICS OF ZENER DIODES

Two distinct physical mechanisms are concealed behind the Zener diode component:

### 2.1. The Zener Effect

In Zener diodes with relatively low breakthrough voltage levels below about 6V, the Zener effect is the triggering factor (the electrons "tunnel through"

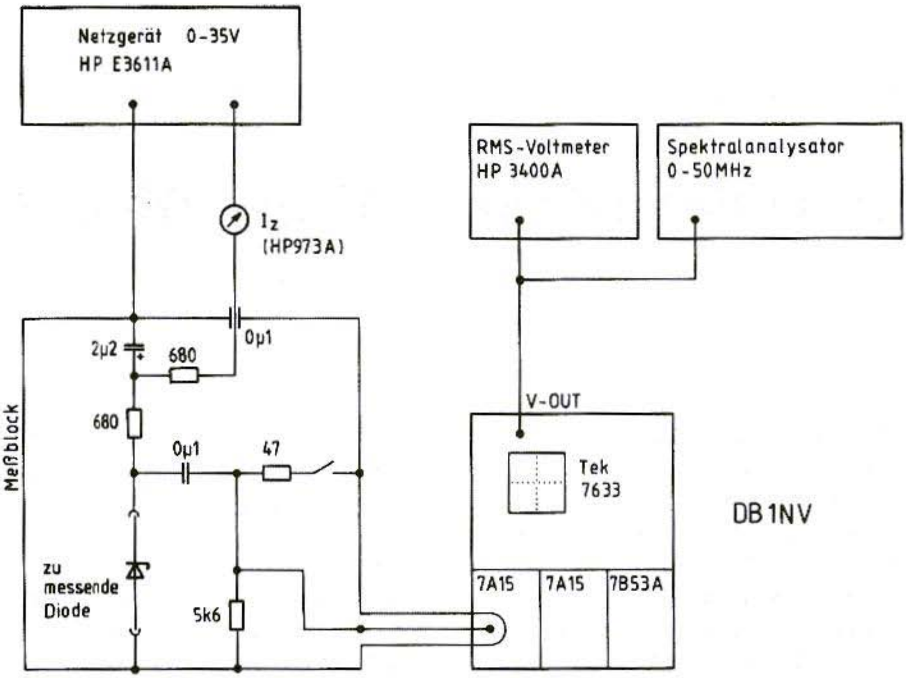



**Fig.1: Measuring Rig for Measuring Zener Diode Noise**
***Netzgerät* = Power supply, *Spectralanalysator* = Spectrum analyser,**
***Meßblock* = Measuring block, *Zu messende Diode* = Diode to be measured**

the barrier layer, as in the tunnel diode). These Zener diodes display a negative Zener voltage temperature cycle, and the bend in the barrier characteristic is not particularly sharp. To put it another way, the differential internal resistance (i.e. the reciprocal value of the characteristic gradient) of the Zener diode is rather high. For a current of 1mA and a Zener voltage of 2.4V - 5.6V, it can lie between 100Ω and 400Ω. Values of around 100Ω are typical for a Zener current of 5mA.

## 2.2. The Avalanche Effect

Zener diodes with breakthrough voltages exceeding 6V use the avalanche effect, or avalanche breakdown. Here electrons originating spontaneously in the barrier layer are so strongly accelerated through the barrier voltage which is applied that they can knock electrons out of other atoms, which are now accelerated in their turn. The result is an electron avalanche, which we observe as current (I hope the physicists among our readers will forgive the simplified representation without band models). The avalanche breakdown has a positive temperature coefficient and the differential internal resistance is considerably lower. For a current of 1mA and for Zener voltages of 6.8V - 15V, 30Ω to 200Ω is typical. For a current of 5mA, the current is only 5 - 20Ω.

Zener diodes with a breakthrough voltage of between 5 and 6V are distinguished by two features

Firstly, the differential internal resistance reaches its absolute minimum here, and secondly the Zener and avalanche effects are superimposed. Since the two effects have opposing temperature cycles, the resulting temperature coefficient is almost zero. For this reason, reference diodes with high stability and a low temperature cycle are usually structured for around 6V.

With regard to the noise behaviour of Zener diodes, most instruction manuals will tell you that the noise diminishes as the Zener current increases.

While creating a broad-band amplifier, the author came across the following remarkable facts:

1. Noise values differing by powers of ten can be obtained even from diodes from the same source
2. The avalanche and Zener effects display completely different types of noise behaviour
3. Chaotic relationships can often be observed between diode currents and noise
4. The spectrum distribution of the noise varies with the specimen diode and the current

**Measurement Bandwith 300 Hz - 15 MHz**

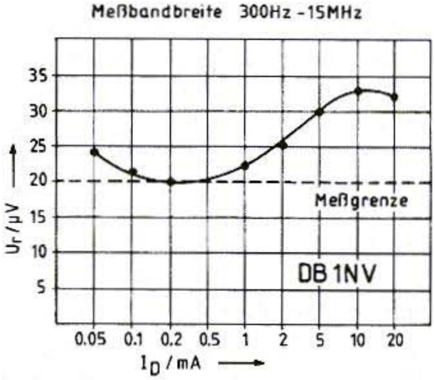



**Fig.2: Noise Behaviour of "real" Zener Diode BZX83C4V3**
***Meßgrenze* =Measuring limit**

## 3. A MEASUREMENT RIG TO DETERMINE THE NOISE PROPERTIES

To measure the noise of a Zener diode, you need a well-screened and decoupled measurement rig, as otherwise local medium-wave and short-wave transmitters can be measured instead of the diode noise.

Fig.1 shows the rig set up. A closable tinplate housing contains the Zener diode. The bias voltage is fed into the screening housing through a feedthrough capacitor and an RC module, against low-frequency noise. The bias voltage is provided by a low-noise power pack (HP E3611A). The diode current is measured using a battery-driven multimeter (HP 973A), which avoids parasitic couplings such as often arise in mains-operated multimeters.

The noise voltage is tapped through a high-pass at 300 Hz, and is fed into an oscilloscope with a high-sensitivity vertical amplifier (Tektronix 7633 with

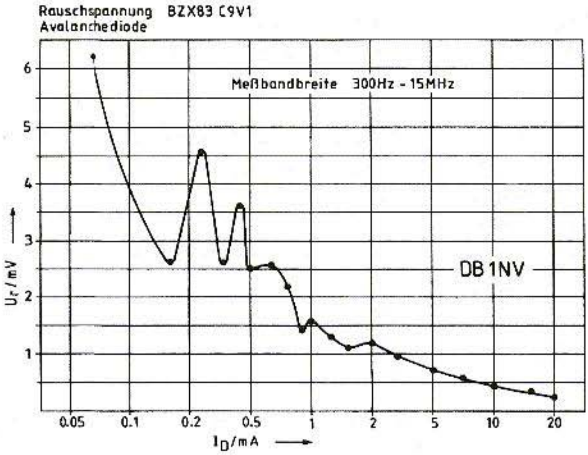



**Fig.3: Noise Voltage of Avalanche Diode BXZ83C9V**

***Rauschspannung*** **= Noise voltage,** ***Meßbandbreite*** **= Measuring band width**

7A15 vertical plug-in unit). The deflection factor lies at 500µV/cm.

The signal, pre-amplified by a factor of 50, is tapped at the vertical output of the oscilloscope and is fed in parallel into a broad-band effective value voltmeter (HP 3400A) and a spectrum analyser (built by the author himself). The 3dB band width of the measuring rig lies at 300 - 15 MHz, and the internal amplifier noise relating to the test object at 20µVeff.

There is an additional 50Ω load resistor in the screening housing which can be connected up, which makes it possible to specify the internal resistance of the "Zener diode" noise voltage source.

Even the first experiments showed that the internal noise resistance of the Zener diodes was only a few Ohms. This confirmed a fact well-known from experience, that the direct parallel wiring up of a capacitor alters the noise of a Zener diode only a little. The source impedance is just too low!

## 4. THE NOISE BEHAVIOUR

The group of "real" Zener diodes (BZX83C4V3, BZX83C5V6, BZX79C5V6, ZPD4.7, ZF5.6) with Zener voltages of 5.6V and below, all displayed noise voltages which increased with the diode current. For some specimens, saturation occurred at currents exceeding 10mA, or the noise even decreased again.

The readings for the ancient specimen ZF5.6 were no different from those for newer types. For current levels varying between 100µA and 1mA, the noise voltages were below 20µV, i.e. within the internal noise limits of the measuring rig. For current levels of between 5 and 20mA, the maximum noise voltage which could be measured was app. 100µV.. The spectrum distribution was approximately uniform (white noise). A shallow noise hump could frequently be observed at 5 to 10 MHz, which migrated to higher frequencies as the diode current increased.

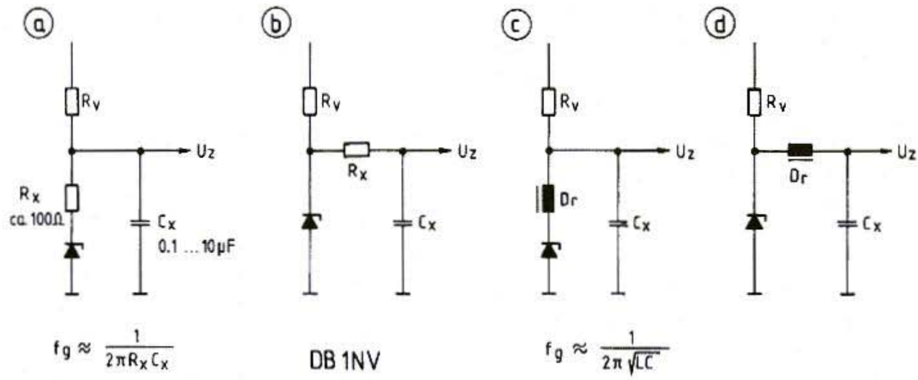



**Fig.4: Filter Circuits for Zener Diodes**

For the avalanche diodes (BZX83C7V5, BZX83C9V1, BZX83C27, ZPD7.5, ZPD10), the textbook cycle for noise voltage plotted against the diode current ensued. At low current levels of 100µA, noise voltages could be measured of 2 - 4mV, which fell back to values of between 150µV and 1mV for a current of 10mA. But the decrease is not uniform. Instead, it displays numerous minimum and maximum values in the current range below 2mA.

The levels and positions of the maximum values are subject to strong scattering effects which vary from one specimen to another. Perhaps a semiconductor specialist among our readers could find an explanation for this.

The behaviour of two transient protective diodes, type 1.5KE6.8, was particularly remarkable. Here, the noise generation set in abruptly at current levels of between 1.2 and 2.4mA!

Figs.2 and 3 show examples of the noise voltage, plotted against the current, for a "real" Zener diode (BZX83C4V3) and an avalanche diode (BZX83C9V1).

Sample LED's of various colours from various manufacturers displayed no measurable noise when measured!

## 5. ADVICE PROPOSALS FOR CIRCUIT DEVELOPERS

When using Zener diodes in low-noise voltage control circuits, or for coupling, operation point setting and potential displacement in amplifiers, the basic rule is as follows - Zener diodes are a low-Ohmic source of noise voltages (internal resistance < 50Ω), with a spectrum which extends up into the megaHertz range.

The noise is not specified in the manufacturers' data sheets, and is subject to strong scattering effects which vary from one specimen to another, with associated chaotic current dependencies.

The following rules can be derived from the measurement results:

1. If the Zener diode can be operated at high levels of current (exceeding 10mA), then the differences between Zener and avalanche diodes are blurred, although the "real" Zener diodes have a tendency to generate less noise.

2. At low current levels below 1mA, the noise from avalanche diodes increases rapidly. Minimum and maximum values arise, which can lead to the curious case in which increasing the current may even increase the noise.

3. Real Zener diodes (U ≤ 5.6V) generate only a little noise, even at low current levels below 1mA, but they have a higher internal resistance, leading to poorer stabilisation properties. They are thus better suited to low-noise voltage controllers, for battery-operated equipment, or for coupling and operation point setting in amplifiers.

   From the point of view of noise behaviour, it can be worth while to connect up several low-voltage Zener diodes in series, instead of using one avalanche diode. The noise from one Zener diode for operation point setting can be amplified directly in a broad-band amplifier, or can modulate the output frequency in an oscillator.

4. Extremely low voltages below app. 3V can more expediently be stabilised using light-emitting diodes operated in the conducting direction. The temperature coefficient is considerably lower than for silicon diodes in a series circuit.

The following table compares typical temperature coefficients.

| Diode type | Temperature coeff. in mV/°C |
|---|---|
| 1N4148 | -2mV/°C or -0.28%/°C |
| IR LED | -1.6mV/°C or -0.1%/°C |
| Red LED | -1.4mV/°C or -0.08%/°C |
| 3V Zener | -2.7mV/°C or -0.09%/°C |
| 5.6V Zener | ±0mV/°C |
| 10V Zener | +6mV/°C or +0.06%/°C |
| 20V Zener | +16mV/°C or +0.08%/°C |

There are "Zener diodes" for low voltages, such as the BZV86 range from Philips, which consist of just 2 or 4 silicon diodes wired up in series, and which are operated in the conducting direction. Their temperature cycle corresponds to that of normal silicon diodes!

5. Since the noise resistance of the "Zener diode" noise sources lies at only a few Ohms, connecting a capacitor up in parallel to reduce the noise is rather pointless. Large capacitors would be required, and moreover the effective series resistance of the capacitor is too high to short-circuit the noise voltage.

   Assistance is provided by an LC or RC filter between the Zener diode and the consumer, which naturally increases the static or dynamic internal resistance of the stabilisation circuit. Fig.4 shows some examples.

## 6. SUMMARY

Zener diodes with a breakthrough voltage exceeding 6.8V are not suitable for generating low-noise operating voltages unless careful after-filtration is provided for and the diode can be operated at high levels of current (app. 10mA).

Zener diodes with a breakthrough voltage below 6.8V (Zener diodes in the narrower sense) generate at least 20dB less noise output, and can thus be operated even at low levels of Zener current (below 100µA), since the noise output generated is largely independent of the current. This is of particular interest in relation to battery-operated equipment.

The noise output is subject to strong scattering effects which vary from one specimen to another, and displays chaotic oscillations in the operating current, particularly with avalanche diodes (> 6.8V).

Since the noise spectrum is approximately "white" up to at least 50 MHz, these effects can be used to construct simple noise generators - e.g. for the well-known aerial noise bridges.

One alternative remains to diodes operated in the conducting direction for the stabilisation of very low voltages - the use of LED's. They are less temperature-dependent, and they generate no measurable noise.

*Andreas Schaumburg, DF7ZW*

# Elimination of Self-Oscillation Points in the DB1NV Spectrum Analyser

**The author has been using a DB1NV DIY spectrum analyser for about 5 years, and it has proved its worth as a universal measuring apparatus.**

**Secondary reception centres (self-oscillation points) have proved to be a particular source of interference, generating signals on the screen where there are normally no signals at all.**

**The origins of these undesired composite products are investigated in more detail below, and measures are described to remedy the situation.**

## 1. ORIGINS OF COMPOSITE PRODUCTS

The basis for any improvements in the tried and tested concept of the spectrum analyser (SA) is a precise knowledge of how self-oscillation points arise.

In his description [1] of the high-frequency/intermediate-frequency assembly, the author went into the problem of secondary reception centres, which could more relevantly be described as self-oscillation points, in some detail (on p. 50).

I tried all the measures suggested there on my own SA, but without any convincing success. The structure of the high-frequency / intermediate-frequency assembly did not exactly correspond to the article, since original helix filters are difficult to obtain. The Neosid triple helix filters used in the first intermediate frequency (type 7.3E/C, article no. 00514990) gave a first intermediate frequency at 511 MHz, with the first local oscillator oscillating at 500 MHz. By means of an additional block circuit, the image frequency could be reduced by more than 70dB, even with these two filters. To process the measurements from the oscillation points, the first mixer (SRA220) on the printed circuit board was shut down, and the first intermediate frequency was fed through an SMA jack in the assembly at 511 MHz.

A test transmitter could now be connected up and its frequency varied over a wide range. To obtain a reference line at 511 MHz, a level of approximately -30dBm was needed

The test transmitter's output level was then raised to 0dBm, and its frequency was again varied, into the GHz range. Secondary reception centres could be discovered above about 2.5 GHz, where the condition: N*500 MHz ± 10.7 MHz is fulfilled. At approximately 3.5 GHz, there was a signal app. 40 dB above the noise at the above input level.

The second mixer (IE500) mixed not only 510.7 MHz but also various harmonics with an oscillator frequency of 500 MHz at a second intermediate frequency of 10.7 MHz, in accordance with the above formula.

Normally the two helix filters would prevent these high frequencies from reaching the second mixer at all. However, spurious resonances and undesired couplings are unfortunately responsible for this cross-talk.

If the first LO now oscillates, for example, at 875 MHz, then the fourth harmonic, which is generated in the first mixer, descends onto the secondary reception centre at 3.5 GHz and generates a oscillation point.

The author has operated a normal 500 MHz mixer with an LO frequency of 500 MHz in the laboratory and measured the conversion loss for harmonic mixing at up to 10 GHz.

100 to 200 MHz was set as the initial intermediate frequency. At 3.5 GHz, the conversion loss was app. 30dB, and at 8.5 GHz about 50dB. These measurements show that a normal 500-MHz mixer is still thoroughly capable of mixing and generating self-oscillation points, even at far higher frequencies, although with higher conversion loss.

## 2. ELIMINATION OF OSCILLATION POINTS

The first mixer was taken out of its existing tinplate housing and connected to SMA plugs by means of dual screened coaxial cable. A specimen from ANZAC was used as a mixer, in a metal housing, with SMA jacks (or Mini Circuit ZFM-2; ZFM-150). This had only a measure of success, as the harmonics from the first LO were still reaching the intermediate-frequency module.

A glance at the manuals for commercially available equipment was of assistance here. For example, in a 20 year old spectrum analyser design (HP 8555), between the first and second mixers, HP had inserted, in addition to the resonant cavity filter at 2.05 GHz, a coaxial low-pass filter with a limiting frequency at 2.5 GHz. This suppressed the inevitable spurious resonances of the resonant cavity equipment in a reliable fashion, so that self-oscillation points no longer occurred.

Such a commercial coaxial low-pass, with a limiting frequency of 1,300 MHz,

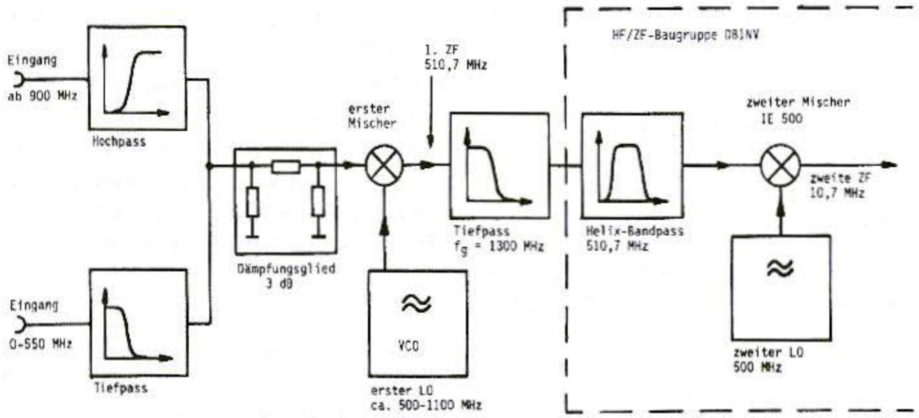



**Fig.1: Block Wiring Diagram of Modified Spectrum Analyser**
***Eingang*** **= Input,** ***Ab*** **= From,** ***Hochpass*** **= High pass,** ***Tiefpass*** **= Low pass,** ***Dämpfungsglied*** **= Attenuator,** ***Erster Mischer*** **= First mixer,** ***Erster*** **= First,** ***Ca.*** **= Approximately,** ***ZF*** **= Intermediate frequency,** ***HF/ZF Baugruppe*** **= High-frequency/intermediate-frequency assembly,** ***Zweiter Mischer*** **= Second mixer,** ***Zweite ZF*** **= Second intermediate frequency,** ***Zweiter*** **= Second**

could be found in a surplus store, and was also incorporated into the test apparatus. A low-pass with a limiting frequency of 600 MHz would have been better, but one was not available.

The selection requirements are not very stringent. An attenuation level of 25 to 30dB is sufficient, since without a low pass the oscillation points lie at max. app. 15dB above the noise. But no serious attenuation breakdowns should occur below 10 GHz, which makes it difficult simply to build the low pass for yourself. This measure made it possible to get to grips with the problem of oscillation points. As long as only one mixer is housed in each tinplate housing, the screening of the tinplate housing is completely adequate.

The next problem was the background noise of the analyser, which was still not uniform over the entire frequency range.

Due to adaptation problems, the conversion loss of the mixer oscillates, which can be traced back to the precision of level. A mixer should see app. 50 Ohms at least one port, which could be obtained at the mixer input with an attenuator of app. 3dB. True, this reduced the sensitivity of the analyser, but created an almost uniform background noise without any ripple effects over the entire frequency range.

On its own, without an attenuator, the combination of high pass and low pass at the input of the analyser can not come close to providing a constant 50 Ohm adapted load for the mixer over the entire frequency range. The precision of level was measured as being ± 2 dB, at a constant input level of -30dBm.

Fig.1 shows the outline wiring diagram of the improved spectrum analyser.

To monitor the dynamic range of the altered analyser design, a two-tone signal with the frequencies 145 MHz and 155 MHz was fed in at the input. The level of 2* - 22dBm generated intermodulation products of - 88dBm, which had a distance of ± 10 MHz from the two carriers, and lay just above the background noise. The inter-modulation distance which corresponded to the dynamic range of the analyser was thus 66dB, a typical value for mixers in the low-level class.

Thus my DIY apparatus was again improved quite a bit.

## 3.
## THE INPUT RING-MIXER

Putting a suitable ring mixer into the housing with SMA jacks proved to be more of a problem than initially expected. Unfortunately, the mixers referred to above, from the firm of Mini-Circuits, were not easy to find.

In a search for alternatives, I obtained several mixers of up to 26 GHz and tested their serviceability in expanding the frequency for the spectrum analyser.

The low-cost M21L mixer proved to be usable up to LO/RF = 3 GHz. The measurements were taken using the built-in version, with short coaxial cables soldered to its connections. The conversion loss produced a ripple effect of app. 10dB ±2dB. The ripple effect increased strongly above 3 GHz, so that the mixer could be used in an analyser design only up to app. 1.5 GHz.

The frequency cycle of the intermediate-frequency connection is in the uncritical area. An intermediate frequency up to 1.7 GHz would have been feasible, although the intermediate-frequency connection of the mixer was specified for use only from DC up to 1 GHz. 0.2 MHz was measured as the lower limiting frequency at the high-frequency connection.

On most ring mixers, the intermediate-frequency and high-frequency connections can be swapped over without any need to allow for any marked increase in conversion loss. With a spectrum analyser, it is actually useful to connect the input to the intermediate-frequency connection of the mixer, since in this way you can measure down to levels which are almost all the way down to DC voltage.

The first intermediate frequency was still above the highest input frequency for an SA with carrier wave mixing. This trick is used even by prestigious manufacturers. The intermediate frequency goes directly to the diode quartet of the ring mixer, with the high-frequency and LO connections being connected up through a ferrite repeater. The leakage inductance of the repeater restricts the maximum operating frequency. The minimum frequency of the corresponding port is determined by saturation of the ferrite core. The directly coupled intermediate-frequency connection thus always has a broader band.

The MDC123 mixer from MA-COM is suitable for an analyser design going up to 2 GHz. Although the high-frequency / intermediate-frequency and LO connections are all specified as from 10 MHz to 3 GHz, the LO can go up to 4.5 GHz without the conversion loss exceeding 10dB ±2dB ripple. The manufacturer gives the conversion loss as 8dB in the data sheet. The minimum input frequency at the high-frequency connection is 0.6 MHz. With signals of LO = 8 GHz and HF = 4.5 GHz, the conversion loss is only 16 dB, although the mixer is then being operated far outside the manufacturer's specifications. The MDC123 is supplied in a metal housing with SMA jacks, which provides for interfaces of exactly 50 Ohms in use.

A mixer from the firm of Marki-Microwave, specified as going up to 26 GHz, was measured last. Due to a lack of suitable test transmitters for this high frequency range, I had to improvise with Gunn oscillators and frequency synthesisers.

This meant that the manufacturer's specifications could be replicated only approximately.

The mixer can be used to expand a UHF spectrum analyser, for example, as described by DB1NV, up into the 24-GHz amateur band.

An X-band signal source (8.2 to 12.4 GHz) or a suitable frequency synthesiser can be used as an LO through sub-harmonics mixing. Only the odd harmonics can be considered as LO's for a symmetrical ring mixer. The signal from a 24 GHz Gunn oscillator, with a level of 0dBm, was therefore fed to the high-frequency connection of the mixer. The signal at the intermediate-frequency connection was 2 GHz.

With an LO frequency of 8.666 GHz and a level of +13dBm, the conversion loss was 19dB - i.e. a mixing with the third harmonic.

With an LO frequency of 4.4 GHz - i.e. a mixing with the fifth harmonic, the conversion loss was 26dB. The conversion loss remained constant if the LO frequency and the intermediate frequency were changed, so that the values measured remained constant over a wide frequency range, as long as the mixing involved the same harmonic. So a ready-synthesised frequency can be used for measurements in the 24 GHz range, and the intermediate frequency can be altered accordingly.

For mixing where the carrier wave high frequency is 10.4 GHz and the intermediate frequency is 2 GHz, the conversion loss with a reading of LO = 8.4 GHz is app. 8dB, which corresponds to the manufacturer's specifications. The high-frequency and LO connections could be used down to about 1.5 GHz. The amateur radio enthusiast thus has the possibility of expanding a simple UHF spectrum analyser right up to 24 GHz, using a frequency synthesiser or another signal source.

## 4. LITERATURE

[1] A Spectrum Analyser for Amateurs, Part 1
Dr.Ing. Jochen Jirmann, DB1NV
VHF Communications, 3/87, pp. 154-166

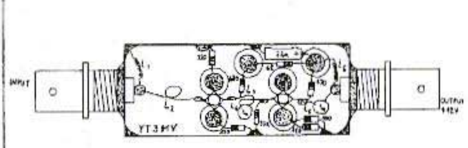
A Spectrum Analyser for Amateurs, Part-2
Dr.Ing. Jochen Jirmann, DB1NV
VHF Communications, no. 4/87, pp.232-242

A Spectrum Analyser for Amateurs, Part-3
Assembly Instructions with Printed Circuit Boards
Dr.Ing. Jochen Jirmann, DB1NV
VHF Communications, 2/89, pp.108-119

## 5. APPENDIX

Measuring equipment used: test transmitters HP8614; HP8614; HP8640 and R&S SMG: spectrum analysers HP141T + HP8552 + HP8555, wattmeter HP432





*Richard A. Formato, Ph.D., K1POO*

# A Genetically Designed Yagi

**Genetic algorithms (GA) are a class of optimisation techniques that mimic natural selection (survival of the fittest).**

**GA are applicable to many types of problems, and they are becoming increasingly useful in antenna design [1], [2].**

**This note describes a genetically designed 3-element Yagi that provides very good performance and illustrates how effective GA can be.**

Unlike deterministic optimisation schemes, GA are based on random selection. A binary-coded genetic algorithm starts by creating a population of *chromosomes* which are random bit sequences (zeros and ones). Each chromosome contains a complete antenna design (in this example, a complete 3-element Yagi). The chromosome is made up of *genes* which are strung together one after another. Each gene corresponds to one of the antennas design parameters.

| Gene # | Name | Length | Min | Max |
|---|---|---|---|---|
| 1 | REF Length | 5 | 0.4 | 0.6 |
| 2 | REF Radius | 4 | 0.0005 | 0.002 |
| 3 | DE Length | 5 | 0.4 | 0.6 |
| 4 | DE Radius | 4 | 0.0005 | 0.004 |
| 5 | DE Separation (from REF) | 5 | 0.05 | 0.3 |
| 6 | DIR Length | 5 | 0.4 | 0.6 |
| 7 | DIR Radius | 4 | 0.0005 | 0.002 |
| 8 | DIR Separation (from DE) | 5 | 0.05 | 0.3 |

**Table 1. Gene Table for 3-Element**

The Yagi gene table appears in Table 1. A design is fully specified by 8 genes: reflector (REF) length and radius, driven element (DE) length and radius, director (DIR) length and radius, and DE/DIR location along the boom. *Gene length* is its length in bits (for example, REF length is 5 bits).

The minimum and maximum values of each design parameter also appear in the table, and all dimensions are in wavelengths (waves). The DE length, for example, cannot be longer than 0.6 wave or shorter than 0.4 wave.

Since each design parameter is a decimal number, not a bit sequence, the actual value of the parameter is computed by *decoding* its binary gene using the following transformation equation:

$$X = X_{min} + \frac{X_{max} - X_{min}}{2^L - 1} \bullet D$$

where $X$ is the decimal value of the parameter, $D$ is the decimal value of the genes binary sequence, and $L$ is the genes length.

To illustrate how this decoding scheme works, consider the 37-bit chromosome that contains the design for the Yagi discussed below:

00101110000110111110010111000101
11100

The DE length is coded in gene #3, which starts at bit #10 and ends with bit #14. The binary sequence for the DE length gene is 00110, and its decimal value is

$$0(2^0)+0(2^1)+1(2^2)+1(2^3)+0(2^4) = 12$$

Since gene #3 is 5 bits long, the denominator in the transformation equation is $2^5$-1=31. The DE length is therefore 0.4+(0.6-0.4)(12)/31 = 0.477419355 wavelengths. Because the computer model used to calculate the Yagis performance inputs the DE *half*-length instead of its overall length, this value is divided by 2 and rounded to 3 places to give 0.239 wave. This decoding scheme is used to evaluate each of the Yagis design parameters. The DIR radius (gene #7), for example, evaluates to 0.0015 wave, and so on.

The genetic algorithm begins by creating an initial population of random 37-bit chromosomes. It then applies the operators of *selection*, *crossover*, and *mutation* to filter out unfit designs while retaining the better ones. Successive applications of these operators create *generations* of antenna designs, with each subsequent generation hopefully containing better designs than the previous one. But, because of the algorithms inherently random nature, there is no guaranty of obtaining better designs. They may actually become worse from one generation to another. Well-designed GA, however, usually produce progressively better designs, at least on the average, and every new run holds the intriguing possibility of producing a previously unseen best design.

The selection operator determines which chromosomes are fit enough to survive to the next generation. Some may be automatically discarded (for example, the worst 10%), while others are typically killed at random, as they would be in Nature. Others may be automatically retained (the best 5%, for example). The algorithm designer is free to implement whatever selection process seems best. The crossover operator mates two chromosomes (parents) to produce two new chromosomes (children), which become members of the next generation. Child chromosomes usually maintain a constant population from one generation to the next, although the population could grow if desired. Each parents chromosome is split at a gene boundary, usually randomly selected, and the pieces are swapped (concatenated together) to form two different chromosomes. This is the primary process by which GA propagate good genes from one generation to the next. Finally, the mutation operator randomly flips a bit here and there with some small probability. This simulates the genetic mutation that occurs randomly in Nature.

In each generation, all of the designs (chromosomes) are ranked from best to worst using a *figure-of-merit* (FoM). The FoM combines various antenna performance measures computed by a *modelling engine*, which is another computer program separate from the genetic algorithm. Individual antenna performance parameters, for example, can be calculated with any suitable antenna modelling program(s).

The FoM used for the Yagi described below is [5(G)+4(FB)-SWR]/10. This particular FoM gives slightly more weight to the main lobe gain (G) than to the front-to-back ratio (FB), and relatively less weight to the input SWR.

The algorithm designer is free to define any FoM that reflects the relative importance of different performance measures, including even non-electrical parameters (such as cost or time to build, or amount of material required, and so on). This feature is a major distinction between GA and deterministic optimisations, which frequently cannot optimise arbitrary FoMs.

Other significant differences are that GA produce *groups* of designs with similar FoMs, instead of the single best design, and they usually require much less computer time than deterministic algorithms.

The genetically optimised 3-element Yagi has the following dimensions (in wavelengths at the design frequency $F_o$):

| | |
|---|---|
| Reflector Length: | 0.530 |
| Reflector Radius: | 0.0008 |
| Driven Element Length: | 0.478 |
| Driven Element Radius: | 0.004 |
| DE Distance from REF: | 0.123 |
| Director Length: | 0.446 |
| Director Radius: | 0.0015 |
| DIR Distance from DE: | 0.106 |

The boom length (sum of DE/DIR separations) is only 0.229, less than a quarter-wave, which is quite short. At 51 MHz, for example, this Yagi is only 53 inches long. The REF, DE and DIR lengths are 122.66, 110.62, and 103.22 inches, respectively, with diameters of 0.37, 1.85, and 0.694 inches. DE is located 28.47 inches from REF, and DIR is located 24.53 inches from DE. It is quite interesting that the genetic algorithm converged to the maximum allowable value for the DE radius, because it is known from analytical considerations that increasing DE diameter can improve Yagi performance substantially [3].

The free-space main lobe gain, front-to-back ratio, input impedance (resistance and reactance), and SWR (relative to 50Ω) are plotted in Figs.1-4, respectively. These parameters were computed over a 10% band centred at the design frequency $F_o$. The azimuth and elevation patterns at Fo appear in Fig's.5 and 6, and the front-to-rear ratio (FR) is plotted in Fig.7.

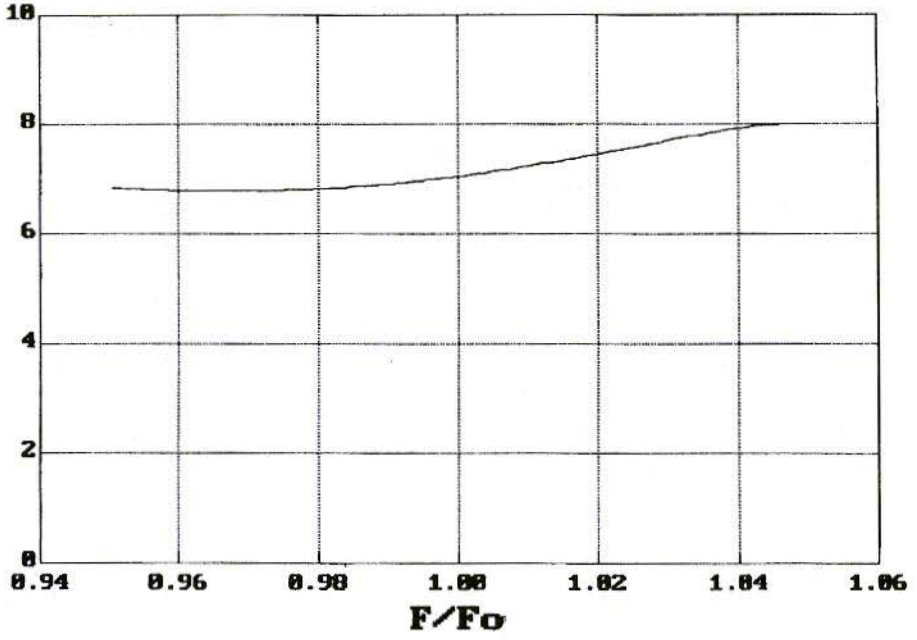



Fig.1: Main Lobe Gain

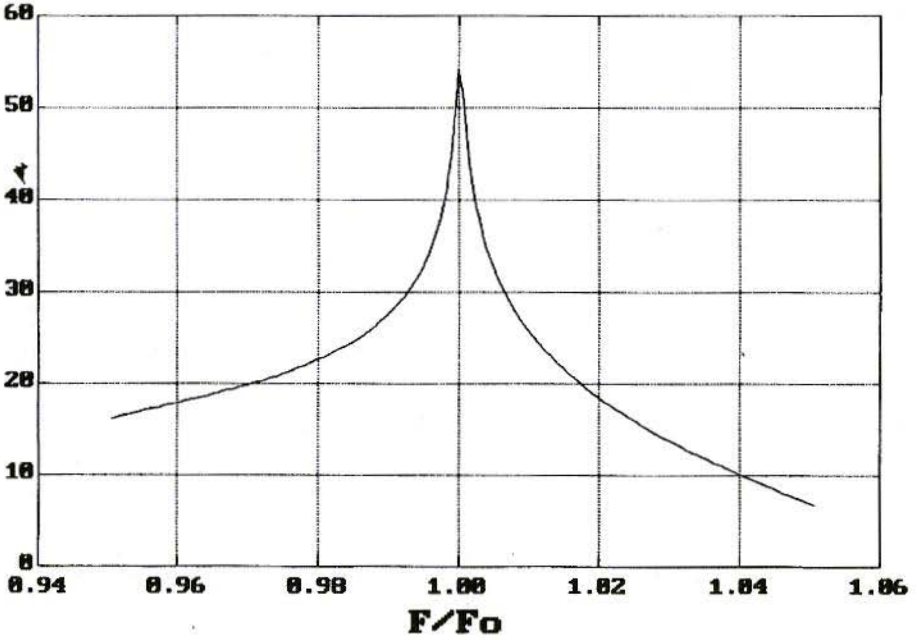



Fig.2: Front-to-Back Ratio

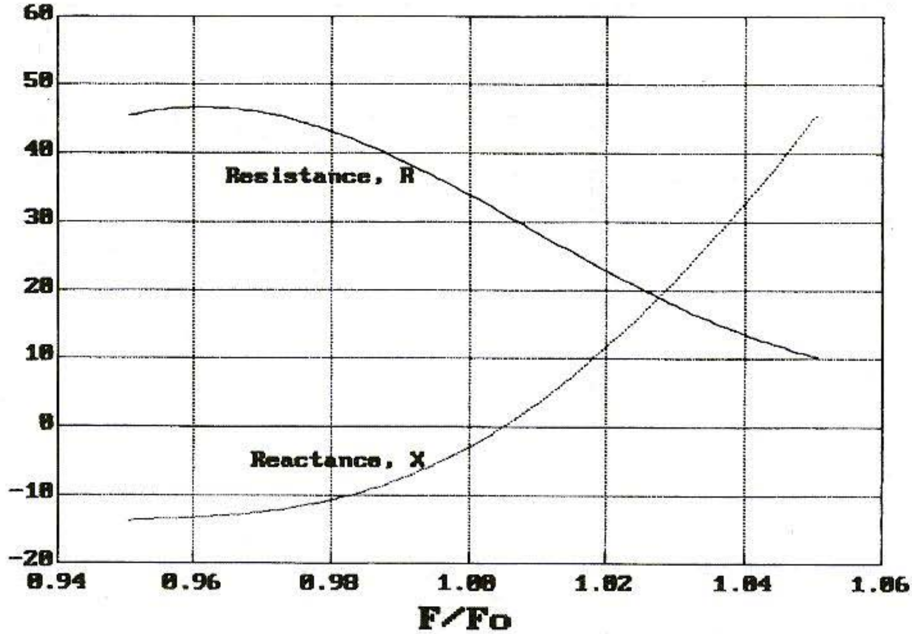



Fig.3: Input Impedance

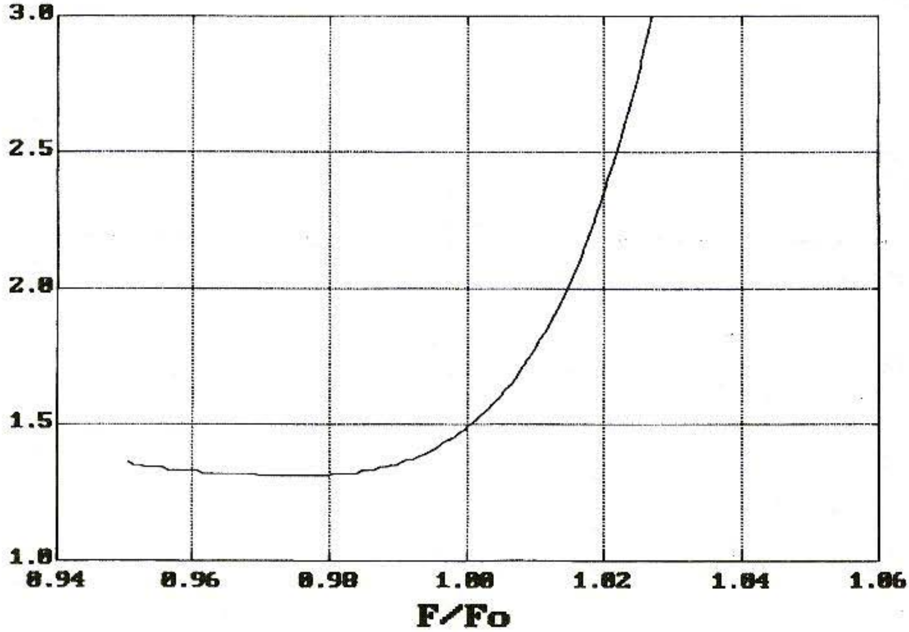



Fig.4: SWR

Key performance measures are:

| | | |
|---|---|---|
| Gain | | 7 dBi |
| FB | | 54.2 dB |
| Zin | | 33.9 - j3Ω |
| SWR//50Ω | | 1.49 |
| HPBW | az | 66° |
| | el | 122° |

The band-centre gain of 7dBi is typical of well-designed 3-element Yagis, and the optimised antennas FB of 54dB is exceptionally good. For comparison, this FB figure is more than 16dB better than the best FBs of typical quarter-wave designs described in W2PVs treatise on Yagi antennas [4] (see especially Fig.2.9). The optimised antenna also exhibits good FB bandwidth, with values exceeding 20dB from 0. 97$F_o$ to 1.017 $F_o$ (4.7%).

The optimised Yagi is nearly resonant at Fo (input reactance of 3 capacitive, which is less than 10% of the input resistance). The SWR is less than 2 from 0.95$F_o$ to 1.015$F_o$ (6.5%). If desired, this antenna can be fed directly with 50Ω coax, eliminating the insertion loss introduced by a matching network or antenna tuner. Of course, a balun should be used to maintain feed system balance (it would be interesting to build this antenna with and without a balun to see how much difference it makes).

For the 51 MHz design, the SWR is below 2, and the FB is greater than 20dB, from 49.47 to 51.76 MHz, a bandwidth of 4.5%. The lower band edge can be shifted up to 50 MHz by increasing the design frequency to $F_o$=51.55 MHz and recalculating the dimensions. Note that the wavelength is computed as 299.7956/F MHz, which is more accurate than the commonly used formula 300/F MHz.

The optimised Yagis azimuth pattern (E-plane) has a characteristic 2-lobe structure with a deep broadside null. The half-power (-3dB) beamwidth (HPBW) is 66°. The rear lobe is about 22dB down, which is quite low. The elevation pattern (H-plane) has a single, broad lobe with HPBW=122°. FR is the ratio of maximum main-lobe gain (E-plane) to the highest sidelobe level for the rear lobe between 90° and 270° azimuth. It is plotted against normalised frequency in Fig.7. FR is above 20dB from about 0.982$F_o$ to 1.018$F_o$, resulting in a bandwidth of 3.6%, which is quite good.

Comparing the genetic antenna to a similar, deterministically optimised 3-element Yagi [5] shows that the genetic antenna is actually better. The genetic Yagi is smaller, has more than 1% additional FR bandwidth, and requires no matching network (simpler, less expensive, easier to maintain, less likely to fail). Both antennas have nearly the same theoretical forward gain, with the genetic Yagis being only slightly lower. In the real world, however, the genetic Yagi probably has *higher* forward gain, because the unspecified matching network in the deterministic design introduces losses that are not included in its theoretical forward gain figure.

The genetically optimised, 3-element Yagi is a very compact antenna that provides excellent performance. This example illustrates that GA can produce very good antennas indeed. Genetic algorithms are easily implemented on a PC and can provide significant advan-

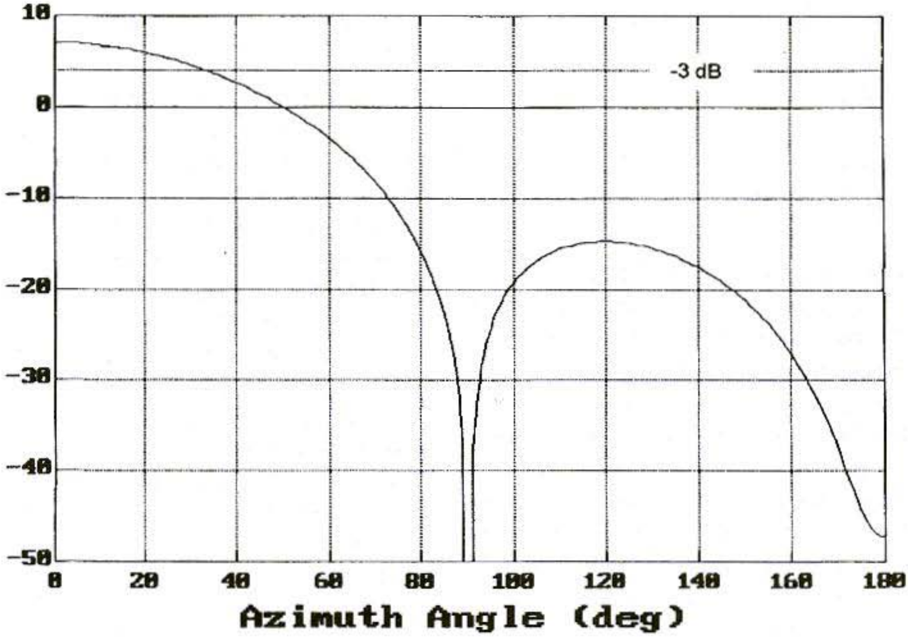



**Fig.5: Azimuth Pattern**

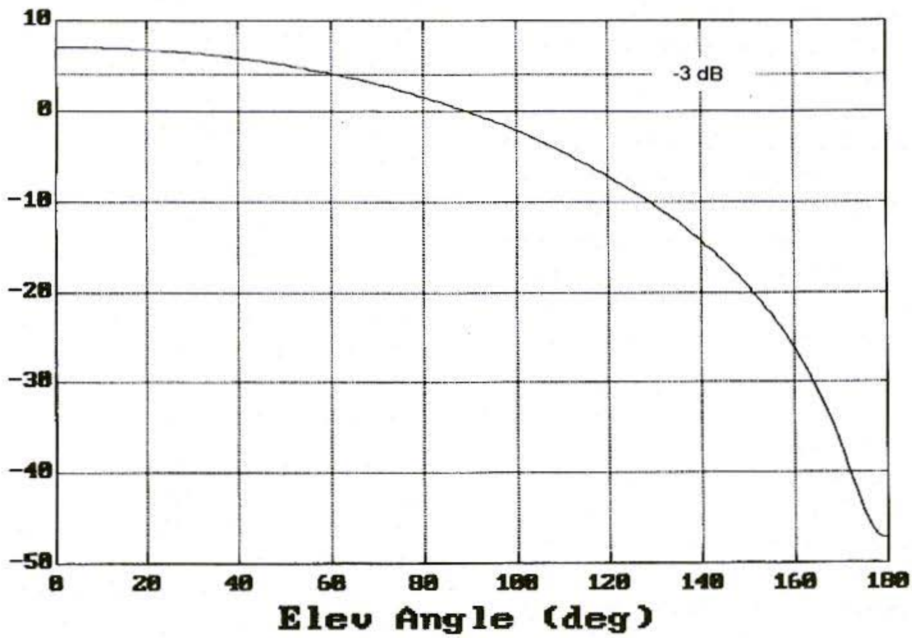



**Fig.6: Elevation Pattern**

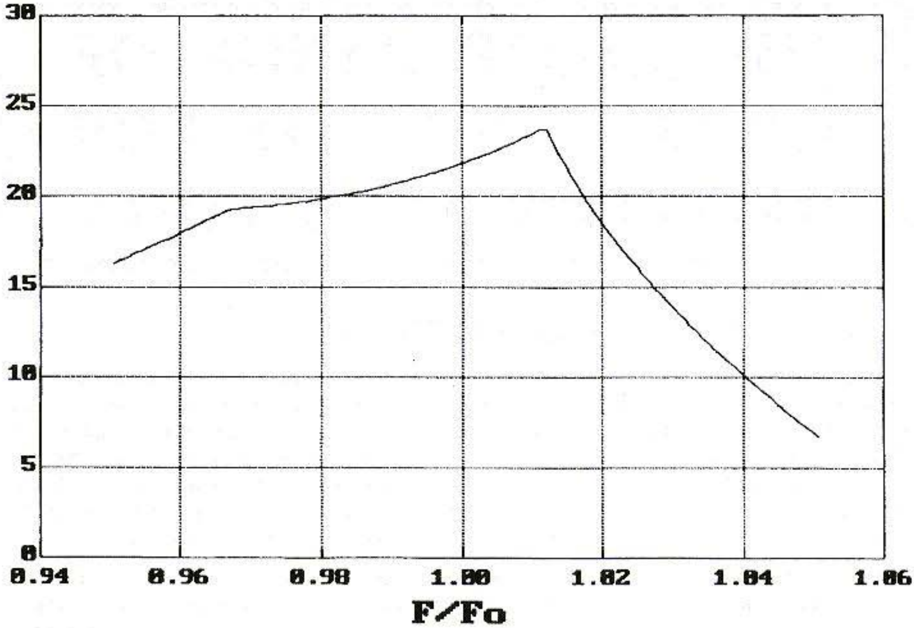



**Fig.7: Front-to-Rear Ratio**

tages over deterministic techniques. It is likely that we amateurs will hear more and more about the genetic design approach, and it certainly merits serious consideration by hams who are interested in antennas.

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[4] James L. Lawson, W2PV, Yagi Antenna Design, ARRL, Newington, CT, 1986

[5] The ARRL Antenna Book, 17th edition, ARRL, Newington, CT, 1994 (see design #320-16, Fig.10, page 11-8).

## 3cm LNB's

Low Noise Block receive converters have been the main contributive factor that has opened up the 3cm band to ATV and other modes. With the low receive noise figures now available transmissions are no longer restricted to clear line-of-sight paths. Test transmissions have conclusively shown that over the horizon transmissions by various propagation modes are now possible. None line-of-sight transmission by means of scatter from rain clouds has also been achieved. Equipped with suitable antenna systems etc. operation well in excess of 100Km is readily achievable. Circular input types will fit directly onto standard offset satellite dishes. No feed horn is required as it is integral with the LNB. Types available as of April 96. Please check on availability as some types can be in short supply.

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