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A Publication
for the Radio-Amateur
Especially Covering VHF,
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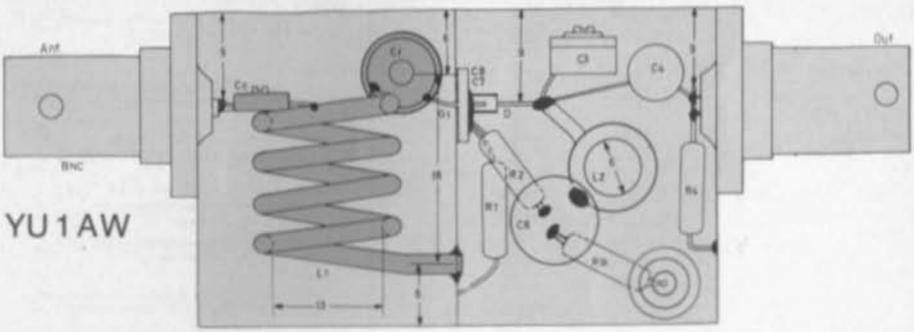
VHF communications

Volume No. 19 · Winter · 4/1987 · DM 7.00



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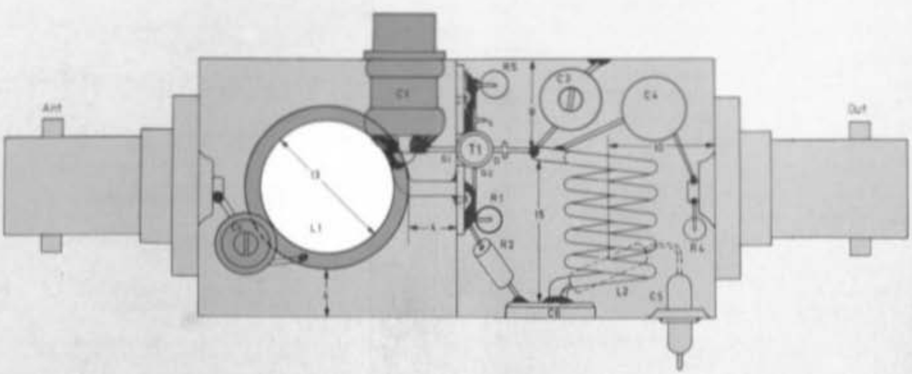

Low Input Losses Yield Very Low Noise Figures

VHF communications

A Publication for the Radio Amateur
Especially Covering VHF, UHF, and Microwaves

Volume No. 19 · Winter · Edition 4/1987

Published by: TERRY BITTAN oHG, P.O.Box 80, Jahnstraße 14, D-8523 BAIERSDORF, Fed. Rep. of Germany, Telephone (9133) 47-0 Telex 629 887, Telefax 0 91 33-47 18, Postgiro Nbg. 30455-858

Publishers: TERRY BITTAN oHG

Editors: Corrie Bittan, Colin J. Brock (Assistant)

Translator: Colin J. Brock, G 3 ISB / DJ Ø OK

Advertising manager: Corrie Bittan

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 published in Spring, Summer, Autumn and Winter. The 1987 subscription price is DM 25.00 or national equivalent per year. Individual copies are available at DM 7.00 or equivalent each. Subscriptions, orders of individual copies, purchase of PC-boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative, or – if not possible – directly to the publishers.

Printed in the Fed. Rep. of Germany by R. Reichenbach KG Krelingstr. 39 · 8500 Nuernberg.

We would be grateful if you would address your orders and queries to your representative.

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Verlag UKW-BERICHTE, Terry D. Bittan
POB 80, D-8523 Baiersdorf / W. Germany
Creditanstalt Bankverein, WIEN Kto. 17-90.599
PSchKto WIEN 1.169.146

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ISSN 0177-7505

Contents

This edition completes the nineteenth year of VHF COMMUNICATIONS and we hope that you had many informative hours in reading the magazine during the past year. We also hope, due to your appreciative assistance, to be able to increase the circulation of VHF COMMUNICATIONS to a more satisfactory level in order to ensure the continuation of the magazine. Please renew your subscription for 1988 **now** to be sure to get your next copy in time.

The staff of VHF COMMUNICATIONS wishes all readers a **happy and prosperous New Year.**

The editor

UKW**berichte** Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133 47-0. For Representatives see cover page 2

Roland Barchet, DK 2 LT

Converting the TELECAR TS 160 into a 2m, 80-Channel Amateur FM Transceiver

The Telefunken FM transceiver TELECAR TS 160 is now a little out-dated and in the process of being replaced by more modern equipment. It has therefore become possible for the radio amateur to obtain one of these solidly built transceivers and with a small total outlay, convert it into service as the main station equipment or for mobile use. Besides moving its band of operation from 160 MHz down to 144 MHz a synthesizer is installed, thus converting the TELECAR into a modern FM transceiver for amateur purposes.

For this purpose, the 80-channel handheld, described in VHF COMMUNICATIONS 1/1986 by DL 5 NP, has a very suitable modern synthesizer which has been borrrowed for this project. Günter Prokoph, DL 5 NP had forseen the possibility of the synthesizer being used for other projects and established a dividing line on the PCB DL 5 NP 001. The numerical frequenccy indicator shows 00 = 144.00 MHz up to 79 = 145.975. An offset of 600 kHz is available for relay operation and the IF is 10.7 MHz. Fig. 1 shows the completed equipment.

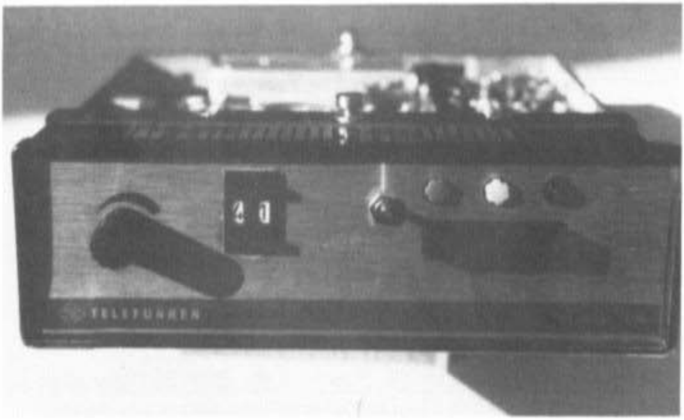


Fig. 1:
Front view of the modified equipment

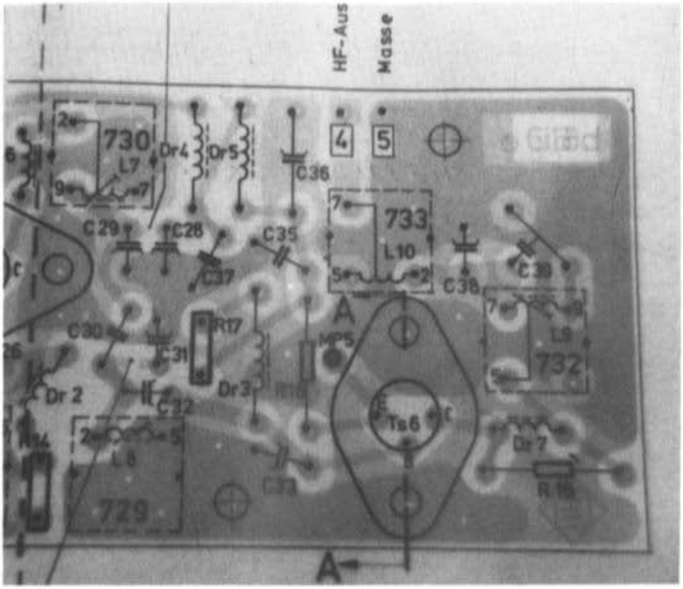


Fig. 2: Driver component layout. The panel is sawn in two at the dotted line

1. REMOVING THE UNWANTED COMPONENTS

First of all the two housing covers are removed and the front panel gain control and channel selection knobs removed. Now take off the escutcheon panel. The switches 'Rauschsperre' (mute) and 'Lautstaerke' are allowed to remain untouched.

Following that, all the modules are removed so that only the basic chassis remains. The two screws securing the frame to the back wall of the housing are loosened, as well as the screws along the chassis. It is pointed out that one of these screws is located underneath the largest electrolytic capacitor. When all the screws are out and the green-, black-, red- and yellow-marked plug has been withdrawn the back wall complete with chassis may be carefully taken out.

The cable cleat on the electrolytic is now loosened and unsoldered from the inside and bent out of the way. Now the channel selector switch stops are fully exposed. They are bent up and removed and the unwanted (green) channel selector switch desoldered and removed.

Now the channel indicator lamp holder (i.e. the white lamp) is sawn through from both sides. This position is required for the BCD switch. The square cut-out in the escutcheon is filed out on both sides and bottom in order to fit a two-position miniature code switch. This is fixed into position with a two-component adhesive.

The transmitter-multiplier module printed circuit board is now taken and a saw cut made immediately behind L6 in order that L7 to L10 and Ts 6 (**fig. 2**) may be further employed as a driver. The rest of this board, as well as the sender and receiver oscillator boards, will not be required and can be disposed of. On the send side, all the pin plugs of the sender oscillator and the neighbouring pins of the sender multiplier are de-

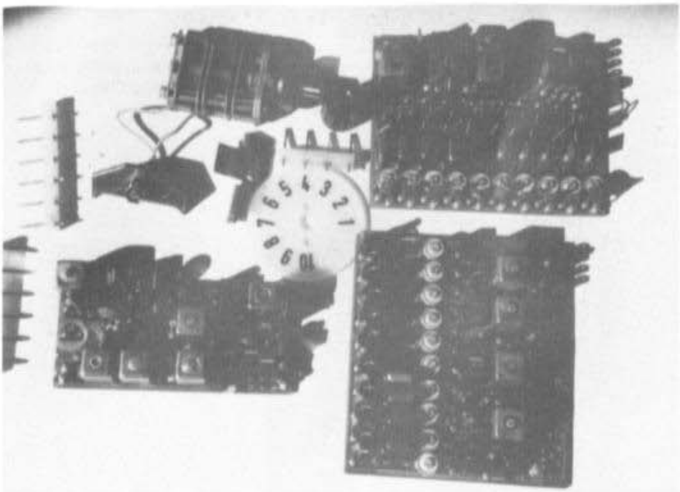


Fig. 3:
The unused parts of the TELECAR

soldered and removed. The space created is required for the synthesizer. All the unwanted parts that have been removed are shown in the photo of **fig. 3**.

The chassis can now be carefully inserted into the equipment frame and all screws fastened firmly. The electrolytic is soldered in together with the plug. With this, practically all the mechanical work has been accomplished.

2. MECHANICAL WORK

The original (unmodified) positions of the modules are shown in **fig. 4** for the receiver and in **fig. 5** for the send side. The parts no longer required are indicated on these diagrams.

An 85 x 105 mm panel is then sawn from paxolin or another similar insulation material which is then fitted into the space vacated by the former transmit oscillator. Before installing it, place the antenna filter in position in order to ensure that there will be sufficient place for it. The paxolin panel is secured to the chassis using the existing screw holes. The antenna filter is then again removed.

The reason for fitting this insulated panel is that the TELECAR TS supply is wired positive pole to chassis and the synthesizer is conventionally supplied with negative pole to chassis. This panel must therefore be installed in an insulated manner.

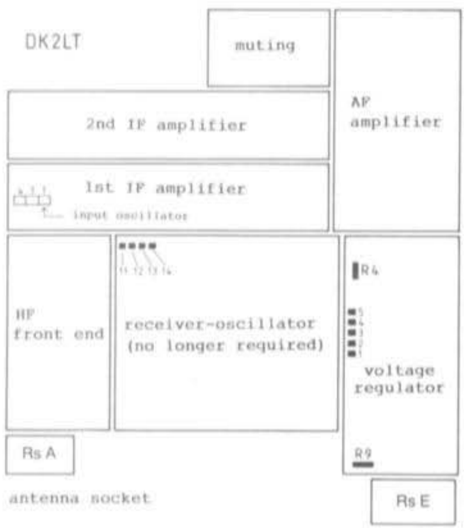


Fig. 4: Module layout of the TELECAR TS 160 receiver section

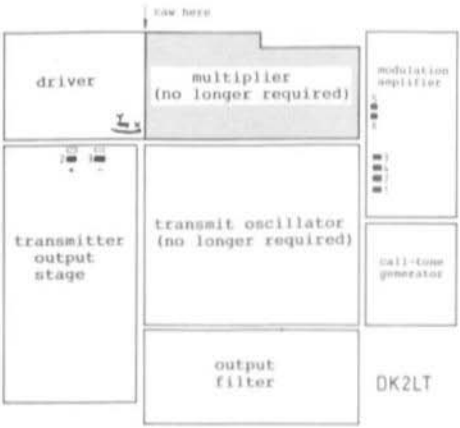


Fig. 5: Module layout of the TELECAR TS 160 transmitter section

3. CONVERTING TO 144 MHz

First the links A, C, E, F, K, and M in the HF input section are soldered in and afterwards the capacitors C2, C4, C11, C14 and C16 are removed and replaced by 10 pF, 10 pF, 15 pF and 15 pF and 15 pF respectively. **Fig. 6** shows the position of the links and capacitors mentioned above. The tuning will follow later with the adjustment of the individual inductor dust-cores.

In the receive audio amplifier, two links must be removed if they have, in fact, been wired in at all. They can easily be seen from the track side of the board. The links set the de-emphasis from 'flat' when removed to 6 dB per octave when installed.

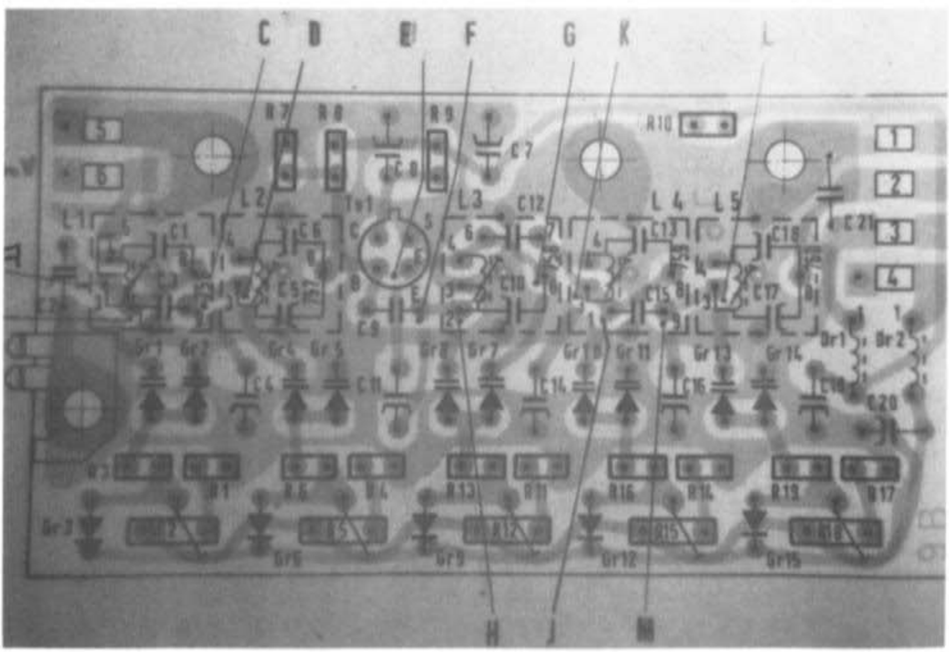


Fig. 6: Capacitors and links of the HF input module

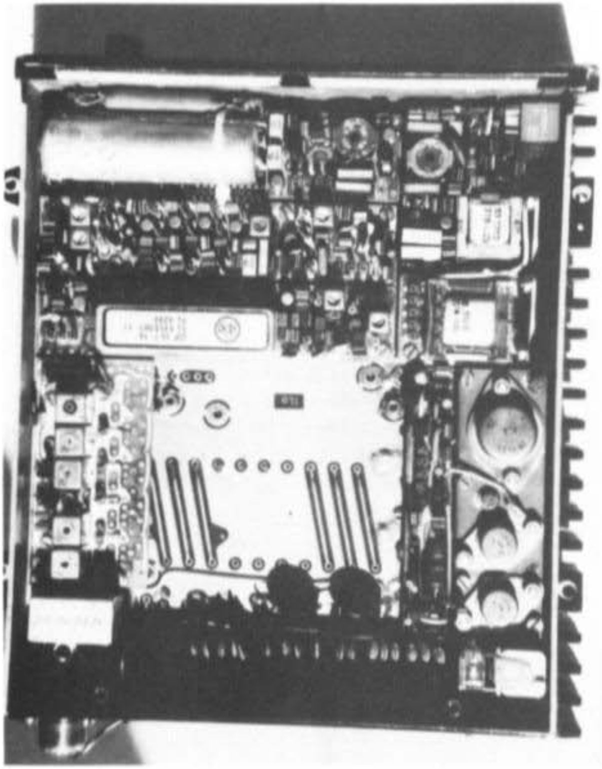


Fig. 7: Completed receive section in the TELECAR TS 160

Two links are then removed from the modulation amplifier in the same manner as for the receive audio amplifier. They can be located also on the PCB trackside. **Fig. 7** shows the receiver part of the TELECAR TS following this modification.

Now two insulated conductors with different colours are soldered to pins 2 and 3 of the transmitter output amplifier (**fig. 8**). Solder then as near to the insulation as possible. These conductors will be used in the 'send' condition to control a separate relay for the voltage supply to the driver and synthesizer output-amplifier.

By following the track of the output of the receive oscillator board (pin 12), the end is found under the modulation amplifier (pin 1 of the first IF stage). A thin coaxial cable (RG-174/U or PTFE cable) is soldered on to it – but only the inner, the screen remaining unconnected to anything at this point. The other end of the coaxial cable will be later connected to the receive oscillator in the PLL carrying a signal of 133.3 to 135.3 MHz. By these means, the first IF of 10.7 MHz is derived and which the TELECAR TS 160 will process further.

Following a visual inspection and then an actual working test to ensure that the afore-mentioned modifications are functional, these modules may be re-installed. As mentioned above, the sender and the receiver oscillator together with the sawn-off greater portion of the transmit multiplier are no longer required. The escutcheon is not yet put on.

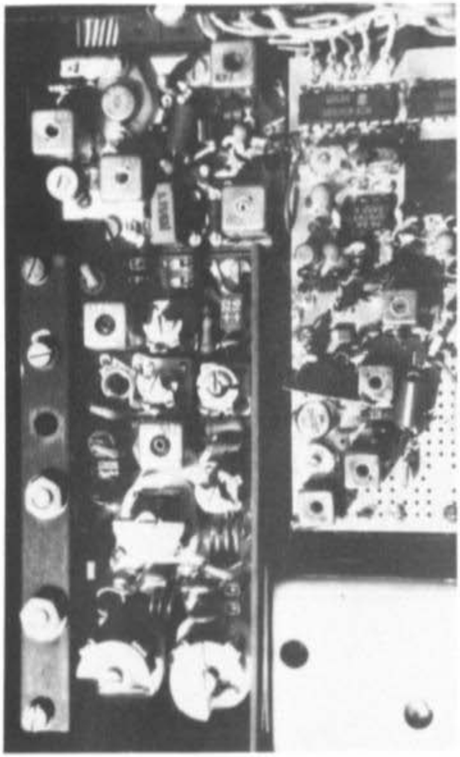


Fig. 8: TELECAR driver and output stage showing also part of DL 5 NP 001 synthesizer printed circuit board

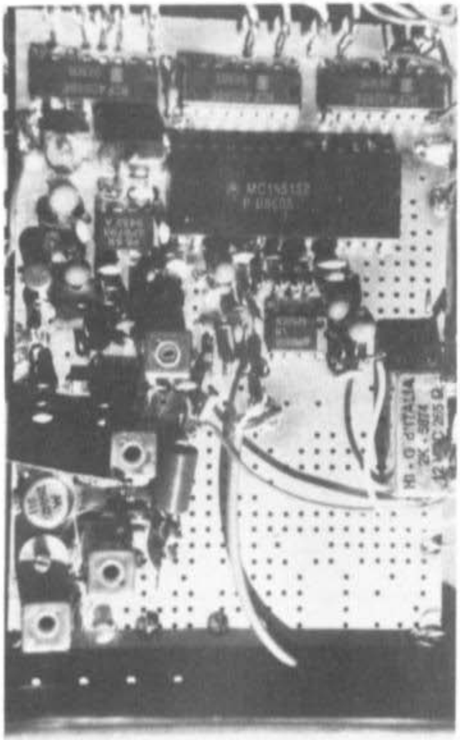


Fig. 9: Completed DL 5 NP 001 printed circuit board ready for installation in the TELECAR TS 160

4. INSTALLING THE SYNTHESIZER

The details of the DL 5 NP 001 module need not be gone into here as this information can be obtained from the VHF COMMUNICATIONS 1/86 which, together with the printed circuit board and components, can be obtained from the publishers. Only the frequency determining portion as well as the driver and output stage are required.

The transmitter is constructed from synthesizer to output stage (T9) in exactly the manner described by Guenter Prokoph. The whole of the receiver together with the modulator and PTT are not equipped. Diode D_9 also is not required. The power supplies for T8 and T9 are introduced via R 49, R 39 and RFC 2. There is a lot of space available on the printed circuit board and it is possible to utilise the (unused) receiver printed tracks. A small relay with two change-over contacts is fitted as indicated in **fig. 9**. This relay switches the supply to T8 and T9 in the 'send' condition. The relay coil is connected to the two leads from pin 2 (+) and pin 3 (−) prepared earlier on the transmitter output stage. A protection diode (1N4148) is shunted across the relay. This, together with other details may be seen in **fig. 10**.

Pin 5 of the modulation amplifier is connected directly to the ground plane of DL 5 NP 001 PCB

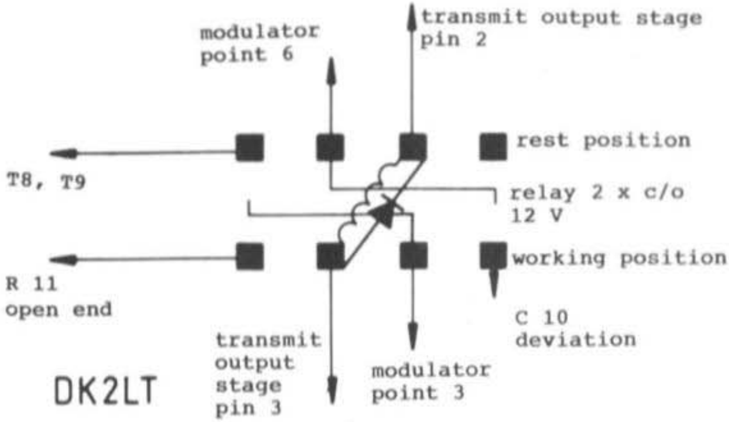


Fig. 10:
Auxiliary relay
wiring plan

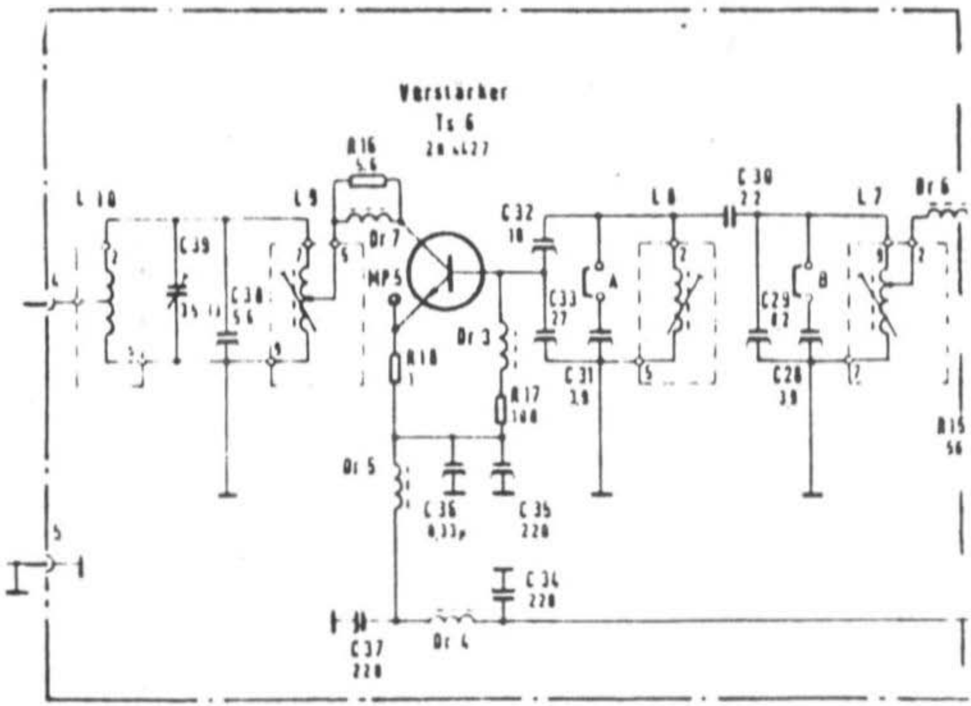


Fig. 11: TELECAR driver stage modifications

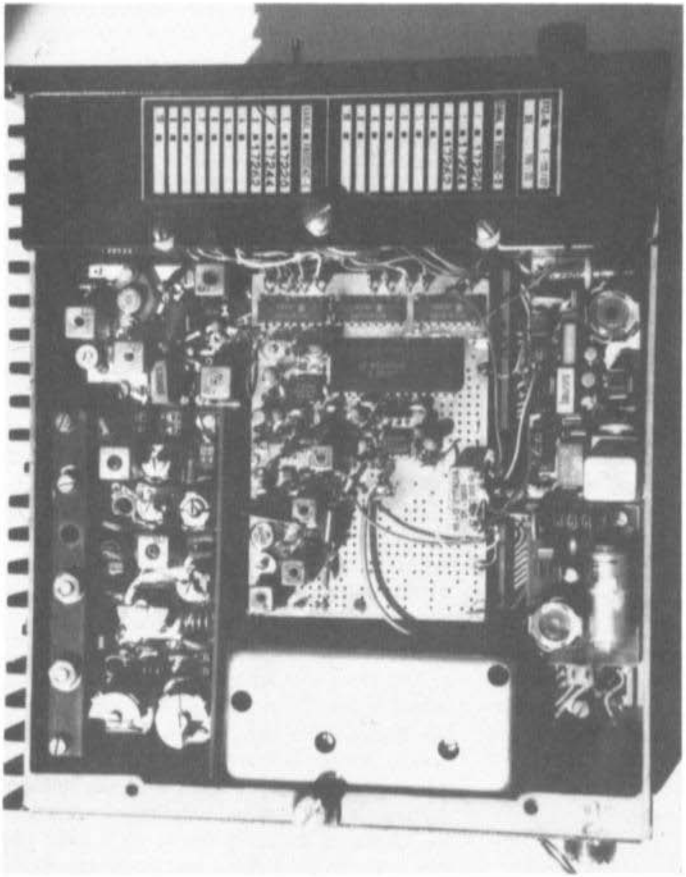


Fig. 12: Completed transmit section modifications

and pin 3 is connected to the + Ve of the board as shown on page 10, fig. 5 of VHF COMMUNICATIONS 1/1986. The connections for the integrated circuits, also shown there, must be carried out as well. The part of the driver to be utilised is connected at point X (fig. 5) with the PCB ground-plane. Here, it should be remembered, the PCB DL 5 NP 001 is mounted so that it is isolated from the main frame, on the paxolin (SRPB) board.

A thin coaxial cable (inner) is soldered to point L7/C58 of the transmit output stage (T9); the coaxial screen is soldered to ground. The other end is connected to point Y (fig. 5) of the remaining part of the TELECAR driver but the screen remains unconnected at this end. The point Y is the supply from L7 in the TELECAR driver (**fig. 11**). So far, all the connections of the 'new heart' should lie to hand. **Fig. 12** shows the equipment in this condition.

5. FINISHING THE WIRING WORK

Now, the BCD switch and the S/D switch is wired as indicated below. The escutcheon plate can now be fitted and the control knobs mounted.

Following a further check, the original TELECAR plug is inserted (it must be ensured that this plug is with the equipment upon purchasing). From the plug comes the battery cable red (+ Ve) and blue (– Ve), the handset together with 'send' switch, and the external loudspeaker connections.

The power supply voltage should be between 12 V and 30 V, voltages for which the voltage regulator in the TELECAR can accommodate. This regulator is short-circuit proof, that is, after the short has been removed, the voltage is restored to normal after a few seconds break.

Upon connecting the supplies, the pressel switch EIN is operated. When the RAUSCHSPERRE button is pressed there should be a white-noise sound in the loudspeaker and LED D3 should illuminate. If this occurs, then the equipment can be switched off again. The coaxial cable, which has previously been connected to pin 1 of the first IF stage, is now further connected to point C 45 – screen to ground.

6. TUNING

The equipment is switched on and channel 40 (i.e. 145.000 MHz) is selected. A strong FM signal, preferably from a commercial signal generator, is now required. This should be heard in the loudspeaker and also when the switch is selected to D – the 600 kHz offset frequency.

If nothing is heard, in spite of increasing the input power from the signal generator, it is highly likely that the PLL has not locked in. Try screwing the core of L1 fully in, and when that doesn't work, fully out until the receiver locks in to 145 MHz.

Now the cores of the receiver module inductors can be adjusted for maximum receiver sensitivity – always reducing the input signal power from the signal generator as the tuning progresses. Then adjust the regulator DC output voltage to 11.5 V with R4 as measured with a voltmeter connected between pins 3 and 4.

Normally, the first and second IF, as well as the other stages, are in their original condition and require no further adjustment. The equipment I modified gave a sensitivity check of 0.4 to 0.7 microvolts for 12 dB SINAD.

Now it's the turn of the transmitter. Operate the hendset pressel switch and the green lamp will come on – if not, check the bulb. The TELECAR TS' send/receive relay should be clearly heard anyway. Also listen for the operation of the small auxiliary relay when changing over to the 'send' condition. if all is well, the transistors T8 and T9 should have a working voltage when the pressel switch is pressed and which returns to zero when the switch is released. Check also, when in the send condition, that the AF modulator has a low frequency modulation voltage at C 10.

The equipment is then set to the send condition and the trimmer capacitor C 57 tuned for maximal output power. Then, all the trimmers shown in **fig. 8** should be tuned to maximise the output power. The power meter should now indicate an output power of 2.5 to 4 watt. The driver indicator (L7) core is then screwed-in but be very careful in the choice of trimming tool to do it with! The tuning should result in a power output maximum. If this does not occur, solder a 3 to 4 pF capacitor across L7 (underneath the board) and then try to adjust L7 core; again for a maximum power output.

Following that, tune L9 and C 39 for a maximum power output and then re-adjust the TELECAR output stage slightly to optimise the output power. The trimmers in the antenna filter are now adjusted for maximum output power. If any of them do not exhibit a definate tuning characteristic, connect the three links which are clearly located beneath the board.

Now regulator potentiometer R9 is adjusted for a 6 W output power. I have achieved 10 W at this point in the procedure but the regulator cooling is not sufficient to allow continuous operation at this power. In anycase, 6 W is the unmodified TELECAR's specified output power.

Both potentiometers in the modulation amplifier are now turned to midrange where experience has shown that a frequency deviation of 4.5 kHz to 5 kHz is to be expected.

Using some high-quality components could result in the TELECAR driver stage being over-driven. This becomes apparent when the equipment input supply current exceeds 2.5 Amps and also perhaps, a stage starts to self-oscillate. This can be overcome by connecting a resistor (of a few ohms) in the connecting lead to point X. A resistor in the PLL driver supply to T8/T9 will bring the working potential down. The power then is not so high and all stages are driven normally.

All the measurements were made with a Schlumberger test-set 4040 together with a model 632-1 spectrum analyser. Inspection of the output spectrum revealed spurious radiations of smaller than 25×10^{-6} W as demanded by the regulations.

If the full range of the 2 m band is desired, all the cores in the HF input stages and those in the transmitter must be re-adjusted accordingly. Both the output power and the receiver sensitivity will be compromised slightly.

Now that the work has been carried out, time can be devoted to actually operating the modified equipment. I have now modified three of them and used them all without any problems whatsoever. They are used in the car, at work and in the family home.

Roman Wesolowski, DJ 6 EP

5760 MHz Power-Amplifier using the YD 1060

A review of a modern 6 cm Transverter

This article should constitute a further contribution for the awakening of activity in the 6 cm band. It describes the construction of a valved power amplifier designed in the coaxial technique (fig. 1) and on account of its performance data represents an excellent alternative to the low-power travelling wave tube (TWT) amplifier.

1. MECHANICAL CONSTRUCTION

In order to minimize the undesirable electrode lead inductances and at the same time to simplify the construction as much as possible, contact

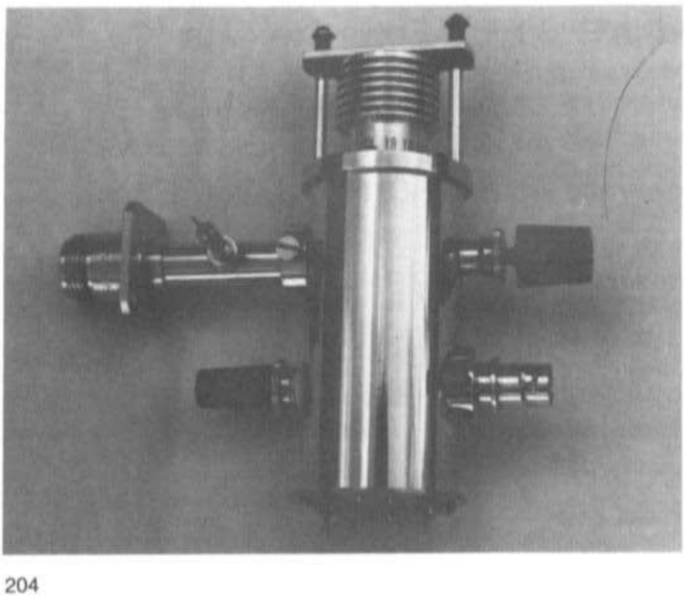


Fig. 1: A completed YD 1060 power amplifier

fingers and complicated coaxial plungers were dispensed with altogether.

The main body of the stage consists of two brass cylinders which have been press-fitted into one-another (**fig. 2**) by means of a tuned ring. In order to achieve a concentric construction, the inner (grid) tube (3) is turned to 13.20 mm (grid ring seating) and externally to 16.60 mm (to suit the distance ring) – all in one working operation on the lathe. This can be seen more clearly in **fig. 3**. The completed ring (4) is now force-fitted on to the inner cylinder with the aid of an engineer's vise and eventually cleanly soldered.

The mechanical construction is commenced with the mounting of the guide sleeve (5) for the output socket (6) and the oppositely located bush (7) carrying the tuner-plunger (8) into the main (anode) cavity. Both parts are force-fitted – if possible – into the main cylinder and soldered. The anode resonator is then screwed-out in order to be able to remove any protruding internal portions of the output socket. Only then is the inner cylinder, together with the fitted ring, forced on to the anode resonator. The feed and tuning arrangements for the cathode resonator are similar to those of the anode resonator inasmuch that the feed-in bush (9) and the tuning screw (10) pass through both cylinders. The requisite details may be obtained from **figures 4 to 6** and the following component list identifies the most important components.

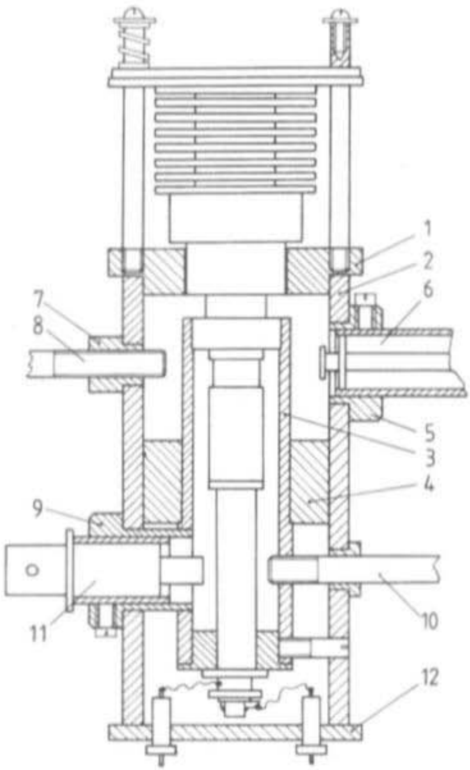


Fig. 2: Main components cross-sectional drawing of 6 cm band power amplifier

Main Components List

1) anode cavity cap
2) anode cylinder (main cylinder)
3) grid cylinder (inner cylinder)
4) distance ring
5) guide sleeve
6) output assembly
7) tuning bush
8) tuning plunger
9) input assembly bush
10) tuning plunger
11) input assembly
12) lower cap
13) cathode contact sleeve
14) cathode tube
15) outer cathode tube
16) heater connecting sleeve
17 a, b) teflon discs
18) pin
19) cathode resonator cap
20) U-washer
21) teflon washer
22) teflon washer

The anode cavity cap (1), which together with the anode ring of the cylinder and the PTFE insulating foil, forming a blocking capacitor, is secured to the main cylinder with six M2 x 10 mm screws. Be careful with the upper surfaces of the main cylinder and the anode cavity cover not to scratch or soil them in any way. The lower cover of the tube, which carries the feedthrough capacitors for the cathode and heater, is now mounted.

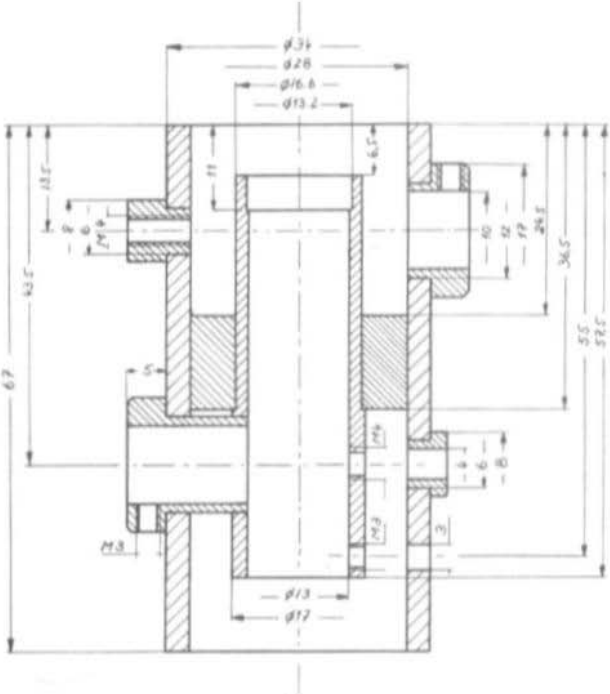


Fig. 3:
The main assembly: Anode and grid cylinders with guide bushes for the input/output sockets and tuning plungers

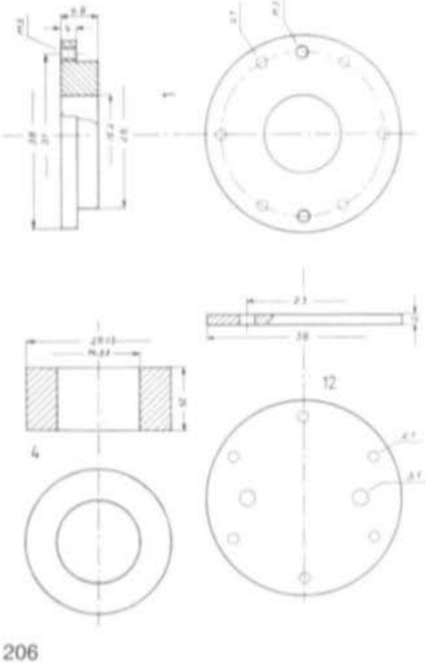


Fig. 4:
Anode cap, end cap, distance ring

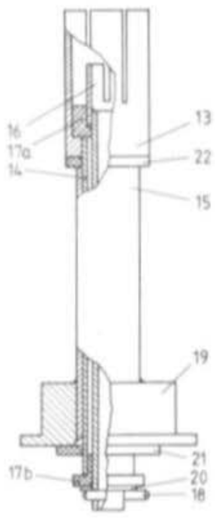


Fig. 5:
Cathode assembly for the YD 1060 6 cm power amplifier

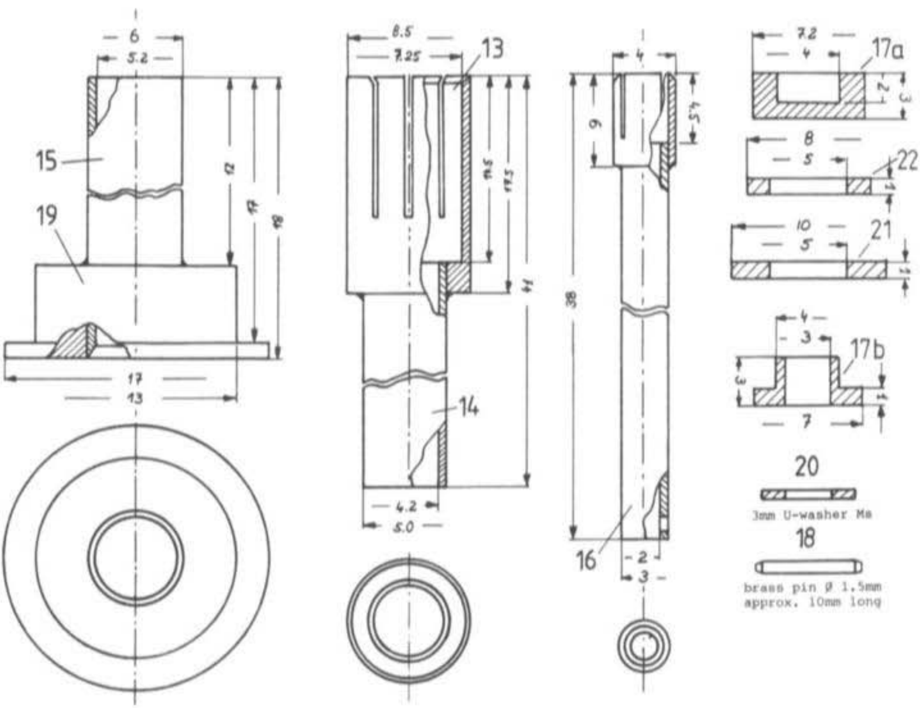


Fig. 6: Cathode assembly piece parts

1.1. The Cathode Resonator

Fig. 5 depicts the detail of the cathode resonator. The cathode contact sleeve (13) is soldered on to the 5 mm dia cathode tube (14) and slotted several times longitudinally. The proprietory tube (all the other tubes up to 12 mm dia can be obtained in model construction shops) is wrapped tightly with PTFE foil and fitted into the outer cathode tube (15) of 5.2 mm internal dia. The heater connecting sleeve (16) is pushed into the two teflon washers (17 a, b), fixed at the end by means of a 1.5 mm dia wire pin through a drilling in the tube, and soldered to the connecting leads. The cathode resonator cap (19) is rubbed with conductive silver solution in order to ensure a good contact. The whole assembly is then fitted into the cathode resonator and secured by side screws.

1.2. Input and Output Assemblies

Both resonators are capacitively coupled. Specially prepared BNC (preferably N) sockets are fitted into 10 x 1 mm brass tubes, the cleaned and polished ends of which protrude into the resonator. The output socket is fitted with a 5 mm disc which protrudes about 1.5 mm deep into the anode resonator. The distance of the coupling shaft (4 mm tube soldered to socket inner) to the cathode cylinder is about 2 mm. **Figures 7, 8 and**

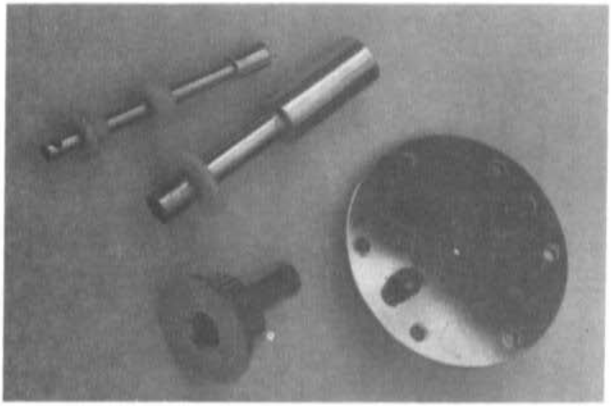


Fig. 7:
Lower cap and associated parts fo fig. 6

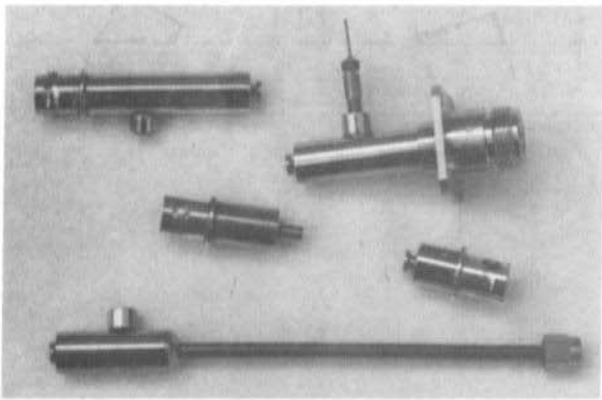


Fig. 8:
Various input and output coupling arrangements

Fig. 9:
Partly assembled resonator with cap and tuning plunger

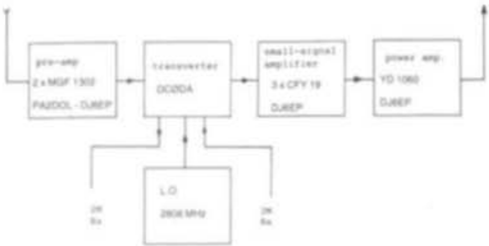


Fig. 10: Block diagram of a 6 cm band linear transverter

9 should show the construction more clearly than any written description.

1.3. Fitting the PA Tube

The anode ring of the tube is wrapped in PTFE-sheeting and screwed tightly into the anode cover. The thin PTFE sheeting must be employed carefully as the mechanical properties could very well change with temperature thus tending to loosen the grip on the tube. It has been the experience that 0.08 to 0.10 mm teflon silk has very good properties in this respect. The grid seating and ring are both coated with conductive silver and the tube, complete with anode plate, is mounted into the assembly and screwed in.

2. COMMISSIONING AND TUNING

The output is connected to a suitable power meter and the heater to a 6 V supply. The working point is adjusted with the assistance of a constant voltage two-pole network arranged in the usual form. With an anode voltage of 400 VDC, a quiescent current of 20 mA should flow.

The amplifier can now be driven, the cathode resonator and the tuning plunger should be iterated for various settings of the depth of the coupling shaft for a maximum anode current. An RF input drive power of 100 mW should cause about 60 mA of anode current to flow.

In the same manner, and under constant supervision of the anode current indicator, the anode circuit is tuned for a maximum output power. If the amplifier gain is too high, i.e. the output coupling is too tight (screwed too far in), self-oscillation will be apparent.

3. THE TRANSVERTER

As the cavity mixer and the varactor multiplier are things of the past, as far as the 6 cm band is concerned, a more modern scheme of things will now be outlined.

In the self-constructed equipment made by the author, a single board transverter was employed in micro stripline technique following the design of DC 0 DA. A home-constructed linear amplifier with two CFY 19s in parallel in the PA delivers the required 150 mW in order to be able to adequately drive the subject valved PA.

A modified, two-stage pre-amplifier, designed by PA 2 DOL, gives a 1.9 dB noise figure to the down-converter mixer, equipped with an MGF 1302 and also an overall gain for the unit of 26 dB. The equipment, shown in **fig. 10**, and given only the briefest of mentions here, was introduced at the Dorsten GHz gathering in February 1987.

4. RESULTS

With the design as described, and an anode voltage of 450 V, 6 Watt of output power was achieved. The anode current was driven from a quiescent 20 mA to 100 mA which, with provision for adequate cooling, allows the constant operation in both telegraphy or SSB.

An inspection of the send output spectrum reveals a suppression of image and local oscillator signals of a further 26 dB by the use of a valved stage. In conclusion I must acknowledge the assistance of Rolf Kueppers, DL 4 JK for the measurements and Horst Lehrke for the photographs.

Eugen Berberich, DL 8 ZX

VCOs using Semi-Rigid Cable Tuned Circuits

Oscillators for the VHF/UHF range present particular difficulties for the designer in order to make tuned-circuits of sufficient Q and stability. In addition, if the oscillator also has to be immune to external influences such as microphony, radiation and stray coupling then open circuits and coiled inductors are inadmissable. Basically, tuned lines with an air dielectric have a high Q but they possess a high sensitivity to microphonic effects and are difficult to fabricate. A good alternative is offered by concentrated line components as well as the printed-circuit, stripline technology.

Fixed frequency oscillators can nowadays employ surface wave resonators, with advantage, to work well into the UHF range. They are only worth while for the amateur if they can be obtained for the specific frequency required.

On account of the above reasons, I am now suggesting that semi-rigid cable (1) can be used for the tuned circuit of a VCO. If the thinnest type is not used, the dielectric is likely to be PTFE (Teflon) and the silvered inner conductor forms a resonator with a relatively high Q. This has been confirmed by test-measurements and later a circuit example was constructed for the frequency range 600 to 800 MHz.

1. Q-MEASUREMENTS

In April 1987 a series of unloaded Q-measurements were made on the various resonators (**see table 1**) using a Boonton 190 AP Q-meter. Of course, the loaded Q is not the only criterion which determines the low-noise working of the oscillator. A very important factor is the loaded Q of the tuned circuit which is determined by the coupling to the active circuit element (transistor). The looser the coupling, the higher the loaded Q.

2. A PRACTICAL VCO

A U 310 FET was chosen as the active element of the oscillator and connected in a common-gate circuit. The relatively high output resistance only

Z Ω	Length mm	Dielectric	freq. MHz	Q --	Remarks
60	150	foam	190	110	
60	200	polyethylene	150	130	similar to RG-213/U
75	200	teflon	190	40	**thin** (Suhner SR2-75)
75	200	teflon	190	110	**thick** (Suhner SR3-75)
50	100	PTFE (RG-402/U)	270	80	**thick** (Suhner SR-3)
50	200	PTFE (RG-402/U)	180	80	
240	150	foam	150	170	symmetrical
75	100	epoxy	190	50	stripline
50	150	PTFE/ceramic	190	50	stripline
---	100	teflon 1mm	270	40	industr. UHF-oscillator
---	100	air	150	250	industr. VHF-resonator
---	---	air	150	300	air-coil 4 wdg. 8 Ø 1 mm silvered wire
---	---	air + VHF core	150	100	as before, but with VHF slug
---	---	air + Ms core	150	150	as before, but with brass slug
---	---	glass rod 4 Ø	150	270	7 wdg. 1 mm silvered wire, 12 mm long
---	---	as before	150	270	as before, but with marmor kitt 1000 (marble filler) dry
---	---	ceramik coil	150	200	5 wdg. 8 Ø

Table 1

lightly loads the resonator. In the same manner, the Q of the frequency determining capacitor must also be considered: modern miniature types are not really suitable for low-noise oscillators, the larger (higher voltage) construction are much more advantageous.

Fig. 1 shows the final circuit in which the power FET P 8002 has been used. The U 310, however, brought better results and it could be directly let into a drilling on the PCB and directly soldered in – the metal housing being connected internally to the gate.

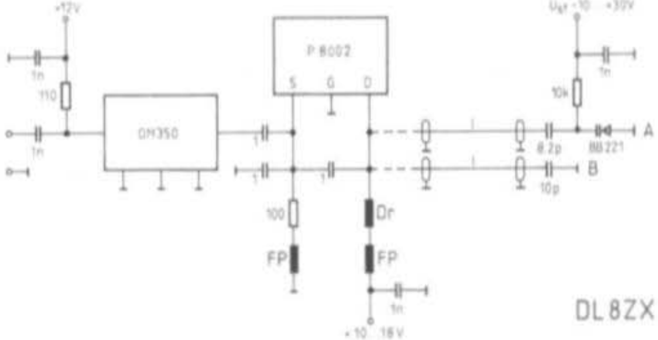


Fig. 1: VCO complete with buffer amplifier for (A) 550 - 800 MHz (B) 550 - 900 MHz; L = 40 mm copper clad cable (Suhner SR-3)

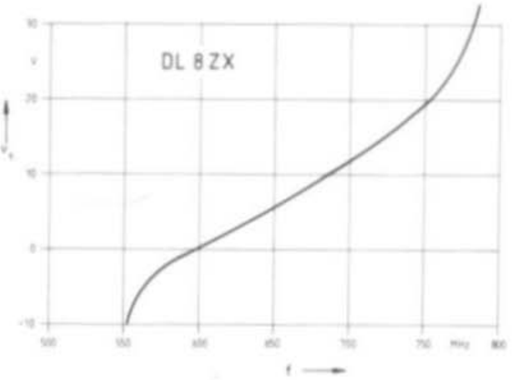


Fig. 2: VCO frequency as a function of the tuning voltage for the circuit of fig. 1.

The oscillator circuit can either have a λ/2 or a λ/4 tuned circuit. At the higher UHF frequencies, however, the λ/4 resonators are mechanically so short as to be impractical and λ/2 lengths are to be preferred (2 to 5). The DC power is introduced via a radio frequency choke (RFC). This component must be chosen carefully in order that it does not cause parasitic oscillations. The ideal point for the DC to be fed in would be the electrical centre of the resonator – this entails making a slot, or at least a slit, in the cable outer. The semi-rigid line should be soldered at several places – preferably along its entire length – to the ground plane of the printed circuit board.

The frequency is coarse-tuned by a trimmer connected to the open end of the λ/2 tuned-circuit, the fine-tuning is effected by a loosely coupled varicap diode. Example VCO A covers a frequency range of 550 to 900 MHz and VCO B from 550 to 900 MHz (**fig. 2**). The output voltage, without buffer stage, and measured in 50 Ω was:

at Vb = 10 V : 150 mV
Vb = 12 V : 200 mV
Vb = 18 V : 500 mV

Tuning across the bands caused level variations of up to ± 3 dB – the tuning voltage varying from – 10 to + 30 V.

The output from the source electrode is fairly rich in harmonics, the second harmonic being only 10 dB weaker than the fundamental. This can be of advantage when a frequency doubler is required. These harmonics can, however, be easily filtered out should a clean output be desired.

The drain output is quite free of harmonics but it is more prone to frequency pulling under the influence of load variations. As this effect may be considered as quite a large disadvantage, it must be mitigated by the addition of a buffer stage to this output connection. This was accomplished here with a wide-band hybrid amplifier OM 350 (or OM 345) from Valvo. With a directly coupled output across a 110 Ω resistor the output impedance drops from 75 Ω to 50 Ω.

Without a buffer stage the UHF output frequency alters by – 500 kHz when the output is terminated via a 1 pF capacitor. The buffer stage reduces the pulling effect very considerably. Further isolation

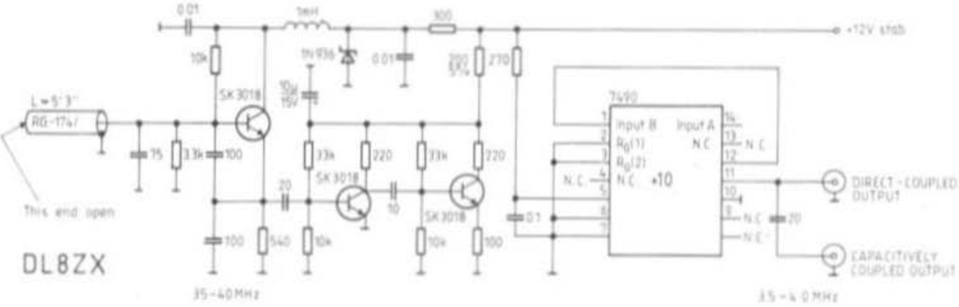


Fig. 3: A 35 - 40 MHz VCO using a 180 cm length of RG-174/U coaxial cable (QST Ed. 8/1980)

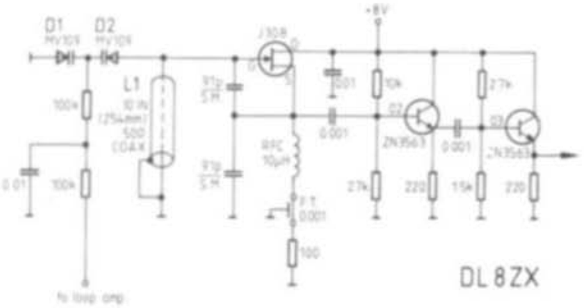


Fig. 4:
A 50 MHz VCO using a 254 mm length of coaxial cable not further specified (QST Ed. 9/1981)

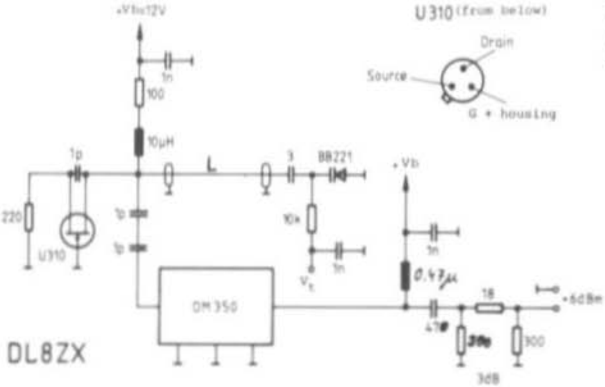


Fig. 5:
A DL 8 ZX VCO tuning from 850 to 1000 MHz

may be obtained by including an attenuator between the oscillator and the buffer amplifier.

This hybrid amplifier has been designed for an output level of 106 dB rel.μV. At this level the intermodulation products are greater than 60 dB down on the carrier level. When using one signal, however, the hybrid amplifier may be drriven to deliver 0.5 V into 50 Ω in order to allow ring mixers to be properly driven. At this level, of course, the harmonic output is also increased accordingly.

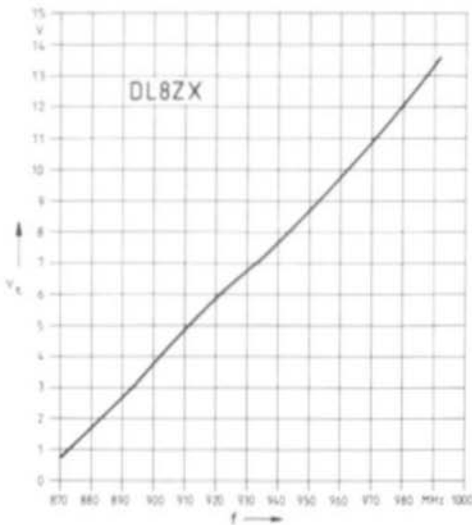


Fig. 6: Frequency as a function of tuning voltage for the VCO of fig. 5

3. FURTHER EXAMPLES

A few years ago the American ARRL amateur magazine QST published two voltage-controlled oscillators using cable resonators. Their frequency of operation is lower (**fig. 3/4**) but they do represent a completion of the range of UHF VCOs developed by myself.

As a further example: –
With a semi-rigid line L of 30 mm long and Z_0 = 50 Ω, a tuning range of 850 to 1000 MHz (**fig. 6**) was achieved. This circuit was constructed on an HF type development printed circuit board using through-plated connections and the **shortest possible connections**. The output power, after the aforementioned OM 350 hybrid amplifier and 3 dB attenuator, was + 6 dBm. The 3 dB attenuator, incidently, was made from chip resistors.

4. LITERATURE

(1) Editors:
Balun Transformers from Semi-Rigid Cable
VHF COMMUNICATIONS Vol. 8,
Ed. 4/1976, P. 221

(2) VHF/UHF-MANUAL 3. Ausgabe, 3.1. - 3.7.
(über Verlag UKW-BERICHTE)

(3) Möhring: Empfangstechnik im UHF-Bereich
Seite 57 - 61

(4) Karl Weiner: UHF-Unterlage,
Gesamtausgabe Kapitel A 2.3.1

(5) Koaxial-Tedroden, Kapitel 4.4.2
Fa. Valvo

Angel Vilaseca, HB 9 SLV

Measuring Wavelengths at Microwave Frequencies – Simply and Cheaply

A difficulty which was always cropping up when using microwave equipment was the problem of knowing the frequency of operation. Of course, the solution to the problem is the use of a microwave frequency counter but for most amateurs this lies well outside the bounds of possibility. A further method is the use of an absorption wavemeter but the construction of such an item of test equipment requires care, precision and experience. Also a microwave frequency meter must be used for calibration when it has been constructed!

I then chose another solution which is simple, gives a three-place resolution and costs practically nothing – the Michelson Interferometer.

1. THE MICHELSON INTERFEROMETER

When two waves from different sources superimpose on each other at the same place, an interference occurs, that is, their amplitudes add. The resultant amplitude depends upon the relative phase difference between the two waves. A constant amplitude is apparent where there is a constant phase difference between the two waves: in other words, they must be **coherent**.

The best way of obtaining two coherent waves is to split a single beam into two divergent beams. The two beams are, from the laws of physics, inherently coherent. The phase difference, as observed from a distant point depends only upon the path length that each beam has to the point where they are made to converge and impinge on each other.

The sketch of **fig. 1** shows Michelson's Interferometer. The beam splitting is carried out by means of a grid which divides the beam into two equal power paths, one directed to reflector A and the other to reflector B. The practical aspects of both grid and reflectors are quite simple and will be dealt with later.

The beam deflected from reflector A is directed back through the grid where it undergoes a further splitting, one part going to a receiver and the other part away to a place of no importance for the moment.

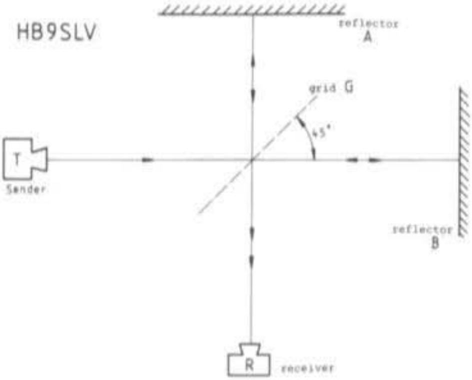


Fig. 1: Principle arrangement of a Michelson Interferometer

Fig. 2: Measurement of the distance between 20 maxima

The beam deflected from reflector B is also divided in the same manner and one of the component beams again directed to the receiver.

Recapping then: a portion of the beam produced by the microwave sender follows the path T-G-A-G-R and another portion follows the path T-G-B-G-R. The receiver will receive a maximum amplitude when both waves are in phase and that occurs when the path length between the two waves differs by a multiple of the wavelength.

If, for example, the reflector B of path T-G-B was moved, but always ensuring that it remained perpendicular to this path-axis, the amplitude at the receiver would go through a series of maxima and minima. The distance the reflector moves between two observed maxima (or minima) amounts to a half wavelength. This may be proved quantitively, see the appendix.

An interferometer was built by the author which worked at a frequency of 10 GHz. At this wavelength – about 3 cm – it is very difficult to measure such a small fraction of a wavelength. This is alleviated by measuring over a distance of 20 maxima – the first one being counted as zero – and dividing by 10, thus effectively increasing the resolution of the measurement of the wavelength (**fig. 2**).

In order to obtain the frequency in GHz, the number 30 is merely divided by the result (in centimetre) found for the wavelength.

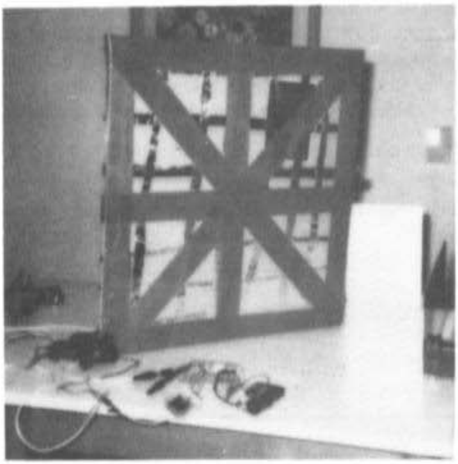


Fig. 3: The grid construction can be clearly seen in this photograph

2. A PRACTICAL INTERFEROMETER

The main problem when measuring with an interferometer is the unwanted reflections from surrounding objects which affect the reading on the receiver S-meter. For this reason it is better to construct the interferometer away, as far as possible, from vertical reflecting surfaces. This could be on a table in the middle of the room or even in the open air.

The grid is the most critical part of an interferometer. After much searching and experimentation the best results were obtained with a 50 cm x 50 cm square of cardboard which, on one side, was criss-crossed with 2.5 cm strips of aluminium foil about 4 cm apart. These strips may be affixed with adhesive tape (**fig. 3**).

The reflectors were made from ready-to-hand materials. One consisted of a large piece of unetched printed-circuit-board, the copper side towards the microwave source. The other was a 50 cm x 50 cm piece of aluminium sheet. Although it is not absolutely necessary, it would be good practice to use reflectors which are of the same material and size.

Small horn radiators were employed for the sender and receiver directional antennas. The sender produced about 10 mW from a Gunn element. The receiver consisted of a mixer with an MA 41453 crystal. The local oscillator signal was derived, also from a similar Gunn oscillator, and fed to the mixer via a 30 dB directional coupler. The IF amplifier was provided by the use of an FM car radio receiver.

The above mentioned microwave parts originated from a few dis-assembled car RADAR warning detectors. The aluminium die-castings were of remarkably good quality. Each had a horn, a 30 dB cross-coupler, a mixer-diode holder, a resonator with a Gunn element, a microwave choke with tuning screw as well as a special absorber for the excess power from the Gunn element. The mixer diode receives only a thousandth of the 10 mW of produced power owing to the 30 dB coupling loss of the directional coupler. The only disadvantage of the equipment is that nothing can be soldered to the aluminium die-castings.

3. MEASUREMENTS

Both the sender and the receiver are switched on and allow some time to thermally stabilize. Also in the interest of a stable frequency, stabilized supplies should be used for all equipment.

The sender and receiver are then both tuned to the test frequency and the system checked for any frequency drift. It is imperative that both receiver and sender remain tuned to each other during the minute or two duration of the test. The selectivity is rather high and therefore any frequency drift can quickly upset the test. High sensitivity is not, however, required owing to the proximity of the test components to each other. It should be therefore possible to use the mixer diode as a detector and thereby eliminate the selectivity. For this the receiver oscillator supply should be disconnected and the current measured between diode and ground. When doing this the usual precautions must be taken to prevent the diode from being destroyed by static charges. These are the same precautions that have to be taken with MOS devices.

The placement of sender, grid, reflectors and receiver is carried out according to the arrangements shown in **figs. 1 and 4**. The distance between the sender and the grid is about 65 cm, and also that between grid and reflector A and reflector B. These distances are not particularly critical.

The reflector B is then moved along axis E-G-B always keeping it perpendicular to the plane of movement. A maximum in the receiver is noted. If the maximum causes an S-meter deflection of more than 60 % FSD, the sender beam should be weakened by some power absorbent material – a stack of telephone books can be placed in front of the sender horn.

The reflector B's position for a maximum is noted for a zero reference and marked on a paper strip

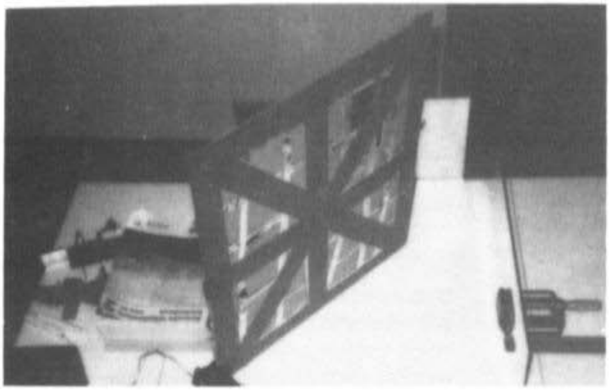


Fig. 4: The author's 10 GHz-range Michelson Interferometer arrangement

which has been previously affixed along the T-G-B axis.

Now the reflector B is again moved along the path T-G-B to the right, keeping the plane of the reflector always perpendicular to the path direction. As previously explained, the S-meter will exhibit a series of maxima and minima as the reflector B is moved. These maxima (the first being counted as zero) are counted and the twentieth marked. The distance to the reference point is now measured and divided by 10 in order to obtain the wavelength in centimetres. The figure 30 is then divided by the test result in cm and the frequency in GHz obtained.

4. APPENDIX

A portion of the beam takes the path T-G-A-G-R and the other portion takes the path T-G-B-G-R. The path-length difference between the two is:

$$\triangle = 2\,GA - 2\,GB = 2\,(GA - GB) \qquad (1)$$

The voltage at the receiver is a maximum when the path difference $\triangle$ is a multiple of a wavelength:

$$\triangle = n\lambda \qquad (2)$$

When reflector B is moved from point B_1 to B_2 – both maxima positions – then:

$$\triangle_1 = 2(GA - GB_1) = n_1\lambda \text{ and}$$
$$\triangle_2 = 2(GA - GB_2) = n_2\lambda \qquad (3)$$

The path-length difference between the two positions is:

$$\triangle_1 - \triangle_2 = 2(GB_2 - GB_1) = (n_1 - n_2)\lambda \qquad (4)$$

That is also a whole multiple of the wavelength λ.

$$\lambda = \frac{2\,(GB_2 - GB_1)}{n_1 - n_2} \qquad (5)$$

λ is included in the wavelength $GB_2 = GB_1$, i.e. the measured distance between the reference and end position of the reflector B.

$n_1 = 0$ and $n_2 = 20$:

In this case equation (5) can be simplified

$$\lambda = \frac{2 \times \text{measured distance}}{20} = \frac{\text{measured distance}}{10} \qquad (6)$$

Example: measured distance = 28.8 cm
wavelength = 2.88 cm
or frequency = 10.4 GHz

Dragoslav Dobričič, YU 1 AW

Pre-amplifier – Pros and Cons

Perhaps the most serious competitor to the subject of antennas in the attentions of radio DXers is the receiver pre-amplifier. Whenever a problem is complex it gives greater scope for misconceptions, the spread of widely accepted so-called facts, pseudo-scientific folklore, all of which tends to cloud the issue in an aura of mystique.

It is common knowledge nowadays that the gain specification alone of a pre-amplifier is insufficient to assess its merit. Gain is easily achieved and an excess of it merely drives the following stages into producing intermodulation distortion under strong signal conditions. Likewise, it is also well-known that the noise producing qualities of a pre-amplifier, expressed by such quantities as noise-factor, noise-figure or noise temperature, are much more important parameters to be considered. The appendix gives the relationship between these quanties.

Now comes the questions: How low must the noise in a pre-amplifier be? Should one disregard cost and get the very lowest noise specified? Also, how much gain – if it is not the highest gain that to be striven for. What is the optimum gain and upon what does it depend?

1. MAXIMUM AMPLIFICATION?

The aim of this article is to point out a few common mistakes apparent in the employment of a pre-amplifier intended to improve a system and to suggest simple methods by which these mistakes may be avoided. The dilemma of how much gain will be unravelled first. Why not simply use as much amplification as possible in order to detect the weaker signals? Well, this obvious solution would be fine if the receiving system (aerial and pre-amplifier) itself did not generate noise. All the pre-amplifier would do is to amplfy the noise along with the signal thus giving higher levels of both but with no greater discrimination between them. In order to obtain a higher signal level with lower noise, it is necessary to minimise the additional noise introduced by the receiving system during the process of amplification.

The noise introduced into the antenna is, however, unavoidable and nothing much can be done about it once the antenna has been erected with the due attention paid to the antenna elevation, location, working frequency and the gain. The signal-to-noise ratio (S/N) at the antenna output

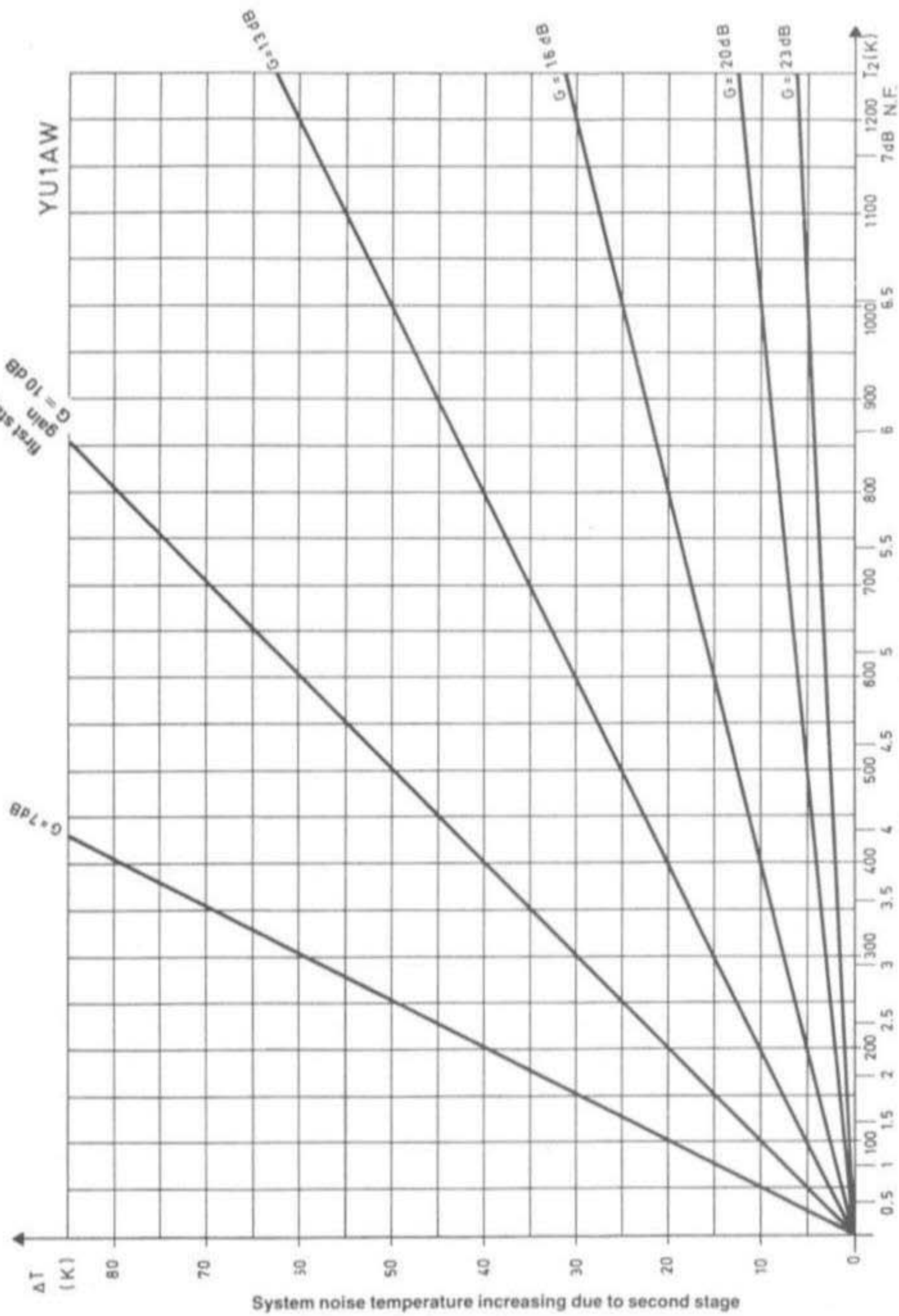


Fig. 1: The influence of the second-stage noise (T_2) on the total receiver noise (ΔT) for various values of first-stage gain (G)

terminals to the receiver, is then, the best that is available for that particular antenna and it is clear that in the subsequent processing of the signal by the receiver, nothing should be allowed to further deteriorate it. This cannot be achieved in practice and the design effort should be devoted to confine the noise to an acceptable level for the prevailing conditions. This is because the pre-amplifier not only amplifies the signal and the noise arriving at the antenna terminals but also adds its own internally generated noise. The signal-to-noise ratio of the output of an ideal amplifier is exactly the same as that at the input. An actual amplifier, unfortunately, always has a lower S/N ratio at its output than is presented to its input and it is logical to conclude that the merit of a pre-amplifier may be assessed by how low its internally generated noise is.

The pre-amplifier gain evidently, has so far played no part in the discussion. The previous discourse is valid only for the case where an ideal amplifier follows the subject pre-amplifier. As this can never be achieved in practice, it can be concluded that the pre-amplifier gain should only be high enough to prevent the noise from the second stage from sharply deteriorating the S/N of the overall system.

This point may be clarified by a numerical example. It will be assumed that the second stage of amplification has the same level of internally generated noise as that of the pre-amplifier. Also, the amplification of the pre-amplifier is 50 times (i.e. 17 dB of gain). The input to the second stage will then be the signal plus noise from the antenna together with first stage's own internal noise – all amplified by 50. The noise contributed by the second stage will then be only 1/50th (2 %) of the overall noise. This can be expressed in the following manner:

$$T = T_1 + \triangle T \text{ and } \triangle T = T_2/G$$

where

T = the overall noise temperature
$\triangle T$ = the noise contributed by the second stage
T_1 = the noise temperature of the first stage
T_2 = the noise temperature of the second stage
G = the gain of the first stage

From this example it is clear that the second stage will only deteriorate the overall S/N ratio by a very small amount. It is also apparent that amplification values of between 20 and 50 for the first stage will surfice to overcome the noise T_2 of the second stage as long as this noise is not excessively high (**see fig. 1**).

When a pre-amplifier is to be added to a receiver, it is the receiver itself which can be regarded as the second stage of the above example. The sensitivity of the overall system will be improved from the mediocre specification of the receiver alone but the gain of the pre-amplifier may not be sufficient to determine the second stage noise if, for example in the extreme case, that receiver employs a high-level ring-mixer in order to improve the intermodulation performance.

An important point has been arrived at and that is the decision to sacrifice pre-amplifier gain in order to preserve the higher level of signal handling properties of the receiver system or whether the circumstances dictate that the pre-amplifier be given a high gain in order that the overall receiver is as sensitive as possible.

There is no universal recipe and the receiving system must be tailored in order to meet the prevailing conditions as seen at the antenna output terminals. The first factor determining these conditions is the noise arriving with the signal noise which cannot be influenced by the operator but must be coped with by the receiver. This noise will now be analysed and each component of it examined.

2. ANTENNA NOISE

In the VHF range the **sky noise** is the **greatest contributor** to the total antenna noise and this is universally proportional to frequency. It cannot be influenced in any way but can be neglected at frequencies above 1 GHz (1).

Above 1 GHz, the **ground noise** is constant but decreases towards lower frequencies owing to the increasing ground reflectivity. But the total noise level is the sum of noise radiated from the earth and sky noise which has been mirrored from the earth's surface. When the antenna is directed skywards, as in earth – moon – earth (EME) or satellite communications, the contribution of this noise is small and is largely dependent upon the distribution of the side-lobes (1). The desirability for an EME antenna, to possess a clean lobe pattern, may now be appreciated.

When, on the other hand, "normal" VHF/UHF communication is carried out over the earth's surface, the antenna receives both ground and sky noise because the antenna lobes are directed to both, sky and earth in about equal amounts.

Two further contributions are: The **man-made noise from large cities and industrial areas** which vary according to location, and **atmospheric noise**. The later is very much smaller at VHF/UHF than at HF and dependent upon the prevailing atmospheric conditions.

The **thermal noise** across the radiation resistance of an antenna may be neglected owing to the very high efficiency of VHF/UHF antennas.

From what has been already said, it may be concluded that at the terminals of every antenna which is directed at the horizon, a noise power may be measured. When the antenna is directed towards the sky this noise power falls. The relationship of this noise to the working frequency is shown in **fig.2**.

Since it became known that the sky born noise fluctuated considerably, minimum values for certain areas were taken, other areas may have random noise power distribution (1). The objective values for antenna noise represent that of the absolute minimum because the urban noise was not taken into account and also there exists the possibility that the skyward directed antenna could be pointed to a particularly noisy part of the sky. The curves, then, can only indicate the minimum noise which may be expected under the most favourable environmental and space conditions.

3. SELECTING A PRE-AMPLIFIER

A simple method of selecting a pre-amplifier would be to estimate the noise arriving with the wanted signal and then aquiring a pre-amplifier which would develop an equal value of self-generated noise. This would result in a 3 dB deterioration in the S/N from the antenna as seen at the output of the pre-amplifier owing to the effective doubling of the total noise power. This may be acceptable, especially when it is to be compared against frequently occurring fades of below – 30 dB.

Using this method of assessment, a satisfactory pre-amplifier selection can be made which would be suitable for normal terrestrial communication purposes, assuming of course, that the receiving system's (pre-amplifier plus receiver) total noise contribution is equal to that of the antenna noise.

It is important for some communications applications to reduce the Δ(S/N) from 3 dB to 1 dB. This necessitates reduction in receiver noise by several times involving a further sacrifice of receiver dynamic range and also entails an extra expense.

Bearing this in mind, it may be concluded that the construction and employment of a pre-amplifier having a lower noise than that of the antenna does not make much sense for terrestrial work. The antenna noise is a relevant factor in the determination of other elements of the receiving system.

In order to illustrate this contention, an example will be taken from amateur practice.

Radio amateurs living in small country towns, away from motorways and industrial plants, can expect an antenna noise temperature of about 1000 K (Kelvin). The author has measured noise temperatures of around 800 K when a 2 m antenna was directed towards one of the quieter regions of Belgrade at about midnight.

Assuming that the amateur radio receiver has a 4 dB noise-figure – a typical enough value for a

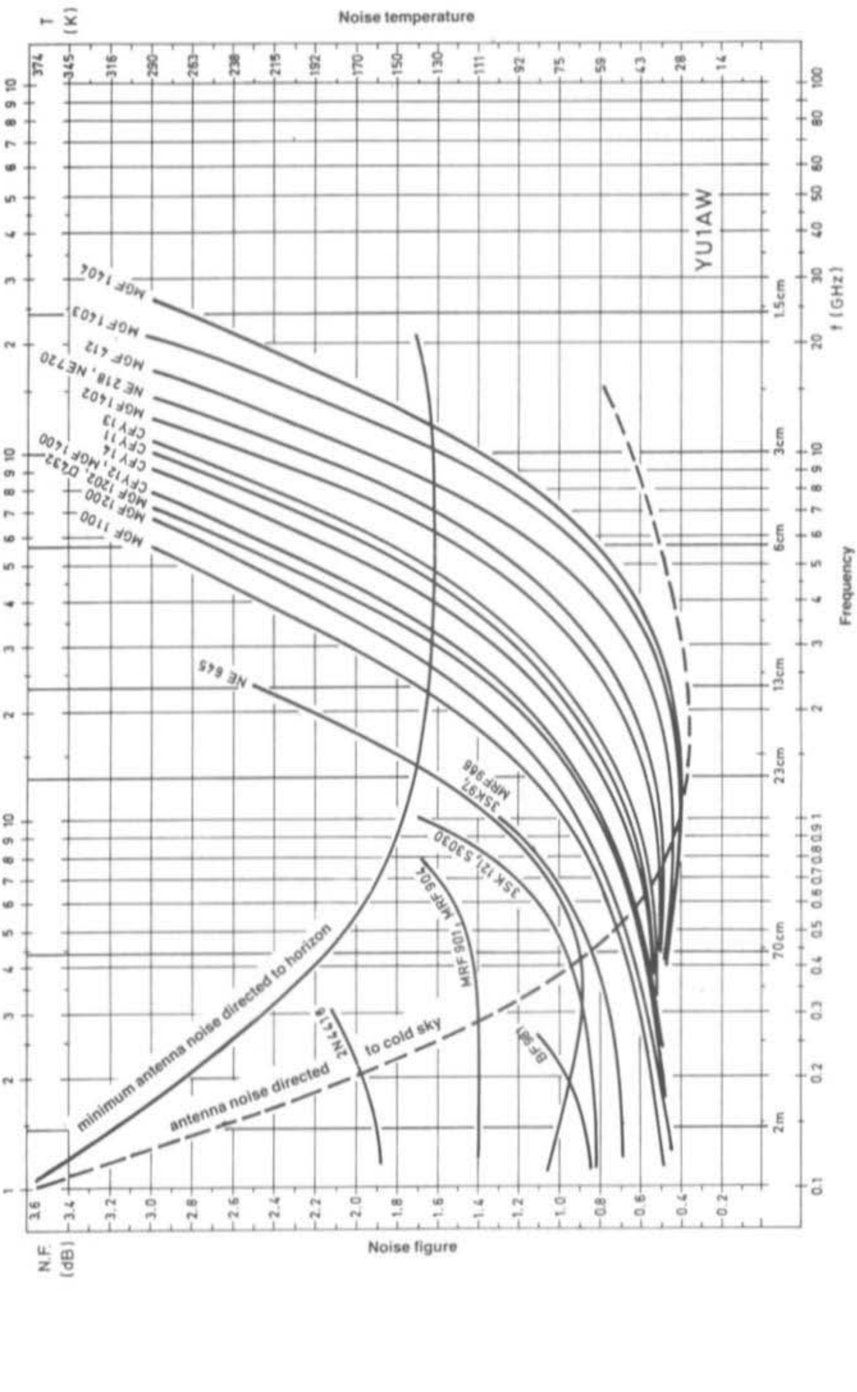


Fig. 2: Frequency dependence of the noise-figure of various well-known transistors and the minimum noise values to be expected from antenna noise.

commercial receiver – and that the antenna coaxial cable has a loss of 1 dB. If a CW or SSB signal, accompanied by noise of 10 dB lower, is induced into the antenna, the signal at the output of the receiver will have a lower value of signal-to-noise – namely 7.9 dB. This is a 2.1 dB deterioration of the signal when compared with the ideal (non-existent) receiving system.

The radio amateur realizes that he has a loss and tries to correct it by purchasing a GaAsFET pre-amplifier with a noise figure of 0.4 dB, a gain of 23 dB and costing a great deal of money – aha! This does the trick and brings the S/N up to 9.5 dB when connected at the receiver and when connected directly at the antenna terminals 9.8 dB. This represents an improvement of 1.6 dB or 1.9 dB in the latter case. The price in performance which has to be paid for this improvement is that the receiver's dynamic range has been drastically reduced making it useless for contest work.

The amateur manages to sell his dream performance pre-amplifier and obtains a much cheaper one with noise-figure of 2 dB and an amplification of 10 dB. He could have modified the input stage of his receiver as it is relatively easy to achieve noise figures in the region of 2 dB.

With this 10 dB pre-amplifier the dynamic range of the receiver is still somewhat reduced but by no means what it was when using the first pre-amplifier. Now, with the improved receiver, the original dynamic-range performance has been retained and a S/N of 8.9 dB at its output.

Was it really worth compromising the dynamic range of the receiver for a 0.6 to 0.9 dB increase in output S/N by using such a high performance pre-amplifier – not to mention the cost of the thing? Our radio amateur did at least realise his mistake and corrected it. The super-specified pre-amplifier failed to bring about the improvement in receiver performance which could have been expected from it – why? **Because no account was taken of the antenna noise.**

Only in space communications, and above all in EME work, is a pre-amplifier with the lowest possible noise figure justified. This is because the antenna is looking at the "cold" sky and the antenna noise is therefore much lower. Also, the wanted signal is mostly hovering at, or even under, the noise level so that an improvement in the system noise figure of 1 or 2 dB can bring about an improvement in reception of between 50 and 100 %.

Let's take a closer look at **figure 2**. It may be seen, that all GaAsFET amplifiers in both the 2 m and 70 cm bands have almost the same noise figures. The difference between the various types is that the expensive ones are specified at microwave frequencies as well (3). It does not make sense to pay twenty times the price for a microwave GaAsFET and then use it at VHF/UHF. All GaAsFETs have noise figures which lie under the 2 m aerial noise spectrum but the really high priced microwave types, on account of their narrow gate structure (0.5 to 0.2 μm), are liable to have a higher noise figure at the lower frequencies than at their specified microwave frequency band.

4. THE EFFECT OF CABLE LOSS

In the aforegoing discussion not much has been said about the effect of the coaxial cable on the receiving system. In reality, however, it is the coaxial cable and not the pre-amplifier which is the first element in the receiving system and always contributing a loss.

It has already been mentioned that the first component has a determining influence on the system noise figure. Generally it increases the system noise figure by the amount of attenuation in the antenna down-lead to the receiver pre-amplifier. Even a short length of low-loss cable increases the system noise figure by a few tenths of a dB. The best pre-amplifiers may not compensate for the use of long lengths of low-grade coaxial antenna down-lead, **see fig. 3**.

It follows, then, that a mediocre pre-amplifier located at the antenna terminals is a much better proposition than a high-performance pre-amplifier located at the end of a long length of lossy cable.

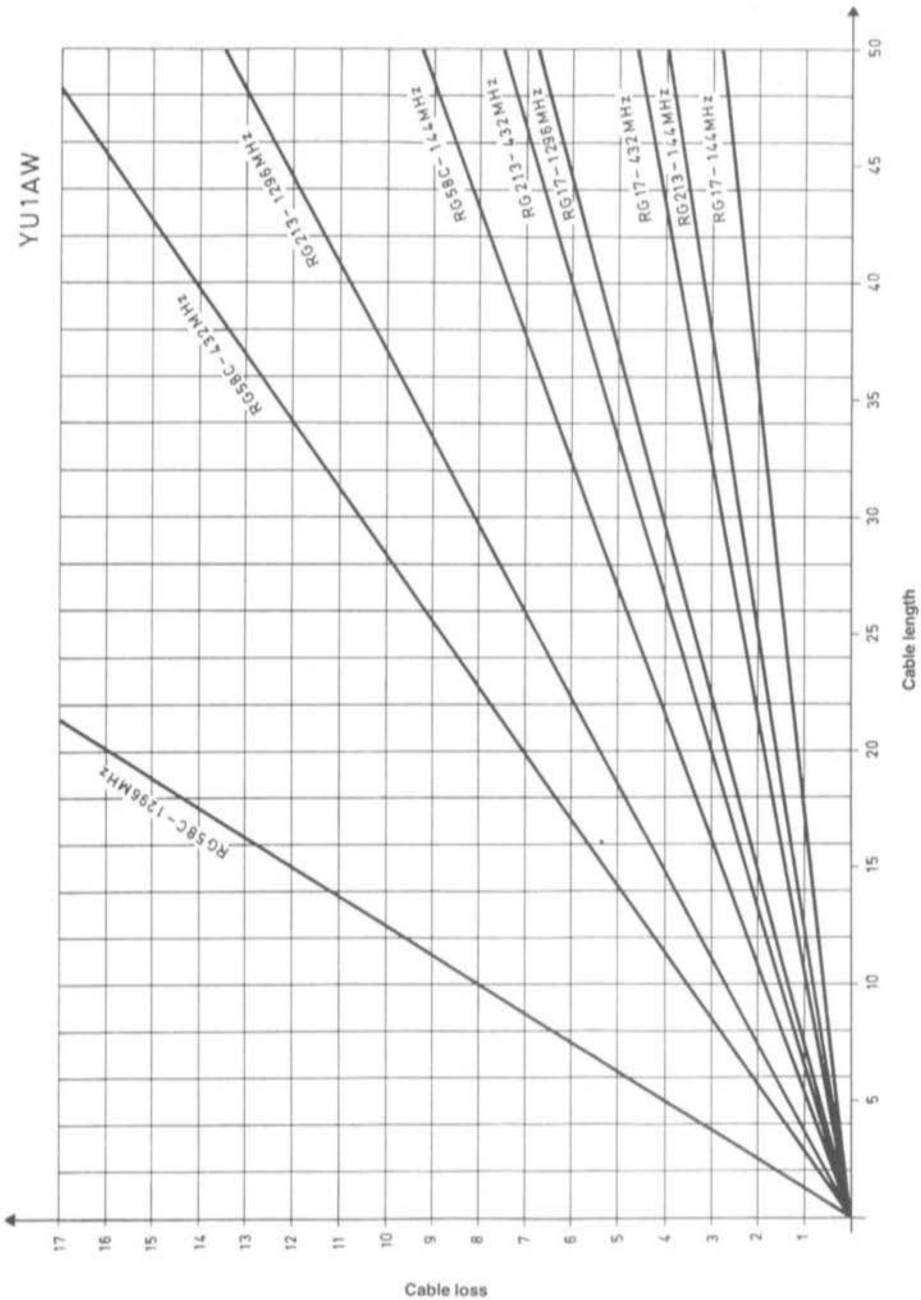


Fig. 3: Cable loss L, as a function of length for a few well-known types and the most popular VHF/UHF amateur band.

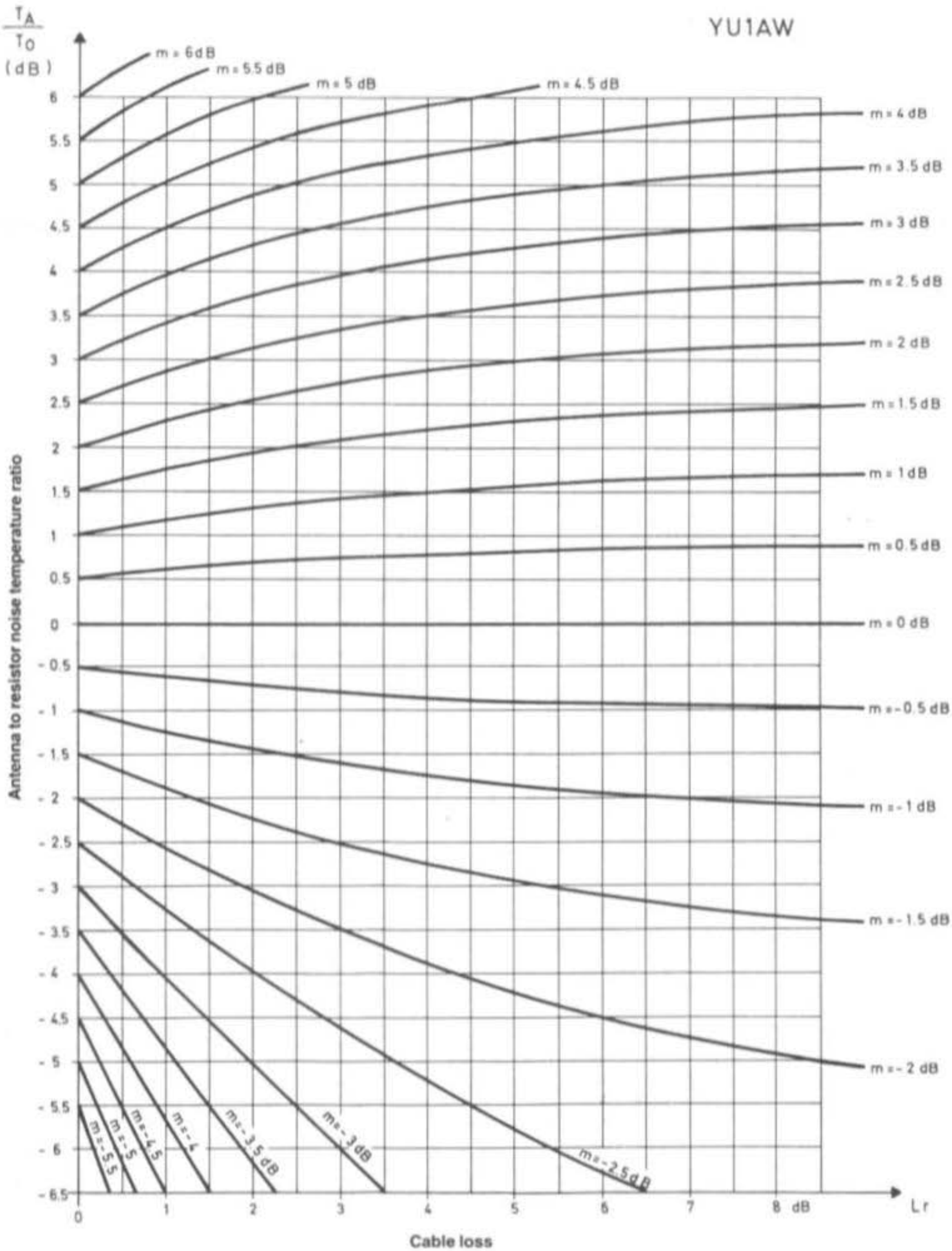


Fig. 4: Correction for the value of antenna noise-temperature T_a depending upon cable loss and the measured value of m.

The noise figure that a pre-amplifier should have for a specified case is then determined by the antenna noise (dependent upon location, antenna elevation, frequency etc.) as well as the allowance for second stage deterioration upon the overall system noise figure. But how can the antenna noise be quantified in order that a pre-amplifier noise figure may be arrived at?

5. MEASURING THE ANTENNA NOISE

The measurement of the antenna noise power is most simply accomplished by measuring the noise voltage across the receiver output and then substituting the antenna with a 50 Ω non-inductive resistor and comparing the two computed powers at 290 K, i.e. room temperature. The output voltmeter can be of any type from an electronic AC millivoltmeter to an ordinary multimeter (2) and (4).

The method is to terminate the antenna input terminals of the receiver/pre-amplifier with a 50 Ω pad of at least 10 dB attenuation, or a metal-film 50 Ω resistor or two 100 Ω MF resistors in parallel. The connecting leads, however, should be as short as possible. Connect the AC meter to the receiver AF output terminals and adjust the receiver gain for a reference voltage. Now replace the dummy 50 Ω antenna with the real thing and note the increased reading. Compute the two readings into decibels i.e.

$$m = 20 \log V2/V1$$

where V2 is the output voltage when the antenna is connected. V1 is the output voltage when the antenna input is terminated.

If there is a length of coaxial cable between the antenna and the receiver system's antenna input, the value of m must be corrected (fig 4).

With the knowledge of the type of cable, the frequency of interest and the cable's length, the cable loss can be arrived at by **fig. 3**. This cable loss (dB) is then used in **fig. 4** to obtain the corrected figure for m by seeing where the cable loss value on the X-axis intersects the computed value for m. The corrected value lies on the Y intersect expressed as a temperature ratio T_a/T_0 (dB) where T_a is the antenna noise temperature and T_0 the 50 Ω termination's noise temperature. This corrected noise temperature ratio T_a/T_0 (dB) is then used in the characteristic of **fig. 5** in order to arrive at the required system noise figure or temperature. This, of course, includes the cable loss, so to find the figure required for the receiver itself the cable loss should be subtracted from it. Now it can be assessed what measure of pre-amplification – if any – is to be provided by the first stage of the receiving system.

6. PRACTICAL CASE

Taking a practical case: Assume that the antenna noise test described above gave a result of m = 3.5 dB. The fact that the antenna noise can be measured using the receiver is proof enough that its noise figure is pretty reasonable and that it is the antenna noise which is the limiting factor.

From the X-axis of fig. 5 it can be seen that the m value of 3.5 dB corresponds to a noise temperature T_a of 650 K. The cable loss L_r (25 metres of RG 8/RG 213) can be obtained from fig. 3 as 2 dB at 144 MHz. With this, the correct value of m can be read from fig. 4, T_a/T_0 = 4.3 dB. Taking this value to the X-axis of fig. 5 gives the corresponding noise temperature of T_a = 780 K.

If it is known that the receiver itself has a noise figure NF = 3 dB, then the overall NF is 5 dB (taking into account the 2 dB cable loss) it can be seen from fig. 5 that the S/N deterioration is only 2.5 dB. It is then apparent that the greater proportion of the 780 K noise temperature is in fact urban noise and that under these circumstances the receiver has a satisfactory noise figure.

If it is required to improve the S/N ratio deterioration from 2.5 dB to 0.5 dB, the corresponding noise-figure is NF = 1.2 dB (found from fig. 5). As the cable has a 2 dB loss, an NF = 1.2 dB can only be achieved by locating a suitable pre-amplifier directly at the antenna terminals.

Taking another example: A quiet situation with low antenna noise as befits a country QTH. The antenna noise tests, this time, gave an m of 0 dB. Using 13 m of RG 213 (i.e. L_r = 1 dB) yields T_a/T_0 = 0 dB from fig. 4. If a S/N degradation of not more than 2 dB is required then the total noise figure (from fig. 5) should be 2 dB. From this must be subtracted the cable loss and the second stage degradation.

Subtracting the cable loss first leaves a receiver system noise figure of 1 dB (= 75 K). This includes the second stage contribution and this has an NF 2 of say 3.5 dB and the first stage gain G1 is 13 dB. Referring to figure 1 gives a second stage noise temperature of 18 K. Deducting this from the total receiver noise temperature i.e. 75 K - 18 K gives the required first stage noise temperature of 57 K or NF 1 = 0.8 dB. The receiving system would be better served if the pre-amplifier was placed at the antenna terminals. The second stage loss then increases by the cable loss i.e. 1 dB plus 3.5 dB = 4.5 dB (530 K). The first stage amplification of 13 dB reduces the effect of the second stage noise $\triangle T$ to 27 K.

As an overall noise figure of 2 dB (= 175 K) is required, the noise temperature that the pre-amplifier must have is 147 K i.e. NF 1 = 1.8 dB.

From this second example it may be seen that for the same output S/N deterioration, locating the pre-amplifier in the most strategic position, will limit the demands upon its performance and thereby its cost.

7. HIGHER FREQUENCIES

At higher frequencies the value for m can become negative especially when the antenna is pointed shywards. Here is an example for 432 MHz.

The measurement of antenna noise yielded a result of m = − 1.5 dB when the antenna was directed at the horizon. The cable loss (6.5 m RG 213) was 1 dB. With fig. 4 T_a / T_0 = − 1.9 dB. Fig. 5 indicates that this corresponds to a noise temperature of 185 K. The second stage noise contribution was held to 1 dB resulting in an overall noise figure of 0.7 dB (50 K).

If the receiver had a noise figure of 5 dB – typical for a proprietry amateur receiver – then the necessary pre-amplifier noise figure may be determined as follows:

With a first stage gain of 16 dB and a second stage noise figure of 5 dB and a feeder loss of 1 dB the second stage influence is found to be 22 K. It will be recalled that an overall noise figure of 0.7 dB (50 K) was required. Now taking into account the 22 K, the first stage NF 1 (i. e. the pre-amps NF) must be 0.4 dB at the antenna terminals. Placing the pre-amplifier in the station at the end of the 1 dB feeder loss would make it impossible to realise the 1 dB S / N degredation limit.

This example shows that at 430 MHz and higher, especially for space communications, the utmost in pre-amplifier noise performance is required. The reason lies in the very much reduced antenna noise at these frequencies.

8. CONCLUDING REMARKS

Pre-amplifier gains of 23 dB, which are so agressively aclaimed in amateur radio magazine adverts, have already been discussed. The receiver subjected to such a high gain input device would shatter into intermodulation distortion unless, of course, that it was used before a high loss (10 - 15 dB) feeder. Then it would represent a good solution as the 23 dB gain at the antenna would bring the overall NF to the same order as that of

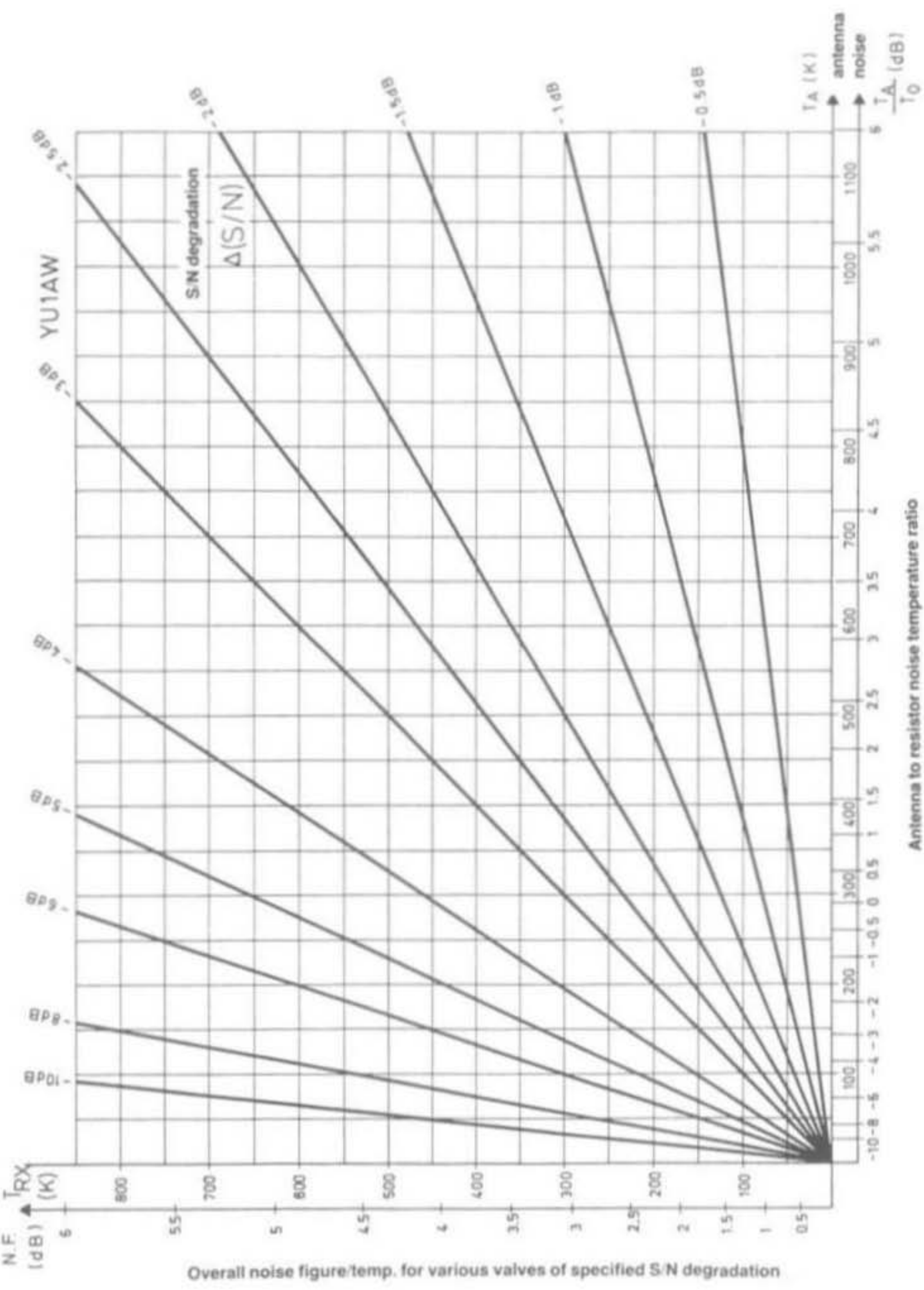


Fig. 5: Signal to noise degradation Δ(S/N) in relationship to the overall receiver's NF and the antenna noise temperature T_a

the pre-amplifier. It is the only case where such a high pre-amplifier gain is useful.

This, of course, begs the question of what such a high-loss cable is doing on an amateur station – especially during periods of transmission! The answer lies in the purchase of low-loss cable and the disposal of the 23 dB pre-amplifier.

Finally, a few more words on the measurement of antenna noise. It is important that during the measurement, none of stages are saturated with noise i. e. the whole receiving chain is working linearly. The receiver mode switch should therefore be turned to CW or SSB and not FM (limiters and saturated detectors). The AGC must be switched off and the RF gain turned down in order to avoid overloading effects. If this is not possible – as some receivers do not have the facility of RF gain control and AGC on / off switches – the results may be acceptably accurate as only low levels are involved in the measurement. The AF indicator must be sensitive to show the noise when the antenna is terminated in 50 Ω. The L. S. should be cut out of circuit and replaced with a terminating resistor (2). Local interference noises from vehicles, arc-welding and on-frequency signals should be avoided and to this end several measurements should be carried out.

The measurement error, even with a linear system, is compromised by the fact that the antenna almost never matches the receiver exactly and the antenna VSWR is higher than unity.

The autor hopes that this article will bring some enlightenment to a subject which has puzzled many amateurs over the years. Finally, the old axiom will again be recalled:
The best signal frequency pre-amplifier is the antenna!

9. APPENDIX

Only an ideal amplifier which generates no noise of its own is able to amplify without deteriorating the input S / N ratio. This is the same thing as saying that the signal-to-noise at the input is exactly the same as that at the output i. e. $S_i / N_i = S_o / N_o$. The noise factor can therefore be defined as that factor can therefore be defined as that factor by which the S / N at the output is more than that at the input of a device, i. e.

$$F = \frac{S_o / N_o}{S_i / N_i}$$

The ideal device amplifier / receiver etc. has a noise factor of unity. Actual devices all have a noise factor of greater than unity. The noise factor F allows the possibility of comparing one receiver's measured performance directly against another's as long as the measurement circumstances remain equal for both receivers. The noise factor F may be expressed in decibels in which case it becomes the noise figure NF.

$$NF = 10 \log F \text{ (dB)}$$

conversely,

$$F = 10 \text{ antilog } NF / 10$$

The noise generated within a receiver / amplifier can be expressed as an equivalent noise temperature T which is considered to be the temperature of a resistance at the input of the device which develops the same noise power at the output as the device itself – the device being considered as noiseless.

$$T = (F - 1)\ 290\ K$$

or

$$T = (10 \text{ antilog } NF / 10 - 1)\ 290\ K$$

Since the noise factor F is referred to the thermal noise at room temperature, i. e. 290 K, then by definition it may be written:

$$F = 1 + T / 290$$

or

$$NF = 10 \log (1 + T / 290) \text{ (dB)}$$

The (fictitious) ideal receiver has an equivalent noise temperature T = 0 K, all other receivers have a noise temperature of greater than 0 K.

10. LITERATURE

(1) D. Dobričič, YU 1 AW:
Determining the Parameters of a Receive System in Conjuction with Cosmic Radio Sources
VHF COMMUNICATIONS, Vol. 16, Ed. 1/1984, P. 35 - 50

(2) D. Dobričič, YU 1 AW:
Amaterske radio komunikacije preko Meseca.
Radio-amater 12/1979

(3) Mitsubishi Application Notes for GaAs-FETs

(4) J. Gannaway, G 3 YGF/D. Holmes, G 4 FZZ:
Some Pittfalls in Noise Figure Measurements
VHF COMMUNICATIONS, Vol. 13, Ed 1/1982, P. 44 - 48

Jochen Jirmann, DB 1 NV

A Spectrum Analyser for the Radio Amateur Part 2 – Concluding

5. THE FIRST LOCAL OSCILLATOR

This oscillator, tuning from 450 MHz to 1000 MHz, has already been described in VHF Communications 4/86 and therefore the details here can be held to a minimum. In this version, shown in **fig. 5.1.**, the oscillator transistor is a BFR 91 and the buffer amplifier an NPN transistor BFG 96, the latter delivering at least 10 mW drive to the following SRA 220 ring mixer. An auxiliary buffer output of 0.5 mW is taken from the twin-hole cored transformer and used for the tuning linearity circuits and for the PLL drive. The components are soldered to the track side of the printed crcuit board, the arrangement being shown in **fig. 5.2.**

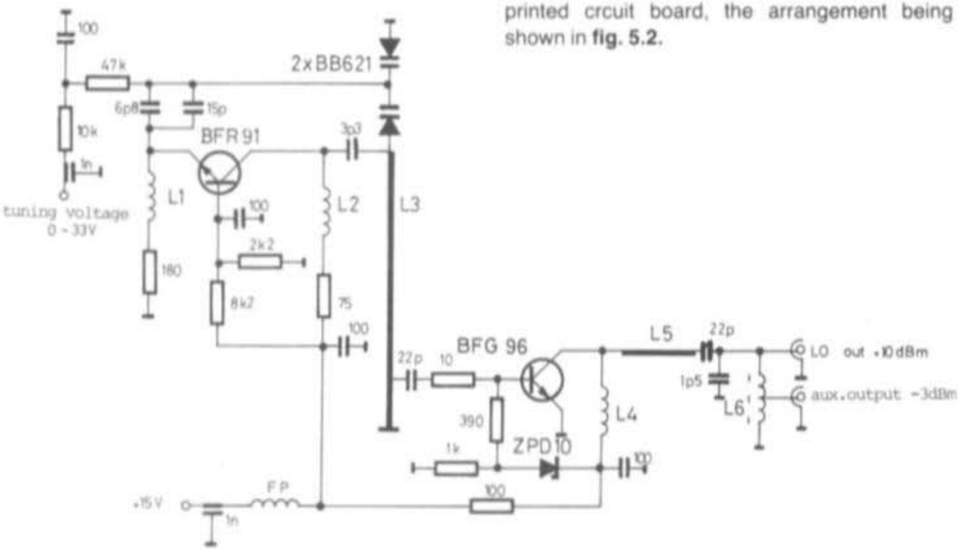


Fig. 5.1.: The first local oscillator having a 450 MHz to 1000 MHz tuning range

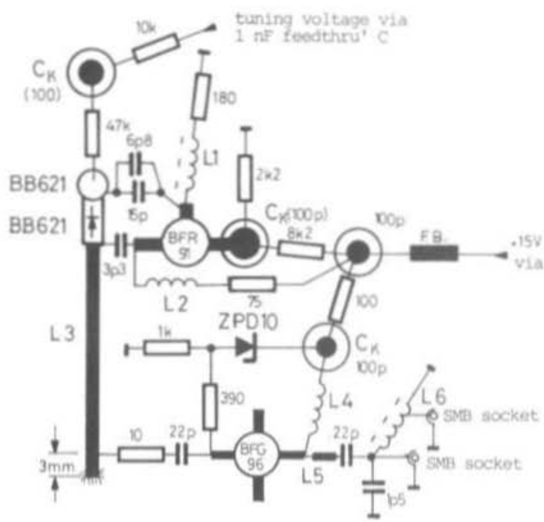


Fig. 5.2.:
The component arrangement of tbe first local oscillator

6. FREQUENCY CONTROL LOOP AND TUNING LINEARITY

The relationship between the tuning voltage and the output frequency of a diode-tuned oscillator normally approximates a square root fuction. In order to obtain a linear relationship between tuning voltage and frequency, it is necessary to employ a resistance-diode network possessing an inverse characteristic, which, when combined with the VCO characteristic, produces the desired linear relationship. The only snag with this arrangement is that each oscillator must be individually adjusted for optimum results.

The employment of a frequency control loop is an alternative method of obtaining the linear relationship between tuning voltage and VCO frequency. A highly linear discriminator is fed from the output of the VCO and delivers information concerning the actual frequency of the VCO. A control amplifier then causes the VCO to return to the set frequency. The disadvantage of this

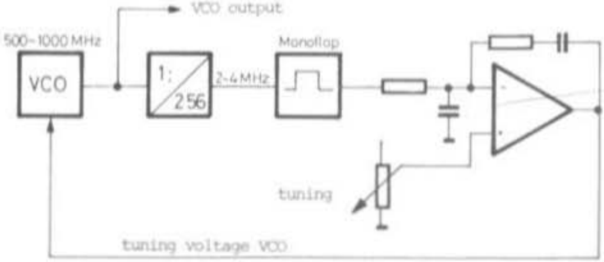


Fig. 6.1.:
Frequency control loop used to linearise the 1st LO's tuning characteristic

method is that the frequency control loop has a finite reaction thus setting a definate time limit to the tuning rate. The diode function generator, on the other hand, suffers no such disadvantage. For the control loop method, a suitable frequency discriminator may be realised using a monostable multivibrator. For every period of input voltage, it delivers an output of constant width and amplitude. The average period of the output pulses determines the output frequency.

As this discriminator can only work up to frequencies of only a few megahertz, the output from the VCO is scaled down with the aid of a simple ECL divider of the type frequently used in commercial TV equipment. The block diagram of the frequency control loop, complete with scaler, is shown in **fig. 6.1.**

The prototype frequency control circuit used by the author is shown in **fig. 6.2.** The VCO main + 10 dBm output goes on to the PLL circuitry and will be dealt with later. The − 3 dBm output from the VCO is fed to a resistive power splitter. It is then buffered by a BFW 92 amplifier before being taken to the SDA 4211 ECL scaler. Following a frequency division by a factor of 256 and a level shift to CMOS, a signal directly derived from the VCO in the range 2 MHz to 4 MHz is available.

The frequency-voltage transducer consists of a monostable 74 HC 123 whose time constants are controlled by a metal-film resistor and an NPO capacitor. The monostable's stability determines the re-setting accuracy and the long-term stability of the frequency control. Since the amplitude of the impulse peaks are determined by the 5 V supply voltage, the high-grade regulator LM 723 is used. This is an order better than the widely used three terminal regulators but the use of a precision regulator such as the REF 02 is not considered to be justified for this application.

The RC network tends to smooth the monostable's output pulses somewhat before they are fed to a low-noise NE 5534 IC amplifier, connected as an integrator or as a PI regulator, which compares the actual with the reference frequency. A transistor amplifier raises the level to that required by the VCO's control voltage

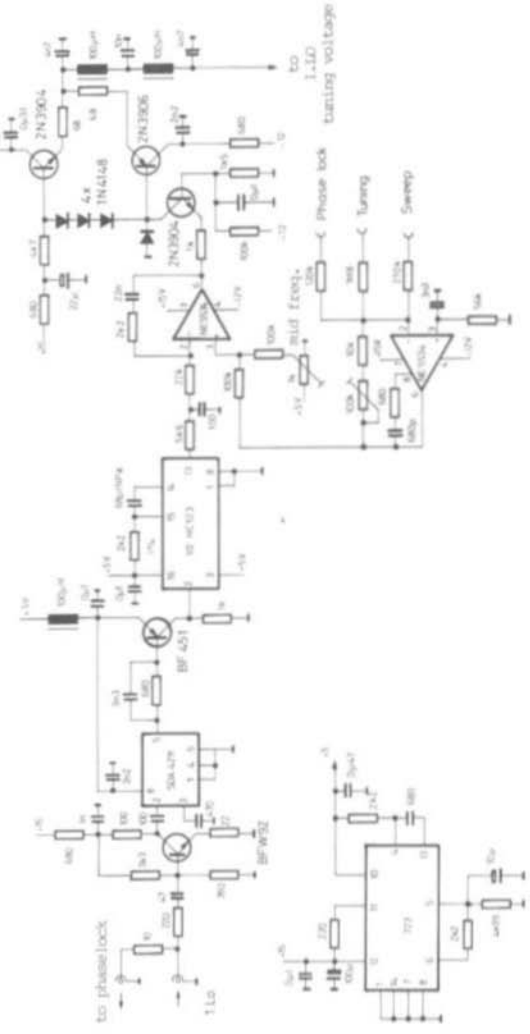


Fig. 6.2.: The first LO's frequency control loop

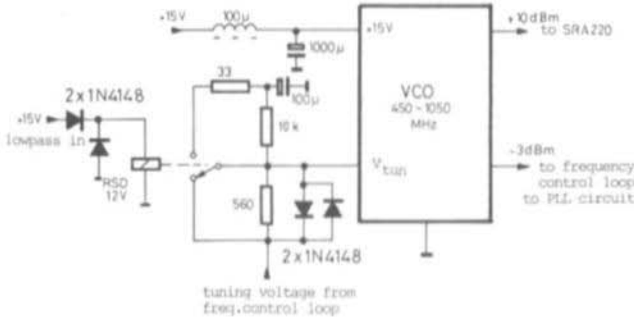


Fig. 6.3.:
The external circuitry of the 1st LO

0 – 30 V and an LC network filters out the residual RF.

Immediately before being applied to the VCO control input, the tuning voltage may be additionally filtered by a relay selected RC network (**fig. 6.3.**). This extra filtering, although improving the noise superimposed on the control voltage, also increases the control loop reaction-time enormously. For this reason it is only activated when the spectrum analyser is being swept at narrow band (200 kHz/div.) from the 2nd local oscillator. The detail is shown in fig. 6.3. The reference frequency for the loop is represented by a weighted sum of the tuning potentiometer voltage, the sweep voltage and the tuning voltage from the PLL circuitry. A second NE 5534 is used here as the summing amplifier. Two multi-turn trimmers adjust the centre frequency and tuning sensitivity.

With the dimensioning as suggested here, the frequency control loop has a 200 Hz bandwidth. This allows a loop reaction time to tuning voltage variations of 1.5 ms. The VCO frequency stability is better than 100 kHz with this control loop and it is, in most cases, largely dependent upon the externally generated tuning voltage.

A static measurement of linearity using a frequency counter and a digital voltmeter reveals practically no errors. The VCO exhibits a residual modulation, i. e. from hum and LF noise, of some 30 kHz peak. Only on the filter bandwidth from 1 to 2 kHz can it be observed that a better VCO stability would be desirable. This theme is persued in the following chapter which introduces a PLL circuit which may be switched to sweep presentations of 200 kHz / div. or less.

7. THE PHASE LOCKED LOOP

When swept displays of under 200 kHz / div. are required, the spectrum analyser presented here sweeps the 2nd LO at 460 MHz whilst the 1st LO is free-running. In order to reduce the effects of the 1st LO's noise modulation, the oscillator is controlled from a crystal frequency reference. The author attempted to use a television receiver PLL synthesizer for this purpose but it proved unsuitable. Test measurements showed that the noise from a television synthesizer was greater than that from the frequency control loop described above. The trouble lies in the large division factor of 64 or 256 in the TV tuner and its inherently low phase comparator frequency.

The PLL circuit is therefore so dimensioned that the smallest possible VCO division factor is used.

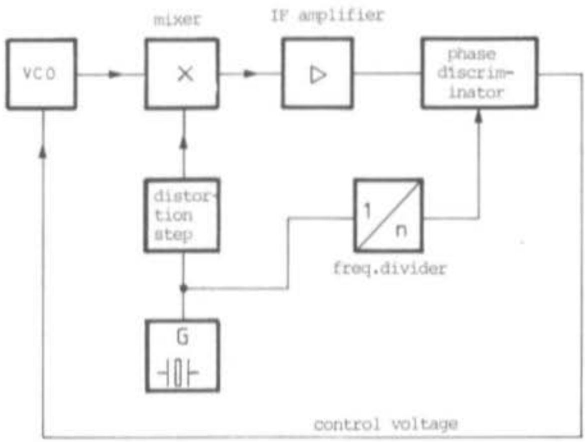


Fig. 7.1: 1st LO PLL block diagram

In the interests of easier operation, a control technique, which allows manual tuning between fixed spot frequencies, is to be preferred. One possibility is a PLL circuit with harmonic mixing as shown in block diagram (**fig. 7.1.**). The VCO signal is mixed with a harmonic-rich crystal oscillator. The resulting intermediate frequency is amplified and taken to a phase / frequency comparator which controls the VCO tuning voltage.

A problem crops up here with the production of the harmonic signals as the spectrum must extend well over the 1 GHz region. The author found that an harmonic spectrum up to about 300 MHz could be achieved relatively easily by using the gate times of TTL circuits. This offers the possibility of dividing the VCO frequency to under 300 MHz and then to arrange the harmonic mixing.

The final PLL circuit is shown in **fig. 7.2.** It may be seen that the input again possesses a resistive divider which passes the greater proportion of the input power to the LO output on the front panel. A smaller portion is amplified by a two-stage transistor (BFR 15) amplifier and taken to the CA 3199 ECL scaler.

The price of integrated UHF circuits has fallen recently (to less than 10 DM) making their use almost mandatory. For a start, the Mini-Circuits MAR-8 is suggested. The input signal is divided by 4 in a reasonable priced (6 DM) CA 3199 frequency scaler which is capable of working up to 1250 MHz. The output from the divider is taken via a low-pass filter to a mixer which is arranged in a sample-hold FET circuit.

A CD 4060 crystal oscillator circuit working at 4 MHz is divided by 16 and presents a 256 kHz square wave into positive going needle pulses about 2 ns wide. The needle pulses gate the FET switch. A further division of the crystal frequency produces a 32 kHz reference for the frequency / phase comparator.

The IF signal is taken at high impedance from the mixer by a FET buffer and then on to the CD 4046 frequency / phase comparator via a 50 kHz low-pass filter. A control voltage from the CD 4046 pin 13, is eventually taken to the PLL after being filtered and buffered. A BF 246 FET switch passes the PLL control voltage on to the phase-lock input of the frequency control loop (**fig. 6.2.**). The 500 Ω preset is adjusted to optimise the PLL sensitivity. A lock-detect signal is also available from the CD 4046 pin 1. This is made to drive an LED located on the front panel, marked "LOCK", and signals the correct functioning of the PLL.

When initially switched on, it was found that the PLL did not lock at or near the selected spot fre-

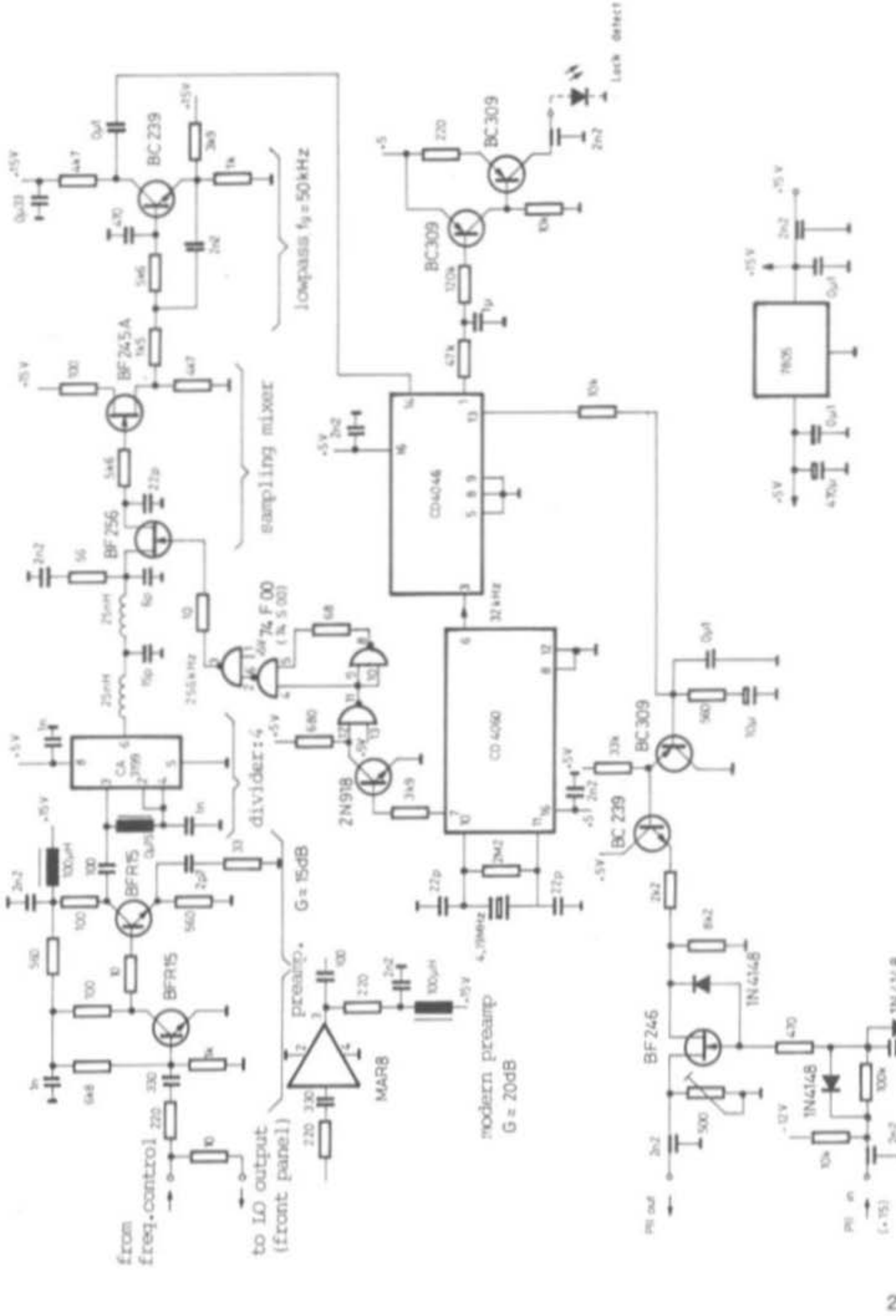


Fig. 7.2.: Detailed circuit diagram of 1st LO PLL

quency but locked at some other random frequency far removed. This error was corrected by arranging the PLL to turn on gradually upon switch-on thus avoiding switching transients. This was accomplished by including an RC network in the gate circuit of the BF 246 switching FET. This has the effect of gradually raising the gate voltage upon switch-on and thereby the PLL loop gain.

Using the circuit proposed, here the frequency-comb, spot-frequencies are spaced by 1 MHz. This presents no problems for the fine tuning between spot frequencies; this being done at 5 MHz by a second oscillator. The frequency stability of the spectrum analyser is only determined by the 1st LO's reference crystal stability and the drift from the fine-tune, second oscillator. When thermally stabilized, the frequency drift of the analyser is less than 10 kHz. In operation, the analyser is first switched to a sweep display of 200 kHz per division, at about the frequency of interest, and then the PLL is switched on. By careful correction of the main tuning, the PLL is brought into lock as indicated by the illumination of the "lock" LED. Now, the desired signal can be fine-tuned and the sweep width further reduced for a narrow-band display. The main tuning should not then be altered otherwise the 1st LO will lock at a neighbouring frequency and the display lost. If a display of 500 kHz per division, or greater, is required the PLL is switched off automatically.

8. OTHER MODULES

A few other circuit details must be mentioned in order to complete the description of a practical spectrum analyser. Some modules e. g. the power supply delivering + 15 V, + 35 V, − 12 V, − 24 V, the saw-tooth generator for the sweep voltage and deflection amplifiers have been deliberately omitted as these items are heavily dependent upon the type of display tube which has been utilised.

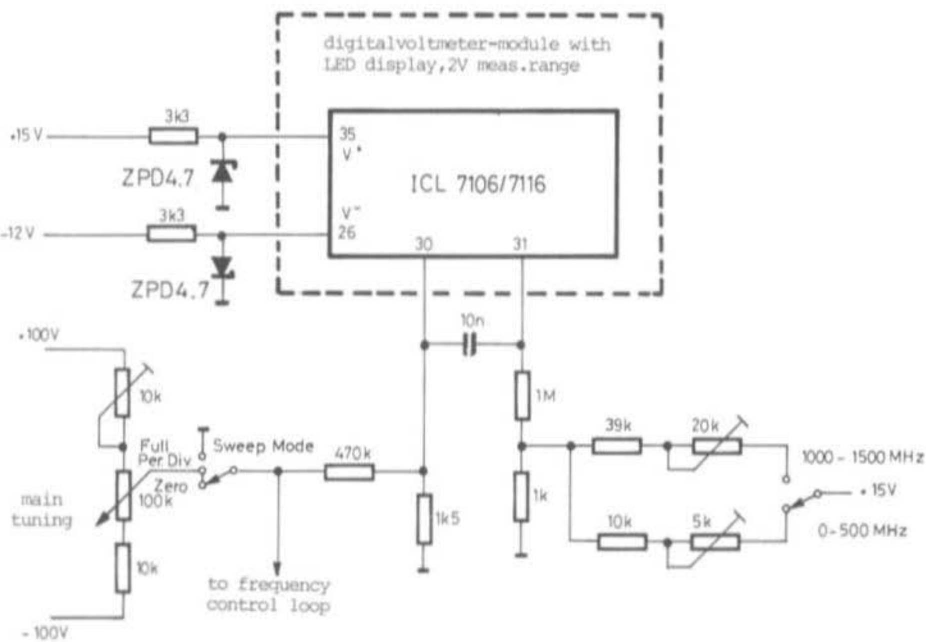


Fig. 8.1.: Arrangement of the tuning voltage and frequency display

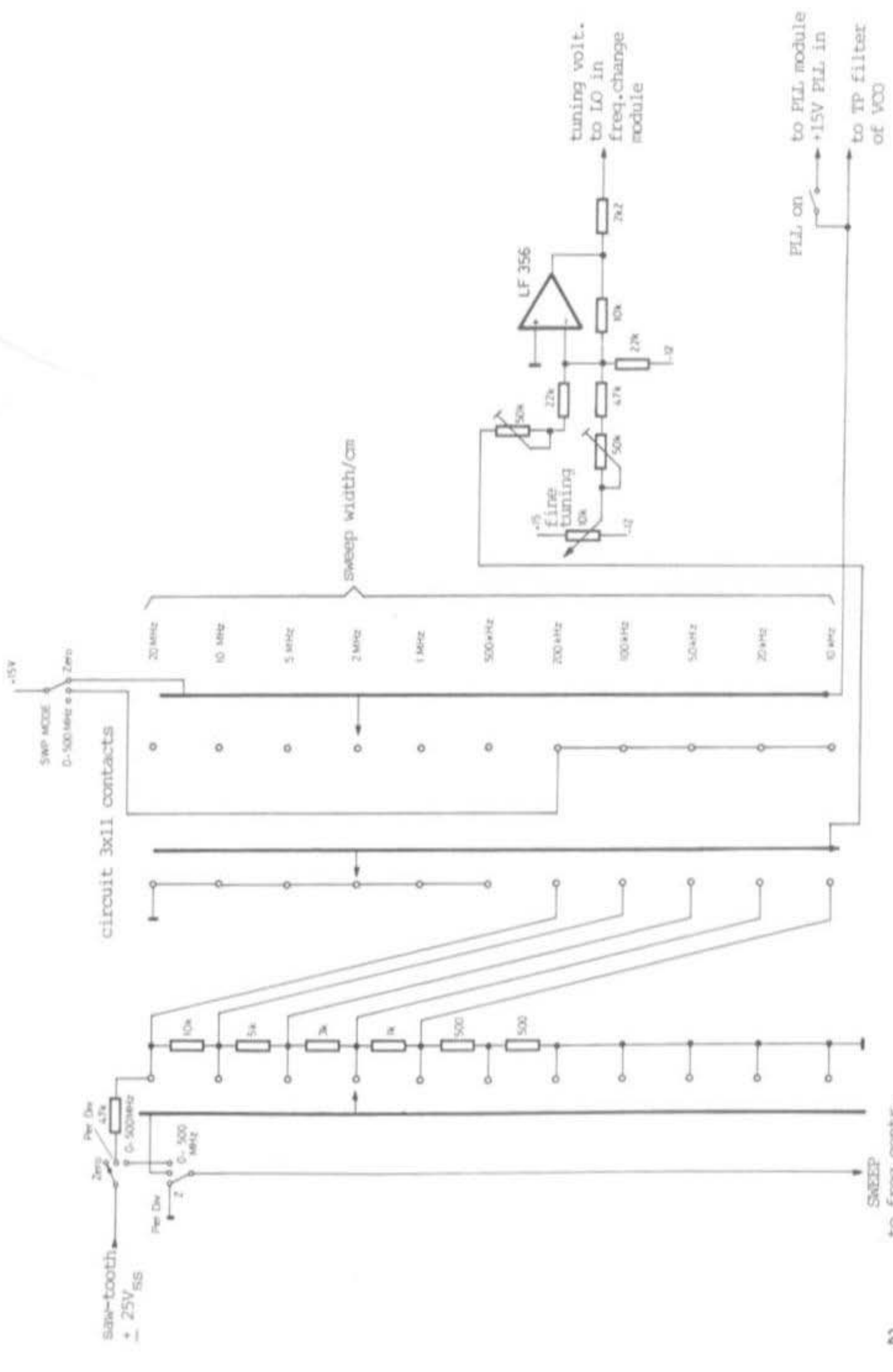


Fig. 8.2.: The switching logic for the tuning voltages of the 1st and 2nd LOs

The circuit details to be considered now are the production of the tuning-voltage supply, the digital frequency indicator and the support module for the display sweep-width. A 100 kΩ, 10-turn potentiometer serves as the tuning potentiometer. The tuning voltage, in the author's version was drawn from a highly stable ± 100 V supply which was available in the instrument. This voltage range is not strictly necessary as the tuning range can be altered by suitable dimensioning of the summing resistors in the frequency control loop circuit to suit the available tuning voltage supply to the tuning potentiometer. The main thing to watch, is that the supply itself is absolutely stable and free of hum and noise.

A digital voltmeter can be used as the frequency display which is actually measuring the tuning voltage. The author used a 3 1/2 digit voltmeter with an LCD display and a 2 volt range. This used the well-known CMOS module ICL 7106 or ICL 7116 by GE-Intersil. This unit is equipped with a floating differential indicator and has a consumption of only 2 mA! As opposed to the usual 9 V battery supply, a mains derived source is used in a full-wave circuit to two Z-diodes delivering ± 4.7 V w.r.t. ground.

When constructing this item, or using a modified commercial digital voltmeter, two points should be considered. If the module originally used 9 V batteries then the minus (pin 30) is strapped to ground (pin 32). This connection is to be interrupted.

This IC sometimes exhibits problems with common-mode voltages which drive the integrator into a premature run-down condition. This has the effect of the display indicating up to a certain input voltage and then saturates, or sticks, at that indication. This effect can be countered by increasing the value of the integration resistance at pin 28 or increasing the value of the integration capacitor at pin 27.

Figure 8.1. shows the external circuitry used by the author to support the ICL 7116 in the area of the signal input. By means of a selector switch, either one of two receive input ranges may be selected 0 to 500 MHz or 1000 to 1500 MHz and correctly indicated. Operating this switch can also be caused to switch in the appropriate high and low-pass filters before the first mixer, but the author did'nt do it in this manner in the prototype instrument. The frequency indication has an accuracy of a few MHz which is sufficient for most purposes. If a higher accuracy is desired, it can be calibrated with a spectrum generator.

The sweep voltage, employed to wobbulate the oscillators of the spectrum analyser, is a symmetrical about zero, saw-toothed wave-form of 25 V peak to peak. The frequency of the waveform is variable between 2 Hz and 25 Hz. The generation of the various scan widths is given in **fig. 8.2.** By means of a stepped divider sweep widths of 10 kHz/div. and 20 MHz/div., in a 1-2-5 sequence, may be selected. For scan widths over 500 kHz/div., the 1st oscillator is deviated with an attenuated portion of the sweep voltage applied to the frequency control loop circuit. When smaller scan widths are selected, only the second oscillator is wobbulated. Also, the sweep-voltage and the voltage from the fine-tune potentiometer are summed in an operational amplifier and taken to the 2nd LO's varicap diodes in the IF converter module. A correction of the tuning characteristic was found to be unnecessary. In this mode of operation, the control speed of the frequency regulator loop is reduced in order to attenuate the 1st LO's noise modulation. In addition, the PLL stabilization can be switched in.

By means of a further switch a display over the whole 500 MHz range can be selected – the mid-frequency being fixed. Another facility is that the whole wobbulation system can be switched off in order to observe the modulation on a single signal.

9. CONSTRUCTIONAL HINTS

There are many and various ways by which the described modules may be connected together to form a complete spectrum analyser. The most elaborate way is to make the display and power supplies in order to build a self-contained instrument. Life is also made easier when a ready-to-hand oscilloscope can be pressed into service for the display. The oscilloscope's time-base can then be used for the spectrum analyser's saw-tooth waveform requirement. Most oscilloscopes have an external connector for the time-base voltage but if not, one can be easily fitted. Also, with any luck, the oscilloscope's power supplies will also supply the analyser's power requirements as well.

The owners of an oscilloscope having a plug-in unit capability are even better off. One of the plug-in units can be stripped and the chassis used to build the spectrum analyser circuits. The oscilloscope can then be used for its original purpose, or for the spectrum analyser facility, as desired. The author chose this method using a Hewlett Packard 140 A oscilloscope main frame and one of the little used plug-in options being turned into an analyser plug-in. The main frame could deliver + 250 V, + 100 V, − 12 V and − 100 V. The + 35 V and − 24 V supplies required by the analyser were easily derived from the +/− 100 V oscilloscope supply, using resistors and zenner diodes.

The + 15 V supply had, however, to be a little more elaborate because of the 500 mA consumption. One possibility, offered by all older plug-in oscilloscopes, is the employment of the 6.3 VAC heater voltage which is brought out at the main plug for the plug-in unit. Using a small laminated transformer, a full-wave rectifier and a three-terminal regulator, the necessary voltage may be made available.

The HP 140 does not have a deflection amplifier, i.e. the deflection card of the CRT lies directly at the plug-in's plug. Two deflection amplifiers are necessary therefore but the bandwidth need only be about 50 kHz. The author used for this purpose a ready-to-hand board which also contained a saw-tooth generator and a blanking amplifier which suppressed the trace return.

It is not necessary to go into further details as they are largely dependent upon the type of display device used by the author. Other equipment must use an individual method for utilising it to combine with a home-made spectrum analyser. Before attempting to modify an oscilloscope for this purpose, it is recommended that the oscilloscope handbook is given a thorough persual. Details on deflection amplifiers may be found in

Fig. 9.1.: The author's spectrum analyser

volume 4 of the series "Professional Electronics" published by Francis Verlag.
The author's prototype instrument is shown in **fig. 9.1.**

10. ADDITIONS

The spectrum analyser as described may be given a few further facilities. A few suggestions are given as follows:

– An additional VCO tuning from 1000 MHz to 1500 MHz will enable the input range to be extended from 500 MHz to 1000 MHz. The ECL divider, in the frequency control loop and in the PLL, are well able to work at these frequencies as tests have shown.

– A frequency converter or an external harmonic mixer for the microwave bands can be connected.

– A sweep generator having a 60 dB dynamic range and a tuning range of 0 – 500 MHz may be made by synchronising another oscillator so that it is swept in step with the spectrum analyser's 1st LO.

– Using high and low-pass filters, automatically switched with the input range switch, reduces the possibility of receiving spurious signals in the display trace.

– The use of a better, i.e. noise reduced, local oscillator will improve the instrument's dynamic range. This is extremely important when observing a number of closely arrayed signals with large level differences between them.

It may be seen that the spectrum analyser offers many possibilities for individual expansion schemes and for experimentation for improved performance with the basic circuits. The purpose of this article, as stated at the start, was to generate the necessary interest to begin a project of this kind and also to dispel the unfounded fears attending its complexity.

Dragoslav Dobričič, YU 1 AW

Low-Noise 144 MHz Pre-Amplifier Using Helical Tuned Circuits

The introduction of field effect transistors (FETS), especially those made from gallium-arsenide (GaAs), in low-noise pre-amplifiers was a significant step forward in a field hitherto dominated by parametric and MASER amplifiers. With the advent of EME and satellite communications, amateurs are beginning to improve their receiving systems with pre-amplifiers. Before, however, low-noise transistors are employed for this purpose, there are one or two things which have to be taken into account.

It is widely believed that it is sufficient just to bring a GaAs-FET into operation in order to obtain the noise figure published in the databook for the device. The reality is, however, a little different owing to the losses in the input tuned circuits. These losses have the effect of directly increasing the noise figure of the amplifier. Even using an excellent transistor, the pre-amplifier can be rendered worse than useless if due attention is not paid to the input circuit losses.

1. A LITTLE THEORY

A transistor, intended for low-noise pre-amplifiers, can only achieve the specified noise figure if certain working conditions are exactly adhered to. First of all, the important parameters, drain-source voltage (U_{DS}), gate-2 source voltage, (U_{G2S}) and drain current (I_D) should all be precisely defined. How important, above all, the drain current is can be seen in **fig. 1**.

A further important condition is that the transistor input "sees" a definate impedance Z_{NF} at which its noise is at a minimum. With most FETs, and especially with GaAs-FETs, this impedance Z_{NF} differs widely from that at which the transistor

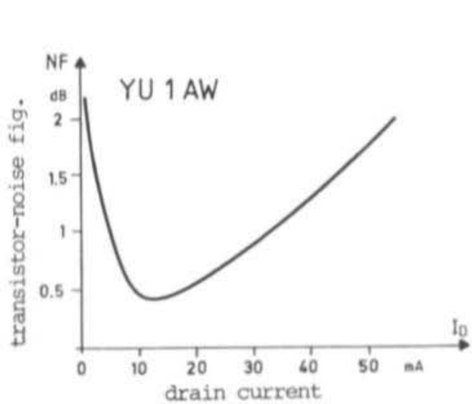


Fig. 1: FET noise figure as function of drain current I_D

transistor type	loss min. (dB)	L 1 turn	C_c pF	C_t pF	U_b V	U_{DS} V	U_{G2S} V	I_D mA	R_s Ω	R 1 Ω	R 2 Ω	R_D Ω
BF 981	0.1	4	4.7	2	12	10	4	10	10 - 50	10 k	15 k	150
MGF 1200 MGF 1400	0.15	5	2.7	3	5	3	–	10	* 100	–	–	100
CFY 13 CFY 14	0.15	5	2.7	3	5	4	–	10	* 100	–	–	10
CF 300	0.36	5	1.6	2	6	5	2	10	* 100	10 k	15 k	10

Table 1: Element values for pre-amplifiers

* approximate value (see text)

delivers a maximum power transfer. That is, the noise matching reduces the transistor's amplification but the lower noise figure is by far the highest consideration. A high amplification, indeed, can be positively disadvantageous (1).

The setting of the correct transistor working potentials and currents is normally no real problem, but a few things have to be observed. The problem of arranging the gate to look into an impedance Z_{NF} is more difficult because of the

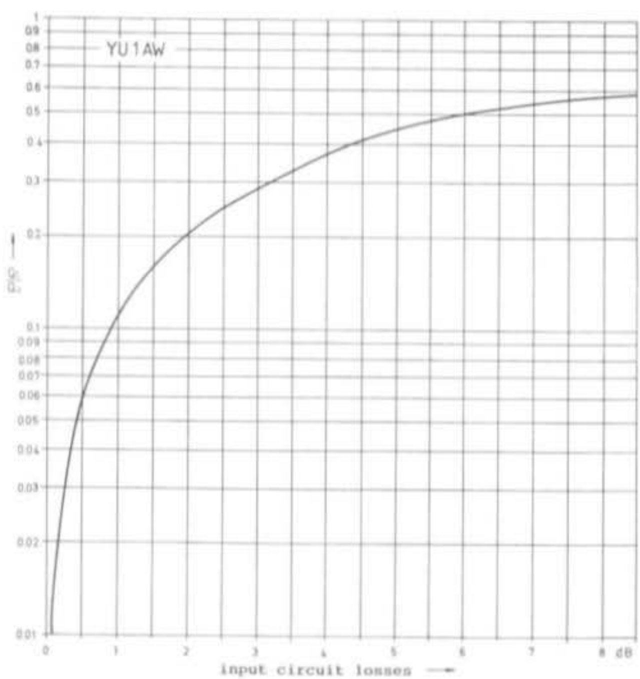


Fig. 2:
Circuit losses (dB) as function of Q_l / Q_u ratio

transistor	MGF 1200				BF 981					CF 300		
input circuit	1.	2.	3.	4.	5.	6.	7.	8.	9.	10.	11.	12.
Q_u	307	709	225	113	307	709	225	213	165	307	709	154
Q_l	11.2	11.2	17.1	17.1	4.5	4.5	4.5	4.5	6.3	29	29	29
min. loss (dB)	0.24	**0.14**	0.69	**1.43**	0.13	**0.06**	0.18	0.19	0.34	0.86	0.36	**1.8**

Table 2: Comparative data for different published pre-amplifiers

losses entailed in transforming from the antenna impedance (normally 50 Ω) to Z_{NF}.
The losses L of a matching circuit are determined by the relationship of the loaded Q (Q_l) and the unloaded Q (Q_u).

$$L = 10 \log \left(\frac{1}{1 - Q_l / Q_u} \right)^2$$

It may be seen from this formula that the losses are smallest when Q_l / Q_u is as small as possible, i.e. Q_l must be as small as possible and Q_u must be as high as possible (**fig. 2**)

Q_l and Q_u are determined from physical and technical characteristics. By the correct choice of circuit parameters and a careful construction, the optimum value for Q_l and Q_u can be achieved and thereby the losses for a particular application minimised.

Q_l should then be as small as possible. The lower limit is determined by the transformation ratio $n = Z_{ant}/Z_{NF}$. The minimal value for Q_l can then be expressed as: –

$$Q_l\,(\min) = \sqrt{n - 1}$$

Sometimes the value required for Z_{NF} entails a value for Q_l which is above the minimal value according to the above formula. This value of Q is determined by the ratio of imaginary to real parts of the impedance Z_{NF}: –

$$Q_d = X_{NF}/R_{NF}$$

The important consideration when designing pre-amplifier input circuits is that the value for Q_l is determined by that required for correct noise impedance match and will be generally higher than Q_d or Q_l (min). The increasing of Q_l owing to Q_d would explain why many transistors work better in certain frequency bands than in others. The correct choice of transistor is determined then, upon the Q_d at the working frequency. For example, on account of the large relationship of Q_d at low frequencies (below 1 or 2 GHz), the majority of the well-known GaAsFETs are ineffective in the 2 metre band. A larger value of Q_d, and with it Q_l, increases the input circuit losses such that the minimal value for noise-figure cannot be attained. The result can be worse in practice than using much cheaper transistors (**table 2**).

Q_u depends upon the circuit itself, its arrangement, the number, type and manner of operation of the components used. Q_u is higher if the individual components employed also have a high Q, that is, they are of high quality and the component-count is kept to a minimum.

Intrinsic losses and radiation losses are two factors which must be kept under control. In most proprietry pre-amplifiers, the inductors are mounted upon a printed circuit board and usually have significant radiation losses. Good insulation materials are a must. Air is, without doubt, number one, PTFE (teflon) and high-frequency ceramic can also be recommended.

The resistance of conducting materials must be held to as low a level as possible in order to minimise losses by skin-effect. For this reason, higher-conductive metals, such as copper and silver, are used and also metals of a non-oxidising nature (gold, platinum). These measures help to achieve high-Q values in circuit elements.

At lower frequencies, the inductor losses are the chief cause of the problem as they, in practice, determine the Q_u of the resonant circuit. At higher frequencies, the losses are considerably reduced by the use of strip-lines and coaxial resonators which mean higher values for Q_u.

Also the relationship between inductance and capacitance (L-C ratio) influences the value of Q_u. Using a higher value of inductance and a lower value of capacitance for a given resonant frequency, means that the unloaded Q (Q_u) is higher.

The minimum number of elements, required to undertake an impedance transformation, is two. The circuit is known as an "L section" network and consists of a capacitor and an inductor. This circuit has the disadvantage that the loaded Q (Q_l) cannot be independently variable but is fixed for a given transformation ratio.

Three elements in a transformer circuit, allow n and Q_l to be selectable independently from one another and within certain limits, may be varied thus making the tuning easier. From the many possible circuits using three elements, the Pi and the parallel tuned circuits are the ones most frequently used.

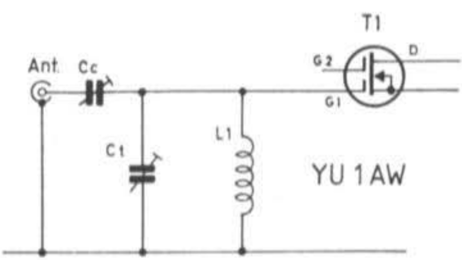


Fig. 3: Parallel input circuit with capacitive loading control

Additional losses can occur owing to the chokes and resistances used for the working point setting of the pre-amplifier. This applies also to the capacitors for coupling and decoupling. For this reason it is best to incorporate tuned and transformation circuits into the active devices' electrode supply leads.

Bearing these points in mind, the input circuit of **fig. 3**, using a capacitive coupling to the antenna, may be regarded as satisfactorily fulfilling these requirements. Using striplines or coaxial lines, instead of coils, results in an increased Q_u but at 144 MHz, the line would be more than 30 cm long and therefore unpractical. Normal coiled inductors have a Q which is directly proportional to the coil diameter and the square root of the frequency for a given inductance. This means having fewer turns but an unwieldly large diameter to achieve the necessary Q_u.

The solution is the helical circuit, a kind of coaxial resonator with a spiral-formed inner conductor (2, 3). This arrangement results in a decreased propagation factor for the circuit, and thereby the length of the $\lambda/4$ or $\lambda/2$ line element, quite considerably. At the same time, all the advantages concerning low losses are retained thereby endowing the helical tuned circuit with a Q which is many factors higher than a coil tuned circuit of similar dimensions. Also, as shown in fig. 2, the losses are considerably lower with a reduction in the relationship Q_l/Q_u.

Helical circuits are employed in proprietary receivers but either in an actual or a conceptually incorrect manner. Conceptually: the input circuit was designed for a high selectivity before the first transistor was mostly using a three-stage helical filter with loose coupling to obtain a high Q_l and a narrow bandwidth. Because Q_l was nearly as high as Q_u, the filter exhibits a high loss (8 to 12 dB) and the resulting noise figure was miserable, as a series of tests have confirmed. These filters were therefore removed in order to

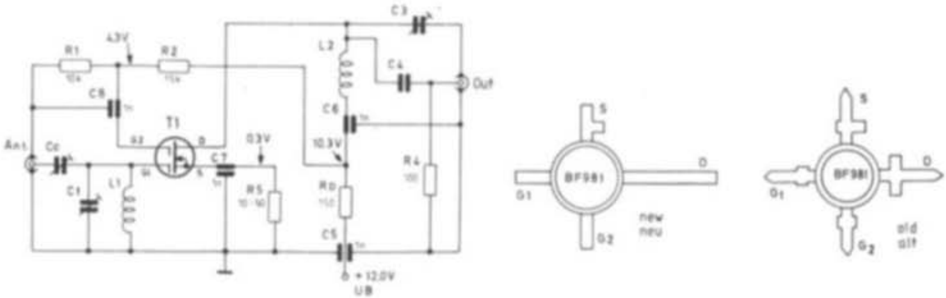


Fig. 4: Circuit schematic of pre-amplifier using BF 981

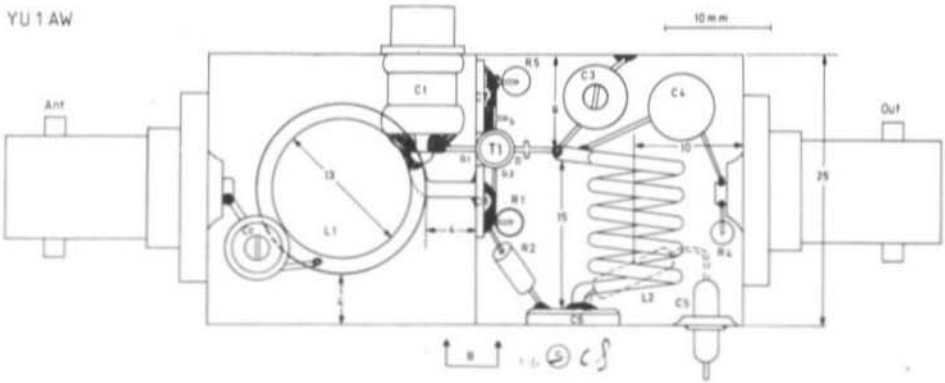


Fig. 5: Plan-view [A-A] of pre-amplifier constructed from fig. 4

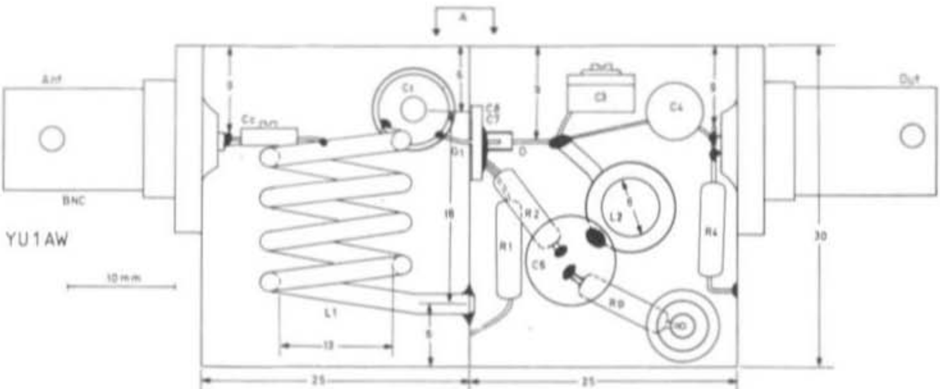


Fig. 6: Side-view [B-B] of pre-amplifier

bring the noise figure down to an acceptable value of 3 dB – normal for the transistors used.

In the pre-amplifier to be described, the helical circuits are employed in quite a different fashion. The ratio Q_l/Q_u is made small by using a large value for Q_u and thereby sharply reducing the losses. In other words, the replacing of a coil by a helical tuned circuit and arranging for the Q_l to be the same, the ratio of Q_l/Q_u is more favourable. The improvement thereby obtained is considerable and can be seen in the results of **table 2**.

The construction and adjustment of helical circuits are really simple and, if care is taken over the relationship of the geometric dimensions, not at all complicated. This is important because the desired results are only achieved when the construction is easily reproduceable.

The author designed the pre-amplifier for his EME-station. The GaAs-FETs MGF 1200 and CFY 13, as well as the MOSFET BF 981, are employed in circuits which were computed with the aid of a computer program "YU 1 AW MATCHNET". All these circuits were superior to previously published designs and confirmed the theoretical considerations outlined in this article. The input circuit losses were reduced by about 0.1 dB.

The author has now checked over 25 published designs with this program and obtained some pretty shattering results. The input circuit losses lay between 0.2 and 1.6 dB, which doe not include additional losses caused by radiation, poor feed-in or incorrect coil-tapping, unsuitable layout, drain-current too low, etc. Some of the examples can be found in table 2. Many of the pre-amplifiers tested claimed noise figures which lay under the figure for the input circuit losses! This shows that many designers neglect the negative influences of the input circuit and give the noise figure of the pre-amplifier as that published for the specified noise-figure for the transistor in its data sheet.

2. MOSFET PRE-AMPLIFIER

All these findings were applied to the construction of a pre-amplifier which would be capable of outperforming all existing designs inasmuch that all the conditions outlined earlier would be fulfilled. Only by using helical tuned circuits, it is possible to construct a 2 m-band pre-amplifier with a noise figure which only exceeds that of the transistor used by 0.1 dB. The construction using the MOSFET BF 981 (**fig. 4**) yields an unloaded Q (Q_u) of about 700 and a loaded Q (Q_l) of 4.5. The input loss is then computed to be 0.1 dB.

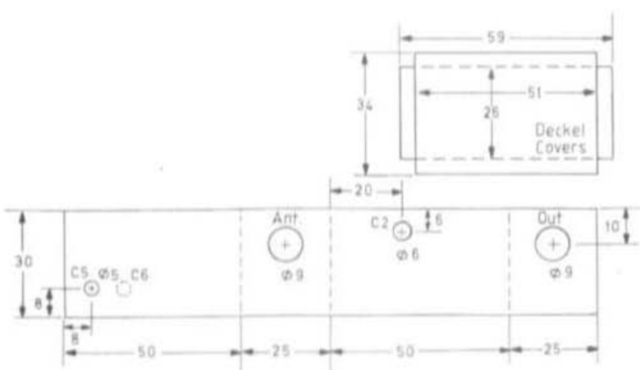


Fig. 7: Material working dimensions for pre-amp. box

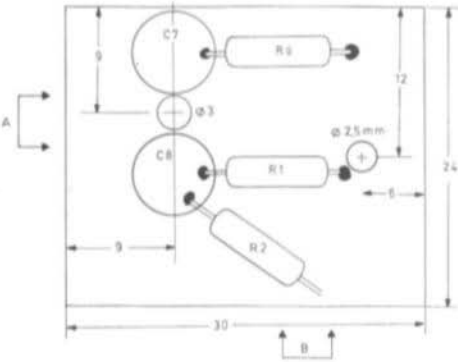


Fig. 8: Details of screen-mounted components

Figures 5 and 6 show the arrangement with the necessary dimensions given for two planes. The input and output socket utilise BNC flange receptacles soldered directly to the metal face. In the helical resonator enclosure, the tuning capacitor C_t and the antenna coupling capacitor C_c are to be found. The other compartment contains the drain tuned circuit suitably damped by resistance R4 (1) in order to reduce the amplification and at the same time to improve the matching to the following receiver. The coils are mounted perpendicularly with respect to each other in order to reduce the possibility of unwanted coupling to a minimum.

The pre-amplifier is housed in a container made from 0.5 mm thick brass or copper (**fig. 7**) and the input circuits separated from the output circuits by a metal separating screen. This dividing wall has a 3 mm hole, drilled in it through which the gate 1 lead is passed. As shown in **fig.8**, on each side of this hole are soldered chip capacitors. These are connected to the source and gate 2 and also the resistors which set the working point. The supply voltage is introduced by means of a soldered-in feed-through capacitor.

The housing is sealed with soldered-on lids, both above and below the helical tuned circuit so that the amplifier can function in stable conditions and not to be subjected to interference fields. The layout is such that fitting and subsequent removal of the two covers should not adversely affect the amplifier performance. The adjustment of this pre-amplifier is quite simple.

First of all, the drain current is determined by measuring the PD across the drain resistance R5. If the drain current is smaller than 10 mA or greater than 15 mA, the resistance R5 must be changed accordingly.

Following this adjustment, the pre-amplifier is connected between antenna and receiver and the trimmer C_c adjusted to that indicated in **table 1**. Alternatively, a suitable fixed capacitor may be also soldered in and the other trimmers tuned for maximum amplification using a low-level signal source.

If a noise generator, noise comparator (4) or noise test-set (5) is available then the optimum adjustment for C_c and C_t can be quickly found. Otherwise an approximation's method must be employed. This consists of increasing the value of trimmer C_t off the point of maximum gain until the output has dropped some 2 to 3 dB. This is approximately the tuning for minimum noise figure, verified by the author on a number of examples.

It should also be mentioned that it is very important to employ only the best-quality trimmers with air as the dielectric or perhaps ceramic or PTFE.

Owing to the low-loaded Q of the tuned circuit, the bandwidth of the amplifier is relatively high so that no retuning is necessary across the whole of the two-metre band.

With R4 shunted across the output, the gain is reduced to some 16 dB. If a gain adjustment is required this resistor may be raised to increase the gain or C4 lowered.

The only element whose value cannot be definately given is the source resistance R_s. This lies between 10 and 50 Ω according to the value of the transistor's I_{DSS}. This production spread in tolerances is quite normal for FETs.

Noise-figure measurements carried out on several selected BF 981 (about 0.8 dB) showed that the pre-amplifier noise figure lay only 0.1 dB above that for the transistor alone. This verified the theory and justified the whole effort. All the expectations were fully achieved and therefore no improvements or modifications were necessary. A few of these amplifiers were used for EME and MS work and others are in constant use with some keen tropo-DX-ers.

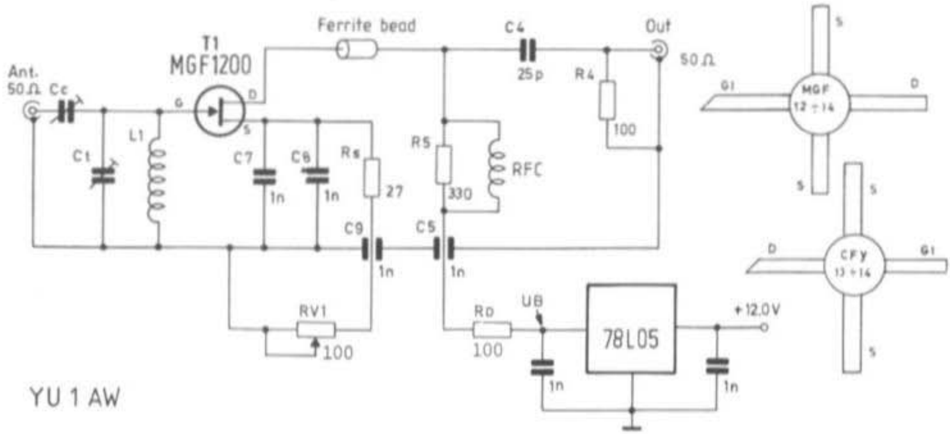


Fig. 9: Circuit schematic of pre-amplifier using GaAsFETs

3. A PRE-AMPLIFIER USING GaAs-FETs

Several other pre-amplifiers were equipped with one or two-gate GaAs-FETs, the MGF and CFY series and also the CF 300. The latter was very hard to match in the two-metre band but was much better in the 70 cm and 23 cm bands in this respect and was very much easier to work with.

The circuit diagram of an MGF 1200 pre-amplifier may be seen in **fig. 9**. The input is identical with that in fig. 4 except that L1 here is only five turns. The output circuit is non-resonant.

All GaAs-FETs are very difficult to match at low frequencies and are only conditionally stable at frequencies under 2 GHz. Conditionally stable means that they can, under certain circumstances, start to self-oscillate especially under conditions where maximum gain is being extracted. Conditionally stable pre-

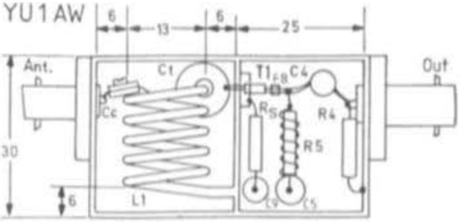


Fig. 10: Side-view of pre-amplifier constructed from fig.9

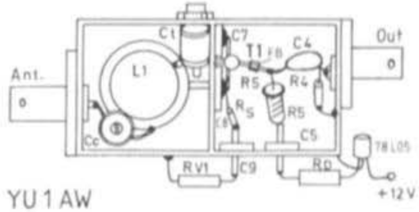


Fig. 11: Plan-view of pre-amplifier constructed from fig. 9

amplifiers are only stable in a very limited band of the input and output impedances and for this reason the peak values for gain and noise figure are often unobtainable. A resistance loading in the drain circuit reduces the superfluous amplification and brings the FET in a stable working region. A ferrite bead slipped over the drain lead also helps to improve the stability.

The amplification may also be adjusted in this amplifier by varying the values of R4 and C4. The drain current I_D may be adjusted by the preset potentiometer RV1 which is mounted outside the screened housing. This method is convenient and can also be used for the BF 981 pre-amplifier described earlier.

The only difference between the MGF and the CFY types (besides the lead arrangement) is the drain source voltage V_{ds}. The MGF types require 3 V and the CFY types 4 V. R_D must have a value of 10 Ω.

The results achieved also conformed very well with the calculations. The input losses lay at about 0.15 dB – a little higher than the BF 981 amplifier. The reason for the higher loss lay in the unfavourable input impedance of the GaAsFET. This resulted in a higher value for Q_l so that, for a given value of Q_u, the losses have to be higher.

The input losses with the CF 300 could only be minimised to 0.36 dB and they could be very much higher if the proper care is not taken.

The importance of the input circuit quality is made evident in table 2. It contains a comparison of a few common input circuits extracted from general literature and completed with a few proprietary pre-amplifiers.

3.1. Components List

C_t:	1.5 - 5 pF trimmer cap. (Johanson)
C_c:	1.5 - 6 pF ceramic or PTFE cap.
C 3:	2 - 10 pF ceramic or PTFE cap.
C 4:	25 pF disk cap.
C 5, C 9:	1 nF feedthrough cap., solderable
C 6, C 7, C 8:	1 nF disk cap. without connection
RFC:	12 turns Cul wire, 0.2 mm on R 5
L 1:	D = 13 mm; L = 18 mm; silvered wire, 2 mm Ø
L 2:	D = 6 mm; L = 15 mm; n = 5 Cul wire, 1.5 Ø

4. SUMMARY

Upon the basis of theoretical proof and practical research, it has been shown that it is feasible to employ low-noise transistors more efficiently in low-noise pre-amplifiers. Helical resonators instead of normal coils result in a higher unloaded Q

3.2. Explanation to Table 2

No.	Type	D (mm)	L (mm)	n	d (mm)	Note
1	coil	14	18	5	2	helical coil without shield
2	helical res.	14	18	5	2	resonator 25 x 25 x 30 mm
3	coil	9.5	12.7	5	1.2	published in magazine
4	coil	4.8	12.7	9	0.6	published in magazine
5	same as no. 1					
6	same as no. 2					
7	same as no. 3					
8	coil	9	18	6	1	published in YU-VHF-Magazine
9	coil	7	15	6	1	published in magazine
10	same as no. 1					
11	same as no. 2					
12	coil	6	12.7	8	1	published in magazine

(Q_u). By careful arrangement of the input circuit and the choice of the transistors, very low values of loaded Q (Q_l) were achieved and thereby very low input matching circuit losses. The losses result directly in a low-noise figure for the pre-amplifier. The above described pre-amplifiers represent, in this respect, a worthwhile improvement upon previously published designs.

5. LITERATURE

(1) Dobričič, D., YU 1 AW:
Pre-amplifier – Pros and Cons
VHF COMMUNICATIONS Vol. 19,
Ed. 4 / 1987

(2) Grossmann, D., DJ 4 RX:
Simple, Compact PA Stages for 2 Metres
VHF COMMUNICATIONS Vol. 2,
Ed. 1 / 1970, P. 45 - 55 and
Vol. 2, Ed. 2 / 1970, P. 111 - 122

(3) Kestler, J., DK 1 OF:
Two-Metre Receiver Front-End
VHF COMMUNICATIONS Vol. 17,
Ed. 4 / 1985, P. 241 - 251

(4) Compton, J. R., G 4 COM:
An Alignment Aid for VHF Receivers
Radio Communication 1976, Ed. 1 (January)

(5) Dohlus, M.:
A Home-made Automatic Noise Figure Measuring System
VHF COMMUNICATIONS Vol. 15,
Ed. 1 / 1983, P. 2 - 11 and
Ed. 2 / 1983, P. 66 - 83

BRIEFLY SPEAKING...

Another modification to the YU 3 UMV picture memory (VHF COMMS 4/82 + 1/83)

The resolution of the picture, especially in black/white and white/black transition zones may be improved by altering the capacity value of the component on pin 6 (I 107) ADC 0804 from 100 nF to 10 nF. The resistors pins 4 - 11 (I 105) 4007 also, are changed from 150 Ω to 15 Ω. The improvement is'nt spectacular but it can make the difference between legible and illegible text. The legibility of text, using 256 pixels per line is, in any case, unsatisfactory.

Willy van Driessche, ON 6 VD

GB 2 EC
Newport Amateur Radio Society

GB 2 EC is the callsign used by N.A.R.S. as part of our preparations for the Royal Welsh National Eisteddfod which is to be held in Newport from 30th July 88 – 6th August 88. Club members will hold GB 2 EC on a monthly rota from October 1987 until the Eisteddfod begins in July 88, a total of 10 stations. GB 2 EC will be active on H.F. and V.H.F.

The awards for working our special event callsign will be popular, judging by the response we had on July 1987, when we first put the callsign on the air. 10.000 QSL cards have been printed for this purpose.

We are organising a photographic competition which is open for every one. For competition details please contact N.A.R.S. via Box 33, Newport, Gwent, U.K.

GW 4 IED for N.A.R.S.

Ceramic resonators for Microwaves

In a new brochure from **Siemens**, on a double-side page, the dielectric resonator and the coaxial ceramic resonator are briefly explained with the aid of two main diagrams. Also the various types are tabulated. There are dielectric resonators for between 0.95 and 15.6 GHz as well as coaxial ceramic resonators for between 900 and 2400 MHz. It is worth mentioning that the ceramic resonators are available with a temperature coefficient of between – 3 ppm/K and + 9 ppm/K.

Order No.: B4-B3696

Modification of the ON 6 VD and DK 3 VF picture store for SSTV, FAX and WEFAX

This describes a general improvement to the FAX reception which allows both S-N or N-S picture information to be read into the memory, independently of each other and either normal or reversed image. For this facility, the circuit around I 213 and I 214 (74LS161) must be modified (see VHF COMMS. 1/83 fig. 14 and 1/86 figs. 4 and 5). Pin 1 of I 213 and I 214 is connected to the + 5 V line via a 2.2 Ω resistor. The connections between pin 1 of I 213, pin 1 of I 214 and Pt 207 must be broken. Also the Right/Left switch should be disconnected from Pt 204 (this connection can remain open). Now make a new connection from pin 1 of I 213 to the Right/Left switch.

Willy van Driessche, ON 6 VD

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 4/1987 of VHF COMMUNICATIONS

DK 2 LT	**Converting the TELECAR TS 160 into a 2 m, 80-Channel Amateur FM Transceiver**		**Art.No.**	**Ed. 4/87**
PCB	DL5NP 001 /	double-sided, drilled	6948	DM 31.–
Parts	DL5NP 001 /	TELECAR-TS 7 ICs, 5 transistors, 4 diodes, 1 crystal, 4 coil kits, wire, 1 core, 1 ferrite bead, 2 trimmers, 1 miniature relay, 2 mini-code-switches, 1 m coaxial cable RG-174/U, 33 resistors, (0204!), 34 capacitors (diverse types)	6220	DM 225.–
Kit	**DL5NP 001 /**	**TELECAR-TS with the mentioned parts**	**6221**	**DM 256.–**
Semi-rigid cable 50 Ω (Suhner SR-3) per cm				DM 0.50
YU 1 AW	**144 MHz LNA using Helical Tuned Circuits**		**Art.No.**	**Ed. 4/1987**
Kit	YU1AW LNA	144 with following parts: undrilled MS metal strips 30 mm wide, 2 BNC panel sockets, 3 wires, 1 ferrite bead, 1 Johanson trimmer 0.8 - 10 pF, 2 SKY trimmer, 2 ceramic, 3 chip and 2 feedthro' caps, 10 resistors, 1 trimmpoti, 1 5 V regulator and **1 BF 981**	6031	DM 50.–
		as above, but (instead of BF 981) with **1 GaAs FET**	6032	DM 75.–

UKWberichte Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133 47-0. For Representatives see cover page 2

Solar Power Supply for the Radio Amateur

The environment-friendly energy production by means of the photovoltaic cell is also meaningful for radio amateur applications. The operational time for portable and mobile use is overwhelmingly dependent upon the available battery power and the transmit power.

In order to operate with higher powers, or for longer periods, one is obligated to buffer the batteries with cost-free solar energy. The range of application of the solar module is very large:

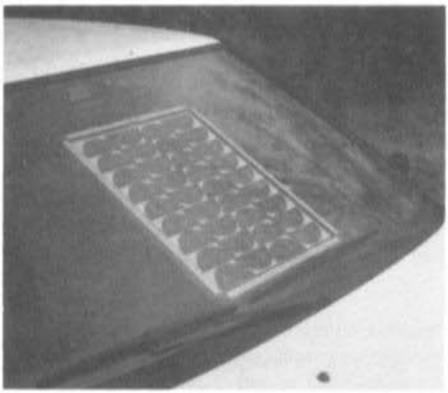
* For example, a handy solar module can be taken along with a small battery in a haversack on a portable expedition. The solar module can then be simply laid in the sun and used as a buffer for the transmitting session.
* When camping in tents, the module can be used for charging batteries for both radio and lighting purposes.
* For a real field-day, i.e. without portable generators or provision of a mains supply, one, or more, solar module may be utilised outside or on the tent.
* The mounting of a large solar module on the roof of a caravan offers advantages for caravan enthusiasts: it can hardly be seen from ground level, it acts as a thermal insulator for the caravan roof and also produces energy without incurring any space penalties.
* It is just as advantageous to use a large solar module for a motor car merely by attaching it to a redundant luggage rack or ski carrier. The installation will produce so much power, even when transmit powers in the KW region are being used, that there will still be enough power to start the car afterwards!

The mono-crystalline silicon discs are embedded in EVA-plastic which allows the cell to thermally expand and contract, and at the same time to protect it from the ingress of water.

The light sensitive side of the cells are covered with a highly transparent, tempered glass screen which protects the modules from both mechanical and meteorological extremes. The back of the module comprises a multi-layer coating of plastic with an aluminium core and a PVF-plastic coating. This makes the module both corrosion and water proof.

The frame of the module is made from anti-corrosion proofed, extruded-aluminium profile. The module is thereby rigid, shockproof, UV-resistant and distortion free.

Ready-to-use modules using 17, 35 and 50 W mono-crystalline solar cells are available from UKW-Technik.



Technical data	Siemens solar modules		
type	**SM 50**	**SM 36**	**SM 18**
Rated power	50 watts	35 watts	17.5 watts
Voltage (OC)	21.5 V	21.5 V	21.5 V
Current (SC)	3.1 A	2.4 A	1.2 A
Ambient temp.	– 40 to + 50 degree C.		
Dimensions (mm)	980 x 460	1014 x 414	524 x 414
Weight	6.3 kg	5.8 kg	3.2 kg
Art.No.	**3220**	**3218**	**3219**
Price:	**DM 908.–**	**DM 680.–**	**DM 459.–**

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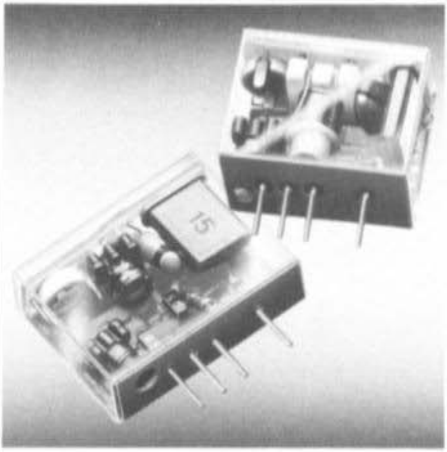

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