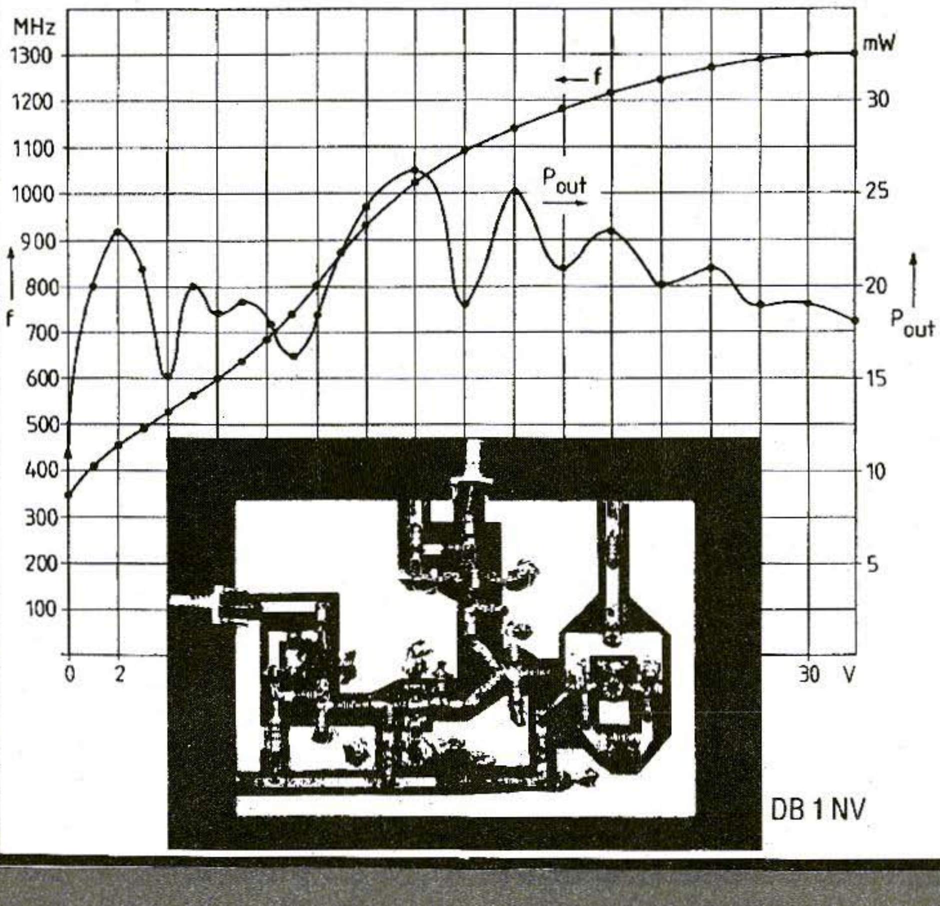
A Publication for the Radio Amateur Worldwide

Especially Covering VHF, UHF and Microwaves

VHF COMMUNICATIONS

Volume No. 24 . Winter . 4/1992 . £3.50

BROADBAND VCO's using MICROSTRIP TECHNIQUES



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Thank you for your continuing support for VHF Communications Magazine
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Dr. Ing. Jochen Jirmann, DB1NV

Broadband VCO's using Microstrip Techniques

The VCO used previously in the spectrum analyser over the frequency range 450 to 1000 MHz was tricky to make mechanically, which led me to look for a simpler solution. From this arose a circuit concept that, according to the components used, is suitable for the frequency range 400 to 1900 MHz and offers guaranteed tuning over significantly more than one octave.

The frequencies 450 and 1450 MHz are no longer the limits and the spectrum analyser can be enhanced to provide two overlapping frequency ranges of zero to 1000 MHz and 900 to 1900 MHz. Two further variants of the VCO achieve 400 to 1250 MHz and 800 to 1900 MHz. The oscillator module is thus suitable for sweep generators and similar applications

The first variant covers the 70cm band and several microwave frequency-multipliers. The second version covers the 23cm band as well as the Meteosat and satellite TV frequencies.

The main limiting data of the three variants are:

* Tuning range: 450 to 1450 MHz
  selectively: 400 to 1250 MHz
  and with alternative components: 800 to 1900 MHz
* Tuning voltage: 1 to 30V
* Operating voltage: 15V
* Output power: > 10mW
* Output stability: better than 6dB

1. CONSTRUCTION BASICS

In order to produce a VCO with a tuning range greater than one octave certain preconditions must be fulfilled.

- ❑ The capacitive tuning range of the tuning diode must be sufficiently large (for one octave this can be calculated as a capacity variation of 1:4).

- The oscillator circuit must be secure over a wide frequency range and operate with no great changes in amplitude.

Until a few years ago tuning diodes out of TV tuners were all that the normal amateur could use, with a capacity variation of from 1:3 to 1:6. On account of the inevitable capacitive effects of the circuitry, one octave was the maximum tuning range achievable.

With the introduction of hyperband cable TV tuners and satellite TV tuning diodes have been produced with a capacity variation of 1:9 up to 1:19. For the VCO presented here two types, made by Siemens (2) and Valvo, were selected following tests:

* the BB619 with a maximum capacity of 37pF and a minimum of 2.7pF
* the BB811 with a maximum capacity of 9pF and a minimum of 1pF.

Both diodes are enclosed in a dual-ended plastic surface mount device (SMD) package style, which exhibits the same kind of low self-inductance (approx. 2.5nH) as the BB621 in mini-MELF glass package used before. Resistive loss is a comparable 0.55Ω.

Self-resonance (series resonance) can be calculated from the inductance and the capacitance of the diode as

$$f_s = \frac{1}{2\pi} \cdot \sqrt{\frac{1}{LC}}$$

Above the series resonance the inductive component of the diode becomes more significant, in other words it behaves like a tunable inductance. Expressed another way, self-resonance is the frequency at which the diode-tuned oscillating circuit can produce the most with an external inductance of close to zero. Practically achievable frequencies lie below this figure since an oscillator circuit inductance of zero is naturally not achievable in practice.

For the BB619 we have, according to capacity, self-resonance from 520 MHz up to 1.9 GHz, and for the BB811 this lies between 1.06 GHz to 3.2 GHz.

A maximum achievable figure of merit (Q) for the circuit can be calculated from the resistive loss and the capacity of the diode (assuming zero loss in the external components). This is:

$$Q_{max} = 1/(R_v \cdot 2\pi fC)$$

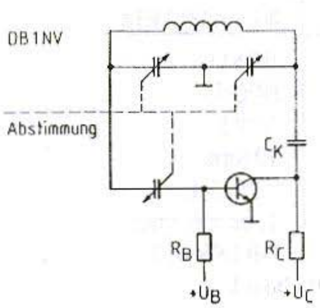

Fig.1: Colpitts Oscillator with Phase Correction

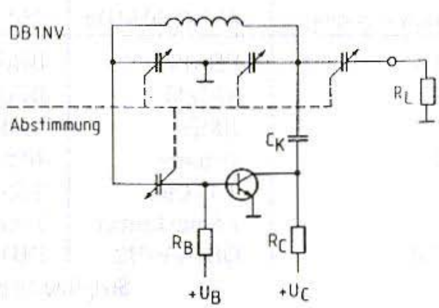

Fig.2: Colpitts Oscillator with Variable Output Coupling

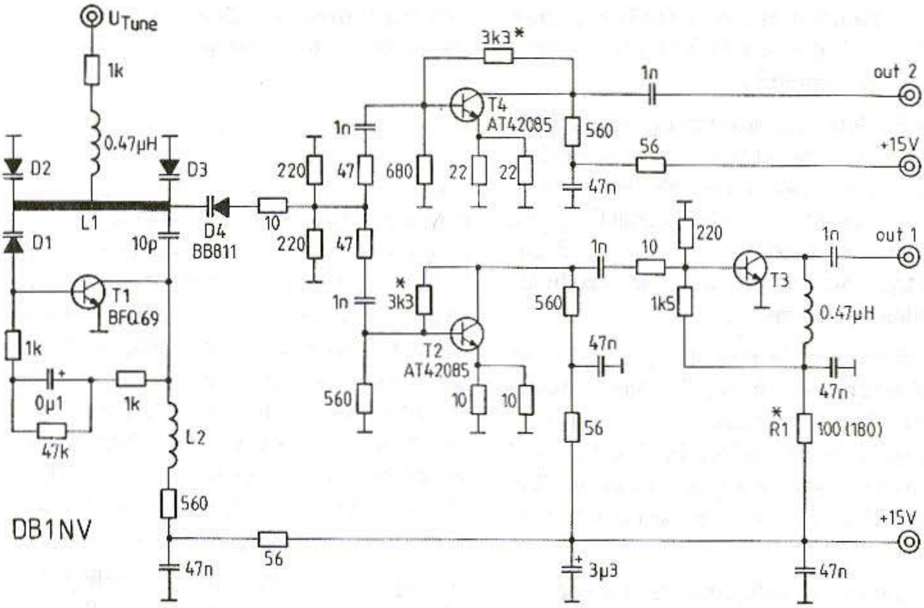

Fig.3: Microstrip VCO for 400 up to 1900 MHz (* = components with leads)

With a measured frequency of 500 MHz we have for the BB619 a range of values from 16 to 215 and for the BB811, from 64 to 580. As can be seen, at maximum diode capacity at the lower end of the tuning range we can only count on a restricted circuit performance and should expect only average noise characteristics from the oscillator.

At low frequencies it is no problem to ensure feedback free of oscillation over a wide frequency range since the phase shift in the active element (the transistor) is

Frequency range	400-1250 MHz	450-1450 MHz	800-1900 MHz
D1-D3	BB619	BB619	BB811
TR3	BFG96	BFG65	BFG65
R1	100Ω	180Ω	180Ω
L2	15 turns 0.35 CuL 1.8mm former	10 turns 0.35 CuL 1.8mm former	10 turns 0.35 CuL 1.8mm former
PC B	DB1NV 012	DB1NV 012	DB1NV 013
L1	Stripline on printed circuit board		

CuL = enamelled copper wire

Table 1: Components of the VCO module which vary according to frequency range

constant at zero degrees (base circuit) and 180 degrees (emitter circuit). In the UHF range this phase shift is extremely frequency-dependent and in common transistors lies between 70 and 120 degrees. With broad tuning ranges it is therefore impossible to avoid the need to compensate for the phase shift of the transistor according to frequency.

A simple and unproblematic solution is the Colpitts oscillator in the emitter circuit, with a variable capacitor to achieve phase correction in the base circuit of the oscillator transistor. Circuit principles are given in Fig.1.

In the amateur literature this type of oscillator was first described by YT3MV (1) for a satellite TV receiver and is best suited for wide tuning ranges. The circuit has a defect in that the oscillator power increases sharply with frequency whilst at the lower end it barely suffices. This can be rectified with a simple modification to the circuit by using a further tuning diode to make the output coupling from the collector of the oscillator transistor variable and thus reduce the output with rising frequency.

The principle is shown in Fig.2, where the tuning diodes are represented as variable capacitors with common tuning.

Having sorted out the oscillator, the circuit can be enhanced with three further functional elements: a resistive splitter to divide the oscillator output symmetrically followed up by two buffer amplifiers. The single stage amplifier is for connecting a frequency counter or a frequency control loop, whilst the two-stage follower provides a level around 20mW for driving the mixer or measuring object.

2. CIRCUIT AND COMPONENT VARIATIONS

The circuit of the VCO module valid for all variants is shown in Fig.3 and the component options in Table 1. The core of the circuit is formed by the oscillator around TR1, a BFQ69. Tests showed that transistor types like BFQ69 or BFR91 were best for this application, whilst still "better" transistors like the low-cost Avantek AT41485 or AT42085 were subject to interfering oscillations between 2 and 5 GHz.

The actual oscillator circuit comprises L1, a stripline inductance, and the tuning diodes D2 and D3; the tuning voltage of from 1 to 30 volts is taken to the "cold" point in the middle of L1 via an SMD choke and a damping resistor.

The diode D1 represents the variable base coupling of the oscillator transistor, whilst the collector (by way of 10pF) lies at the other end of the resonant circuit.

The operating position of TR1 is set by collector resistance and a resistor ahead of the base to around half the power rail voltage (approx. 6 to 8 volts).

An electrolytic capacitor shunts a large part of the base resistor for audio frequency, which reduces the noise sidebands of the oscillator. In this connection the trimming of the coupling diode D1 to the base potential of TR1 instead of ground is not ideal but was chosen to simplify the circuit.

Diode D4 represents the variable coupling out from the circuit, working as a combina-

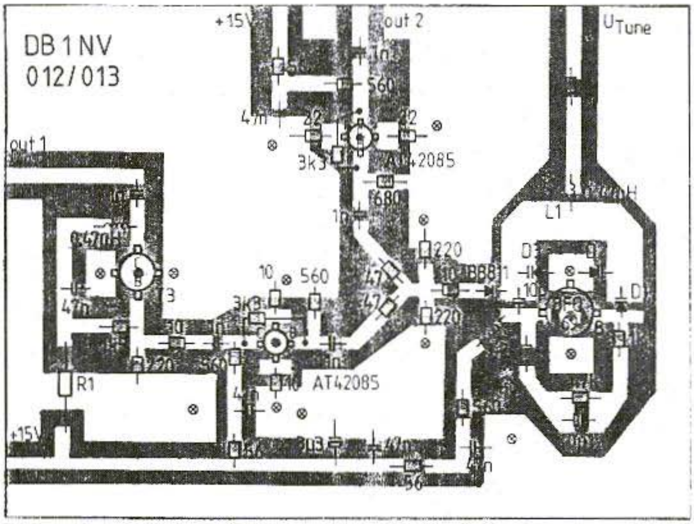

Fig.4: Component Layout of the Microstrip VCO DB1NV 012 and 013. The difference between the two lies in the size of L1. Through contacts are marked ∅

tion of attenuator and resistive splitter, the outputs of which lead to buffer amplifiers.

The amplifier with TR4 for driving a counter or frequency regulation loop is equipped with the highly efficient amplifier AT42085 already mentioned.

A combined series and parallel negative feedback stabilises the operating point and smooths the frequency tracking. At the counter output (Output 2) a power level of around 1 milliwatt is available.

The amplifier for the main output (Output 1) with TR2 and TR3 has a first stage almost identical to that using TR4 but the second stage has, according to variant chosen, either a BFG65 or BFG96. This produces an output power of around 20mW.

The circuitry has no special characteristics.

3. CONSTRUCTION AND COMMISSIONING

Before starting construction, the frequency range to be covered (and hence the PCB version) must be decided. For the two lower frequency ranges we use DB1NV 012, for the upper range it is DB1NV 013. The components for the elements that vary according to frequency are indicated in Table 1 following. The remaining components are in the parts list.

3.1 Parts list

TR1	BFQ69 (Siemens)
TR2, TR4	AT42085 (Avantek)
D4	BB811

(continued overleaf)

3.1 Parts list (continued)

All resistors:	mini-MELF or chip, size 1206. Resistors marked with a star are miniature types with normal leads.
All ceramic capacitors:	chip, size 0805 or 1206.
Electrolytics:	SMD tantalum electrolytics 35V or tantalum bead.
RF chokes:	SMD chokes, e.g. Siemens SIMID01, 0.47uH

Also:

Tinplate housing 74 x 55.5 x 30mm

2 feed-through capacitors 2.2nF

1 feed-through capacitor 22pF

2 miniature connectors SMC or SMB

3.2 Component assembly

Given some experience with SMD components and the overlay diagram Fig.4, the placing of components on the PCB should not pose any problems. First drill holes about 0.9mm through the board at the positions indicated and connect the upper and lower sides with tinned copper wire or proper hollow rivets. In the case of TR1 and TR3, the emitter vanes can be used for through contacts. Next fix all the SMD components, finally the three resistors with leads close to TR2, TR3 and TR4 as well as choke L2.

If the VCO module is to go in a case of its own, a standard tinplate case (size 74 x 55.5 x 30mm) can be used. All RF connections are made with miniature RF jacks (type SMB or SMC). The tuning voltage is led in through a 22pF feed-through capacitor, the feed voltages via 2.2nF, but these values are not critical.

3.3 Commissioning

Commissioning requires two power supplies (15V and 0 - 30V), a frequency counter and a power meter. The 15V power supply provides the operating voltage, the adjustable one the tuning voltage. The counter goes on the counter output and the power meter on the main output.

After connecting these voltages the VCO should be tuned over the relevant range, checking for no gaps in the oscillation or abrupt changes in output power. The measurement curves achieved by the author shown in the next section should serve as a guide to the correct values.

Poor tuning behaviour indicates the RF chokes should be checked; self-resonance generally occurs because of poor construction forms and can be corrected by swapping the chokes for others. However, the author has also come across half-defective oscillator transistors, which gave no problems DC-wise but produced wild oscillations in the GHz region.

3.4 Test results

Figures 5 to 7 show in each case the output power and frequency of the three oscillator variants versus tuning voltage. As can be seen, the desired results mentioned at the beginning are largely achieved.

At the same time the phase noise of the oscillator was investigated. In this connection the author is grateful to Dr Prokoph DL5NP for the measurements made with a phase noise test station type 3048A by

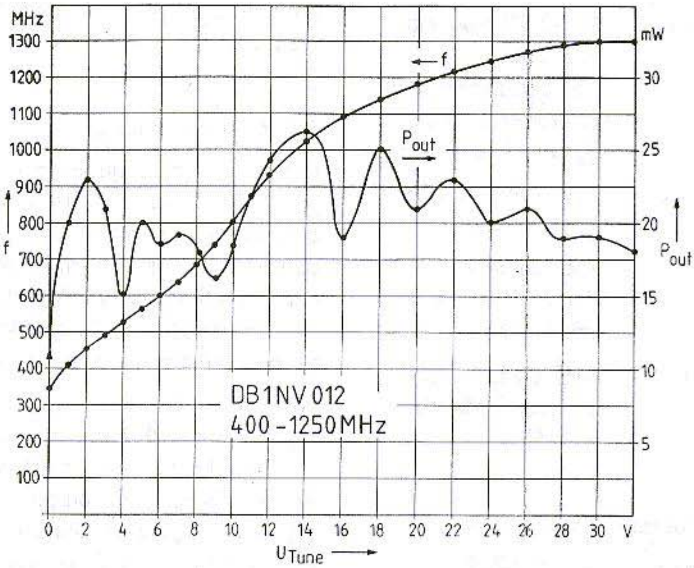

Fig.5: Frequency Tracking and Output Power versus Tuning Voltage for a 400 to 1250 MHz version

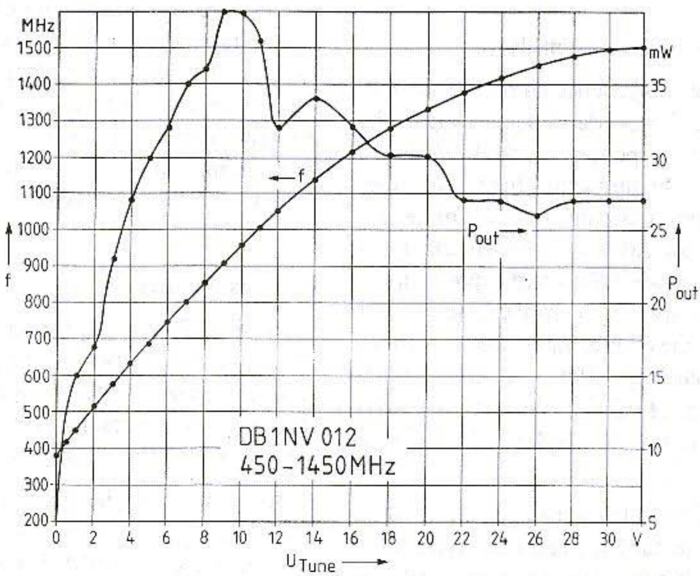

Fig.6: Frequency Tracking and Output Power versus Tuning Voltage for a 450 to 1450 MHz version

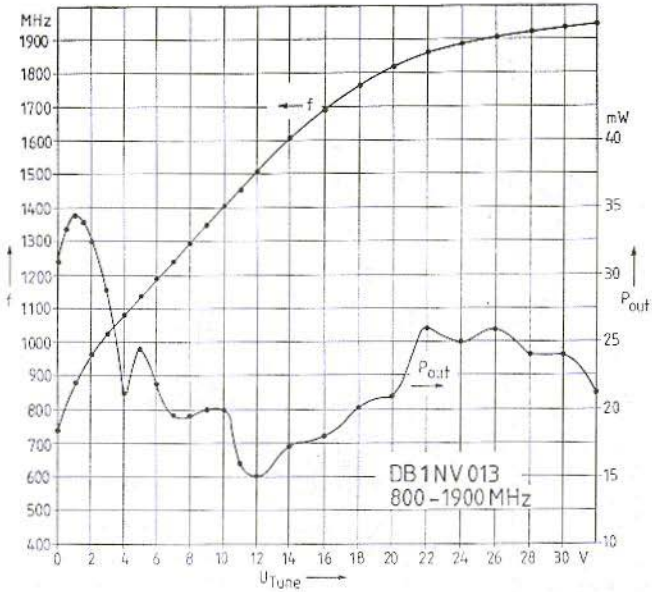

Fig.7: Frequency Tracking and Output Power versus Tuning Voltage for a 800 to 1900 MHz version

Hewlett-Packard. This entailed the use of extremely stable noise and hum-free power supplies to avoid influencing the measurements. The results for an operating frequency of 570 MHz are shown in Table 2.

In practical applications of the VCO in a spectrum analyser it is apparent when measuring neighbouring channel power in RF synthesisers (25 kHz steps) that extended waves 65dB down can be detected if the narrowest crystal filter bandwidth of circa 1 to 2 kHz is selected. Calculation in this case gives a dynamic range of 72 to 75dB. That this is not achieved in practice is because the power supply of the VCO in the analyser is still not clean enough.

Frequency distance from Hz	Phase noise dBc/Hz
100 Hz	-30
1000 Hz	-60
10 kHz	-90
25 kHz	-105
100 kHz	-115
1 MHz	-135
5 MHz	-150

Table 2: Phase noise of the broadband VCO

4. CONNECTING THE VCO MODULE TO THE VCO/PLL GROUP

Since the new VCO is significantly larger than the oscillator originally used in the setup DB1NV 007, direct substitution is not an option. Connection can be made as follows.

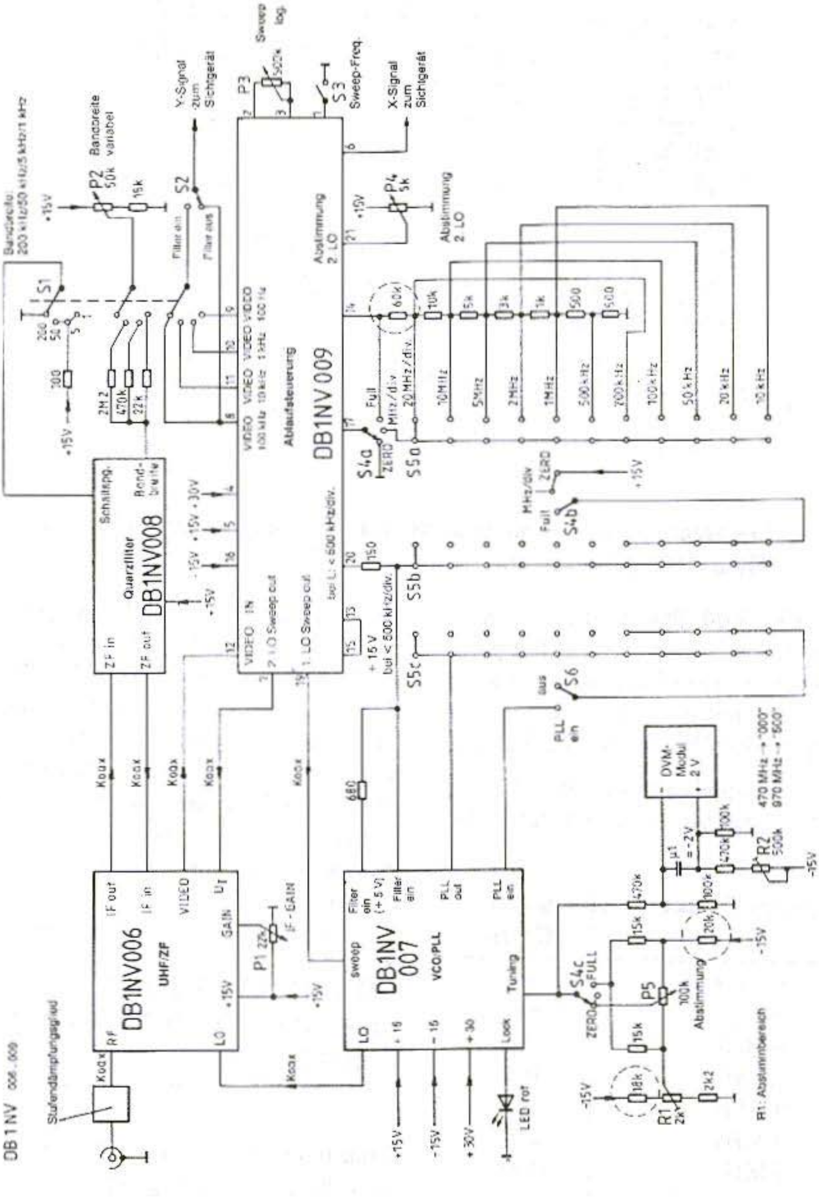

Fig.8: Values of Components to be changed in the Spectrum Analyser are circled

First connect the main output of the VCO to the first mixer of the RF/IF unit DB1NV 006. The existing VCO in the PLL group can be disconnected by unsoldering the supply and tuning voltages and taking these to the new unit. Take care with the routing of the tuning voltage line in order to avoid interference; screened cable is recommended.

The counter output of the VCO is led through screened cable to the PLL unit, with the inner conductor shunted to ground via 47Ω and taken via 470pF to pin-2 of IC2 (SDA4211).

To adjust the frequency regulation loop to the new tuning range, the time constant of IC6 (74LS221) needs to be reduced. So the capacitor between pins 14 and 15 is reduced from 68pF to 47pF.

Finally three resistors in the interconnection of the units are changed as shown in Fig.8 to ensure the new tuning ranges are properly covered. The digital voltmeter used as frequency indication must be recalibrated. This completes the alterations to the analyser.

Verdict: The new broadband VCO concept has been proven over a year in the author's analyser and sweep generator with flying colours. The phase noise of the new oscillator is perforce somewhat greater, which is noticeable in the poorer resolution of closely spaced (< 50 kHz) signals of widely differing levels. This drawback is more than offset by the continuous tuning ability from 0 to 1900 MHz.

To reduce noise further, a new PLL loop is under development by the author, which together with significantly reduced noise will offer quasi-continuous tuning of the oscillator in 50 kHz increments. In this way the analyser will offer definable tuning and will also be capable of use as a precision test receiver with precise frequency indication.

5. LITERATURE

(1) M. Vidmar YT3MV: TV Satellite Receive System Part-2: Indoor Unit VHF COMMUNICATIONS 1/87, pp. 35-56

(2) Siemens-Datenbuch "Tunerhalbleiter" 1990.

Matjaz Vidmar YT3MV

A 1Mbyte SRAM Card for the DSP Computer

In previous issues of this magazine (1) we have described a computer for digital signal processing - the DSP Computer - together with its operating system, the latter being held in permanent CMOS memory.

Originally 256 kilobit static RAMs were used since these were the most widely available memory chips. Thirty-two such chips, spread over four cards, gave 1 MB of storage, allowing convenient picture reception and signal processing on the DSP computer.

Since the computer was designed from the outset to address up to 15 MBytes there is an immediate benefit from increasing this memory capacity.

A first step in this direction was doubling the storage by means of a piggy-back arrangement described by Heinz Kriegelstein (2). The second step is the 1 MByte card described now.

Unfortunately it took longer than expected for large SRAMs to come down to a reasonable price; the first 1 Mbit SRAMs did not reach the market until 1990 and prices then began to drop slowly.

These new memory chips are generally arranged as 128 eight-bit words and are offered in a variety of 32-pin package styles.

The following list of producers known to me of 1 MB CMOS RAMs is bound to be incomplete but will serve as a guide for purchasing.

Supplier	1 MB SRAM
Sony	CXK581000P
Asahi Kasei	AKM628128LP
Hitachi	HM628128LP
Mitsubishi	M5M51008P
NEC	uPD431000CZ

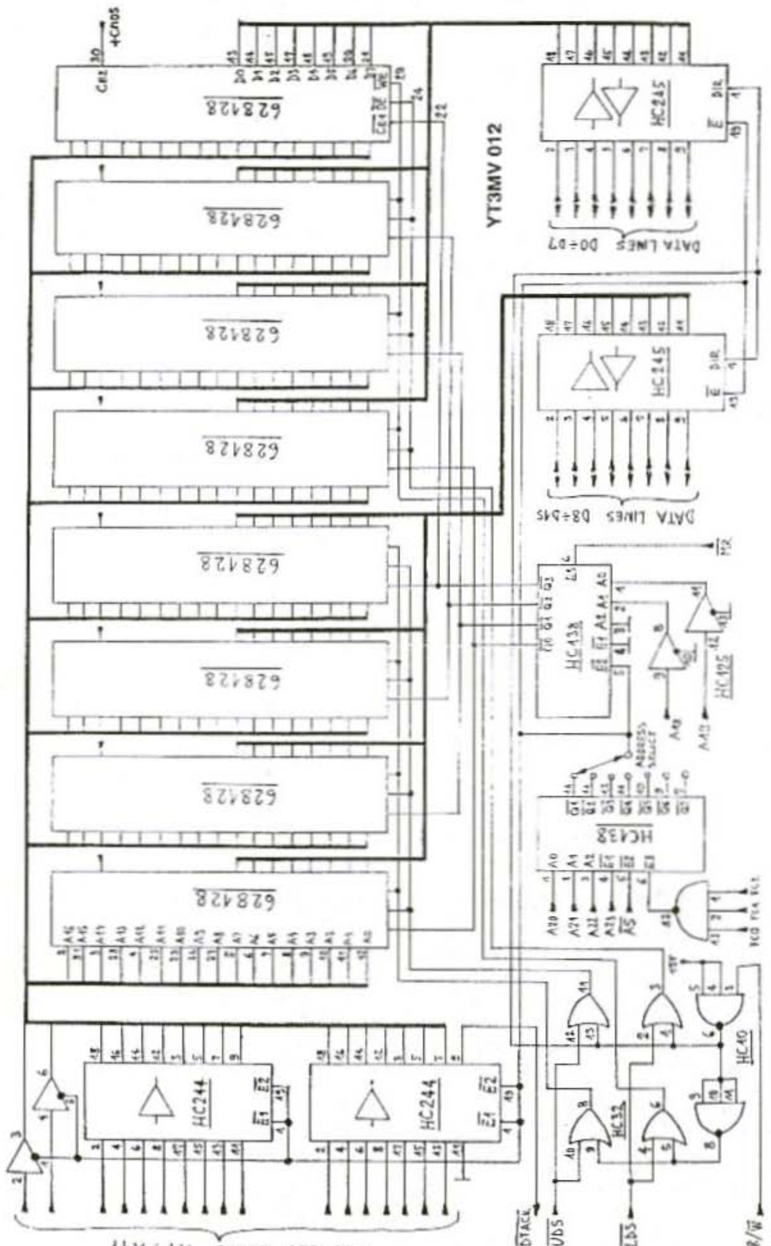

Fig.1: The 1 MByte SRAM Card for the DSP Computer

These integrated memory modules are available in 32-pin DIL packages and in several different SMD styles. In contrast to the previous 256k memories (32 kByte, 28-pin DIP), these have two additional address lines and an extra chip select input (active high).

The 32-pin package also has one unallocated pin, presumably for compatibility with future 4 MB memory chips. The version in DIL cases all have pin-to-pin compatibility.

In order to use the new 32-legged 1 Mbit SRAMs in the DSP computer it was necessary to design a new PCB. This new memory card (YT3MV 012) carries eight of these memory chips, a buffer for the address and data buses plus the necessary address decoding and control logic.

Fig.1 shows the circuit of the new 1 MByte card, which is quite similar to the old 256 kByte one, apart from the simpler address decoder and buffering.

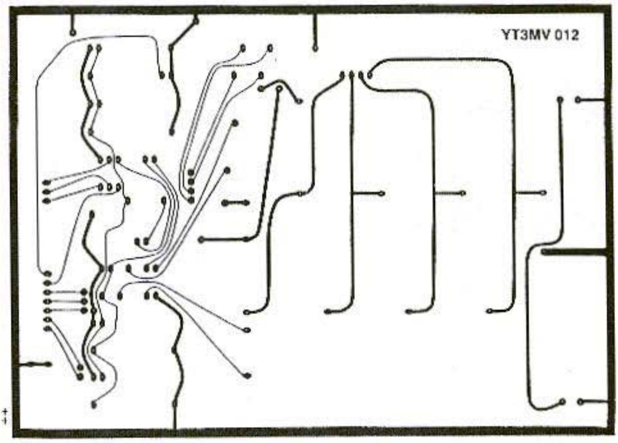

Fig.2: Track layout on the component side of PCB YT3MV 012

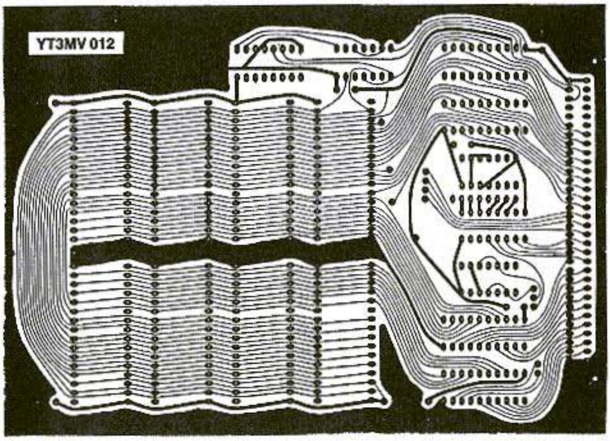

Fig.3: Track layout on the lower side of PCB YT3MV 012

Arrangement of address ranges and wire links

Location	Address range	Chip and pin
1	100000H - 1FFFFFH	HC138 Q1 - pin 14
2	200000H - 2FFFFFH	HC138 Q2 - pin 13
3	300000H - 3FFFFFH	HC138 Q3 - pin 12
4	Return line	HC244E - pins 1, 19, etc.
5	400000H - 4FFFFFH	HC138 Q4 - pin 11
6	500000H - 5FFFFFH	HC138 Q5 - pin 10
7	600000H - 6FFFFFH	HC138 Q6 - pin 9
8	700000H - 7FFFFFH	HC138 Q7 - pin 7

The wire links are inserted between location 4 and the location corresponding to the address range selected.

Another similarity is the programming of the start address of the 1 MB memory blocks with wire links; there are seven different start addresses possible from 100000H to 700000H.

To allow the operating system of the DSP computer to make effective use of the memory, the address range of each new card (and any old ones) should form together one continuous memory range.

The new memory card YT3MV 012 has the same dimensions as the other cards in the DSP computer: 120mm x 170mm. The layout of both sides is shown in figures 2 and 3. The component overlay is given in Fig.4.

The new control logic is entirely similar to that of the old system (a '138 chip merely replaces a '125). The scheme does, however, only cater for ICs of the 74HC (or 74HCT) series; 74LS types should not be used.

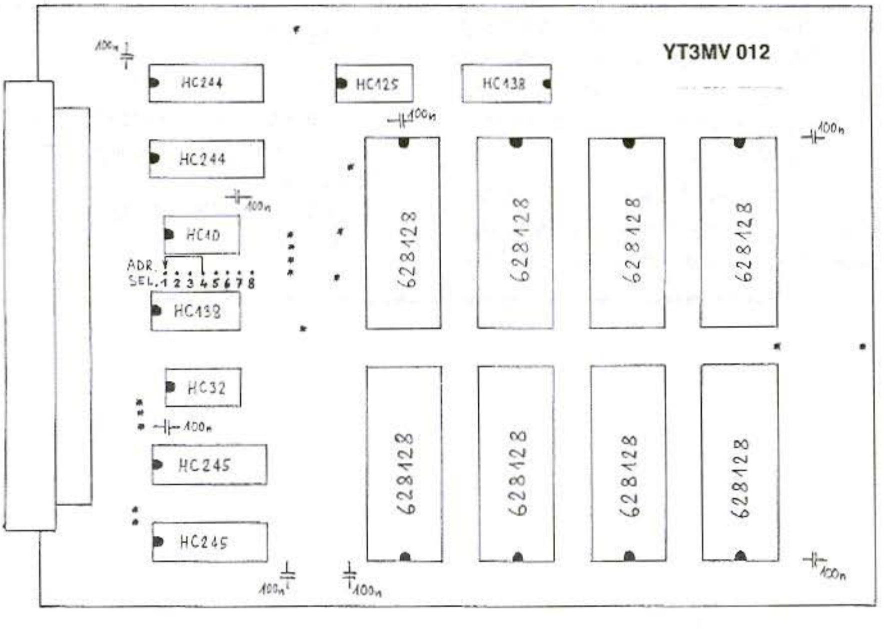

Fig.4: Component Layout of the 1 MByte SRAM card YT3MV 012

The prototype of the new memory board uses eight chips of the type AKM628128LP-10 by Asahi Kasei; it has worked at a CPU speed of 12.75 Hz for more than half a year. An access time of 120ns should therefore suffice for CPU speeds of 10 MHz and 100ns for 12 MHz.

The new 1 MB memory card should be tested in exactly the same way as was described for the older 256 kB card. First use the operating system command "W" to write something in a couple of memory locations and then read it back. Then set a new memory range (operating system command "N") and finally attempt to load some programs from disk and run them.

Suitable locations for the card in the computer are restricted by the bus card - a maximum of four memory cards can be used at a time. This means total memory can be no more than 4 MBytes using four of the new cards, less if some older ones are re-used. If a 1 MB card is not fully populated with chips, then the highest addresses must be started (in pairs!) and the corresponding start address given with the command "N".

In conclusion I would like to mention that these new memory cards have been designed to accommodate 4 Mbit SRAMs with minimal modification, once these chips become available. The address buffers for them are already provided.

LITERATURE

(1) M. Vidmar YT3MV: Digital Signal Processing Techniques for Radio Amateurs
Part-1 VHF COMMUNICATIONS 2/88, pp. 76-97
Part-2 VHF COMMUNICATIONS 1/89, pp. 2-24
Part-3 VHF COMMUNICATIONS 2/89, pp. 74-94
Part-4a VHF COMMUNICATIONS 3/89, pp. 130-137
Part-4b VHF COMMUNICATIONS 4/89, pp. 216-227
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(2) Heinz Kriegelstein: Simple Doubling of the Data Storage Capacity of the DSP Computer
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Doppler Radar in the 10 GHz Amateur Band Part-2

3. SELECTION OF AN SHF RADAR SETUP

Several choices are open, offering varying complexity and results.

3.1 The "Gunnplexer"

This device, once widely used, is based upon a circulator (Fig.18). Power derived from a Gunn oscillator reaches the antenna via a circulator, with a small component reaching the diode mixer as a heterodyne signal. Received signals reach the mixer practically unhindered by the circulator. The desired output signal is the difference between the received and the local oscillator (LO) signal.

In this case the less than perfect directionality of the circulator is a desirable feature, since the LO works as a transmitter and at the same time makes available a heterodyne frequency. The reverse attenuation of the circulator is around 20dB so that even with several hundred mW transmit power the mixer diode is not threatened. In addition the circulator protects the LO from the outside world, which is good for stability.

A frequency drift of, for example, 10 MHz at 10 GHz is just one part in a thousand; this also means that any error in the Doppler frequency is also only 1 in 1000. There are no other errors, since the transmit and heterodyne frequencies are produced by the same oscillator.

With a 40cm dish antenna we achieved ranges of 200 metres.

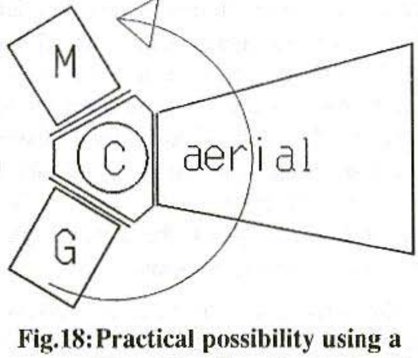


Fig.18: Practical possibility using a Circulator (i.e: Gunnplexer)

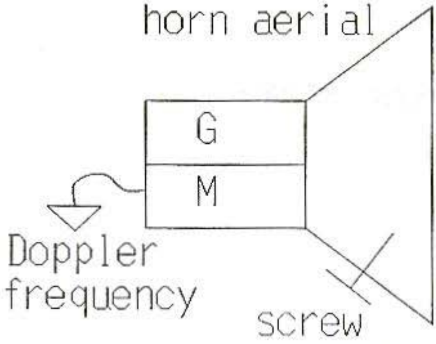


Fig.19: Another alternative: a Burglar Alarm. The minute antenna rules out large radar ranges

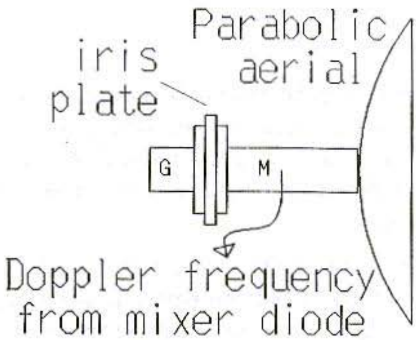


Fig.20: A 'Blast-through' Mixer for broadband FM is a good choice for Radar experiments

3.2 Burglar alarm

The setup sketched in Fig.19 was used some years back in microwave burglar alarms. Use was again made of the Gunn oscillator in the upper module, the transmitter power feeding a horn antenna. A screw in the wall of the horn radiator reflected a small part of the transmitted power into the lower module (the receive mixer). Regrettably half the receive capability was lost in the transmitter.

At the same time, the lack of isolation meant that nearby environmental influences could cause significant drifting, especially with movement directly in front of the antenna (which was what the setup was designed for). In our case the high level of drift led to inaccurate Radar measurements. Another disadvantage is that this assembly cannot be attached to standard waveguide, which makes application with a dish antenna problematic.

Using the horn antenna supplied, very short ranges are achieved and then without great accuracy.

3.3 The "blast-through" mixer

This arrangement was very popular in amateur radio circles some years back since it is easy to make at home and gives good results. A Gunn oscillator is separated from the rest of the assembly by an iris (a plate with a small round hole). According to the diameter of the hole and the thickness of the sheet, a greater or smaller part of the oscillator output reaches the antenna through the mixer (Fig.20). This (low) power is the output signal and at the same time the heterodyne signal.

Being a non-linear element, the mixer diode produces the Doppler signal - an audio frequency signal equal to the difference between the transmit and receive frequencies.

The method of operation is certainly very similar to the Gunnplexer but the transmit output is significantly lower; a few mW instead of 40mW! With an iris of 6mm to 7mm diameter made of 1mm brass sheet and a Gunn oscillator, the approx. 5mW produced is not optimal but adequate for the mixer diode. If we make the iris larger we certainly get more mixer current flowing but the oscillator also loses stability. With high powered Gunn oscillators the mixer diode can even be damaged.

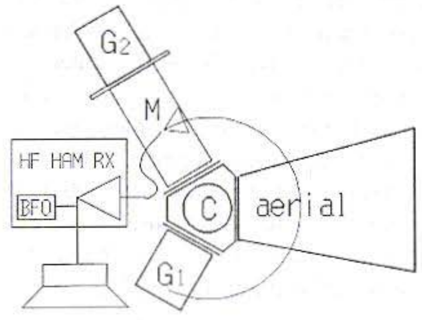


Fig.21: An attempt to increase range by using an IF of 30 MHz was abandoned due to frequency unsuitability

The transmit power will not exceed a few mW with a blast-through mixer. In the past we managed with this amount in wideband FM to span 130km and more, from one Swiss mountain top to another. Yet the range for Doppler Radar is no more than 100 to 150m, as our practical tests demonstrated.

3.4 Higher IFs

In the setups described so far the incoming signal has been converted in the mixer diode down to the audio frequency range.

Down there is also where the strongest phase noise effects of the oscillator appear, and Gunn oscillators are notorious for phase noise. The phase noise intrudes on the audio signal and reduces the sensitivity of the system.

One probable improvement would be to use a GaAsFET oscillator stabilised with a dielectric resonator (5) giving lower phase noise, whilst on the other hand it would be interesting if introducing an IF would increase sensitivity.

In our FM transceivers we used an IF of 30 MHz. So we experimented with the setup shown in Fig.21. The first oscillator G1 produces the transmit frequency f at the highest possible level, the second oscillator being adjusted to f + 30 MHz or f - 30 MHz.

The signal reflected by the target lies between f + 8 kHz and f - 8 kHz, depending whether it is moving towards or away from us, and at what speed. Behind the mixer the signal therefore lies between 29.992 and 30.008 MHz. In the receiver it is amplified several score times and, if we switch in the BFO, we get an audio frequency tone in the loudspeaker. We can adjust the BFO so that this tone amounts to a few Hz when the target is receding at maximum speed from the Radar device; when the target is static the audio frequency is 8 kHz, and when it is approaching fastest the tone is 16 kHz (provided the receiver bandwidth can handle this).

If the target's direction of movement is not of interest, we can adjust the BFO so that the tone is around zero when the target is static.

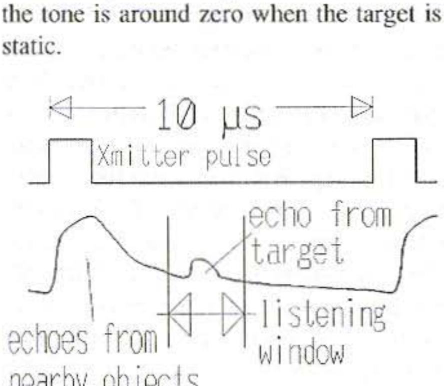


Fig.22: A Pulsed Transmitter and a brief time window can avoid receiver overload and clutter arising from reflection from nearby objects

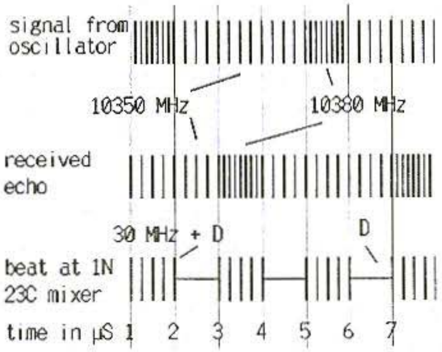


Fig.23: Frequencies in a Frequency-Inverted Continuous Wave Radar based on the 'Blast-through' Mixer. A low frequency is produced in the Mixer Diode between the received pulses: this is the Doppler Frequency

When it now moves, the audio frequency will rise in either case up to a maximum of around just 8 kHz - which corresponds already to a speed of about 450km/h.

The sensitivity of this setup should exceed that of those described previously by a wide margin and to give an example: a bird that flew away a few score metres already gave a very strong signal! We calculate that the range with this setup for a target with a Radar cross section (RCS) of one square metre is several hundred metres. Yet, as we know from our endeavours with the Radar formula, the range can barely exceed 1km - even if we had a noise-free receiver.

This setup using an IF has, however, two serious deficiencies.

Firstly there is breakthrough from the power oscillator G1 through the circulator into the receiver and, small though it is, it exceeds the receive power by several orders of magnitude. This leads to what we have come to know in VHF contests - desense. The idea of using a totally separate transmitter with its own antenna does not help either because reflections from nearby objects and/or side-lobes of the antenna will still cause sufficient coupling of transmit power back into the receiver that blocking effects arise.

Secondly there is the thermal drift of the Gunn oscillator. The calculation of the Doppler frequency depends on a precise 30 MHz difference between transmitter and receiver. This, however, is difficult if not impossible to maintain. Just breathing on one of these oscillators will cause a frequency change of a couple of kHz!

It is possible that the introduction of dielectric resonators (5) and/or PLL circuits could bring the required stability. If only we could get the two oscillators to drift at the same rate, the Doppler error could be reduced to negligible proportions, as we saw in the Gunnplexer.

3.5 Blanking the transmitter using window techniques

A further possibility for improving things lies in keying the transmitter on and off and operating the receiver when we expect an echo from the target (Fig.22). The receiver would then maintain a defined window. During this window period the transmitter would be turned off and the near-target echoes would already be dying away.

How will the Gunn oscillator react to having its power supply keyed on and off? The frequency will presumably have to stabilise each time after switching on and

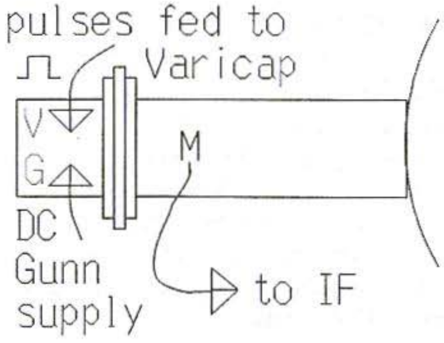


Fig.24a: **Another way of avoiding Rx overload by the Tx pulse is Periodic Frequency Inversion, which would allow an IF of 30 MHz, which wouldincrease sensitivity**

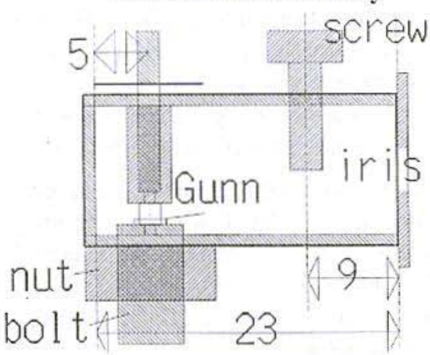


Fig.24b: **Side view of a Frequency modulated Gunn Oscillator Dimensions in mm.**

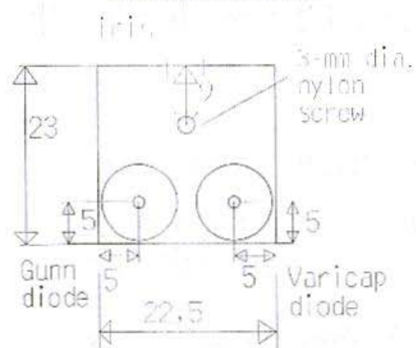


Fig.24c: **Top view of the same oscillator as in Fig.24b**

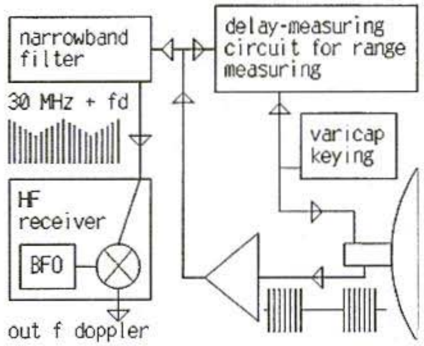


Fig.25: **Block Diagram of a CW Radar with Frequency Inversion based on the 'Blast-through' Mixer**

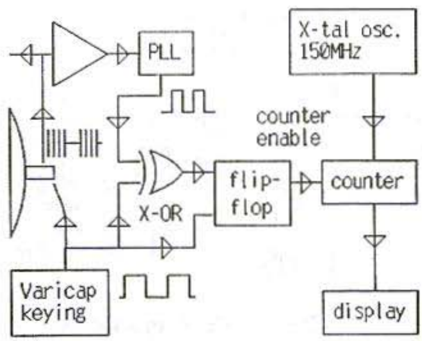


Fig.26: **Distance Measurement: the Counter is started with the first Pulsed Transmitted and stopped at the beginning of the Received Pulse**

would not be constant throughout the pulse. However, this will not do for Doppler measurements.

A PLL will not lock in within the space of a couple of microseconds.

If we achieve the keying of the transmitter with a PIN diode its reverse attenuation becomes a critical factor. If we attempt mechanical keying, for example with a

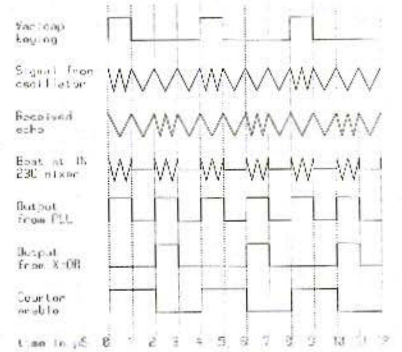


Fig.27: Timing Diagram from Fig.26

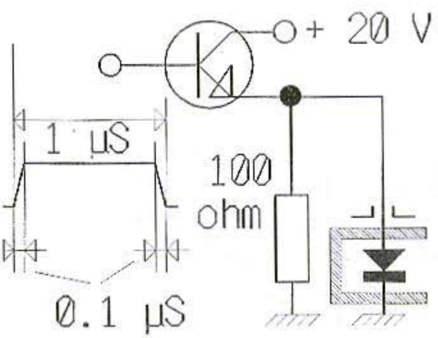


Fig.28: Pulses for the Capacity-Diode must have steep edges to enable the Microwave Frequency to be switched between the two values as rapidly as possible

propeller, the rapidly rotating rotor blades would bring about some "interesting" doppler effects.

3.6 Frequency keying

A better scheme would doubtless be to leave the transmitter G1 running all the time, but shift its frequency. It then transmits only short pulses with the frequency f and for the rest of the cycle is detuned by several score MHz by means of a varicap diode. This also avoids overloading the receiver while listening to echoes (Fig.23).

We carried these thoughts across to the blast-through mixer by equipping it with a varicap diode (Fig's.24a, b and c) supplied with keyed direct current. After this we are still left with no more than an oscillator and the problems of differential drift of the two oscillators remain.

With the knowledge that c = 300m in 1 microsecond, we know that our signal needs 2 microseconds to reach and return from a target 300m away. So in the time interval t we will switch a 20V impulse of 1 microsecond duration onto the varicap diode of the Gunn oscillator. This shifts the frequency of the oscillator by, for example, 30 MHz upwards, for one microsecond, then it reverts to its original frequency. We assume a mark-space ratio of 25 per cent.

Thus, we see on the IF side of the receive mixer a series of bursts with the frequency 30 MHz + doppler frequency. The period of the burst takes 2 microseconds which corresponds to a pulse repetition frequency (PRF) of 500 kHz.

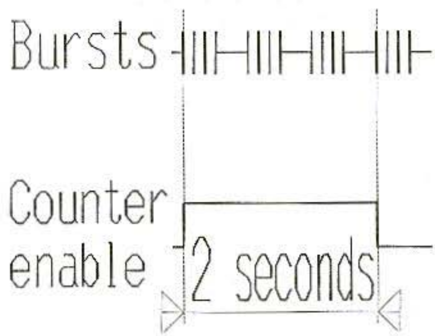


Fig.29: To increase accuracy the Counter can be enabled for several bursts

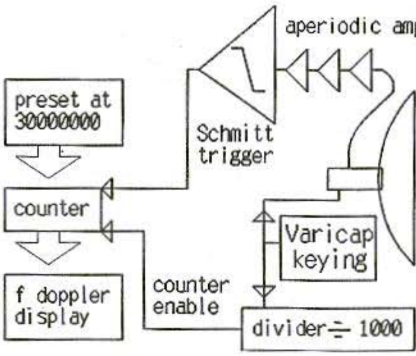


Fig.30: A Counter replaces the HF Receiver for speed measurement

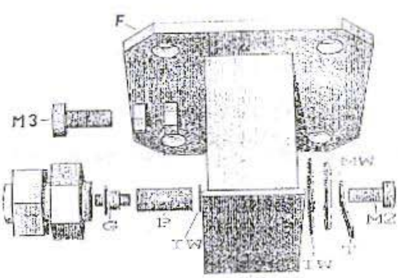
This signal is now amplified in a low-noise 30 MHz amplifier, passed through a narrow-band 30 MHz filter and cleaned up in this way, is passed to the HF receiver. It is still 30 MHz + the doppler frequency with a residual 500 kHz AM, whose level depends on reflections in the filter and its bandwidth. After mixing with the BFO signal we are left for evaluation an audio frequency that is exactly the doppler frequency. Fig.25 shows this scheme schematically.

3.7 Concepts for measuring distance gand speed

With the setup described above distance measurement can be carried out by evaluating the time between the pulse to the varicap diode and the reception of echoes. The principle is shown in Fig.26.

With the leading edge of the pulse that keys the varicap diode a counter is started; it is stopped again by the leading edge of the pulse from the EXOR gate (Fig.27).

With the knowledge that the distance r = t *c/2 it can be calculated that a measured time period of, say, 2 microseconds implies a distance of 300m. To get a figure-perfect display we let the counter



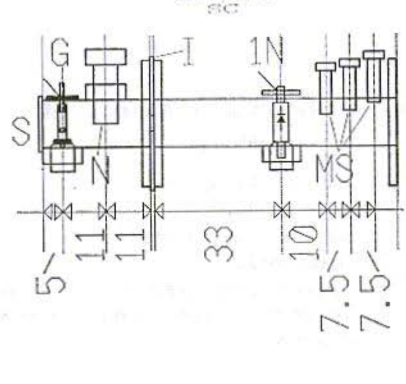


Fig.31a: The Gunn Oscillator in 3-D
Fig.31b: Dimensions of the Oscillator and Mixer

F = Flange
M3 = Metal or Nylon Screw, matching the bearing nuts
G = Gunn Element with M5 nut and bolt for adjustment
P = Metal bolt (copper or brass)
IW = Insulating Washer, inner and outer (PTFE)
SC = Waveguide Short-Circuit
T = Solder Tag
M2 = Steel Screw, M2
MW= Metal Washer
S = Short Circuit (on the plate soldered on the end of the waveguide
N = Nylon Screw, 3 to 6mm in diameter
I = Iris Plate (brass, 1mm thick with 6mm hole in the centre)
1N = 1N23 Mixer Diode
MS = Matching Screws (brass, 3mm diam)

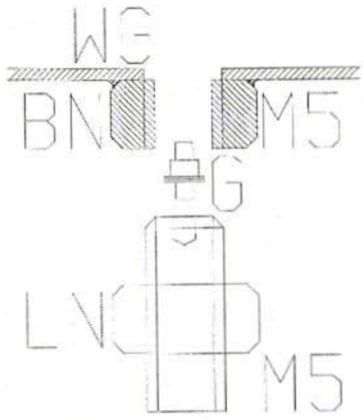


Fig.31c: Mounting details of the Gunn Element

WG= Waveguide (inside)
BN = Bearing Nut (M5, steel) soldered to the Waveguide
G = Gunn Element, larger flange grounded
LN = Nut (M5, steel)
M5 = Support Peg (M5, steel) bored axially for the end of the Gunn Element

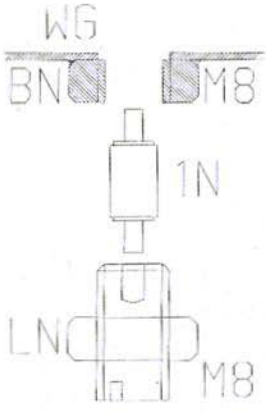


Fig.31e: Detail of the Mixer Diode inset

WG= Inside of Waveguide
BN = Bearing Nut (M8, steel) soldered to the Waveguide
1N = Mixer Diode
LN = Nut (M8,steel)
M8 = Support Peg (M8, steel) bored centrally for the Diode

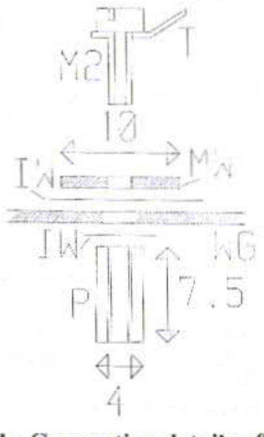


Fig.31d: Connection details of the Gunn Element

M2 = M2 steel Screw
T = Solder tag
MW= Metal Washer, 10mm outside diameter
IW = Inner/outer PTFE Insulating washers
WG= Inside of Waveguide
P = Round peg of brass or copper. Length 7.5mm, outer diameter 4mm, bored axially and tapped M2

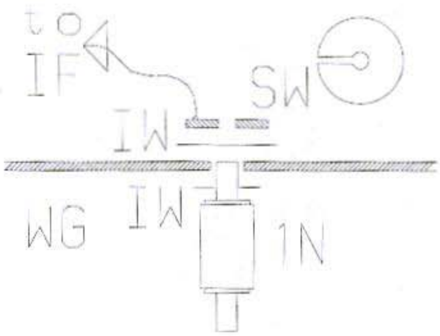


Fig.31f: Detail of the Mixer Diode connections

WG= Inside of Waveguide
IW = Insulating Washers (inner and outer PTFE)
1N = Mixer Diode
SW = Special Washer, fastened to the Diode connection and connected to the AF or IF preamplifier

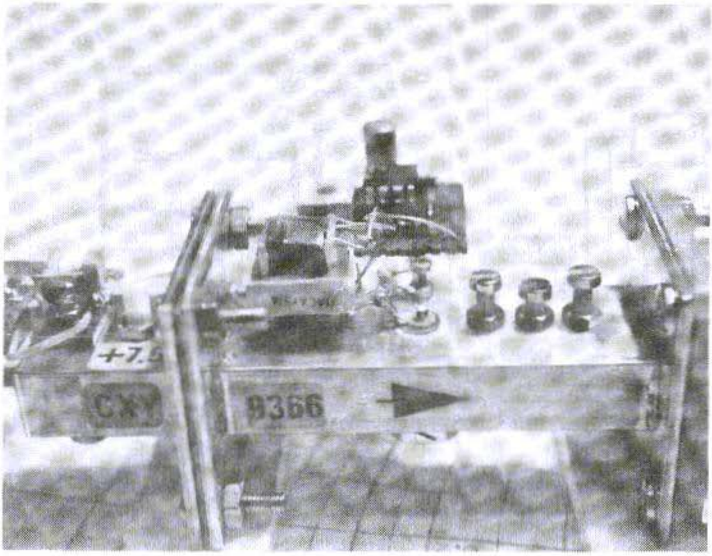


Fig.31g: Our SHF unit

count the pulses of a 150 MHz crystal oscillator: in 2 microseconds we get 300 pulses meaning that we can calibrate the counter display directly in metres.

If everything functioned perfectly - which of course it won't do - we could use this

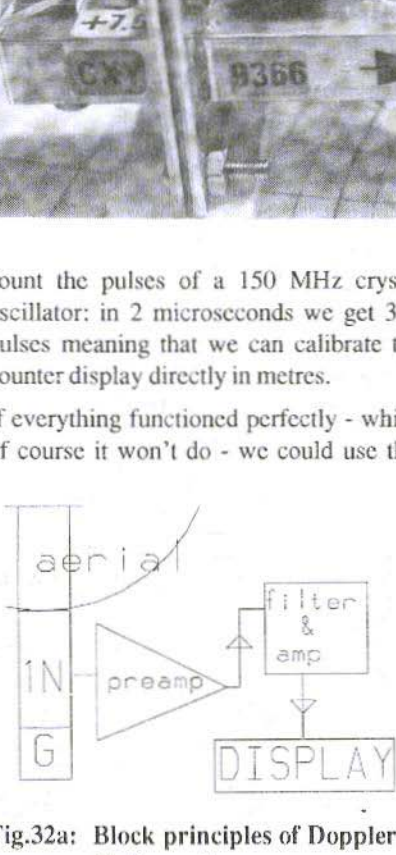


Fig.32a: Block principles of Doppler Radar system

Aerial =IKEA Dish with Penny Feed
1N = Mixer, connected to Antenna and oscillator by Waveguide
G = Gunn Diode
Preamp = see Fig.32b
Display = see Fig.35

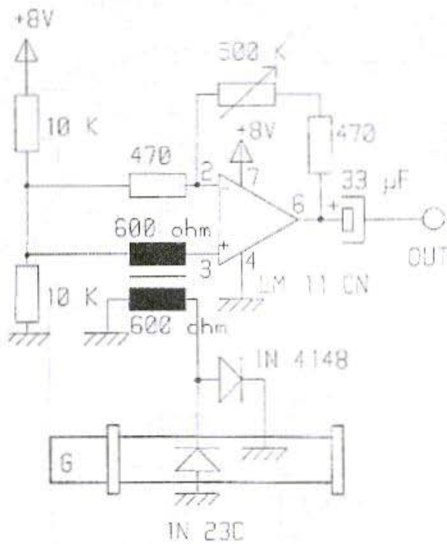


Fig.32b: Circuit of the Audio Frequency Preamplifier. The 1N23 is very sensitive to static electricity and is protected by the Switching Diode.

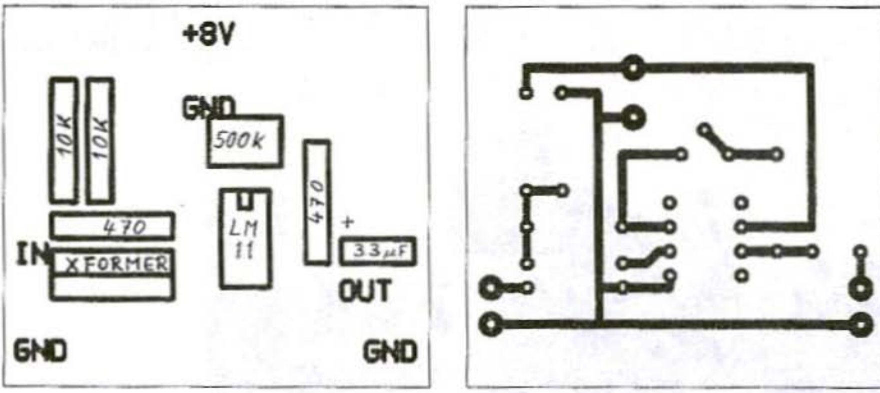


Fig.32c:Printed Circuit Board and Component Layout of the AF Amplifier

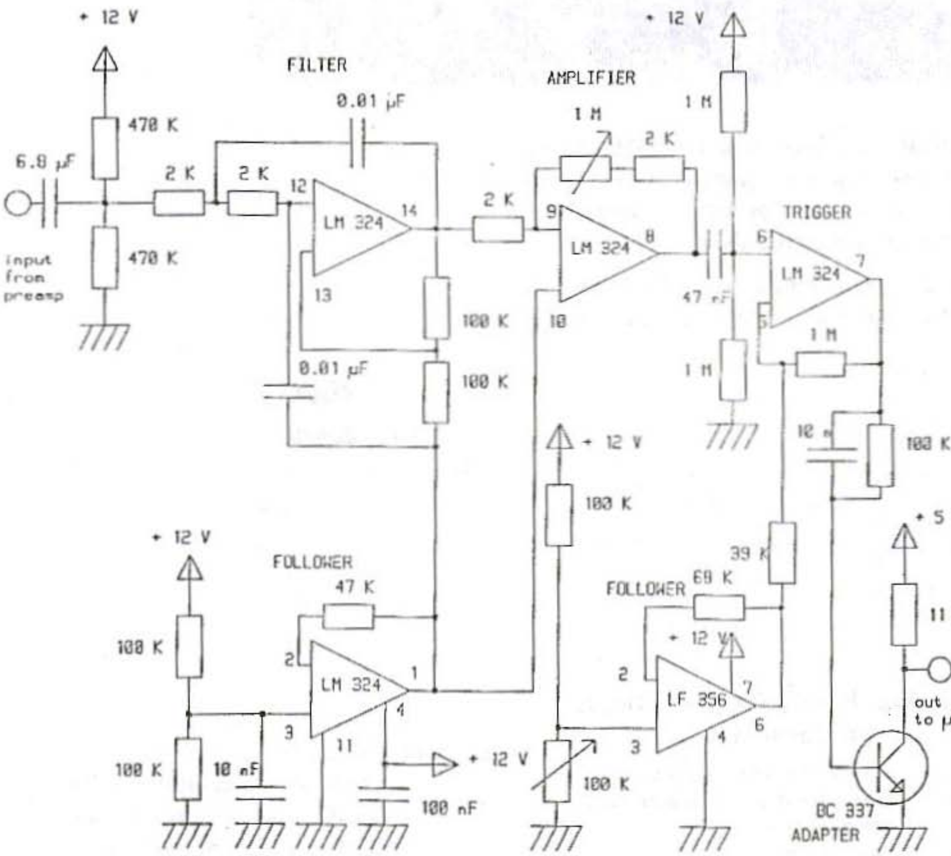


Fig.33a: Filter and Trigger Circuit

setup to measure distances from zero up to 500 metres. A lower pulse repetition frequency would probably give no additional range, even with a better 30 MHz IF unit barely more than 500 metres could be expected.

What we need to investigate now is how fast we can achieve frequency switching at 10 GHz. The 20V, 1 microsecond pulse for the varicap diode must have an internal resistance as low as possible in order that the pulse edges are as steep as possible (Fig.28).

We hope then that the 10 GHz oscillation (that's still 10,000 oscillations in one microsecond) follow on directly from the change in capacity, so that bursts as clean as possible are sent. During reception, when the signal is mixed down to 30 MHz, we shall only be counting 30 oscillations in 1 microsecond; 30 is not many.

The amplifier and filter circuit must be optimised with care if accurate results are to be achieved. We could spread out the frequency count over several periods, to increase the accuracy. The count result must be multiplied by two then, in order to compensate for the 1μs gaps between the 1μs bursts (Fig.29).

This measure would make the critical filter between IF preamp and HF receiver superfluous (Fig.30). The aperiodic amplifier with a necessary amplification of 100,000 to 1,000,000 would not be entirely uncritical, however. The frequency counter would count for 2 seconds and each time be preset by 30,000,000.

If we now measure the speed of a car travelling at 100km/h, it will have moved on 56 metres during the 2 seconds measuring time. It would be advisable then to select a shorter measurement window. In the deliberations that followed, an opportunity for this occurred.

Our setup can be simplified in two ways:

a) If the doppler frequency lies between

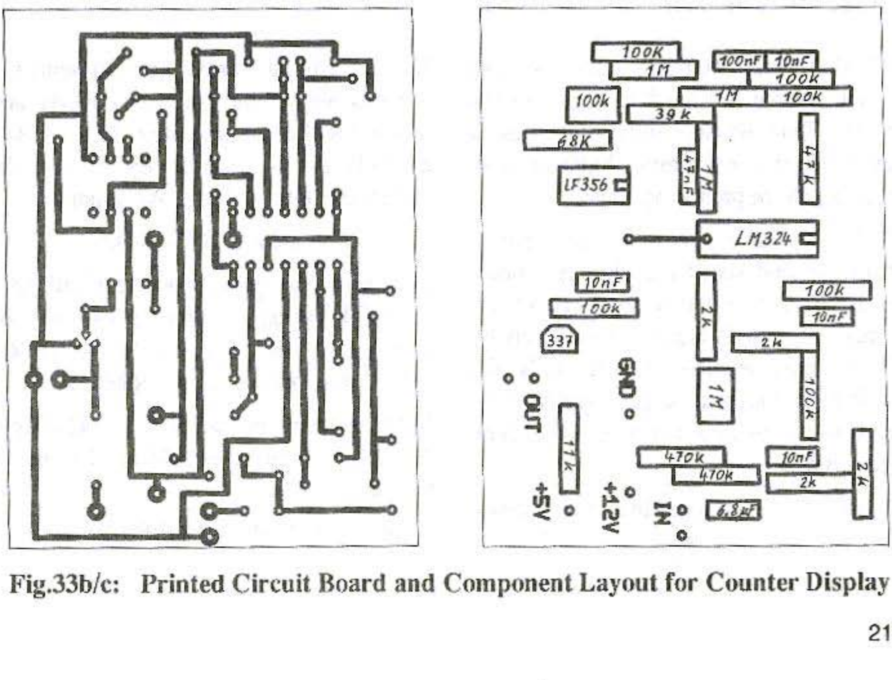


Fig.33b/c: Printed Circuit Board and Component Layout for Counter Display

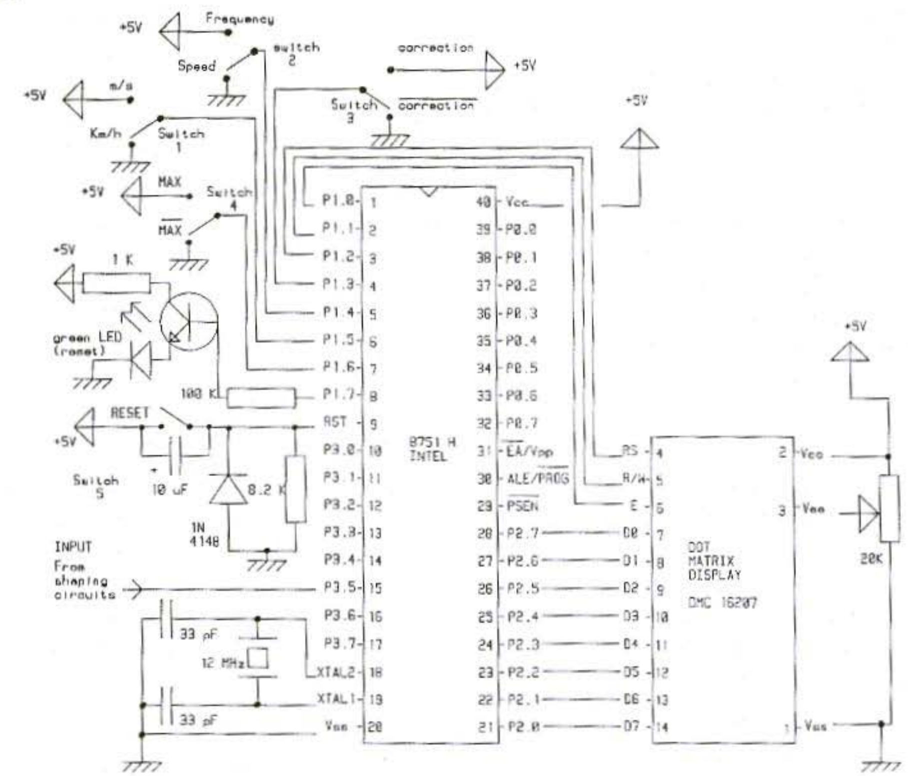


Fig.34a: Processor/Counter Display

30.000000 and 30.018000 GHz we can omit the three left-hand positions of the display. Our display, which only has to cover the five remaining digits, doesn't even have to be preset any more.

b) As we have seen that the doppler frequency and speed are directly proportional to each other, we can count the number of periods during an intentionally chosen time window and multiply the result by a constant, so as to produce the speed in the desired unit of measurement. The constant is:

$1.463 * 10^{-2}$ for a speed in metres/second

or:

$5.266 * 10^{-2}$ if km/h are desired.

If the counting period now measures 2 seconds we get the frequency in Hz and this gives the simplification that we can multiply this 2 seconds by one of the factors mentioned above. We count:

$$2s * 5.266 * 10^{-2} = 0.1053s$$

and the display gives us the result directly in km/h! During this short period the car driving at 100km/h has only moved 2.9 metres, and this is quite acceptable.

If we wanted the display in metres per second, the counter would need a gating time of:

$$2s * 1.463 * 10^{-2} = 0.029626 \text{ seconds.}$$

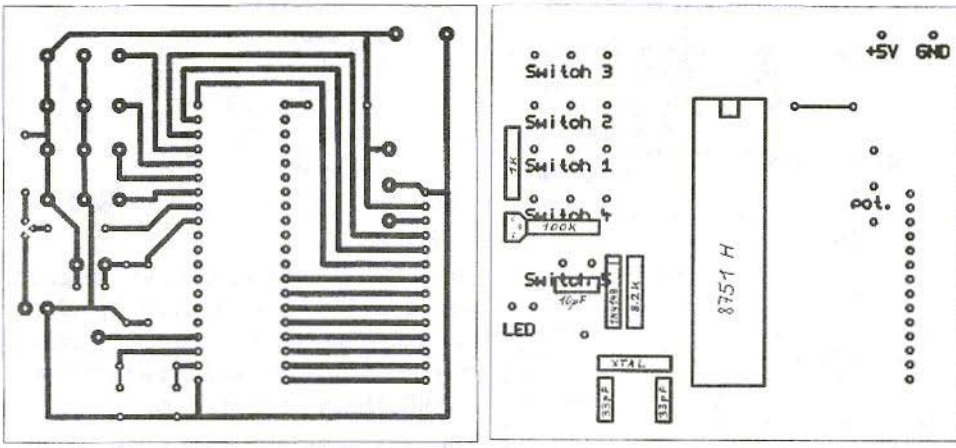


Fig.34b/c: **PCB and Component Layout for Counter and Display**

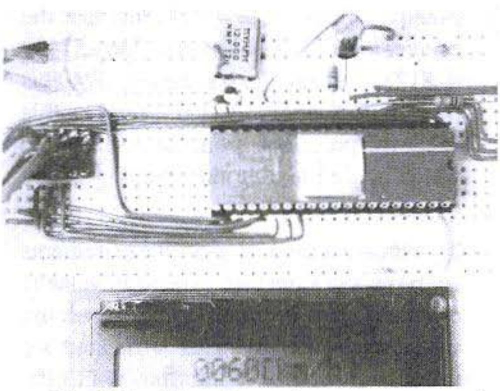


Fig.34d: **Doppler radar processor and Display**

4. CALCULATING SPEED

The dot-matrix LCD used (Fig.34a) needs to be driven by ASCII commands such as CLEAR, cursor position and others. For the calculating tasks we use an Intel 8751 microprocessor, which works with 8-bit words. With 8 bits we can count up to 255 (decimal), meaning that a doppler frequency of, say, 2000 Hz would be

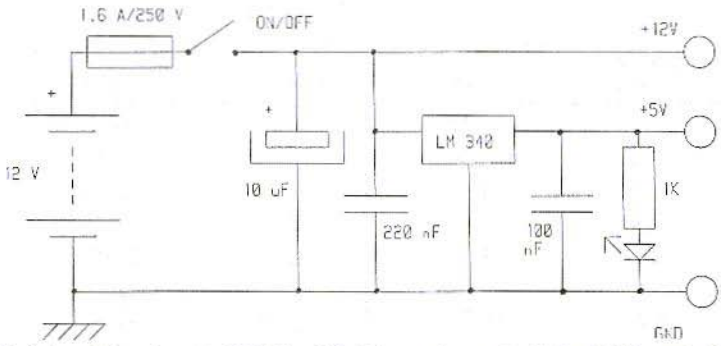
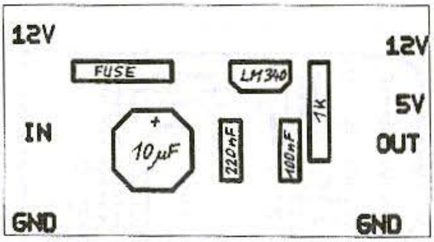


Fig.35a/b/c: **The simple PSU for Fig.34a and overleaf the PCB and Layout**



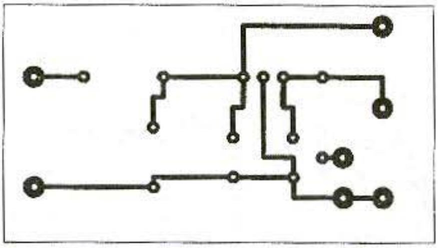


Fig.35b

Fig35c

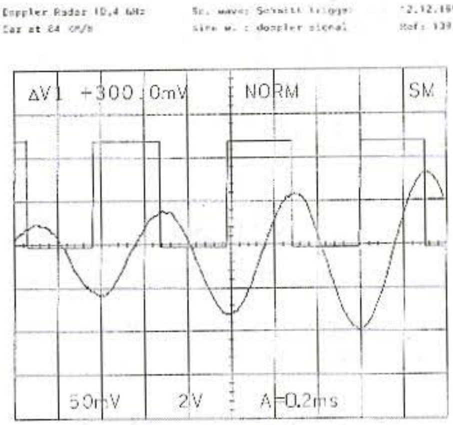


Fig.36: I/P and O/P signals of the Schmitt Trigger

too large. Fortunately, however, the timer embedded in the microprocessor uses 16-bit words.

The selected algorithm counts the incoming doppler pulses over a predetermined time span. With a Radar transmit frequency of 9.4 GHz (without doppler shift) the gate period is 15.947ms if we require speed in metres per second. If we use a different transmit frequency, the program must be modified for the new gate times.

The timer in the processor must be loaded with the complement of the gate time to 65536 ($=2^{16}$). That means, taking an example of gate time = 57409us, that the counter must be loaded with 65536 - 57409 = 8127. It then counts during 57409us from 8127 to 65536. When it reaches this value the processor produces an interrupt which is used to stop the counting of the doppler impulses.

To increase accuracy, eight measurements are taken and from these the mean value is calculated and displayed. The maximum values of all measurements carried out are stored and recalled if the time is insufficient for eight measurements. We assume that the highest value is the best and display this.

F input (Hz)	Display (km/h)	Calculated value (km/h)	Display (m/s)	Calculated value (m/s)
10	0	0.52	0	0.14
50	2	2.59	0	0.72
100	5	5.18	1	1.44
500	25	25.9	7	7.20
1000	51/52	51.8	14	14.40
5000	259	259.1	71/72	71.97
10000	518	518.2	143/144	143.95

Table 2

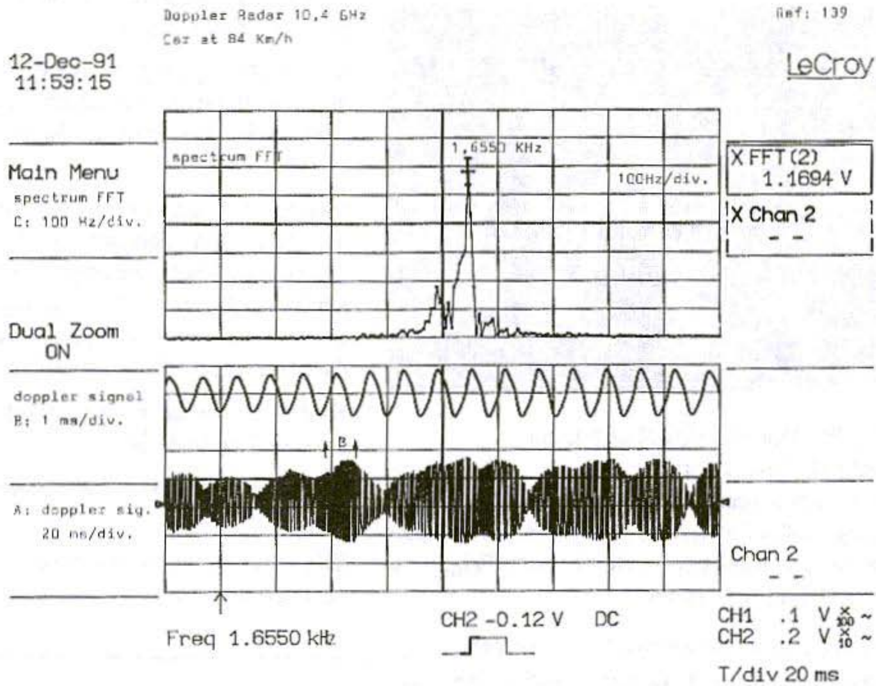


Fig.37: Lower curve shows Doppler signal of a car at 86km/h. Signal variations are clearly visible. Above this is represented extract B; the signal is deviating from a sine wave format. The upper display shows Fast Fourier Transform with a main frequency of 1.6550 kHz

5. PRACTICAL CONSTRUCTION

For our practical trials we wanted to use something simpler. An unkeyed blast-through mixer, very similar to the broad-band FM Gunn transceiver we already had, turned out to be good enough for a range of up to 100 metres.

Figures 31a to 31f show sufficient mechanical details for radio amateurs with some experience of these techniques to see what we have in mind.

The basis is a piece of R100 waveguide with internal measurements 22.86mm x 10.18mm. The Gunn element employed in the oscillator is a CXY11C. Adjoining this but separated by an iris window is the receive mixer with a 1N23C (Fig.31g).

The dish antenna used is by Sivers Lab and has a diameter of 36cm. This should give an antenna gain of 25dB at 9.5 GHz.

Fig's.32a to 32c show the entire electrical arrangements and the low-noise audio frequency preamp in circuit and layout. An LM11CN is used here.

Fig's.33a to 33c show the following stages: filter and amplifier, again in circuit and layout. After adequate amplification

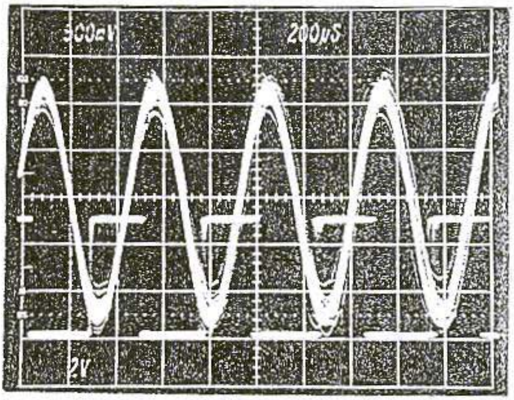


Fig.38: Doppler signal of a train
2030 Hz / 105,2 km/h
The high Radar cross-section value produces a steep (900mV/division) and low-noise Doppler signal; the pulses from the Schmitt Trigger are correspondingly clean

the signal is transformed into a rectangular form in a Schmitt Trigger. Finally this signal is passed to pin 15 of an Intel microprocessor which carries out the counting and calculation (Fig's.34a to 34d). The circuitry for the stabilised power supply is shown in Fig's.35a to 35c.

5.1 External connections

Five switches and one LED provide the user interface. They have the following operations:

Switch 1: toggling the display between km/h and m/s.

Switch 2: toggling between displaying doppler frequency and speed. When frequency is displayed switch 1 has no effect.

Switch 3: switches in a correction that takes into account the 33.6 degrees angle between the road and the Radar direction. If this is angle is neither zero or 33.6 degrees, then the software must be modified.

Switch 4: puts the highest value of the series on the display.

Switch 5: resets the microprocessor.

The LED is illuminated each time the micro is reset.

6. PRACTICAL TESTS

On the laboratory bench we were feeding in known frequencies directly into the low-frequency evaluation circuitry and obtained the speed results indicated in Table 2. As we can see, the accuracy of the counter is within plus or minus 1km/h.

6.1 Sensitivity

Table 3 shows the minimum input voltage needed at the preamplifier to produce a stable signal at the output of the Schmitt Trigger (Fig.36).

As can be seen, our direct conversion receiver is fairly deaf! Of course higher receiver sensitivity would increase the range of our Radar - this has been discussed already.

Frequency (Hz)	Amplitude (mV_{pp})
3200	1.4
7000	2.0
8400	3.0
11000	10
14000	100
20000	1000

Table 3

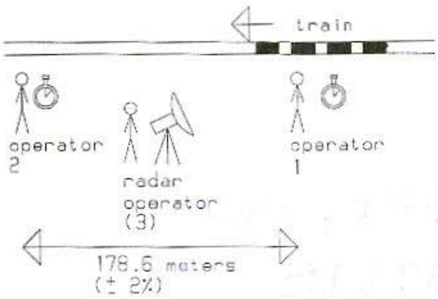


Fig.39: Free space test and Calibration setup

The screen shots (Fig's.37 and 38) show how the system works in practice. In addition the doppler signals produced by trains and cards are recorded at audio level on tape and later shown on an oscilloscope in the laboratory. It is particularly interesting to note here that real doppler signals are not as clean as the ones produced artificially in the lab, but on the other hand, they are not spurious readings.

Practical checking of accuracy and range was achieved as in Fig.39. When a train comes operator 1 signals the exact time the locomotive passes and operators 1 and 2 start their watches. When the locomotive passes operator 2, he gives a signal and they both stop their watches. The average of the two times is taken and operator three reads off the Radar display.

In three tests with varying speeds we found inaccuracies of from 4 to 6 per cent, the variations in the stopped time included. The maximum distance at which we could measure the train was around 100 metres. Possible improvements have already been discussed. If there is no Radar signal on the input, amplifier noise reaches the trigger stage and the display shows spurious readings. This spurious display could of course be suppressed with further software development.

To close we would like to offer cordial thanks to Pierre-Alain Glardon for his intensive and excellent co-operation in this project.

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Dr. Ing. Jochen Jirmann, DB1NV

Active Antennas for the Frequency Range from 10 kHz to 50 MHz

Active receive antennas permit broadband reception from long waves to the 6 metre band and are widely used in the commercial field (e.g. aboard ships). In amateur circles, however, they have acquired a reputation as "intermod generators" and their advantages are not recognised.

The aim of this article is to spread knowledge of some basics on the theme "active antennas" and stimulate people to conduct their own experiments, while keeping circuitry within limits.

For a long time the author has been investigating broadband receiving antennas for the long to short wave bands and first considered a damped vertical antenna of the type made commercially by Telefunken and described in (1).

Stimulated by an exercise in CQ-DL (2), some trials were made of active antennas, the results of which are described now.

1. RADIO PROPAGATION AND INTERFERENCE LEVELS IN THE REGION BELOW 30 MHz

In amateur radio practice it is customary to use the same antenna for transmitting and receiving, since we have all learned that "a good antenna is the best RF amplifier"! In the case of transmitting this is unreservedly correct; field strength at the receiving location increases by orders of magnitude with a correctly set up antenna with a corresponding directional diagram. On the receiving side matters look different and need to be considered from another perspective.

In the frequency range above, say, 100 MHz the quality of reception (i.e. signal-to-noise separation) is determined chiefly by internal disturbances (noise figure of the first RF stage) and antenna downlead

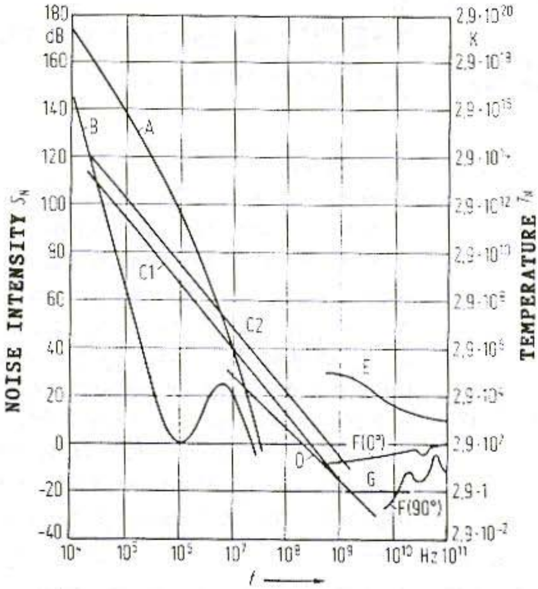


Diagram 1:
Average values of the various Noise Intensities and Temperature compared with Frequency

A = Atmospheric Noise (maximum value)
B = Atmospheric Noise (minimum value)
C1 = Industrial Noise (rural receive site)
C2 = Industrial Noise (urban location)
D = Galactic Noise
E = Quiet Sun
F = Noise from Hydrogen and Oxygen
G = Cosmic background Noise at 2.7K

attenuation. Under these circumstances a good antenna, which couples plenty of energy from the wave field, does make a real improvement in sensitivity. In addition, individual sources of interference can be nulled out with a directional antenna.

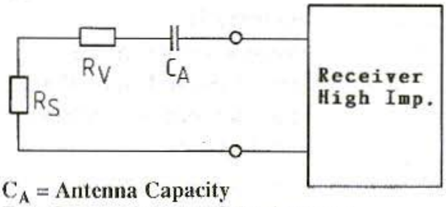
At frequencies below 100 MHz there is a steep rise in the interference level with falling frequency, as shown in diagram 1 from (3). In this the interfering components atmospheric noise, man-made noise and galactic noise (from outer space) are compared with frequency. As can be seen, the noise figure from external factors lies well above the internal noise of a receiver.

Since these sources of interference come in from no single preferred direction, they cannot be nulled out with directional antennas. With good antennas it is often the case that the combined level of wanted and interfering signals is too high for the receiver and the user must insert an attenuator ahead of the receiver input.

2. THE OPTIMAL RECEIVE ANTENNA

At frequencies below 100 MHz the task of a receiving antenna is thus not the production of a large signal but of a signal with the greatest possible separation from interference.

A transmitting antenna that is fixed on the roof of a house in a good spot for radiating almost always pulls in interference from the electricity mains. This leads to specific results in reception. If we recognise the fact that the interference field strength in the near field of a source of interference falls off with the square of the distance, then a remote receive antenna in the garden will certainly produce a lower level of signal but better separation from interference. With common shortwave receivers having a noise figure of 10dB, a whip



C_A = Antenna Capacity
R_v = Loss resistance of the Antenna
R_s = Radiating Resistance; this can be approximated for an Antenna of Length l to:

with $Z_0 = 377\Omega$

Fig.1: Equivalent Circuit of a Short Antenna

aerial about one metre long has so much receive efficiency that it will overcome the external interference level. A whip of this kind is easy and inconspicuous to erect.

Against far-field interference from terrestrial or atmospheric sources that kind of antenna is certainly no better than other designs, but a significant proportion of atmospheric disturbance arises from discharges from the air onto exposed antenna parts. In this respect a small antenna mounted low down is advantageous.

3. THE CONSTRUCTION OF AN ACTIVE ANTENNA

An antenna element that is short in comparison to the wavelength has an input impedance which can be described as a capacity (in the order of magnitude 10pF) with a radiating resistance of some milliohms to ohms, as shown in the equivalent circuit (Fig.1). For matching this to the cable and receiver input impedance of 50Ω a matching network is necessary. Following the rules of conventional RF technology, one would first make the antenna impedance real and then transform it to 50Ω.

As Fig.2 shows, the antenna capacity is first tuned with a loading coil to series resonance and the resulting real resistance transformed to 50Ω with the help of a transformation element (low-pass half-element). In practice the two inductances in series are included in the same coil. It is clear that this network operates only in a narrow band around the resonant frequency and must be changed each time the frequency is altered. The matching network is also suitable for transmitting use.

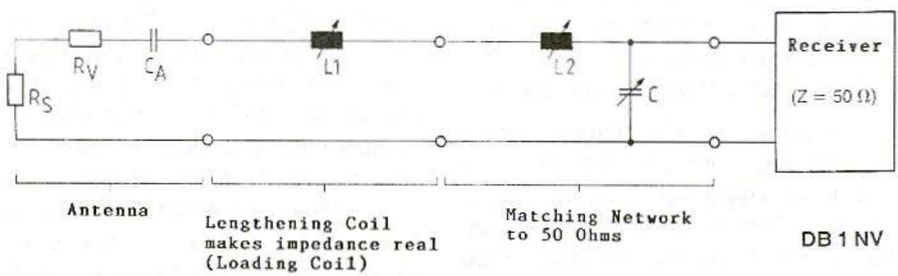


Fig.2: Matching a Short Antenna

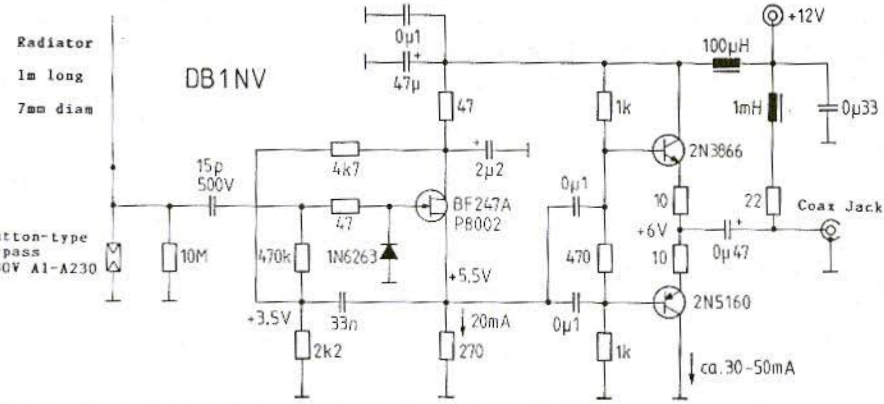


Fig.3: Circuit of an Active Antenna for VLF to VHF (* only required for remote voltage feed via coaxial cable)

The alternative is to take the no-load voltage of the whip antenna in a high impedance, low capacity fashion with an impedance changer and undertake the impedance transformation in a broadband manner using active components. The advantage is that one can make a receive antenna in this way that works broadbandedly from the longest waves to the lower end of the VHF region without tuning elements.

Since the impedance changer is working with signals across the whole frequency spectrum, it must operate extremely linearly to avoid producing intermodulation within the antenna. This is also no justification for the poor reputation that active antennas have in amateur circles. Most of the active antennas forced on amateurs have too small a dynamic range and include a preamplifier with 10 to 20dB gain following the impedance changer, which has to operate with the full signal spectrum and further decreases the dynamic range. In this there are close parallells with preamplifiers in the VHF/UHF region, where their effectiveness is judged by how far they bring up the S-meter and not how much they improve the signal-to-noise ratio.

4.
A SIMPLE ACTIVE ANTENNA

In order to gain an impression of the power capabilities of active antennas, the author developed some simple antenna electronics that consisted of just a barrier-layer FET and a push-pull emitter follower as cable driver. The circuit is shown in Fig.3. The antenna element used is a piece of 1/4'' diameter brass pipe about a metre long, but these dimensions are not critical. A 10 megohm resistor conducts away weak static build-ups. With heavy overload, for instance with a thunderstorm close by, a gas discharge lightning arrestor, wired in shunt, will strike: this is a so-called button spark-gap which conducts at 230 volts. Gas discharge arrestors are significantly lower in capacity than equivalent transient protection diodes and therefore load the antenna to a lesser degree. The arrestor type used on the prototype (A1-A230 by Siemens) is normally used in telephone systems as an over-voltage protection device and should therefore be simple to find. [Translator's note: a similar component is used inside telephone master sockets in Britain for high-voltage surge protection; Astralux of Horsham is a supplier.] As regards the effectiveness and necessity of this protection measure the author has not yet reached any conclusions.

The RF passes via a 15pF coupling capacitor to the gate of a barrier layer FET BF247A or P8002 wired as a source follower. By varying the coupling capacitor it is possible to influence the signal level within some limits. This coupling capacitor must be able to withstand at least 500 volts so that during voltage surges the arrestor can strike before the capacitor breaks down. The operating point of the source follower is tightly determined by DC feedback to around 20mA drain current; a bootstrap circuit achieves a high input resistance without the use of high-value resistors.

A protection diode guards against negative surges on the input that would drive the Gate to Source path into breakdown; with positive surges the Gate to Source path conducts and self-limits the over-voltage. This circuitry is copied from oscilloscope preamplifiers, where it is commonly used.

In this connection it's worth recalling that a MOSFET offers no advantages in this situation since MOSFETs are significantly "noisier" in the audio and low HF regions than barrier layer FETs.

The source follower supplies a push-pull emitter follower made up of complementary transistors 2N3866 and 2N5160. The quiescent current of the emitter follower lies in the region of 30 to 50mA, so common or garden small signal transistors would be overtaxed here. A 50Ω source impedance was dispensed with, since the input of the receiver connected afterwards would normally provide this. Completing the circuitry are a remote power feed splitter and some components for smoothing the power supply to the two stages.

In view of the low component count, construction of the active antenna can be made on a scrap of experimenter's printed circuit board having a continuous ground plane. The only thing to watch is a low-capacity path for the Gate of the BF247A, since any capacity at this point creates a voltage divider between the capacitance of the antenna and the input capacitance of the source follower, reducing the voltage available. The author enclosed his prototype in a round plastic container 40mm in diameter and 50mm long, that happened to be available; a fixing for the aerial whip and the coax socket was also provided. To make the whole arrangement waterproof, the complete assembly was placed inside a length of 50mm diameter cold water piping.

5. TEST RESULTS AND PRACTICAL EXPERIENCE

Since testing antennas in true far-field conditions exceeds amateur test capabilities, the following substitute measurements were undertaken.

- → The frequency range of the active section was determined by connecting a test generator directly to its 50Ω input; this showed a lower frequency limit of 7 kHz at -3dB and an upper frequency limit of 40 MHz at -3dB.
- → Signal limiting set in at an output voltage of 5 volts peak-to-peak, corresponding to an output power of around 60mW.
- → Output power for harmonic suppression of 60dB was determined at 10 MHz as 10mW.
- → The output power for 60dB intermodulation suppression at test frequencies of 10 MHz and 12 MHz was determined as +5dBm.

The two last measurements should be treated with caution since these tests were probably more of the harmonic and/or intermodulation suppression of the two test generators.

After connecting the antenna to the author's Kenwood R2000, the following tests were made to check the correct dimensioning of the antenna.

The whip was removed and with the receiver switched on, power was fed to the active antenna. This should give rise to a slight rise in noise level. This is a sign of correct functioning of the preamplifier and adequate sensitivity of the receiver. With older, less sensitive receivers a (selective) preamplifier can be called in to lift the antenna signal above that of the internal noise of the receiver.

The second stage was to re-connect the whip and here too, a slight increase of noise should be audible on a clear frequency. This indicates, as described above, that external interference is stronger than internal noise. If this doesn't occur, check all the voltages against those shown on the circuit diagram.

Since the antenna is designed for remote powering via the feeder co-ax, it is possible that undesirable interference from the power supply feeding the co-ax could be coupled into the antenna input; this would be detected by noticing a high noise threshold when the PSU is on but the active antenna is disconnected. In most cases (the R2000 included), an internal power source of 10 to 15V within the receiver can be used for this purpose, being led via a 1mH choke to the antenna connector. You should, however, check first whether there is DC on the antenna cable and if so, fit a blocking capacitor in the RF path.

For commissioning, the antenna was placed on the roof peak of the garage, about 10 metres from the nearest house. In practical tests on the HF amateur bands this antenna gave results practically identical to normal passive antennas, even if pure S-meter readings were more decisive.

This antenna is also suitable for long wave to short wave broadcast reception, which was noticeably less subject to interference than a long wire strung up in the vicinity of the house.

Even reception of various time signal and navigation beacons in the long wave region (such as DCF, LORAN-C and DECCA) come in at signal levels of some millivolts. A DCF or LORAN-C standard frequency device could be driven by one of these antennas.

The author hopes this article will stimulate readers to experiment for themselves and would be pleased to hear of their results.

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MES-FETishism II

The dual-gate MES-FETs have the advantage over GaAs triodes of very little feedback (although this cannot be neglected totally). We now present some amplifiers which make use of the special qualities of MES-FETs: not just reduced feedback but also low capacitance, high input and output impedance, high limiting frequency of transconductance and low noise in the range 1 MHz to 1 GHz.

Using the example of a preamplifier, it will be shown how the feedback effect can be reduced even further. All circuit examples use the selection of operating points described in the first part of this article series. With the addition of a few passive components other operating points can be chosen, if (for example) it is necessary to reduce current.

1. NON-AMPLIFIERS

When forward transconductance is not significantly larger than 20mS, amplification in 50Ω systems is at best unity. If one makes the input and output termination arrangements shown in Fig.1, then it is just 0.5 (-6dB). Even so, an amplifier of this kind still has some utility. It can for instance be used to shield the input of a circuit from any interfering voltages at the output of the previous stage.

The use of this kind of isolation stage are well known. Decoupling of oscillators from load variations and digital signals, input suppression of radiation from inter-

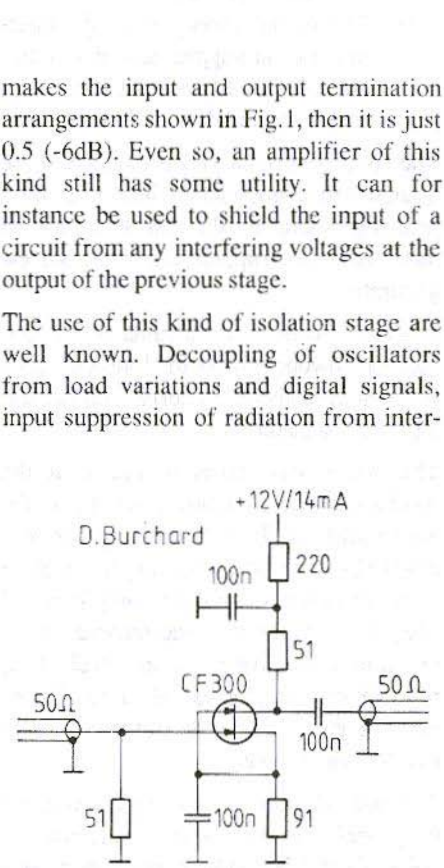

Fig.1: Isolating Amplifier with high isolation and 50Ω matching I/P & O/P

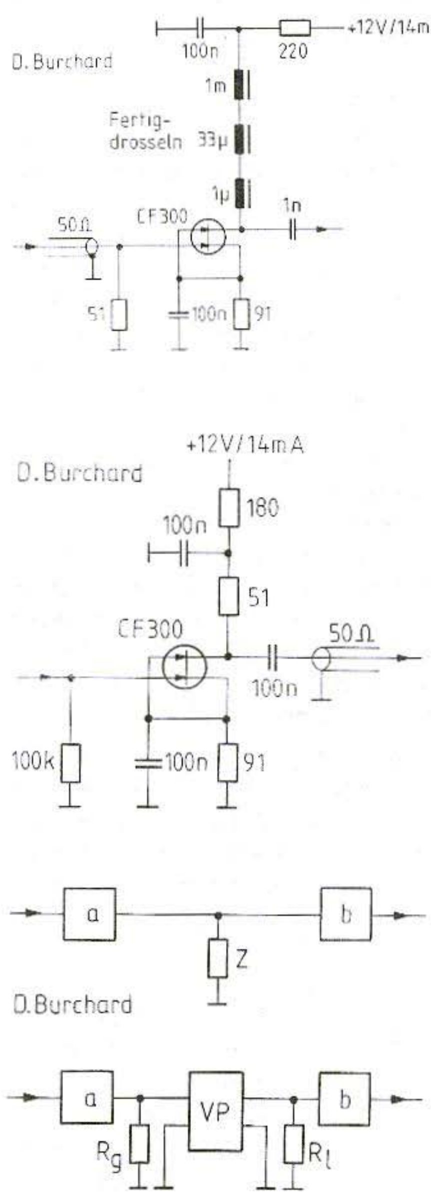

Fig.2: RF Current Source (a) and RF head (b) together with their adaption to 2-pole and 4-pole measurement (c)

Fertigdrosseln = ready-made chokes

ference voltages received from an antenna, decoupling of two signal generators being used in intermodulation tests and so on.

MES-FETs are called in when systems are working close to the limit, often the upper frequency limit. In a circuit such as Fig.1 these qualities can generally be determined without the need for network analysers or computer programs. The predominant determining factor with MES-FETs is the capacitance between input and output (C_{G1D}), on the input (C_{G1S}) and on the output (C_{DS}), as well as the frequency tracking of the forward conductance. If C_{G1S} is 1pF, then the upper frequency limit of the input is 3.2 GHz. That of the output is somewhat higher because C_{DS} is 0.6pF. The feedback capacity C_{G1D} is so small at 30pF that the apparent resistance due to capacity is only detectable above 1 GHz. Even there it is only a hundredth of the system resistance. With amplification factors below 1 the Miller effect plays no part either.

In reverse things look different. Then the feedback capacity causes a reduction in isolation with increasing frequency. With the previous figures we can reckon on an attenuation of 40dB at 1 GHz, in other words a reverse amplification of -40dB. An oscillator decoupled with an isolation stage like this would be decoupled from the load 34dB better than using a 6dB attenuator, assuming the same power available to drive it. The isolation defined in this way is thus

$$\mathrm{Isol/dB} = G_V/\mathrm{dB} - G_R/\mathrm{dB} \qquad (1)$$

At low frequencies the figures are even more impressive: Isol = 54dB at 100 MHz and 74dB at 10 MHz.

Silicon dual-gate MOSFETs have similarly low C_{G1D}, even among bipolar transistors one can find types designed specially for common base circuits, with a few tens of pF. And with "steam radio" valves, these values were far lower, for instance less than 7pF with an EF80. What do we gain then? The significant improvement lies in the much higher limiting frequencies of conductance: 50 MHz for the EF80, over 500 MHz with bipolar transistors and over 2 GHz for MES-FETs.

A gate circuit with MES-FETs, on the other hand, is not to be recommended as an isolating amplifier. This is because C_{DS} then becomes significant as feedback capacity and the isolation would drop by about 25dB.

The amplifier in figure 1 operates linearly (1dB compression) with 1V peak-to-peak input voltage, equivalent to $0.36V_{eff} \cong +4dBm$. According to the rule of thumb,

$$IM\,3 = P_{-1dB} + 15dB \qquad (2)$$

for S-shaped characteristic curves, we can expect its third order intercept point (IM3 point) to be +19dBm on the input.

Further applications are shown in figure 2. An RF current source (a) and an RF probe head (b), which facilitate measurements on high value resistance objects, are connected and properly matched with customary 50Ω measurement devices ahead of and behind them.

With the current source it would be possible, for instance, to stimulate a crystal with noise in order to discover all its resonances; the RF probe head maps out test voltages in the oscillating circuits. The combination of both (c) makes possible impedance measurements over a wide range. With a crystal, for example, both series and parallel resonance can be determined. Measurements on filters with abnormal and varying terminating resistances become possible, while these terminations are simply made connected parallel to the current source and RF probe. By using neat, careful construction arrangements it is possible to measure and sweep up to several hundreds of MHz. Relatively low impedance objects, too, such as SAW filters can be investigated up to 1 GHz.

A high-quality RF probe with times 1 amplification, negligible input capacity and very high bandwidth is shown in Fig.3. This gives sufficient spare amplification for effective inverse feedback. The Source and Gate 2 are connected in such a way

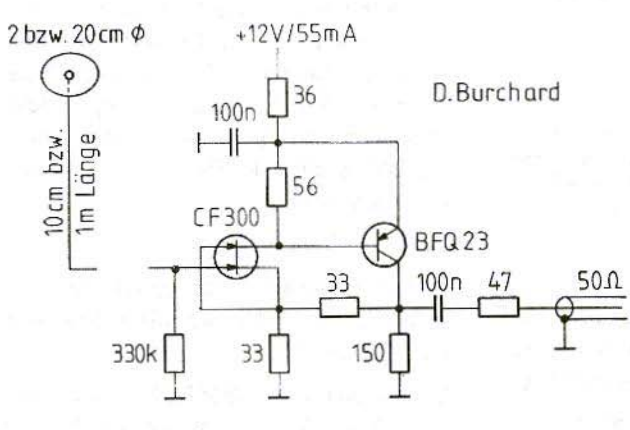

Fig.3: A top-quality Head with very small input capacitance, together with an Antenna for precise Field-Strength measurements in the HF and VHF region

Antenna:
Diameter 2-20cm
Length 10cm-1m

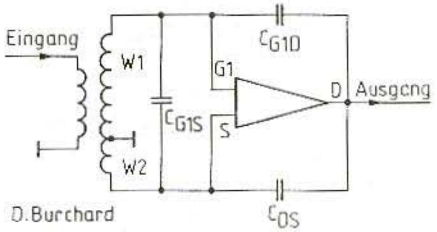

Fig.4: Neutralisation Bridge for MES-FET Amplifier

Eingang = I/P; Ausgang = O/P

that their capacities on Gate 1 become ineffective. With the transistors indicated, miniature components and "dead bug" construction techniques a flat frequency curve can be achieved up to over 500 MHz. Using surface-mount components (use the CF930 and BFT93) over 1 GHz can be achieved. This is against a declining reserve of amplification.

A piece of wire 10cm long with top-loading capacity at the input of the RF probe, as sketched in Fig.3, produces an active antenna for the VHF region and more, which can be put into service for field strength measurements on account of its constant conversion factor. For the HF region (3 to 30 MHz) the wire can be 1 metre long, which will give tenfold sensitivity.

2. PREAMPLIFIER FOR THE TWO METRE BAND

The quality factor of low noise mentioned at the outset has not been mentioned again up to now. We get low noise figures with MES-FETs, even in non 50Ω systems, because its equivalent noise resistance lies around several hundreds of ohms. Connor (2) gives a rule of thumb for the smallest noise figure achievable with common source circuits:

$$F_{min} = 1 + 0.52 \frac{\omega \cdot C_{G1S}}{S} \qquad (3)$$

For the type CF300 we get astonishing figures:

F_{min}	f
0.01dB	10 MHz
0.07dB	100 MHz
0.14dB	200 MHz
0.34dB	500 MHz
0.66dB	1 GHz
1.23dB	2 GHz
3.08dB	5 GHz

Up to now figures like these have not been achieved in practice and at the lower frequencies with MES-FETs. Nor is it worth attempting frequency records here. The calculated value for 1 GHz seems to be not too far removed from the actual one. The discrepancy becomes larger with decreasing frequency. At least below 200 MHz a sub-1dB noise figure is not generally necessary. It appears to be very difficult at low frequencies to achieve good input matching for F_{min}. The manufacturers of MES-FETs are somewhat reticent about further figures, so there's a broad field for individual research both with and without a computer.

Successful designs, by Angle (1) and VHF Communications editors (3), are not considered why they are proportioned and exactly how. Here I shall first broach some

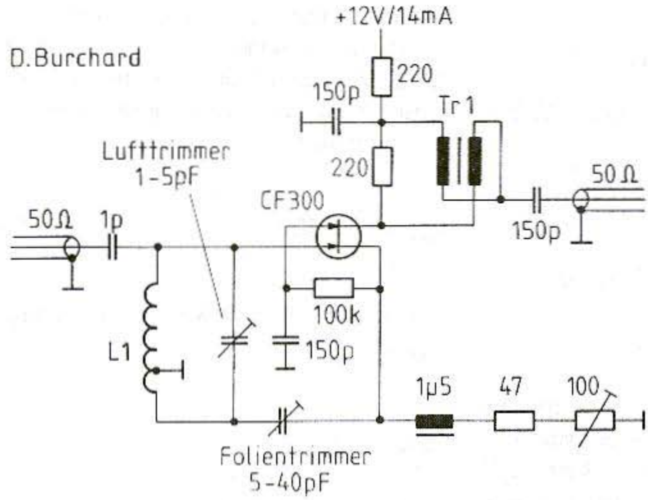

Fig.5: Circuit for Preamplifier for the 2m band using Interbase Circuitry

Lufttrimmer = Air Trimmer; Folientrimmer = Foil Trimmer

aspects important to me, then explain as well why the selected circuit came about. And naturally the test results achieved will be documented and discussed.

In today's radio landscape a preamplifier is always a large signal amplifier as well. The narrower the bandwidth of the input circuit, the more it is burdened with strong, out-of-band signals. So bandwidth needs to be balanced against noise figure. Strong interfering signals inside the band can only be tamed by good signal handling (high IM3 point). We need to select a circuit and an operating point that can handle a high level of overload. MES-FETs turn out to be worse here than V-MOS triodes with an operating point at I_D = 0.5A (usable up to around 200 MHz), because of their typical drain current of just 10 to 20mA. In the third part of this series power amplifiers will be discussed and this aspect will be examined further.

Noise and power **tuning** of the input circuit do not generally coincide. This is almost always a sign that feedback from the output is influencing the input. The small feedback capacity of MES-FETs is noticeable if one has to use a high degree of input transformation for good noise matching. Neutralisation creates a remedy. This has already been used by Laufs (4) and was the state of the art in the valve era. With MES-FETs self-neutralisation sets in as shown in Fig.4 whenever the following is valid:

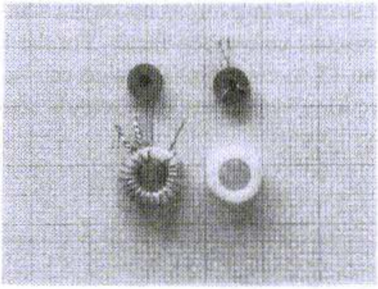

Fig.6: Toroidal Cores Tr1 (upper) and L1 (lower) wound and unwound for the circuit of Fig.5.

L1 = Teflon ring 9.5 x 6.5 x 4mm, 1+20 turns
Tr1 = Ring Core 6 x 2 x 2mm made of Fe810 3+3 turns 0.1mm CuL, drilled non-inductive

$$\frac{W_1}{W_2} = \frac{C_{DS}}{C_{G1D}} \equiv 20 \qquad (4)$$

Noise and power **matching** also seldom coincide. People have therefore discovered the reciprocal rule of matching: instead of the correct matching of an antenna cable, a correct source impedance is supplied. Standing waves in the cable and the additional losses caused by them are still present of course but at least energy is no longer moving back and forth in the cable; the reflected energy is re-radiated by the antenna. During measurements the amplifier, connected to the signal generator, behaves as if it was connected to an antenna.

With a substitute antenna, having a bandwidth narrower than the input circuit, the amplifier behaves completely unpredictably! The remedy is an inter-base circuit, which in its design agrees completely with Fig.4, although in general it demands a different w_2/w_1 behaviour. By means of additional capacitors parallel to C_{G1D} or D_{DS} simultaneous neutralisation **and** coincidence of noise and power matching can be achieved.

For an after-amplifier the same reciprocal matching rules should hold good. In the simplest case we give the preamplifier an ohmic Drain resistance of 50Ω If the degree of amplification turns out too small a broadband transformer is necessary. For frequencies up to 500 MHz this can be made from ferrite ring cores and windings drilled through them, giving whole-figure step-up ratios. If this ratio is 2:1, then the Drain resistance is 200Ω(use 180 or 220) and the amplification is doubled. If the ratio is 3:1, then the Drain resistance has to be 450 (430, 470) Ω and the amplification is threefold. You shouldn't go much higher because the Drain resistance will exceed the optimum otherwise and the overload ability will be lost. A tuning circuit on the Drain only makes matters unnecessarily complicated. You don't know what bandwidth the receiver connected has.

The preamplifier was eventually designed as shown in Fig.5; it was built and thoroughly tested without any further optimisation work, apart from checking its self-neutralisation. A couple of special points remain to be mentioned. Coil L1 is wound on a teflon ring as a toroid. That gives a similar figure of merit (200) as a cylindrical air coil, though with much less stray radiation. Screening between input and output circuitry is not needed. Both circular cores are shown in Fig.6.

The feed leading to the Source has measurable inductance, which is augmented by stray inductance of the input coil. Both are series-compensated by the foil trimmer.

If a signal is now introduced at the output (Fig.7), we can then determine the reflection factor ρ on the output side, also the backward amplification G. ρ turns out to be 0.23 over a very broad range of frequencies. To turn this round into VSWR we use the formula

$$\mathrm{VSWR} = \frac{1 + \rho}{1 - \rho} \qquad (5)$$

producing a VSWR of 1.6. The reverse amplification is close to -50dB, already above the upper limits of the test set-up. Without self-neutralisation the figure would be -20 to -30dB.

Forward measurements are made with two different resolutions on the frequency axis.

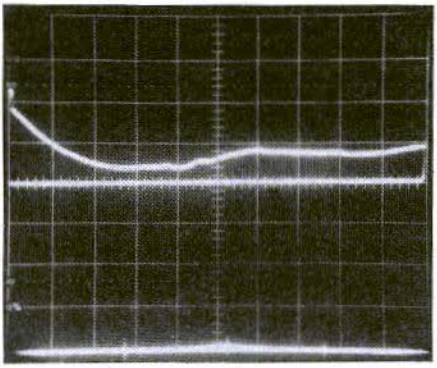

Fig.7: Output reflection Factor and Reverse Amplification

Y1 = Reflection Factor, $\rho = 0$ is highlighted, $\rho = 1$ is 3.5 div's higher
Y2 = Reverse Amplification, reference line -50dB is highlighted, 5db per division
x = 50 to 250 MHz, markers at 50 MHz

Fig.8a gives the forward amplification at the middle of the band as 19dB, the 3dB bandwidth as 4 MHz and ρ less than 0.45 within the 1dB bandwidth., i.e. VSWR equal to or less than 2.6. The best value of ρ (value of VSWR) at the centre of the band is 0.17 (1.4). The values appear to be satisfactory and no further effort was made to improve power matching. According to Fig.8b, amplification dwindles monotonously fast, though less steeply, at the upper frequency end due to the capacitive inward coupling. The peaks visible at 50, 75, 120 and 180 MHz arise from the harmonics and sub-harmonics of the broadband sweep generator used.

Finally noise measurement was investigated, using a laboratory receiver with 7dB noise figure and an active AM demodulator connected behind the preamplifier. The result can be seen in Fig.9. From the relationship of the noise voltages measured (2.6 : 1.95 div) we can calculate F_{total} = 1.1dB and for the preamplifier alone F_{preamp} = 0.9dB. That is still not astonishingly low but not a bad result considering at the same time the isolation achieved of nearly 70dB according to equation (1) and the insensitivity to virtually any mismatching on the input or output.

The total amplification is given by the transformation in the input circuitry and

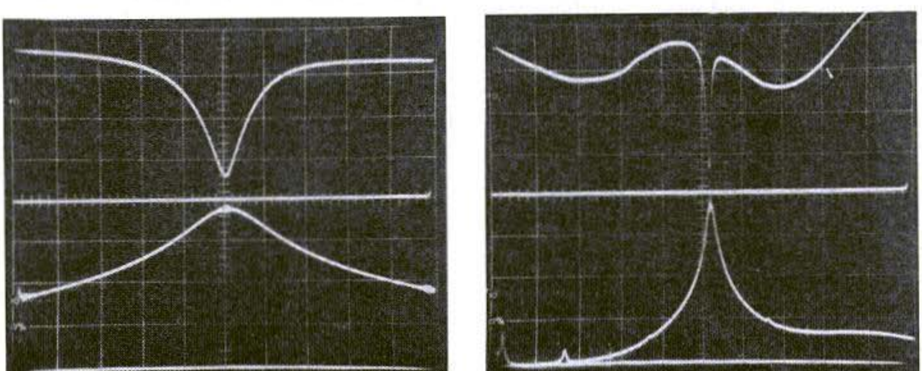

Fig.8: Input Reflection Factor and Forward Amplification

Y1 = Reflection Factor, is 0 is highlighted, = 1 is 3.5 divisions higher
Y2 = Forward Amplification, 5dB per division, reference line equals Input Voltage
X = (a) 140 to 160 MHz, markers every 10 MHz; (b) 5 to 250 MHz, markers every 50 MHz

$S.R_D.1/ü$ together. Since the second term is close to 1, the first must be approaching 9. The point for 1dB compression therefore lies around 35mV input voltage (-16dBm), as would be expected from its linear controllability. The IM3 value is recorded as -8dBm at the input, which represents a noticeable variation from equation (2). Both measured values, P_{-1dB} and IM3, are dependent on each example constructed and on the operating point chosen. A trimpot resistor in the Source line might make optimisation easier.

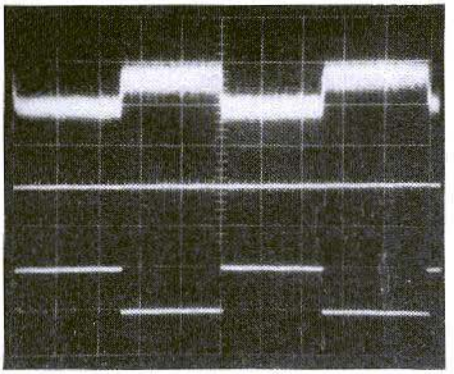

Fig.9: Noise measurement with Noise Source 0dB ENR with ultra-linear DC rectification and low-pass filtering with f_o = 1 kHz

Y1 = 0.2V per division, rectifier zero point is highlighted
Y2 = 5V per division, logic switching signal of the noise source, L ≡ ON, H ≡ OFF
X = 10ms per division

3. OSCILLATORS

Considering their low capacitance and high upper frequency limit, one might expect MES-FETs to be exceptionally well suited to building oscillators for the GHz region.

Sadly this is not the case. Two characteristics of MES-FETs stand in the way: their string scintillation noise and the significant dependence of capacitances on the voltages. The scintillation noise (also known as 1/f noise) sets in already below 1 MHz and in the audio frequency ranges reaches values around ten times higher than barrier-layer FETs and bipolar transistors.

Unwanted phase deviation arising from modulation phenomena in the oscillator achieves correspondingly high values.

Admittedly a CF300 has been used in an oscillator by Reuschle and Shah (5) but without any indication on the selection of the of device or how the problems just mentioned were overcome. My own measurements gave for example at 1.5 GHz an interference frequency deviation (de-emphasis 50µs, CCIR468 values) of many kHz, whilst bipolar devices produced a few hundred hertz.

There are, however, applications where this interfering deviation is not a problem or can be eliminated (TV and intercarrier techniques).

The voltage dependency of capacitance is clearly connected with the small geometric structure of the MES-FET and the formation of Gates as Schottky diodes. We need to stabilise therefore the operating voltages and ideally the operating points as well as possible by using high positioned Gates and long Source resistors, as I have previously recommended here, when building oscillators.

Since test circuits have proved themselves to be more temperature-sensitive than equivalent bipolar set-ups, I have not carried on this work, so no tested examples can be demonstrated here.

4. CONCLUSION

There is a range of applications where using MES-FETs gives better results, extending the frequency region upwards by an octave or more.

At low frequencies, too, they show advantages where very low input and output capacitance is called for. MES-FETs have theoretically very low noise figures, which in the centre frequency ranges are not achieved in practice. The strong 1/f noise makes their use difficult at low frequencies and in oscillators.

5. LITERATURE

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VHF Communications 2/82, pp.77-80

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Funkschau 21/89, pp. 61-65, Franzis-Verlag, Muenchen

Wolfgang Schneider, DJ8ES

SSB Transceiver for 50 MHz using 50Ω Modules

Part-1

(Revised version of the presentation given at the 1991 Weinheim VHF Convention)

Among the various popular operating modes in amateur radio, SSB-DX is one of the focal points. On the frequencies available from 1.8 MHz up to well into the GHz region this mode in particular offers many opportunities for home-brew construction.

Based on good results using 50Ω modules, the author has realised an old dream. This 50 MHz transceiver has been carried through unaided from planning, all the way through conceptualisation, to realisation; it therefore earns with justification the title "home-made".

A number of relationships determined the author's development work. Thus the transceiver was not going to serve only on the 50 MHz band but also to drive existing transverters to 144, 432 and 1296 MHz. The concept diagram (Fig.1) should illustrate this point.

On account of its universal conception (that is, using broadband 50Ω modules), the transceiver described is designed not just for 50 MHz. With appropriate modification to the frequency determining parts (e.g. filters), the transceiver can be put into service on all HF bands and also on two metres.

This article does not go into all the minutiae of construction, rather it tries to present the keen amateur with a viable route to homebrew construction. Apart

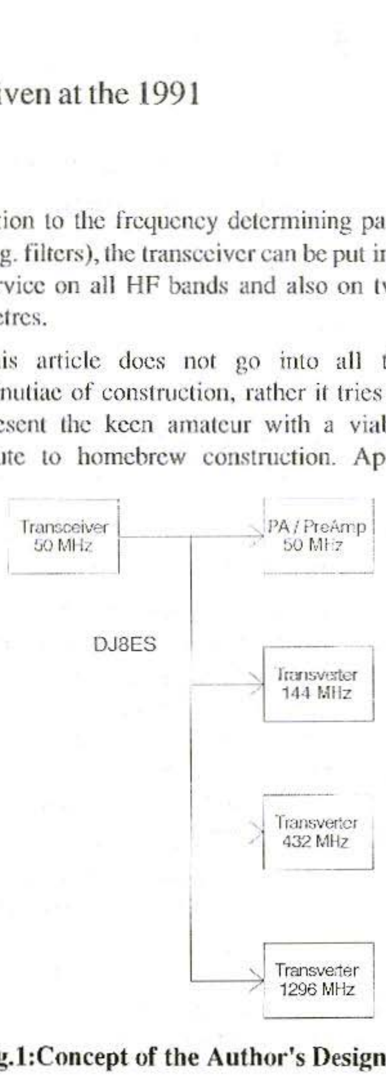


Fig.1:Concept of the Author's Design

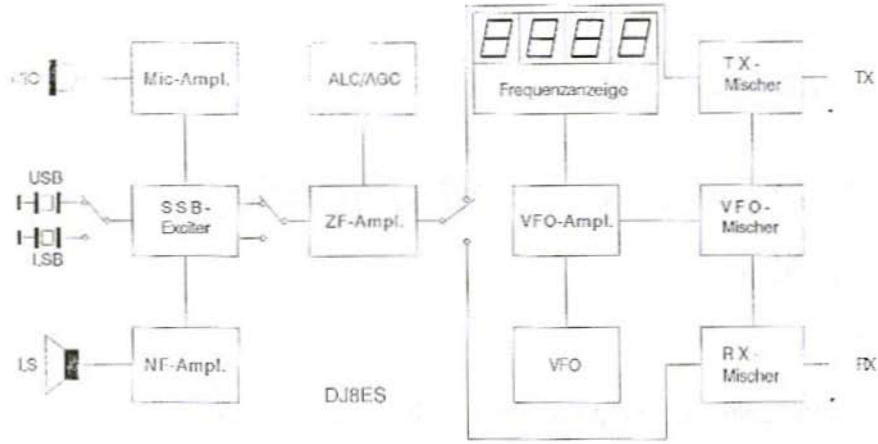


Fig.2: Block Diagram of the 50 MHz Transceiver

Frequenzanzeige = Frequency Display; Mischer = Mixer; NF = Audio; ZF = IF

from a few special components, all other parts should already be in the junk box somewhere.

1. THE CIRCUIT CONCEPT

In the block diagram (Fig.2) the interaction of the individual elements can be seen.

The received signal arriving from the antenna at the RX input is amplified in the RX mixer, passed through a filter and then transposed to the IF level (9 MHz). For that the module requires an oscillator signal tunable between 41 and 42 MHz. In this way the range 51 to 52 MHz is covered.

The oscillator signal needed is provided by the VFO unit. The tunable oscillator works in the region from 5 to 6 MHz. The amplifier following operates as a buffer and at the same time decouples the frequency counter and VFO mixer from one another. Mixed with a 36 MHz crystal oscillator, it produces the desired output frequency of 41 to 42 MHz.

The digital frequency display is executed in eight digits, so the last digit shows hundred Hz units. This counter evaluates the VFO signal directly. To achieve a frequency-accurate display the difference from the wanted frequency (i.e. 45.000 MHz) must be programmed.

The received signal now reaches the IF amplifier. Here the crystal filter limits the bandwidth to the 2.3 kHz necessary for SSB.

For the field-strength display (S-meter) the received signal from the AGC/ALC module is evaluated. At the same time this is used to set the regulator voltage for controlling the IF amplifier.

For demodulation the SSB exciter is employed. In this process, by activating the corresponding crystal oscillator, the desired sideband is selected. At the output of the ring demodulator the audio achieved in this way is passed to a low-pass filter

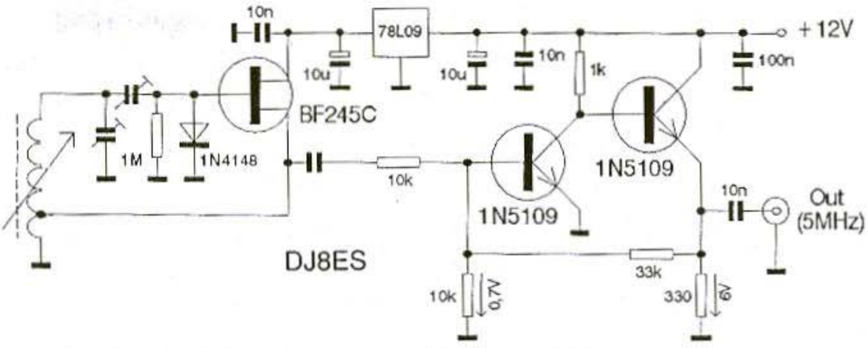


Fig.3: Hartley Oscillator with two-stage Buffer Amplifier

and a two-stage amplifier. The power amplification needed for loudspeaker operation can be undertaken in a separate module.

During transmitting, speech from the microphone is amplified and mixed in the SSB exciter with the desired sideband oscillator. At the output we get a double sideband (DSB) signal. The switchable IF amplifier is used in the transmit path. The crystal filter selects the desired sideband. By using an AGC/ALC circuit we can avoid over-driving the following stages.

The transmit mixer transposes the SSB signal already produced up to 50 MHz. Like the receive mixer, this contains filters and amplifier stages. At the output of the TX mixer we have around 100mW available on the desired frequency.

2. THE OSCILLATOR UNIT

2.1 The tunable oscillator (VFO)

In the transceiver described here an oscillator from a Collins receiver is used. VFOs of this kind or similar ones can be found at radio rallies and boot sales.

The VFO operates with the Hartley circuit; this ensures good stability. The oscillator gets reverse feedback from a tap located about 10 to 25 percent away from the earthy end of the coil.

The higher the slope of the FET used, the less feedback is needed. This should only be enough to ensure foolproof operation of the oscillator. To much feedback leads to instability.

The oscillator is tuned using the core of the coil; this makes it a so-called PTO (permeability-tuned oscillator). The tuning range is about 5 to 6 MHz.

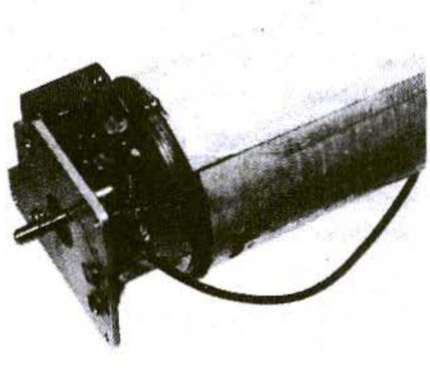


Fig.4: The complete VFO with encapsulated coil for Resonant Circuit

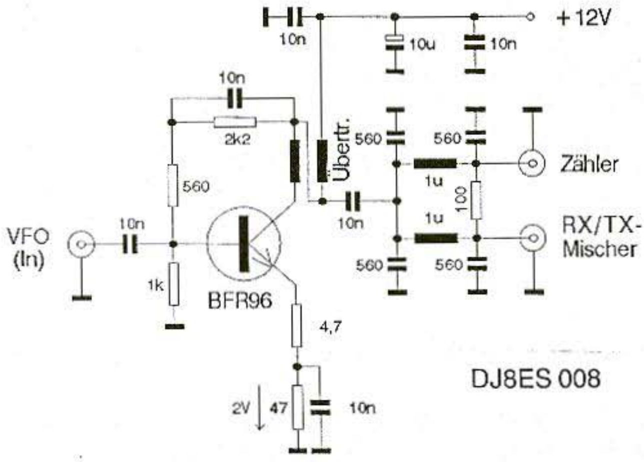


Fig.5: Broadband Amplifier

Mischer = Mixer
Uebertr = Transformer
Waehler = Counter

To decouple the following stages a two-stage push-pull amplifier is used (Fig.3). This buffer stage ensures feedback-free operation of the VFO.

2.1.1 Construction hints and Commissioning

As the photo shows, the complete oscillator and buffer stage are built up "in mid-air". The feedthrough capacitors used for RF decoupling also serve as solder tags. As far as possible, the use of special components was avoided here and in other parts of the circuit; this applies particularly to semiconductors. The low-cost BFX89 gave repeatable results in the buffer.

Using the components indicated, the module delivers an output power of 0.5mW into 50Ω. With a supply voltage of 12V the current drawn is around 20mA.

2.2 The VFO Amplifier with Power Divider

Following the VFO is a single stage broadband amplifier. In the frequency range around 5 MHz (F_{VFO}) this delivers an amplification of about 20dB. At the output the VFO signal is split in a Wilkinson divider for the frequency counter and the VFO mixer. There is a signal of around 10mW at the disposal of each of these units.

The broadband amplifier (Fig.5) will work up to 300 MHz with the components indicated. Notable in the circuit are the inverse feedback and a transformer wound non-inductively. In the section on VFO mixers we describe the winding and connection of these transformers.

The circuit is stable across the whole frequency range. Input and output both exhibit a true 50Ω impedance. These values result from the 4.7Ω and 560Ω feedback resistors and the 4:1 transformer. Details on exact dimensioning can be found in items (2) and (5) of the literature sources (see next issue).

The Wilkinson divider at the output of the module divides the signal equally between the two outputs. It needs to be dimensioned for the VFO frequency. With a coupler of this kind we get good decoupling automatically (more than 30dB) between the two outputs.

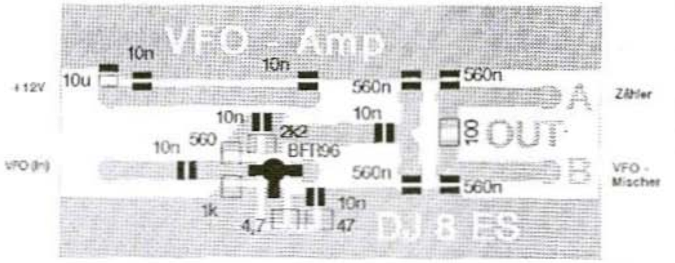


Fig.6: Component Layout with the Surface-mount parts on the track side of thePCB

2.2.1 Construction hints

The circuit is constructed on double-sided epoxy printed circuit board material with the dimensions 34 x 72mm (Fig.6). In general all the PCBs have been designed so that they can fit into commercially available tinplate cases. All of these used have a height of 30mm.

For RF the individual units are connected using thin coax cable (e.g. RG174). To save space and cost, plug connectors are not used. The amplifier is powered through a teflon feedthrough.

2.2.2 Component list

1 x BFR96 (Siemens) or similar

2 x Neosid ready-made filters, 1uH BV5048 (yellow, green)

Resistors and capacitors are surface-mount types.

2.2.3 Commissioning

The current requirement of the amplifier stage at 12V is around 45mA. Two volts are dropped in the 47Ω resistor. Any wide deviation from these values mean you should check the transistor used and the winding of the transformer.

There's plenty of scope for making errors here!

2.3 The Digital Frequency Display

The frequency counter is an eight-digit one and is made from two four-digit counter modules of the type ICM7217A by Intersil (Fig.7). These ICs are designed for seven-segment displays with a common cathode. The units are programmable for producing a frequency-accurate display; we shall need to use this feature here to add 45 MHz to the amount.

The complete counter module is controlled by a driver unit (ICM7207A). As well as the control signals for the two counter units it provides the signal for the gating time element, made from a 74LS90. The decimal counter is preceded by a transistor stage for changing the signal level. The 9:1 transformer converts the 50Ω input to the high impedance required by the transistor.

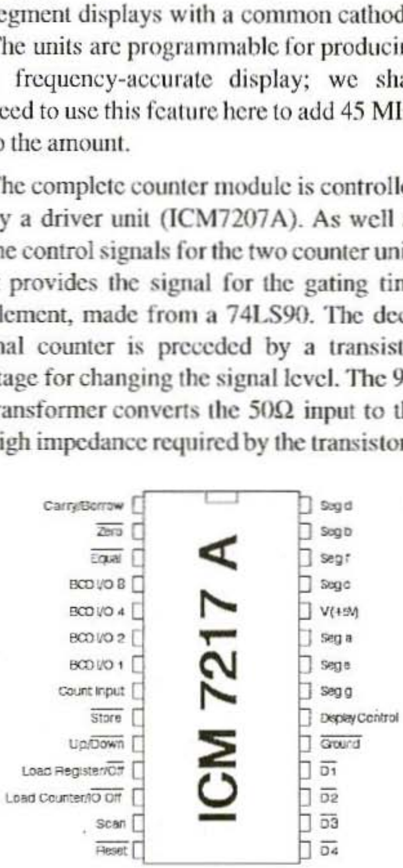


Fig.7: Connections for the ICM7217A

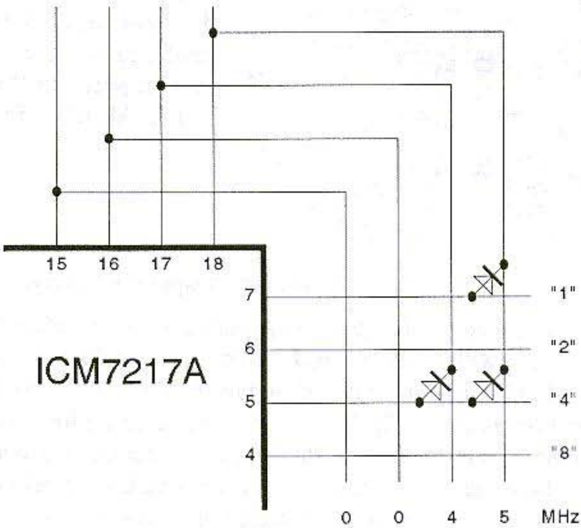


Fig.8: Sample programming arrangement

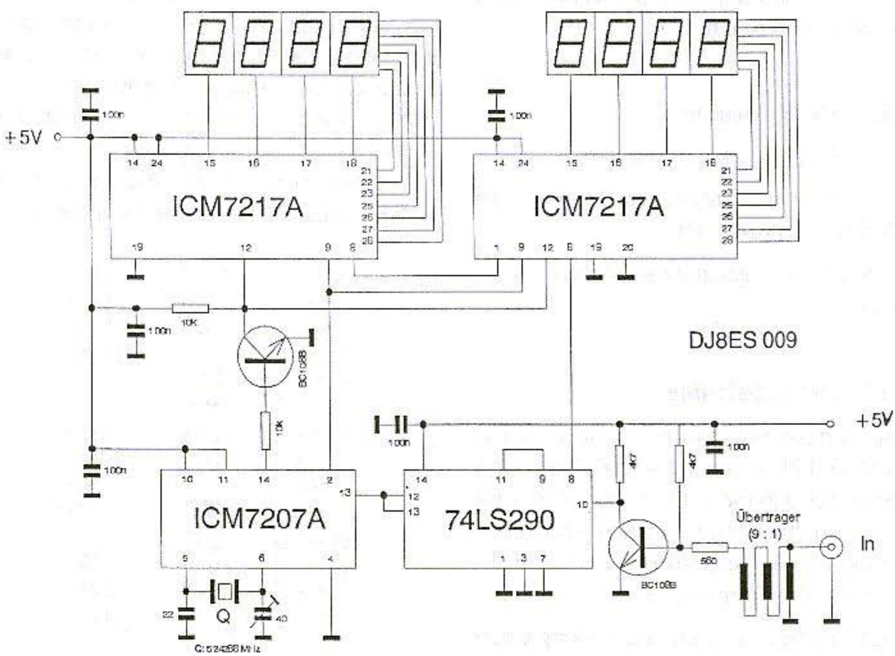


Fig.9: Circuit of the programmable Counter Group; **Uebertrager = Transformer**

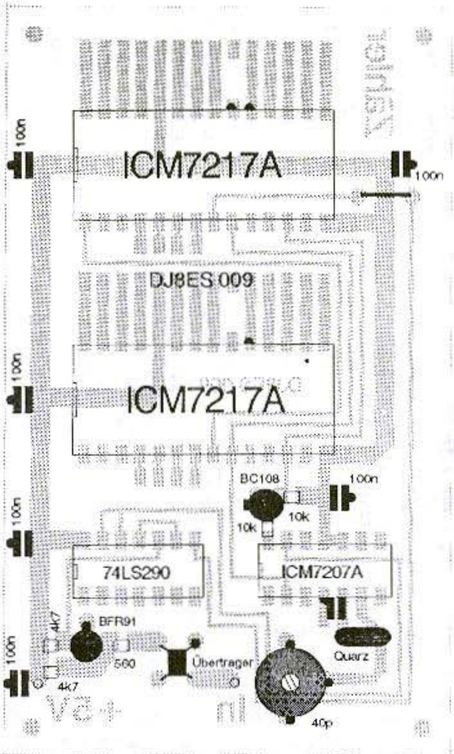


Fig.10: Component Layout of the Counter Module. Surface-mount components are placed on the track side of the PCB

2.3.1 Construction hints

The digital frequency display is spread over two PCBs. As well as the actual counting logic, the eight-digit display is built up separately. This makes it easier for constructors to add additional functions at a later stage. Both PCBs are made from 1.5mm thick double sided epoxy material.

The frequency shift, in our case 45 MHz, is programmed with standard diodes, e.g. 1N4148. Provision is made on the PCB for the corresponding connections. Fig.8 shows an example of this programming. This will be useful in particular when used with transverters as the constructor can always obtain a display which is correct for the frequency.

2.3.2 Components

1 x 74LS901 x ICM7207 AIPD (Intersil)
1 x ICM7217 AIPI (Intersil)
8 x 7-segment displays, e.g: SL1110K
2 x BC108B or similar
Capacitors and Resistors are surface-mount types
1 x trimmer 40pF (10mm spacing, grey)

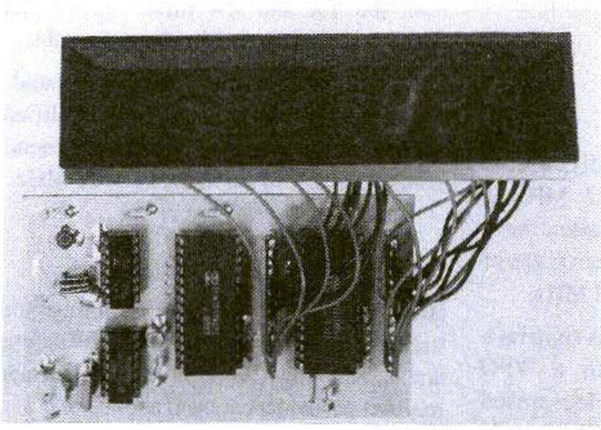


Fig.11: Completed Counter Logic with Display

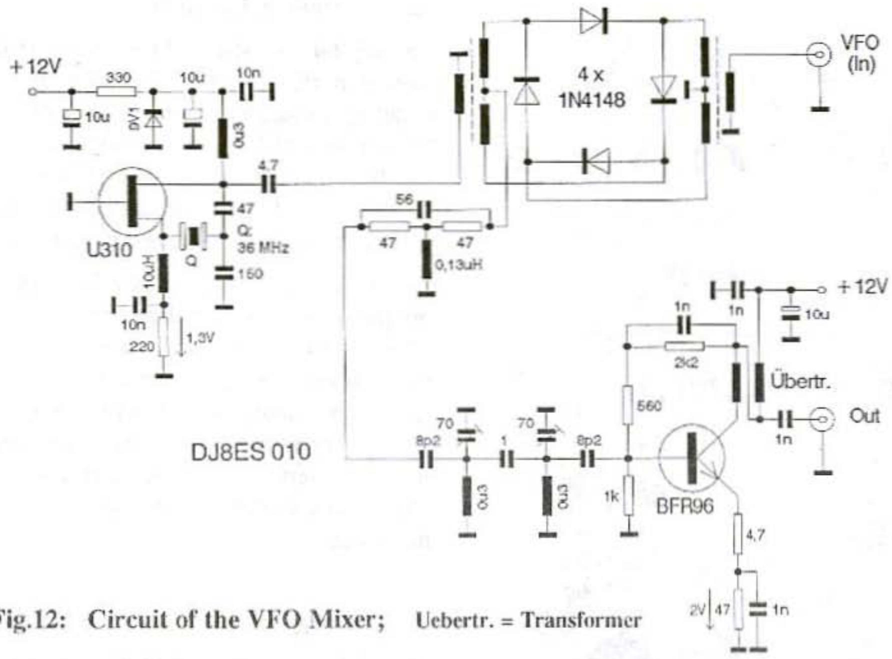


Fig.12: Circuit of the VFO Mixer; Uebertr. = Transformer

2.3.3 Commissioning

After applying the operating voltage of 5V the display will show the frequency programmed, in this case 45.000.0 MHz.

If a defined frequency is fed into the input, this plus 45 MHz should be displayed. Fine adjustment of the counter unit can be undertaken with the 40pF trimmer next to the ICM7207A.

2.4 The VFO mixer

In the SSB transceiver for 50 MHz a VFO signal in the range 41 to 42 MHz is required. Taking into consideration the 9 MHz IF used, this will enable coverage of the six metre band from 50 to 51 MHz.

One possibility of producing this signal is a super-VFO. The signal from a VFO working normally around 5 MHz is mixed with a fixed frequency and is then ready for use in the desired frequency range. The crystal oscillator is equipped with a U310. This well proven circuit guarantees a stable, low-noise signal. On 36 MHz it gives approx. 10mW at the ring mixer. The mixer is made up from transformers. The diodes used are 1N4148, which work just as well as the special VHF mixer diodes. In both the TX and RX mixer stages this opinion was backed up by the test results.

Following the two-stage broadband bandpass filter, the mixer signal is amplified further. At the output we thus have around 20mW, tunable in the range 41 to 42 MHz.

2.4.1 Construction hints

Broadband transformers in a number of varieties are used in this circuit. Winding and connecting this kind of transformer requires particular care and attention.

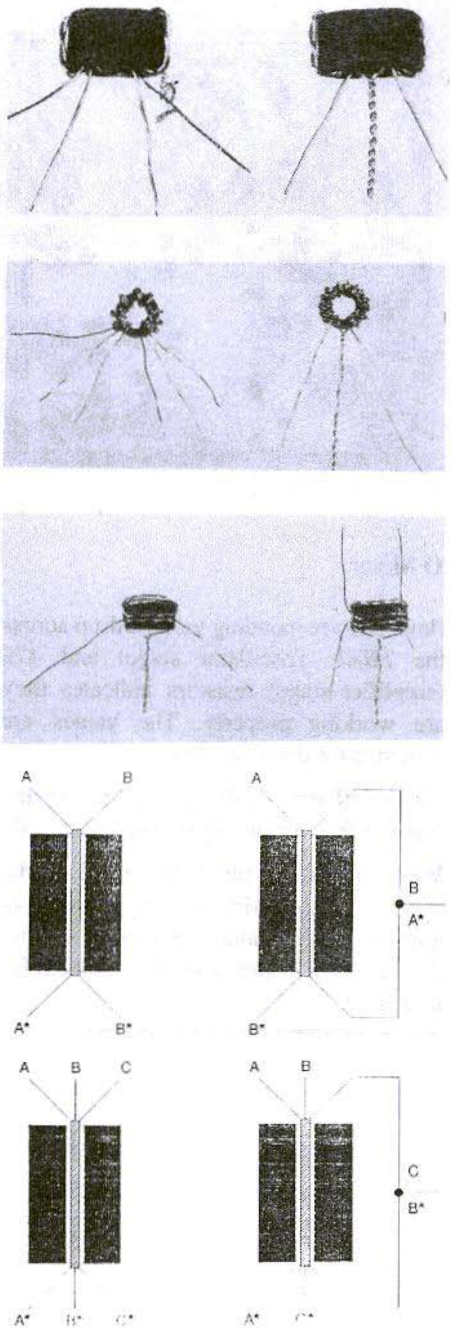


Fig.13: Sample Bifilar (non-inductive) and Trifilar Transformers

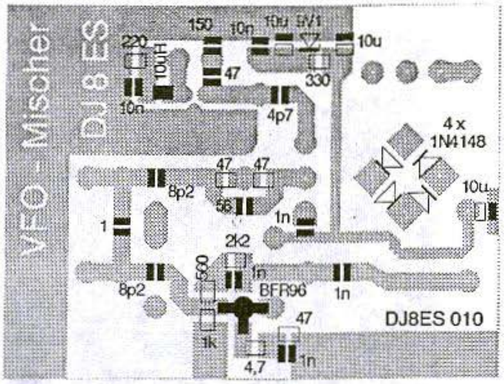
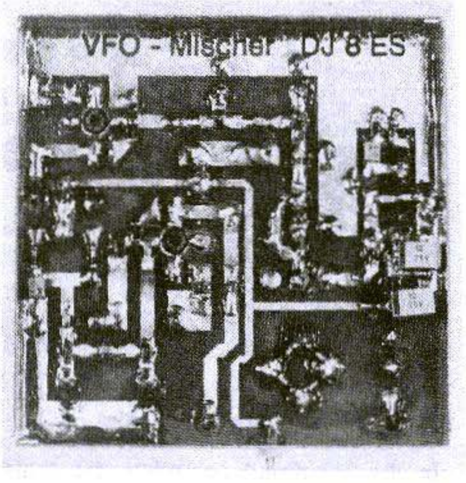


Fig.14: Component layout of the Mixer PCB (groundplane side)

In the broadband amplifiers the transformer comprises a bifilar (two-wire, non-inductive) winding (6 turns on a 5mm ferrite bead). The bifilar winding is best made as follows. Take two equal length pieces of 0.2mm diameter varnished copper wire A and B and twist them carefully around each other using a hand drill. The necessary number of turns of this double wire is threaded onto the core and connected according to the directions given in Fig.13. In the ring mixers the transformer consists of a trifilar (three-wire) winding. Construction and connection is just like the bifilar winding. The PCB for the VFO mixer has the dimensions 53.5 x 72mm. A normal tinplate case is used.

2.4.2 Components

1 x U310 (Siliconix)
1 x BFR96 (Siemens) or similar
1 x zener diode 9V1 (surface-mount)
4 x 1N4148
1 x Neosid ready-made filter 0.13uH,
BV5063 (blue, orange)
1 x Neosid ready-made filter 0.3uH,
BV5049 (yellow, white)



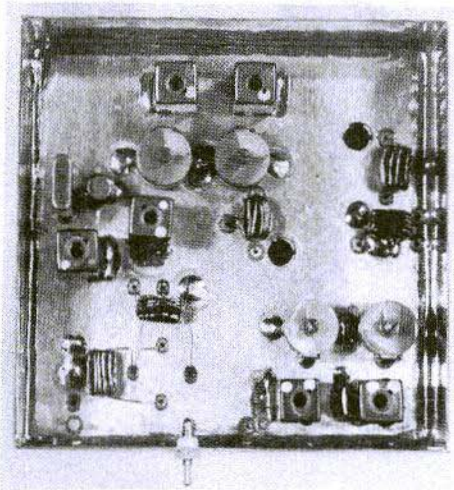


Fig.15: Constructed test sample of the VFO Mixer

2.4.2 Components (cont.)

2 x plastic foil trimmers 70pF (yellow), 10mm spacing.
1 x crystal 36 MHz, series resonant. Holder HC-18U or HC-25U
Capacitors and resistors surface-mount types

2.4.3 Commissioning

After applying the operating voltage (+12V) a current of around 55mA should flow. A corresponding voltage drop across the 220Ω (oscillator stage) and 47Ω (amplifier stage) resistors indicates they are working properly. The values are shown in the circuit diagram.

The oscillator should be adjusted to the nominal frequency using the resonant coil.

When driven by the VFO signal in the region of 5 to 6 MHz an output power of approx. 10mW should be achieved. After this the twin filter should be tuned for maximum.

MATERIAL PRICE LIST OF EQUIPMENT

described in VHF COMMUNICATIONS

DB1NV	**Digital Image-Store for the Spectrum Analyser**	**Art. No.**	**Ed. 4/1991**
PCB	DB1NV 010	6477	DM 44.00
Components	Processor P80C31; 12 ICs; 1 Reg. IC, Transistor; Zener Diodes; Silicon Diodes; Chokes; RAM; EPROM DB1NV 010; 4 x 2k Trimpots; Crystal	6478	DM 276.00

DB1NV	**Tracking Generator for the Spectrum Analyser**	**Art. No.**	**Ed. 1/1992**
PCB	DB1NV 011	6479	DM 31.00

F6ILR F6BXC	**A Digital Slow-Scan Television Transmit Coder**	**Art. No.**	**ED. 3/1992**
PCB	SSTVCODE1 (KM Publications)	SSTV1	£ 28.50

DB1NV	**Broadband VCO's using Microstrip Techiniques**	**Art. No.**	**Ed. 4/1992**
PCB	DB1NV 012	6480	DM 33.00
PCB	DB1NV 013	6481	DM 33.00
Components	400 - 1250 MHz 3 x BB619; 1 x BB811; 1 x BFG96; 2 x AT42085; 1 x BFQ69; 2 x 2.2nF & 1 x 27pF Feed-through Cap.; SMC Connectors; 2 x 0.47 H SMD Choke; 1 x housing 74x55x30mm; 1 PCB DB1NV 012	6482	DM 81.00
Components	450 - 1450 MHz as above but: 1 x BFG65 instead of BFG96	6483	DM 81.00
Components	800 - 1900 MHz as above but: 4 x BB811 instead of BB619 and PCB DB1NV 013 instead of 012	6484	DM 85.00

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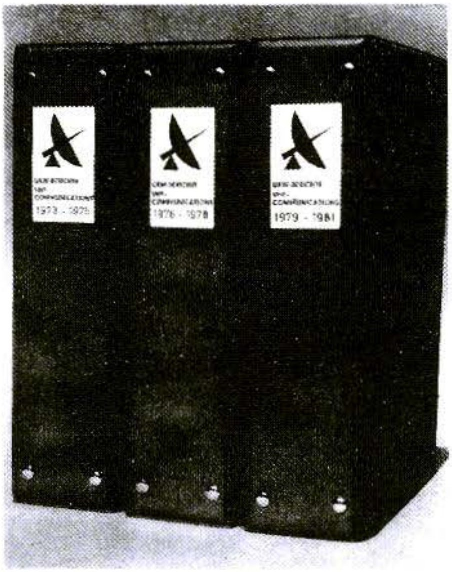

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