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for the Radio-Amateur
Especially Covering VHF,
UHF and Microwaves

VHF communications

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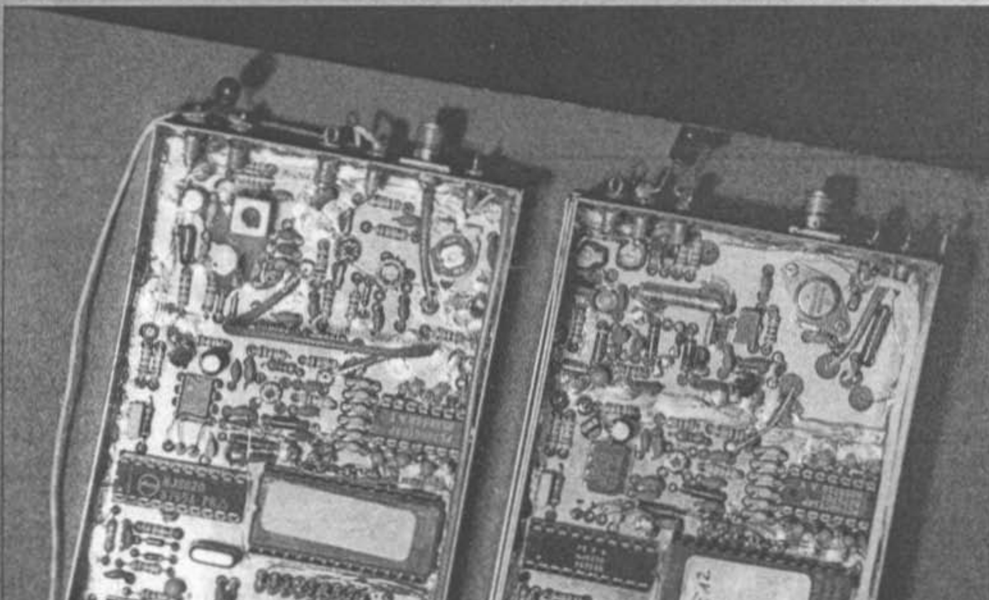

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2 m & 70 cm

VHF communications

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Volume No. 22 · Spring · Edition 1/1990

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R. G. Sanson, ZL 1 TBG

An Injection-Locked Oscillator for the 10 GHz Band

If Gunn oscillators, as commonly used for transceivers on the 10 GHz and 24 GHz bands, are locked to a stable frequency source then it is possible to operate with narrower bandwiths than the usual WBFM. This enables station performance gains of more than 10 dB. The station's operating frequency is also guaranteed. This article describes methods of injection-locking Gunn oscillators to a more stable low-power reference signal.

1. EXPERIMENTAL

In early tests, it was found that an external reference signal of as little as – 20 dBm is sufficient to achieve locking of a normally free-running Gunn diode oscillator. The lock range was around 0.5 MHz, improving to over 5 MHz when the reference source is 0 dBm or more. When locked, the Gunn oscillator is very difficult to FM modulate. Instead, the reference signal can be modulated with NBFM, WBFM or possibly FMATV. The Gunn oscillator will faithfully track this signal resulting in excellent linearity and stability.

A point to note is that with the onset of injection-lock, the output power of the Gunn oscillator decreases, becoming a minimum at the point where its natural running frequeny is set to the same as the reference signal frequency. Monitoring the output power level by means of mixer diode current of a typical transceiver set-up allows easy adjustment and confirmation of injection-locking.

Use of a reference source on a subharmonic frequency, or a 'comb' frequency generator reference also appears possible. This could be useful for locking on 24 GHz or constructing a synthesizer for these bands.

2. CONSTRUCTION

The mechanical arrangements can be seen in the sketch (**fig. 1**). The method of feeding the reference signal to the Gunn oscillator is not very critical. It can be injected at the cavity or outside of the cavity. A 2 - 3 mm probe can be fitted through the wall of the waveguide line from the oscillator. This can be fashioned from some semi-rigid coaxial cable or a panel mount SMA connector soldered to the waveguide. Alternatively, the injection signal can be coupled to the Gunn oscillator with a ferrite circulator or waveguide hybrid junction.

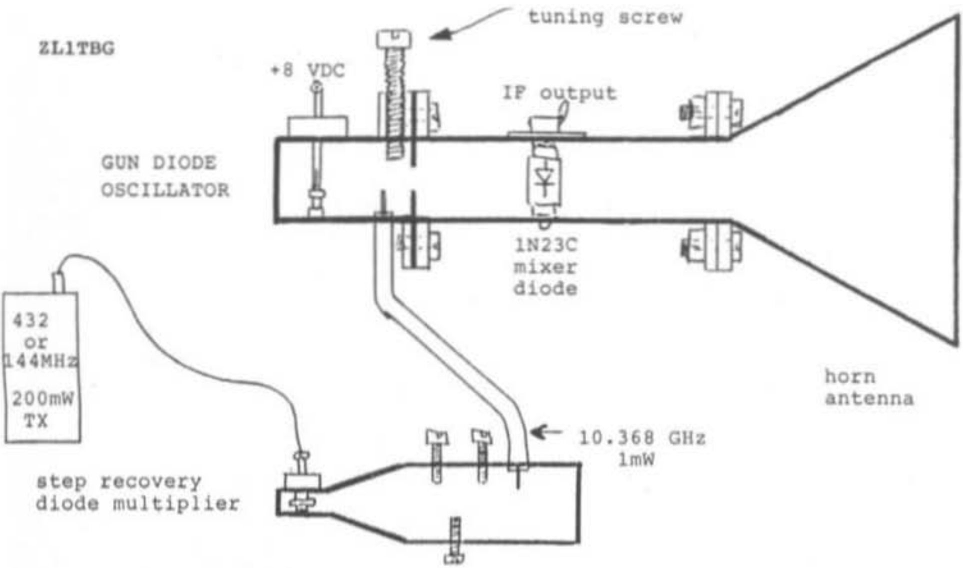

Fig. 1: The injection-locked oscillator's mechanical layout

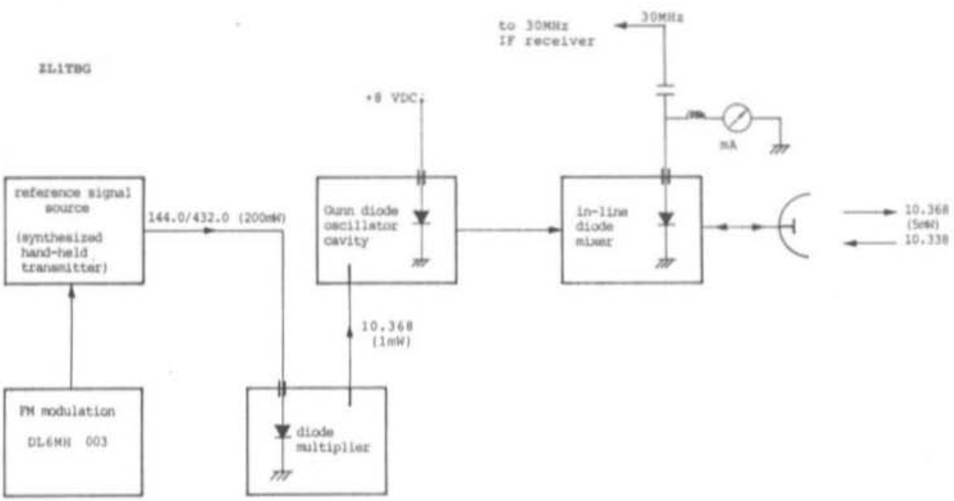

Fig. 2: The injection-locked transceiver's system schematic

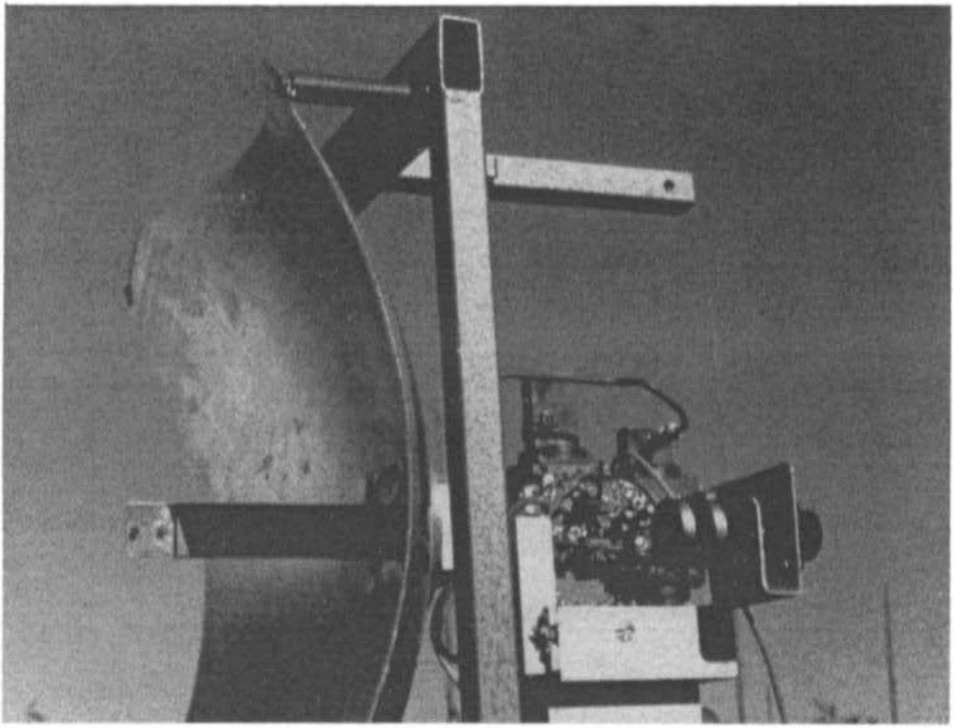

Fig. 3: Photograph of the author's prototype

It is very easy to retrofit injection-locking to an existing WBFM Gunn transceiver set-up. The author's transceiver circuitry resembles the design described in (1); the block schematic is shown in **fig. 2** and the photo of **fig. 3** gives some idea of how it could be mounted directly behind the antenna paraboloid. The tone signal and modulation circuits feed the reference source to provide modulation when transmitting. The Gunn oscillator is powered with a fixed supply voltage (set for maximum RF output power).

3. OPERATION

Injection-lock is established by adjusting the Gunn oscillator tuning so that the frequency passes that of the reference source. When the frequency is close, the Gunn oscillator will jump onto frequency and remain in lock over a reasonable adjustment range. When outside this range, it will jump away to a new free running frequency. Within the locked range there is no 'pulling' or deviation from the stable reference frequency. Locking is observed with either a receiver or by monitoring the behaviour of the mixer diode current (output power).

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A Transceiver for the 10 GHz Band
VHF COMMUNICATIONS Vol. 11,
Ed. 4/1979, p. 208 - 215

Dr.Eng. Jochen Jirmann, DB 1 NV

A Spectrum Analyzer for the Radio Amateur Part 3b: Circuit Options and Ancillary Equipment

6. EXTENDING THE ANALYZER'S CAPABILITY

The spectrum analyzer as described, being a modular system, is capable of being extended if so desired. The modifications and add-on units, tried so far, are as follows: –

- **Calibration Generator:** This generates a 60 MHz crystal-controlled, low-harmonic content test-signal at a power of – 30 dBm;
- **Preselector:** This separates the two input ranges of the analyzer using automatically selected high/low pass filters.
- **Low-Noise HF/IF Stage:** This is a modification of the existing HF/IF module with an additional IF pre-amplifier. This improves the noise-figure of the analyzer from 22 dB to 12 dB. This enables the analyzer to function as a sensitive measurement receiver.

6.1. The Calibration Generator

This unit is built into the analyzer in order to make a stable signal (in terms of both amplitude and frequency) readily available so that the analyzer's linearity and level calibration can be checked. This is quite a simple calibration generator which can take several forms. For example, it should be possible to produce a comb-generator which has a series of signals of equal amplitude spaced from each other by the frequency of the fundamental. As the amplitude calibration of a single-frequency generator is easily carried out, using a calibrated signal-generator, this was the preferred solution.

As indicated by the calibration-generator's schematic (**fig. 16**), the quartz oscillator is a three-terminal capacitive circuit which has a pnp transistor BF 926 (or BF 970, BF 979) as the active element. The output signal is taken from the "cold end" of the tank coil and filtered via an LC low-pass filter before it is attenuated to – 30 dBm by means of a resistive 50 Ω pad. The exact adjustment to an output of – 30 dBm is accomplished by means of a variable emitter resistance. The exact frequency is uncritical, the author used a 60 MHz crystal working in the 3rd overtone.

The few components are easily wired "ratsnest" style into a small screened enclosure, thus obviating the need for a printed circuit board.

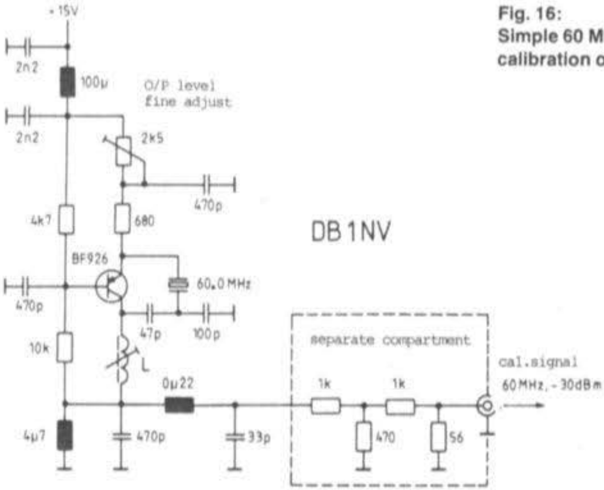

Fig. 16:
Simple 60 MHz, – 30 dBm calibration oscillator

A most important item is, that the attenuator pad should be totally screened in a separate compartment within the unit's housing. This measure will prevent oscillator harmonics from being passed via the pad, directly to the output socket.

Calibration can be carried out, either by means of an RF millivoltmeter with a terminating resistor, or with a professional signal generator having a calibrated, stepped output attenuator. The millivoltmeter allows the – 30 dBm to be set directly. The signal generator is set to – 30 dBm and fed to the spectrum analyzer. The level of the spectral line at 60 MHz is noted and the signal-generator removed. The calibrator is then connected to the analyzer and its output adjusted for the same amplitude as the signal generator's using the preset pot'meter.

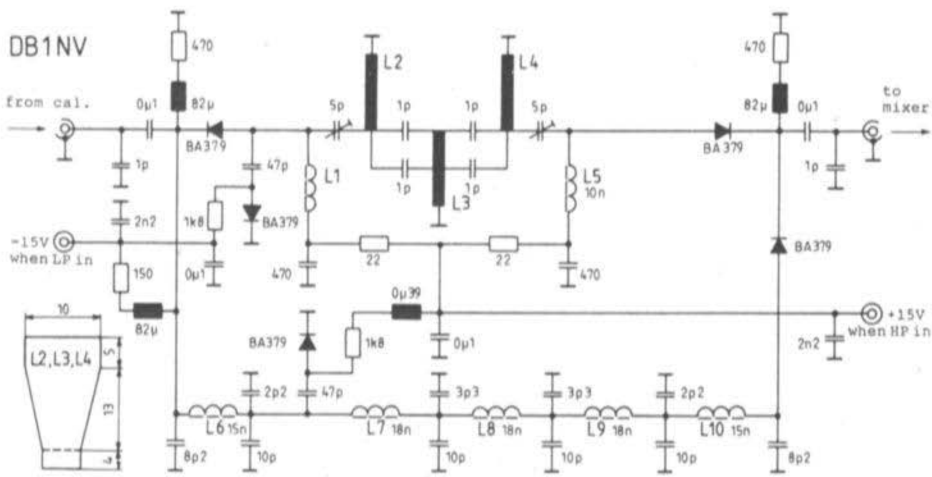

Fig. 17: The preselector is a high and low-pass filter switched by PIN diodes

L1, L5: 1.5 turns 2.5 mm Ø, 1 mm Cu silvered (ca. 10 µH)
L2, L3, L4: brass plate as sketch (ca. 3 µH)
L6, L10: wire 1 mm Ø, 35 mm long (ca. 15 µH)
L7, L8, L9: 2 turns 2.5 mm Ø, 18 mm connector length (ca. 18 µH)

6.2. The Preselector

The spectrum analyzer, so far described, possesses no definite arrangements for the choice of input ranges to suit an (unknown) signal. It can be determined by de-tuning the 2nd oscillator in one direction and noting which direction the observed signal moves on the trace. As this practice is a little haphazard, it was decided to place a low-pass and a high-pass filter before the HF input stage.

These filters are switched with the range switch. The low-pass has a limit frequency of 600 MHz and the high-pass a limit frequency of 900 MHz.

One thing must be made clear, however, the inclusion of a filter in front of the mixer inevitably has a deleterious effect upon the input frequency response of the instrument and also upon its match to 50 Ω nominal impedance. Using matching pads at the input and output of the filter offers some improvement but only at the expense of a still further reduced sensitivity.

Ideally, such a filter would be built with either coaxial or stripline technology. The large dimensions (λ/4), required for the components of such structures, take up a great amount of space and in the case of, the full coaxial technology is constructionally difficult to implement.

Fig. 17 shows a much more realistic approach to the preselector problem. Both low and high-pass filters use, initially, concentrated elements, the values of which can be obtained from manufacturer's tables (Siemens, Telefunken etc.) or calculated with a pocket calculator (programmable) and a filter program from Hewlett Packard. The inductances used in this way, work out to be about 10 nH. It is, of course, axiomatic that parasitic inductances must be kept to an absolute minimum.

The practical realisation of the low-pass filter is relatively simple. Six leadless plate capacitors of 8.2 pF and 10 pF are arranged as shown in **fig. 18** and spaced by ca. 35 mm. They are soldered to a ground plane, e.g. floor of a brass or tin-plate housing, and are used as solder-tag supports for the inductors L6 to L10. The dimension of the inductors may be obtained from the table given in fig. 17. The exact capacitive values are achieved by soldering-in small-value conventional capacitors across the plate capacitors using extremely short connections. The filter constructional layout need not be linear, the

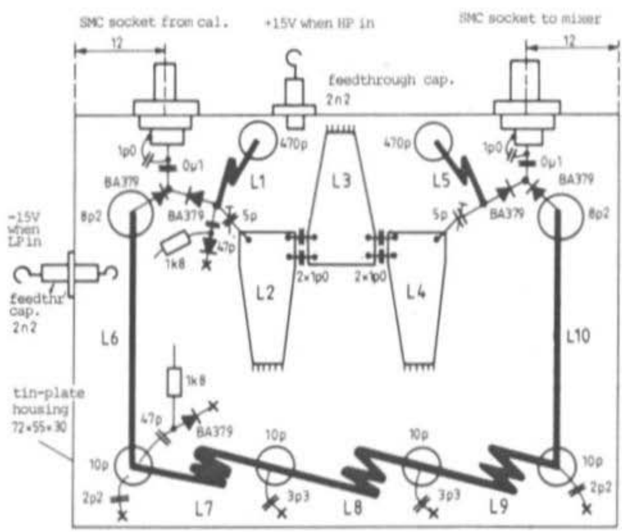

Fig. 18:
Actual sized drawing of the preselector constructional highlights. Only the HF-relevant components are shown

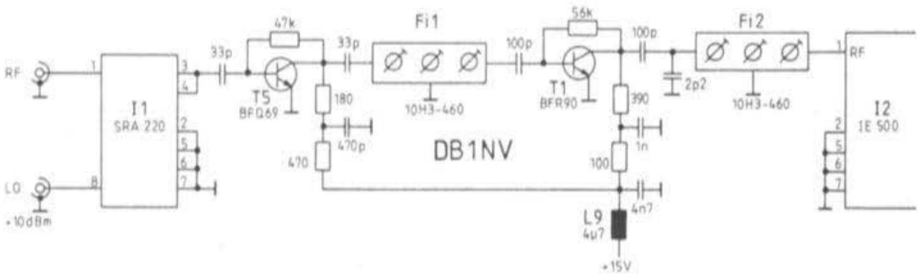

Fig. 19: Modified 1st IF amplifier (extract from fig. 1) f = 465 ± 3 MHz, G ≈ 20 dB

form shown in fig. 18 fulfils the two requirements of having a short path between the input/output connectors whilst, at the same time, having a direct access to the high-pass filter via the change-over switch diode.

The high-pass filter with inductors L1 to L5 represents, actually, the limits of what can be achieved with discrete components, wire and tin-plate. It may be seen in **fig. 18** that L2, L3 and L4 together, form a kind of inter-digital structure and are coupled at their hot ends by two paralleled 1 pF capacitors. These are normal wired types using very short direct soldered connections. Chip capacitors are not as robust mechanically and would break or crack if the housing were to be flexed.

The inductors L1 and L5 serve, on the one hand, to pass DC supply current to the PIN-diode switches and, on the other hand, they form, together with the 5 pF trimmers, an adjustable matching network for the high-pass filter. Now only the alignment stands in the way of a more-or-less "flat" response curve for the filter in the frequency range 1000 to 1500 MHz.

The low and high-pass branches are selected by means of a PIN-diode switch in the signal path. As the attenuation of the BA 379 PIN diodes is fairly low at frequencies around 1 GHz, the input of the unused filters is shunted to ground with a further PIN diode.

Finally, the author would like to point out that this input filter module should be regarded only as a basis for experiment as the behaviour of the high-pass is very much dependent upon individual method of construction. If a suitable sweep generator is not available, or the necessary patience to align it without one is lacking, another solution may be contemplated. The firm Mini-Circuits (Frankfurt) have relatively favourably-priced high and low-pass filters for this frequency range. The price lies between 100 and 150 DM according to whether soldered-in or plug-in

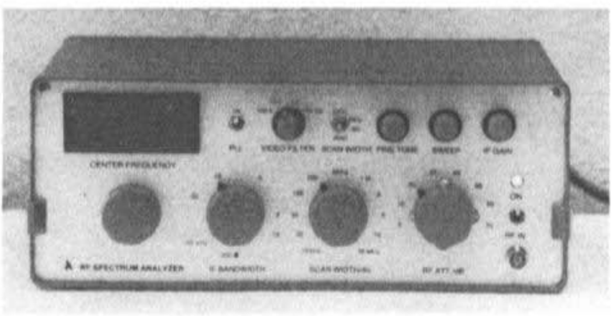

Fig. 20:
A 'sample' instrument

versions are purchased. Switching between them can be carried out with either PIN diodes or, conventionally, with coaxial relays. The author, however, has no experience of their use.

6.3. An IF Pre-Amplifier to improve Sensibility

The original concept used a 465 MHz IF amplifier to compensate for the filter attenuation and to decouple the first from the second mixer. The outcome of this dimensioning is a really poor analyzer noise-figure (ca. 22 dB). Meanwhile, experiments were carried out with a wideband IF pre-amplifier between the first mixer and the first IF filter. A BFQ 69 was used, as may be seen in **fig. 19**. Using SMD capacitors and resistors of the 0204 series, this amplifier can be fitted between the terminal pins of the mixer and the IF filter. For this, the original tracks must be removed and the new amplifier installed on the track side of PCB DB1NV 006. A leadless plate capacitor soldered to the ground-plane side serves as a support point for the supply voltage termination.

Following the installation of the IF pre-amplifier, the sensitivity of the analyzer is improved by about 10 dB. The intermodulation performance was no worse than anticipated. That left the impression that most of the intermodulation is caused by the first mixer.

The higher sensitivity, however, does bring some disadvantages along with it. First of all, the demands upon the RF soundness of both cables and screening enclosures. Secondly, the number of spurious signals, produced in the instrument and visible in the display, increases. Some of these spurious signals rise some 10 dB above the noise floor. They have been caused by the mixing of harmonics from the first and the second mixers. These had previously been buried in the noise floor prior to the introduction of the IF preamp. It really is debatable, whether the increased sensitivity is necessary or whether an add-on selective amplifier would solve many measurement problems in a better fashion.

Manfred Salewski, DC 9 DO

SAT–X
Receiver for the Satellite IF Band 900 - 1700 MHz

With the advent of the television satellites ASTRA and OLYMPUS a new epoch has been opened. Whereas the previous communication satellites could only be properly received with, at least, 1.2 m dishes and special low-noise converters, a dish diameter of only 60 to 80 cm is now required, together with a standard converter. This is sufficient to enable the 16 programmes from ASTRA to be received, all with very good quality!

The minimal configuration for a polarisation plane (= 8 programmes) is shown in **fig. 1**. It consists of the following components: –

1. 60 cm parabolic dish (F = 245 mm, D = 600 mm) with adjustable polar mount
2. Corrugated feedhorn (to be fitted later with dual-plane polariser)
3. Low-noise down-converter (F = 1.5 dB max.)
4. Satellite receiver for the SHF IF frequency 0.9 to 1.7 GHz

The home construction of a parabolic dish antenna with an adjustable polar mount, corrugated feedhorn and low-noise converter is not easy and at today's low prices, hardly worth the effort. The talented radio amateur constructor, however, is in a position to save himself some money and also

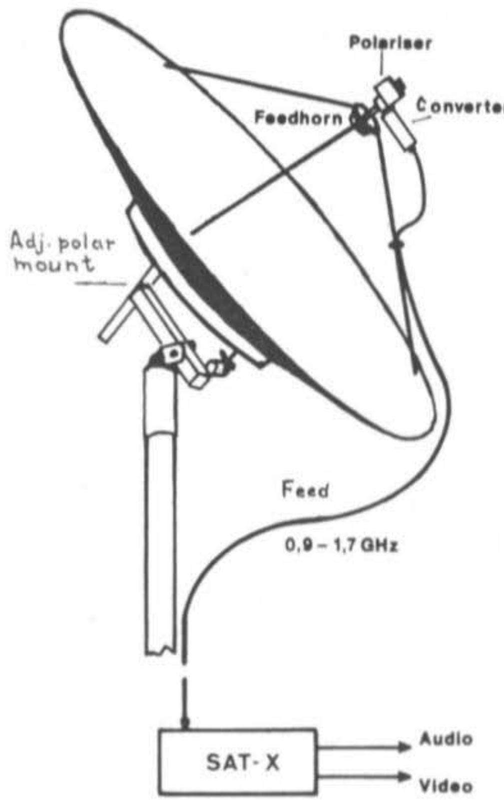

Fig. 1: Out-door unit and satellite receiver

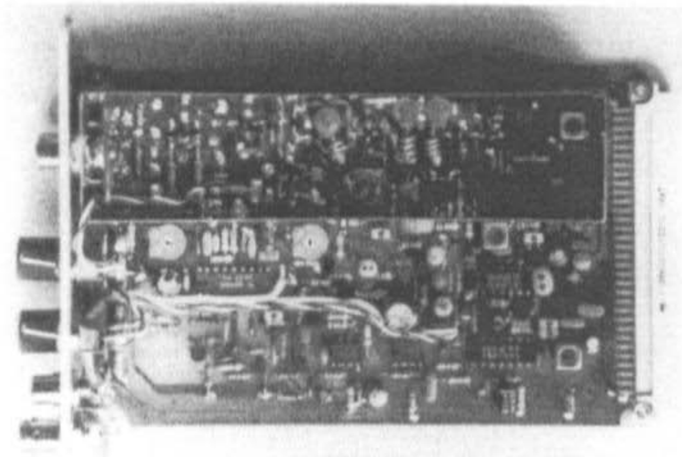

Fig 2: SAT-X receiver (DC9DO 001)

gain experience in the satellite reception technology, by making his own receiver.

For this reason, and also that the author had a good functioning commercial satellite installation for comparisons and test purposes, readily available, he decided to take the risk and make his own SAT receiver. In addition, an ECB bus computer with a ± 12 V operational supply voltage, a PAL/RGB eurocard (home made) together with an RGB monitor were also available. In order to utilise these possibilities to the full and also to keep down the complexity of the project, the following concept was evolved.

1. CONCEPT

Fig. 2 shows the SAT-X receiver built on a two-sided eurocard PCB (DC9DO 001) and supplied via a VG-connector array/ECB bus. When the ECB bus is used, the 35 V tuning voltage must be separately supplied. The construction may be carried out with easily-available components. The circuit concept was kept very simple but this did not preclude the use of a few SMD components and altogether very satisfactory results were obtained. The unit can be operated from a separate DC supply, if desired.

As the critical input tuned circuits use printed inductors and work at a very low impedance, the construction is uncomplicated and does not require much screening and other mechanical work. The tuning is carried out by a potentiometer, over the range 900 to 1800 MHz. Since one of the transponders operates with different sound-carrier separations, or with multiple sound carriers, a facility has been added to switch from "automatic" to "manual". This enables all satellite's radio broadcasts to be tuned in.

2. CIRCUIT DESCRIPTION

Fig. 3 shows the circuit schematic of the complete receiver.

2.1. HF Section

The received signal of 10.9 to 11.8 GHz from the dish antenna and built-in converter(s), is translated to the frequency of the 1st IF, 900 - 1800 MHz in the converter. It is taken via an F-socket to the SHF amplifier IC 1 and amplified in a wide-

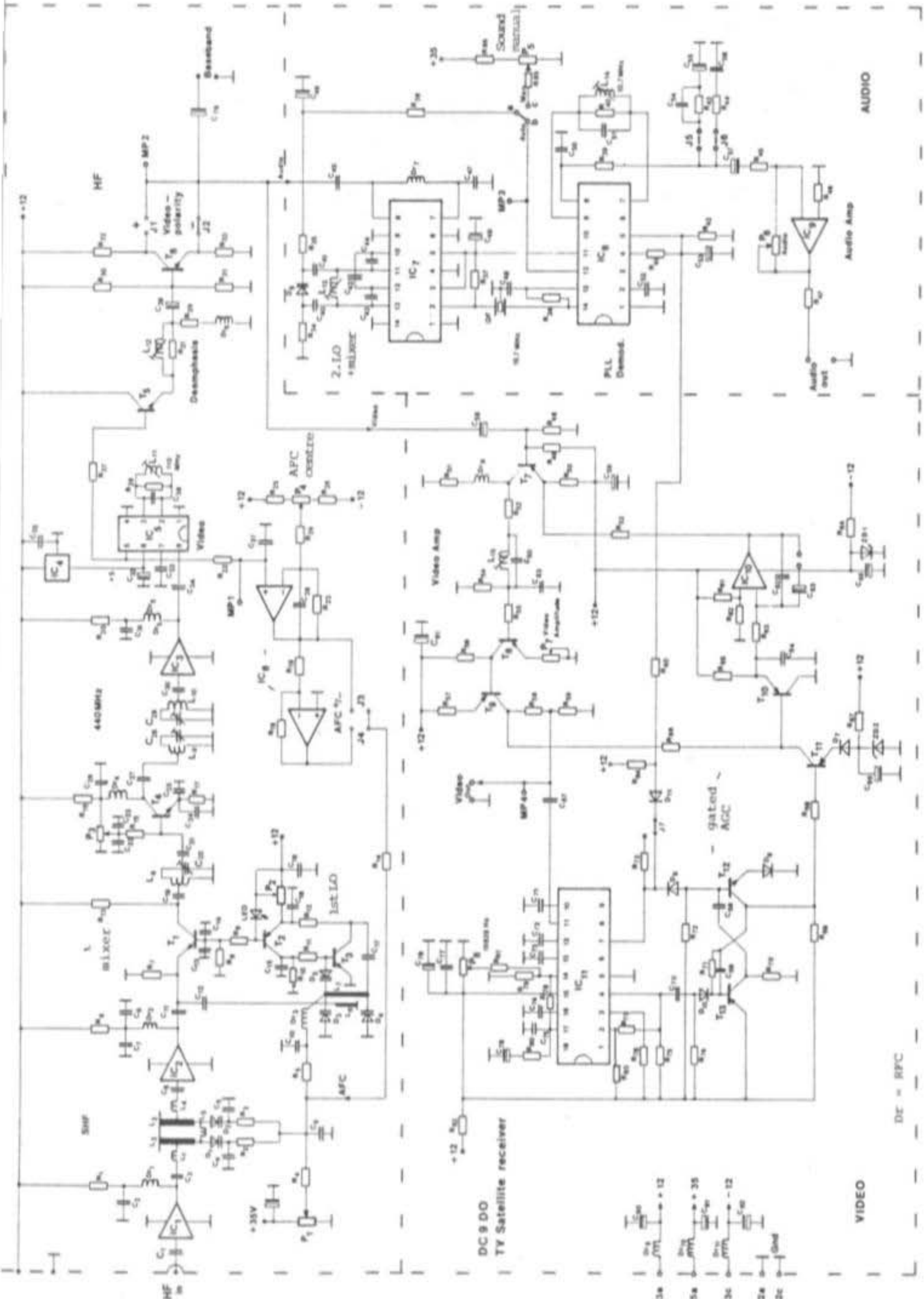

Fig. 3: SAT-X receiver complete circuit schematic

band amplifier. Inductors L2 and L3 form, with the varicap diodes D1 and D2, an input band-pass filter with coupling via L5. The inductors L1 and L4 form a stop filter which rejects all unwanted out-of-band signals. The pre-selected signal is again amplified by IC 2 and then fed via C 11 to the emitter of the mixer transistor. The oscillator is also coupled to this emitter via C 12 and the printed inductor L6.

The SHF oscillator, consisting of L7/D3...D5 and T3, has been taken largely from that suggested by YU3UMV in VHF COMMUNICATIONS 1/1987, in his TV satellite receiver installation article. The combination T2/LED form a constant-current source which determines the working point of both the oscillator and the mixer. Potentiometer P2 serves to set the working point of the SHF oscillator – the temperature drift characteristics of the LED being used to thermally stabilise the oscillator. This oscillator works either 440 MHz above, or below, the 1st IF band, thus producing a 2nd IF of 440 MHz. This high frequency 2nd IF is, however, uncritical in consideration of the attainable amplification and high stability. Other advantages of this high IF amplifier are that it quite easily produces a definite bandwidth of approx. 18 to 25 MHz and also the danger of undesirable spurious frequencies is minimised.

For the reception of communication satellites, such as EUTELSAT, INTELSAT, which work in the frequency range 10.9 to 11.8 GHz, the SHF oscillator should operate above the 1st IF (900 to 1800 MHz) i.e. 1340 to 2240 MHz. Should the oscillator's frequency lie below the frequency of reception, i.e. 460 to 1360 MHz, a double reception of various channels due to the oscillator's harmonics (920 to 2720 MHz) is possible if the input band filter is not aligned absolutely correctly. In addition, where the screening is not adequate, interference with normal TV reception can also be expected because the oscillator is working in the UHF band!

Assuming that the receiver is to be built only for the 16-channel ASTRA satellite (this entails a reception band of 1200 to 1500 MHz) the local oscillator can operate below the input frequency of 760 to 1060 MHz. In this area, results can be obtained using normal transistor types such as the BFW 34 which are used in UHF antenna pre-amplifiers.

At still higher frequencies, transistors with small inter-electrode capacitances and higher limit-frequencies must be employed. Usable results have been obtained by the author with the BFQ 69 and BFW 91. Still better are the AT 41485 and AT 42085 (Avantek) which have a very low noise-figure (1.4 dB) and are favourably priced. They can work in oscillator circuits up to 6 GHz without any problems and have, with a collector current of 20 mA, a limit-frequency of 13 GHz.

Similar considerations apply to the varicap diodes. The stray capacitance at the largest tuning voltages plays a large part here. Normal UHF types, such as the BB 405, will work in the circuit under discussion, but only with the most careful constructional methods will they operate satisfactorily at 1600 MHz. The BB 505 b, or the later BB 801, on the other hand, can be readily utilised.

After the mixer, connected in a grounded-base configuration, follows the 1st IF stage with the first filter L8/C 20 and transistor T4. As the satellite transmits an FM signal and the FM demodulator IC 5 has a very wide dynamic range (5 - 500 mV HF), the provision of AGC can be dispensed with.

If the level is too high at this point, the IF becomes de-tuned and more broad-banded. For this reason, the working point of the 1st IF stage has been made adjustable via P3. When setting the stage gain, a compromise must be found between bandwidth changes during tuning, and sensitivity.

The IF signal is taken via C 27 to the band filter L9/C 26 and L10/C 29 and on to IC 3 and IC 5, the FM demodulator driver stage. The very convenient SL 1452 is used here as it also contains a ÷ 4 scaler and a quadrature demodulator. This module is designed for the frequency range 300 to 1000 MHz at a minimal HF input voltage of about 5 mV. The divider scales the frequency down to some 110 MHz. The quadrature circuit L11/C 36 is tuned to this frequency.

At pin 5 of IC 5, the demodulated base-band signal appears and is taken via R27 to the emitter follower T5. The emitter of this transistor carries the de-emphasis circuit. Transistor T6 enables the base-band signal to be inverted or remain upright in polarity, as required. In normal operation, i.e. $f_0 = f_{in} + f_{IF}$, bridge J2 is completed. The base-band signal is then inverted but the sync. pulses are positive. When, however, the local oscillator lies under the input signal, or with another satellite modulation plan, the bridge J1 is completed and the base-band is rendered upright.

Also, at pin 5 of IC 5, the AFC voltage is taken via R22/C 37 and processed in IC 6. As the stability of the SHF oscillator is, in fact, very reasonable, only a very simple AFC circuit is required. The AFC tuning voltage is generated here, which, according to the base-band signal polarity, is taken via bridges J3 or J4 and is dropped across R14 counter to the tuning voltage. This tends to keep the frequency of the SHF LO stable.

The base-band signal at MP 2 is available for further processing, if required, for example in a decoder or a satellite broadcast receiver. This signal is taken via C 58 and C 45 to the video and audio stages respectively.

2.2. The Video Stage

The signal radiated from the satellite transponder is modulated with a 50 Hz interlaced signal. This is balanced in a gated-control circuit in the video amplifier. First, the video signal (FBAS) is inverted by 180° by T7, taken to the sound-rejection circuit L15/C 60, and then on to T8. Adjusting P7 determines the video output amplitude. Transistor T9 is the FBAS-output driver stage. The output signal is taken via C 67 and IC 11 (TDA 2595). This circuit is a pulse-separator familiar in current colour-TV receivers. Together with the flipflop T12/T13, a small (2 μs) sampling pulse at line frequency is produced at the base of T11 which is coincident with the negative flank of the H-sync. signal. The metred output, including the momentary triangular value, is taken via R66 to the transistor T10 which functions as a synchronous detector. The triangular signal across C 64, together with the interlaced signal (also in phase) is amplified in IC 10, inverted, and then taken to the emitter of the first video stage T7. In this manner, the interlacing is balanced out.

If an unprocessed signal is required at the video output (descrambling operation), C 63 is wired in. This makes available an amplified FBAS signal (without sound carrier), together with an 50 Hz interlace signal.

2.3. Audio Stage

As a result of the differing standards for the satellite program, sound carriers can lie anywhere between 6 and 8 MHz in the composite base-band signal. Up to five sound sub-carriers could be transmitted simultaneously. The sound-carrier signal is taken via C 45 to the balanced mixer IC 7. The oscillator L13 operates in the 16.7 to 18.7 MHz range. The sound-IF of 10.7 MHz is available at IC 7 pin 2.

The sound-IF signal is taken via the ceramic filter QF to the demodulator IC 8. The AGC voltage output pin 12 IC 8 is filtered by R36/C 46/R35 and taken to the varicap D6. This provision changes the sound carrier according to the channel selected. The sound carrier selection can also be carried out manually by S1 before being selected by P5. Bridges J5 and J6 select the characteristics of the sound de-emphasis, either network R43/C 54 or R44/C 56.

Integrated circuit IC 11 contains a coincidence amplifier which switches off the sampled control via pin 7, thus reducing the video gain when there is no usable video signal available. This connection is also used at IC 8 to control the audio muting. The voltage on pins 4 and 5 of IC 8 (TBA 120 U AF amp.) is switched off via D7 thus cutting the sound output. If J7 is open, the muting circuit is switched out of circuit. Finally, IC 9 amplifies the de-emphasised sound to the output level of 300 mV.

3. CONSTRUCTION

Before proceeding with the loading of components on to the SAT-X printed circuit board (**fig. 4**), both the F-sockets must first be soldered into one half of the screening box (screwing them

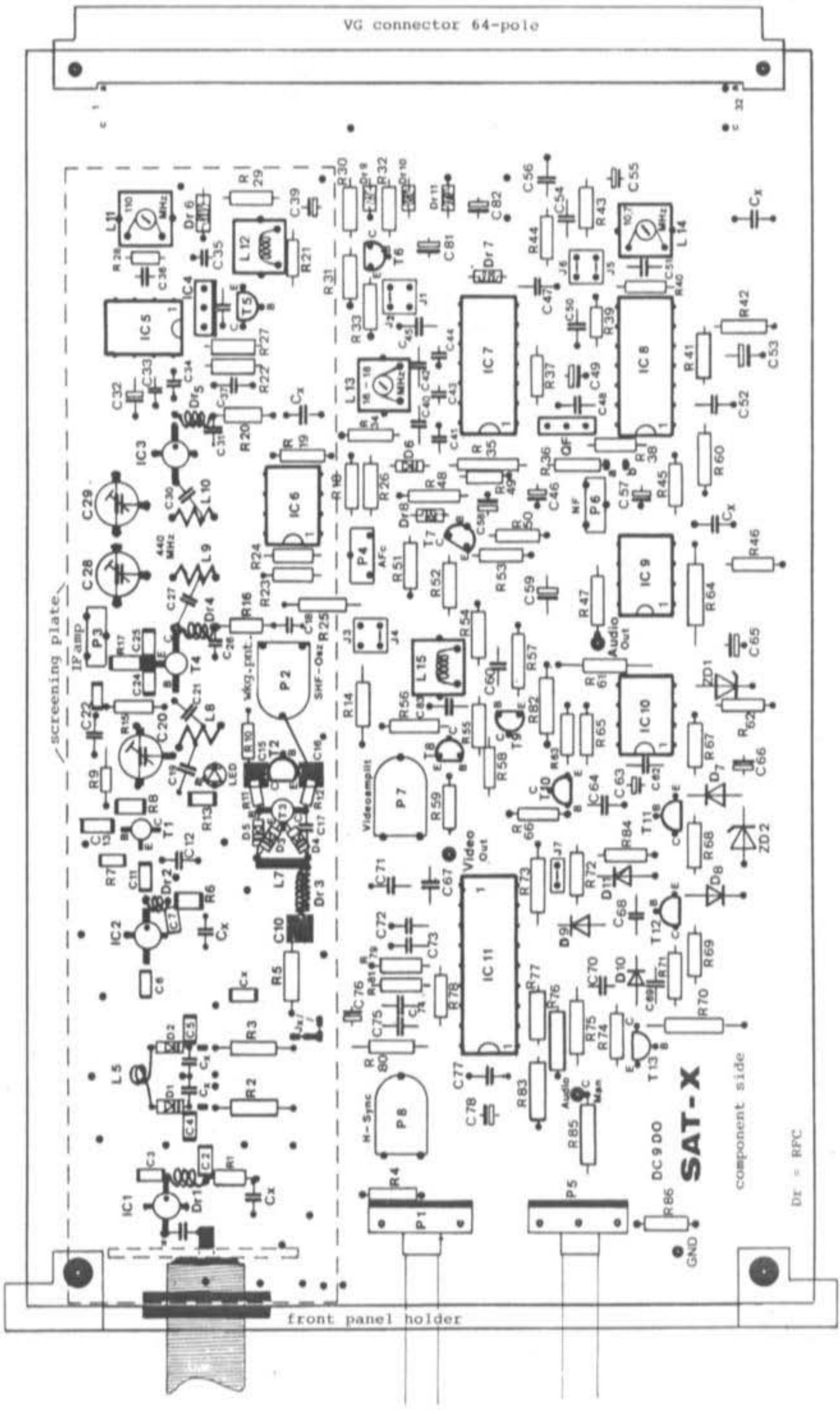

Fig. 4: SAT-X Eurocard, double-sided PCB, component layout plan

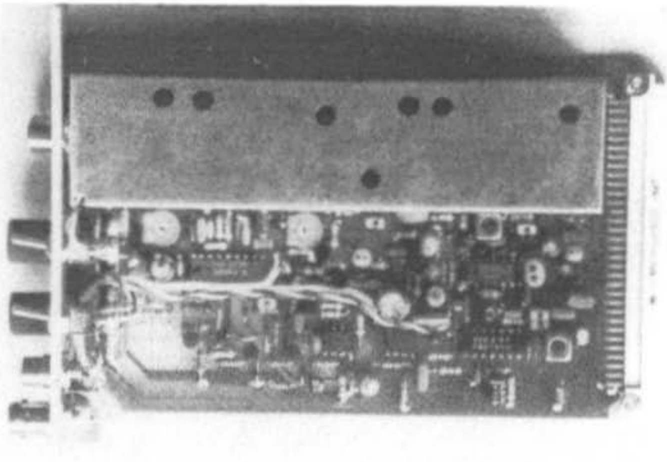

Fig. 5: Completed receiver

in is not good enough!). This tin-plate (**fig. 5**) is firstly soldered on to the PCB with the long side facing outwards. The de-coupling capacitors C 10, 15, 16 follow.

As the construction of tuned circuits above 1000 MHz becomes quite critical, the author has incorporated the input filter L2/L3 and the SHF LO coupling into the PCB. The varicap diodes D1/D2 must be soldered in with the shortest possible leads, otherwise the necessary linearity and sensitivity in the upper frequency part of the tuning range will be compromised (**see fig. 6**).

The inductor L5 forms the coupling element between L2 and L3 and consists of two turns of 0.5 mm CuL wire wound on a 3 mm former. **The golden rule for all the components in the SHF signal path is to keep the connecting leads to the components as short as possible.**

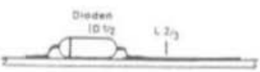

Fig. 6: Constructional detail

Fig. 7: Construction of SHF oscillator

As the circuit is constructed on a normal double-coated epoxy-glass board to keep the costs down, a small amount of "self-supported" construction cannot be avoided. The PCB losses would otherwise be so great, that at these frequencies, the oscillator would not operate cleanly. Therefore, L7, comprising 1 mm silvered wire, 5 mm long (approx.), is self-supporting lying a maximum of 2 mm over the surface of the board and soldered as directly as possible (max. lead length 2 mm) to the ends of the varicap diodes. The ends of R11/R12 and the connections of T3 are also treated in the same manner.

The author proceeded with the SHF oscillator construction (**fig. 7**) in the following manner: – First, T3 has its emitter lead bent at 90° downwards and is pushed through the board. The negative ends of D3/D4 are also suitably bent and also inserted through the emitter hole in the board. Between the free ends of D3 and D4 comes L7. All connections are then soldered, both ends of T3's emitter being soldered. The shortened ends of R11 and R12 are then soldered to the (unsupported) base lead of T3 and C 15. Also, the collector of T3 and C 16 are soldered in a similar fashion. Then follows D5, C 17, RFC 3 and R10. The slider of P2 is then connected to C 16 with a length of wire.

The inductors L8 to L10 consist of 1 mm silvered wire wound on a 3 mm twist drill.

Inductor L8 is tapped one turn from the cold end by capacitor C 19 also 1.5 turns from the cold end by C 21. Capacitor C 27 is tapped into L9 approximately at the centre turn. Capacitor C 30 is tapped into L10 approximately 3/4 turn from the cold end. All coils should be wound in the same direction. The demodulator inductor L11 consists of 4 turns of CuL 0.5 mm wire wound on to a 4 mm dia. coil-former, fitted with a ferrite core and enclosed in a screening can. The windings here must be very tight.

The RF chokes (RFC 1 to RFC 5) consist of 0.4 mm CuL wire (see component plan) wound on a 2.5 mm twist-drill shaft and mounted self-supporting above the board with the shortest possible solder leads.

3.1. Component List

Transistors:

T1:	BFW 34a
T2:	BC 558
T3:	BFQ 69, BFW 91 AT 41485, 42085
T4:	BFW 34a
T5/T12:	BC 548
T6:	BC 547
T7:	BC 328
T8:	BC 875 (Darl.)
T9/T10:	BC 876 (Darl.)
T11/T13:	BC 558

Integrated Circuits:

I1...I3:	MSA 0304 (Avantek)
I4:	7805
I5:	SL 1452 (Plessey)
I6:	TL 082
I7:	S 042 P
I8:	TBA 120 U
I9:	TL 081
I10:	TL 081
I11:	TDA 2595

Diodes

D1...D5:	BB 505 b
D6:	BB 405 b
D7...D11:	1 N 4148
ZD 1:	ZPD 8.2
ZD 2:	ZPD 5.1

LED 3 mm red

Trimmers and pots.

P1:	10 k	Poti 4 mm
P2:	100 Ω	
P3:	47 k	upright
P4:	4.7 k	upright
P5:	22 k	Poti 4 mm
P6:	47 k	upright
P7:	470 Ω	
P8:	47 k	

Tin-plate box 145 x 35 x 30 mm

Bu 1/2:	F panel socket
Bu 3:	BNC panel socket
Bu 4:	Cinch

VG – 64 pole
Front-panel holder for 19" 3HE part front panel
Part front panel 3HE/4cm wide

Coils:

L1...L4:	etched
L5:	2 turns CuL wire 0.5 mm, 3 mm int. dia
L6:	etched
L7:	5 mm long 1 mm silvered wire (10 mm for f_o under IF)
L8...L10:	2.5 turns 1 mm silvered wire, 3 mm int. dia
L11:	4 turns CuL 0.5 mm wire, 4 mm coil former with ferrite core
L12:	RFC (Dr) 33 µH
L13:	18 turns CuL 0.25 mm wire, 4 mm coil former with ferrite core
L14:	10 turns CuL 0.25 mm wire, 4 mm coil former with ferrite core
L15:	RFC (Dr) 22 µH

HF-RFCs:

Dr 1/2:	5 turns CuL 0.4 mm wire, 3 mm int. dia
Dr 3:	10 turns CuL 0.4 mm wire, 2 mm int. dia
Dr 4:	3.5 turns CuL 0.4 mm wire, 3 mm int. dia
Dr 5:	5 turns CuL 0.4 mm wire, 3 mm int. dia
Dr 6:	1.5 µH
Dr 7:	12/15 µH
Dr 8:	33 µH
Dr 9...Dr 11:	100 µH can be replaced by wire bridges
QF:	ceramic filter Murata SFE 10.7 MA

Resistors:

R 1: 220 Ω 1/20 W	R 23: 470 k	R 45: 3,9 k	R 66: 6,8 k
R 2: 22 k	R 24: 3,3 k	R 46: 10 k	R 67: 270 Ω
R 3: 22 k	R 25: 8,2 k	R 47: 1 k	R 68: 10 k
R 4: 27 k	R 26: 10 k	R 48: 18 k	R 69: 2,2 k
R 5: 10 k	R 27: 68 Ω	R 49: 4,7 k 1/20 W	R 70: 2,2 k
R 6: 220 Ω 1/20 W	R 28: 470 Ω	R 50: 150 Ω	R 71: 33 k
R 7: 1 k Chip	R 29: 82 Ω	R 51: 1,5 k	R 72: 47 k
R 8: 1 k Chip	R 30: 18 k	R 52: 560 Ω	R 73: 10 k
R 9: 3,9 k 1/20 W	R 31: 10 k	R 53: 680 Ω	R 74: 6,8 k
R 10: 4,7 k 1/20 W	R 32: 220 Ω	R 54: 1,2 k	R 75: 1,5 k
R 11: 2,2 k 1/20 W	R 33: 220 Ω	R 55: 1 k	R 76: 100 k
R 12: 180 Ω 1/20 W	R 34: 100 k	R 56: 330 Ω	R 77: 47 k
R 13: 1 k Chip	R 35: 100 k	R 57: 33 Ω	R 78: 27 k
R 14: 100 k	R 36: 33 k	R 58: 27 Ω	R 79: 12 k
R 15: 10 k	R 37: 1 k	R 59: 47 Ω	R 80: 680 Ω
R 16: 100 Ω 1/20 W	R 38: 330 Ω	R 60: 4,7 k	R 81: 150 k
R 17: 75 Ω Chip	R 39: 4,7 k	R 61: 39 k	R 82: 33 Ω
R 18: 10 k	R 40: 1 k	R 62: 6,8 k	R 83: 56 k
R 19: 10 k	R 41: 10 k	R 63: 56 k	R 84: 10 k
R 20: 220 Ω 1/20 W	R 42: 47 k	R 64: 180 Ω	R 85: 10 k
R 21: 330 Ω	R 43: 5,1 k (4,7 k)	R 65: 150 k	R 86: 3,3 k
R 22: 330 k	R 44: 680 Ω		

Capacitors:

C 1: 68 p	C 22: 1 n	C 43: 82 p	C 64: 100 n
C 2: 470 p Chip	C 23: 470 p Chip	C 44: 82 p	C 65: 10 - 22 µ
C 3: 68 p Chip	C 24: 470 p Chip	C 45: 27 p	C 66: 10 - 22 µ
C 4: 1 n Chip	C 25: 470 p Chip	C 46: 4,7 µ	C 67: 220 n
C 5: 1 n Chip	C 26: 1 n	C 47: 22 n	C 68: 100 p
C 6: 27 p Chip	C 27: 56 p	C 48: 100 n	C 69: 33 p
C 7: 470 p Chip	C 28: 2 - 12 p Trimmer	C 49: 10 µ	C 70: 180 p
C 8: 1 n *	C 29: 2 - 12 p Trimmer	C 50: 1 n	C 71: 220 n
C 9: 470 p *	C 30: 47 p	C 51: 330 p	C 72: 10 n
C 10: 470 p Plate	C 31: 1 n	C 52: 100 n	C 73: 100 n

C 11: 39 p Chip	C 32: 4,7 μ	C 53: 10 μ	C 74: 4,7 n
C 12: 12 p	C 33: 1 n	C 54: 22 n	C 75: 22 n
C 13: 470 p Chip	C 34: 33 p	C 55: 22 μ	C 76: 4,7 μ
C 14: 470 p *	C 35: 100 n	C 56: 47 n	C 77: 100 n
C 15: 1 n Plate	C 36: 18 p	C 57: 1 μ	C 78: 100 μ
C 16: 1 n Plate	C 37: 100 n	C 58: 47 μ	C 79: 47 μ
C 17: 33 p	C 38: 470 n	C 59: 220 μ	C 80: 220 μ
C 18: 1 n	C 39: 47 μ	C 60: 27 p	C 81: 1 μ
C 19: 47 p	C 40: 390 p	C 61: 220 μ	C 82: 10 μ
C 20: 2 - 12 pTrimmer	C 41: 56 p	C 62: 4,7 n	C 83: 10 p
C 21: 22 p	C 42: 270 p	C 63: 4,7 μ	C x: 1 n

* not strictly required if VHF constructional practices are observed!

4. ALIGNMENT

After the construction of the complete satellite TV receiver and checked for soldering and wiring defects, the alignment can commence. The required test equipment is as follows:

a transistor voltmeter,
an oscilloscope,
an HF-FM generator and
monitor or TV set with an AV input.

First of all, all HF trimmers, coils and adjusters (except P7) are brought to their mid-range position. Pot'meter P7 is turned fully clockwise. All bridges, except J2, are removed. At MP 1, an oscilloscope or a VTVM is connected (note: high impedance point).

Following this, an FM-modulated (at 1 to 10 kHz) 440 MHz signal is applied to the input of I3 via a 2 pF capacitor. Coil L 11's tuning slug is then turned until 2.5 V appears at MP 1.

The test signal is then applied to the base of T4. Trimmers C 29 and C 28 are turned, in that order, until the voltage at MP 1 has been restored to 2.5 V. The HF sensitivity being slowly adjusted by means of P3, to achieve this result.

The test signal is then applied to the emitter of T1 and C 20 adjusted to the aforementioned values. The alignment is continued with the retuning of C 29, C 28 and C 20, in that sequence, checking with an oscilloscope that the symmetry of the AF signal at the video output is maintained.

Providing that the SHF oscillator has been constructed correctly, the LED will illuminate. Potentiometer P2 is then turned clockwise until the LED is just extinguished. P2 is then turned back about a quarter turn.

A correctly tuned unit should indicate a uniformly distributed noise on the monitor connected to the video output.

The SHF signal is then taken from the outer unit to the input of the receiver. Normally at this stage, transmissions should be apparent when the receiver is tuned with P1 but not necessarily with good quality. It is advisable to select the ASTRA satellite for the initial tuning process as its radiated power is very high.

Should no signals be observed, the SHF oscillator should be corrected with P2 for the best oscillatory condition and maximum amplitude. Fine tuning is carried out by tuning the receiver to a mid-band satellite program. The tuning of C 29, C 28, C 20 and P3 is then repeated until the best picture has been received. In the meantime, the video output amplitude must be restored to maximum using L 11. By carefully tuning control P1, the IF can be checked for tuning symmetry. A correctly tuned IF results in the appearance of white flecks when P1 is de-tuned to the right (above) and black flecks when it is de-tuned to the left – by an equal amount.

A weak signal, or bad alignment, results in no noise at all on a received FM video signal – a complete contrast to the AF output! Broken video lines are in evidence as well as the sort of flecks which occur on video recorders. These interferences are particularly prominent upon strongly coloured areas of the picture (high modulation).

When the HF alignment has been carried out correctly, the tuning of the sound IF can be commenced by first of all opening bridge J7. For this, in manual (P5 turned 1/4 from ACW-position), the core of L 13 is turned for maximum audio output. Then, L 14 is tuned for maximum AF amplitude. At pin 12 of I8 there should be a potential of 6 V. By iterating the tuning of L 13 and L 14, the correct alignment can be accomplished. If an FM-signal generator giving an output at 10.7 MHz is available, the alignment can proceed without any trouble. During its progress, the bridges J1/2 should not be completed and the test signal is applied to MP 2. It is to be ensured that the tuning of the first sound carrier occurs at 6.5 MHz. By manual tuning of the tone, using P5 – according to the satellite program – three to five various sound carriers should be received.

The video amplifier does not need any special alignment. With P7, the amplitude at the output is adjusted to approximately 2 V_{pp}. The line-frequency adjustment P8 is adjusted so that upon reception, the sound muting opens cleanly and the tone is heard without distortion (bridge J7 closed).

The author has successfully used components from old disused TV and video-recorder equipment for the construction of several receivers. If this equipment is not older than six or seven years, the chances are very good that the tuner contains SMD components/modules. Many video recorders are also fitted with 1/20 W resistors which are very useful for this construction. A

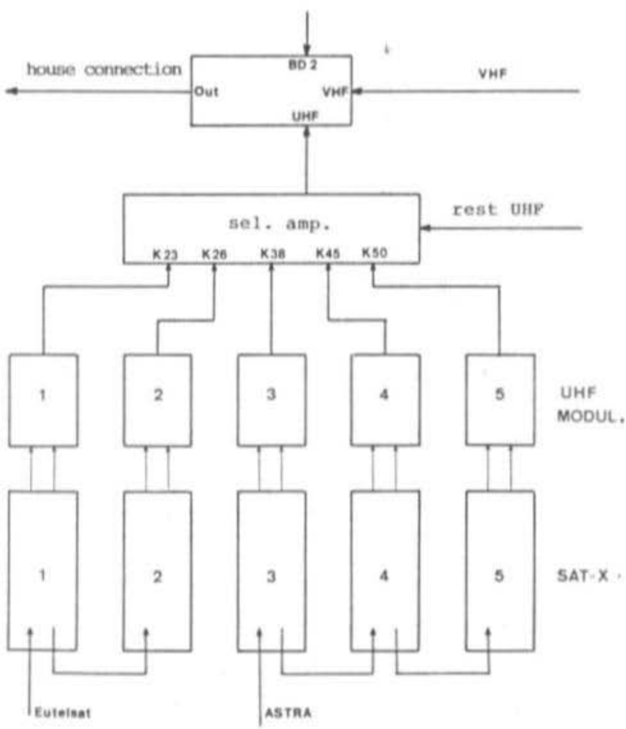

Fig. 8: A small communal installation

further possibility for obtaining suitable components are junk RF boards from electronic dealers.

5. OPERATING EXPERIENCE

The signals at the video and audio outputs are, of course, compatible with standard AV inputs of a normal television set. With a suitable VHF/SHF modulator, the satellite programme can also be coupled into a domestic antenna installation.

In order to check the project's reliability of replication, the author built five of the receivers, as described, using both SMD as well as conventional modules. All circuits exhibited identical behaviour during tuning. Nevertheless, there were some differences in evidence concerning the sensitivity, also there was some tendency for the 440 MHz IF to break into self-oscillation. Therefore, the SMD modules listed in the component complement are to be preferred. Together with five VHF modulators (from Philips video recorders) the author installed the five receivers into a 19" equipment rack. These channels were fed via five tunable antenna preamplifiers into a small 4-user domestic installation (**fig. 8**).

The same programs were simultaneously received on the five receivers using the following satellite stations:

Super Channel
SAT 1 of EUTELSAT 1 F4 and Sky Movies,
MTV Europe and Eurosport of ASTRA 1A.

Although the receivers were separated physically by only 4 cm from each other and without any RF screening (**fig. 9**), there was no evidence whatsoever of mutual interference. However, each individual Sat-receiver was provided with its own 12 V voltage-regulated power supply (7812) in order to avoid coupling via the video amplifiers. For this purpose, the author has provided a 64-pole multi-pole connector with two 12 V connections. When only one is in operation, the connections 13 a and 19 a can be bridged. At pin 15 a the 35 V tuning voltage can be picked up.

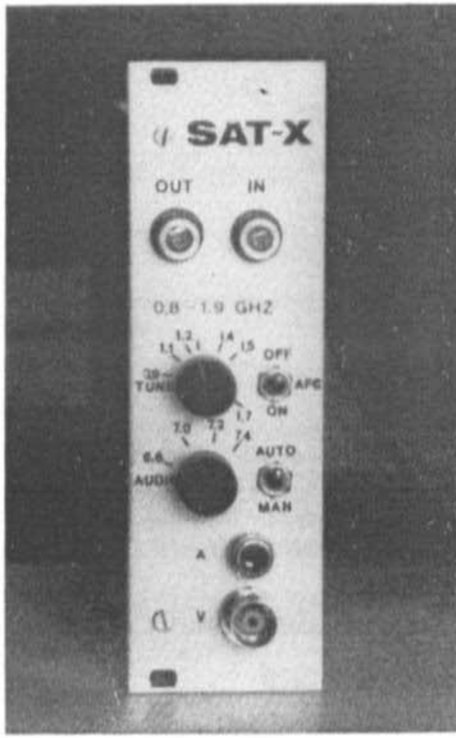

Fig. 9: Satellite-receiver front panel

The reception results behind the house antenna-terminal modulators was exceptionally good. The reception of video text, where present in the transmission, also presented no communications problems.

To complete the construction details, the following points must also be explained:

In the HF part of the conductor side of the PCB, the + 12 V track should be fitted at intervals with additional chip capacitors of approximately 1 nF to earth. This guards against RF coupling taking place via the supply circuits.

The 7 MHz sound carrier can have a very high noise level. If this is the case, check the 440 MHz

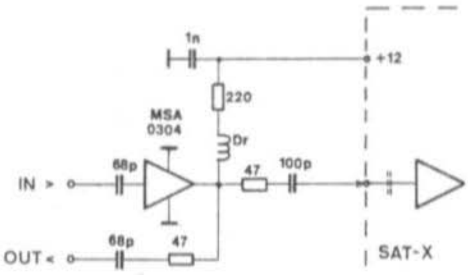

Fig. 10: Wideband pre-amplifier for 800 to 1800 MHz

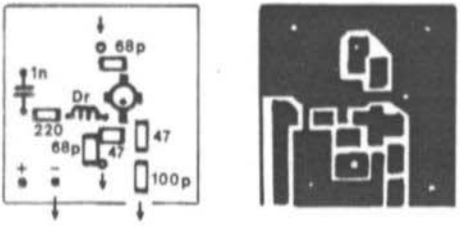

Fig. 11: A small board used for the circuit of Fig. 10

IF circuits. Try reducing the value of R 28 from 470 Ω to 330 Ω in order to widen the IF response.

Should the video carrier be audible in the sound output, an RFC of 8.2 to 10 μH connected across R 37 will bring some improvement.

If the receiver is only required to be operational on one channel, the SHF-input tuned circuit can be removed from the tuning voltage and a 47 k trimmer mounted into the holes provided. Jx is connected to the 35 V tuning voltage. The trimmer is then tuned so that the circuit accords exactly to the frequency of the desired input channel.

6. MODIFICATIONS

To improve the receiver's versatility, the author has developed a further pre-amplifier which is fitted directly behind the input connector. This gives an extra 6 dB gain which enables the antenna cable to be extended for other purposes. This pre-amplifier is also fitted with an MMIC (**fig. 10**): it serves also the purpose of decoupling the receiver from the antenna input cable. The double-sided PCB (**fig. 11**) is fitted on the plain side with input and output connectors, suitably isolated from the copper foil, and the whole module is then mounted directly on to the receiver-antenna input socket and connected to the solder tracks.

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Shortwave Reception – Based on the Thirties Principles Part 1

This article is published owing to its many interesting nostalgic associations, and easy to understand details but not because the editor is of the opinion that our readers necessarily have the desire to construct HF receivers for AM reception. The article should evoke enjoyment and give rise to many a flash of understanding and exclamation of "aha"!

Shortwaves using A3 modulation have been the vehicle for conveyance of news around the globe for more than 60 years. Amplitude modulation has persisted to the present time in spite of more favourable means and will be certainly in widespread use for the foreseeable future. It is for this reason that any investment in terms of both time and money in this technology cannot be trivialized. Also, it may be overlooked that the listener's investment, taken as a whole, is much higher than the broadcaster's. It can be taken for granted, that if the broadcast was transmitted in any other mode than A3, it would simply not be heard by its intended audience. Therefore, it would appear to make sense to improve this method of broadcast reception.

Such improvements can certainly not be made if the receiver is left with yet more knobs and buttons with which to manipulate filters, attenuators, time constants, clippers/limiters and suppressor etc. The optimal receiver for the normal listener should, ideally, have only a means of selecting the required frequency and then, without any further measures, receive the best-possible AF signal consistent with the prevalent radio conditions. This requirement, that even grandma can operate the radio, is the highest demand which can be imposed! Two principles of long standing are to be recalled in this article in order to develop the point. These are: super regeneration and the logarithmic demodulator. Other details will be also considered.

1. PECULIARITIES OF HF RECEPTION

1.1. Frequency Range and Adjacent Channel Selectivity

The shortwave stations considered to be of interest lie in the frequency range 3 MHz to 28 MHz. The so-called tropic bands (under

5 MHz) do not, nowadays, have any relevance in the tropics. The lower frequencies are used mainly for regional reception. Better long-distance reception is to be obtained in the near vicinity of the MUF (Maximum Usable Frequency) whereby an MUF of over 22 MHz is hardly ever experienced. The Citizen Band is, however, viable on the 27 MHz band.

According to several sources, the propagation noise of 50 dB above the thermal noise of a 50 Ω source at 300 K at frequencies below 5 MHz, is caused by atmospheric and galactic variations. At 30 MHz, it is lower at 20 dB. The frequencies between these extremes can be interpolated in a linear fashion. Man-made interference can exceed the propagation noise by some 10 to 20 dB. If the latter is not taken into consideration, the receiver, of 8 MHz bandwidth, must have an internal noise of less than 10 μV (− 87 dBm) at 5 MHz and less than 0.8 μV (− 109 dB) at 30 MHz, if optimum reception is to be achieved with a correctly tuned receiver and a suitably matched antenna.

1.2. Receiving Aerials

The high external propagation noise, in the short-wave regions, makes the employment of optimally cut aerials, for the desired frequencies, superfluous for listening purposes. Since it is much easier to design a receiver with a frequency-independent internal noise, the antenna gain, at the lower frequencies, may be allowed to drop by as much as 30 dB. One such aerial is the CB groundplane. But this is not a provisional aerial. It is designed to match the equipment in use and can, moreover, be installed with standard masts and supporting stays on the house roof right out of the way of man-made interference sources.

The interference signal/noise can be drastically improved by the employment of directive arrays. This is only necessary in special circumstances as such antennas require a lot of real estate.

1.3. Selectivity

The so-called "Luxembourg Effect" (an apparent non-linearity of the transmission medium) has not, as yet, been reported on shortwaves. If stations appear where they shouldn't be, it can be attributed directly to the receiver. These problems have become more acute with the widespread use of semi-conductors and, of course, with the steady rise in the power used by broadcast stations. Break-through from adjacent-channel stations, on the image frequency or on the IF frequency, are easily recognised and the measures to counter them are well known.

Intermodulation between several stations, which are closely adjacent in frequency, cause signals to appear at frequencies where there should be no signal. A receiver intermodulation-free input range of 100 dB is not yet commonly, technically feasible but is essential for the serious listener. Using the above mentioned receiver internal noise of − 109 dBm would mean an IM_3 of + 41 dBm. Such values, according to Oxner (5) are not yet obtainable even with a switched FET mixer.

1.4. Noise Figure and Usable Signal Range

A receiver which is connected to the above mentioned matched antenna and has a noise figure of 20 dB, would worsen the interference s : n ratio by some 3 dB. Employing a much smaller noise figure will not be of the slightest help. Receiver concepts which start at the mixer are nowadays the state-of-the-art technology. Receiver pre-amplifiers are required where short, temporary aerials are frequently employed and also to reduce radiation of the local-oscillator signal.

A3 modulation has no modulation gain. A received signal input power of − 83 dBm is required in order to receive a transmission at a respectable 26 dB s : n ratio at the receiver output. If there is an intermodulation-free range of 80 dB, then the highest input power − 29 dBm must not be exceeded. The usable signal range would be up to 54 dB.

1.5. Low-Frequency Range and Distortion

A shortwave broadcast cannot be a high-fidelity system. A broadcast transmitter separation of 10 kHz only allows a modulating frequency range of up to 5 kHz as a theoretical maximum. Such a limited audio frequency can only transmit speech with an 80 % fidelity, 98 % logatom-intelligibility and 100 % intelligibility. In addition, as there is no such thing as an ideal selection filter, even less than this must be accepted. A bandwidth of only 3 kHz is usually considered as being satisfactory for shortwave broadcast listening. The above percentages then become: 60 %, 88 % and 98 % – a considerable deterioration! Many commercial shortwave receivers do not even have a 3 kHz bandwidth. The reception of low signal-strength signals places a higher emphasis on the logatom-intelligibility than on the intelligibility, especially when foreign languages are being received. That is why even the smallest increase in the overall AF bandwidth is advantageous.

A peculiarity of the human ear is that its mean frequency, as far as speech is concerned, is around 1.6 kHz. A balanced sound impression is attained when lower and upper frequencies lie symmetrically disposed, in a logarithmic manner, about this frequency. Should the upper limit have a frequency of 3.3 kHz, then the lower frequency of the response should lie around 800 Hz. Another mean frequency applies in the case of music; namely 1 kHz. In the above example this would entail a lower limit frequency of 300 Hz. The long forgotten "music/speech switch" in AM radios did have an audio-physiological function.

The modulation signal suffers both linear and non-linear distortion even before modulation of the RF carrier takes place in the radio transmitter. In order to obtain more range and more punch, a high but almost constant degree of modulation is necessary. Measures such as response limiting, dynamic-range compression, pre-emphasis and clipping have to be introduced. After the modulated signal is radiated, it is subject to selective fading caused by the vacillations of the propagation medium which affect both the carrier and the modulation. Further imperfections to the modulation signal take place in the receiver selection and demodulation processes. The human ear is supposed to be able to recognise distortion which is above about 1 % and phase reversals of 180°/octave especially if they occur in the middle of the hearing range. This starts to have a large impact upon speech intelligibility when the distortion rises above about 20 %. With dynamic distortion caused either automatically, or by the hand of the sound engineer, is nowadays so commonplace as to be disregarded by the listener.

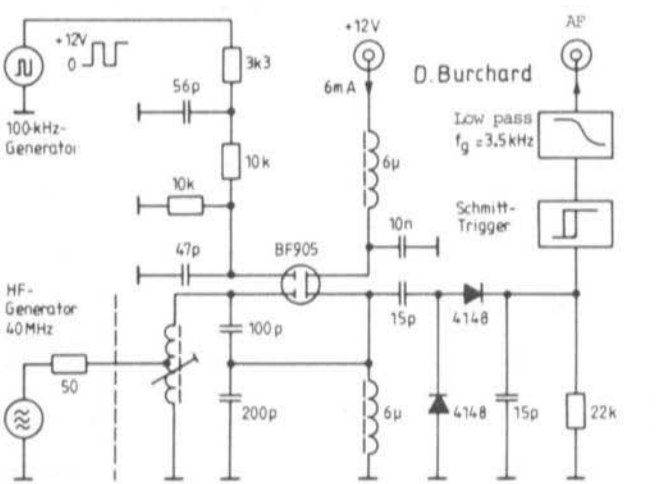

Fig. 1:
Circuit of a logarithmic detector for the examination of its characteristics.
Inductor: 2 + 8 W, 0.35 CuL on a Vogt coil former 3131600000

It may be concluded, that shortwave broadcasting is not the medium for the transmission of music! Enjoyment of this form of entertainment requires no carriers with the speed of light but can be better left to more suitable media. These extend from the gramaphone record, right up to the Digital Audio Tape-recorder (DAT).

1.6. Automatic Gain Control (AGC)

As opposed to A3J (SSB) modulation, A3 has the advantage that it has a reference – namely the carrier – by which the magnitude of the sidebands may be assessed. This facilitates the employment of an automatic gain control device to regulate the variations in received signal strength in order that a near constant level of audio is presented to the loudspeaker. The disadvantages of A3, such as the unnecessary expenditure of energy in the carrier and also the superfluous redundancy of the sidebands, are so well-known that they will not be gone into any further.

The normal AGC takes the form of a negative feedback which is not allowed to be too responsive (otherwise there would be a tendency to demodulate the signal – ed.). It would be advantageous, if a shortwave receiver had a form of detection which is inherently self-limiting in the same manner as that which FM receivers employ – i.e. an AGC system without any time constants. Such a scheme is possible and will be explained below.

Following the foregoing review of a few basic considerations, concrete principles will now be presented by which the reception of HF receivers may be improved.

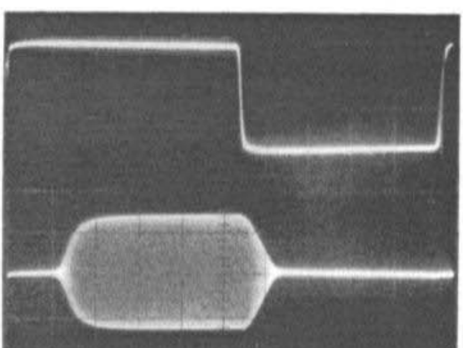

Fig. 2: Input and output waveforms of a super-regenerative detector
X: 1 μs/div., Y1: 5 V/div., Y2: 5 V/div.

2. SUPER REGENERATION

This principle has been known for many decades; a comprehensive description has been made by Barkhausen (1). In the past, super-regenerative receivers, as simple shortwave and remote-control receivers have played a significant role in the development of VHF reception. The circuits were mostly decidedly primitive, the characteristics sparse and the equipment radiated interference signals into other services. With the knowledge from (1), better construction techniques would have been possible. Therefore, what was already known in 1934 will now be repeated:

It has been the experience that every oscillator, sooner or later, automatically starts to oscillate and then settles down to a constant final amplitude. This is the case, even if it is switched on in the absence of electrical perturbations which would give the oscillatory tank circuit an initial start energy. An oscillator circuit for 40 MHz is shown in **fig. 1** with a prospect of its possible use in a shortwave receiver. Via the transistor's G2 voltage, the circuit can be switched on and off (quenched) – the HF generator not being needed for the time being. The diodes connected to the source circuit can also be neglected for the moment.

In the oscillogram of **fig. 2** it can be seen that the switching potential is the upper trace and the resulting HF oscillation is the lower trace. More than a μs elapses before a noticeable HF amplitude is produced. A further μs is needed to obtain a constant amplitude. After switch-off, another μs elapses until the HF energy completely

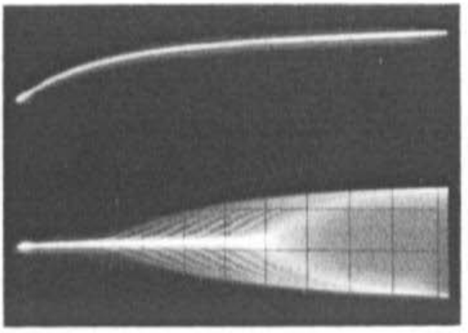

Fig. 3: Build-up of oscillations in a super-regenerative circuit. The start power is in the range − 40 dBm rises to − 150 dBm.
X: 200 ns/div., Y1: 5 V/div., Y2: 5 V/div.

disappears. A precise analysis shows that an important indiscernible sequence of events occurs at the very on-set of oscillations at the switch-on. With the degree of feedback coupling applied in this case, about 3.8 oscillations of the tank circuit are required in order to obtain a rise in amplitude to the value of "e" (1 N = 8.7 dB). The figure 3.8 is the inverse value of the logarithmic decrement of the circuit! The oscillations start in accordance with an e-characteristic, as long as the feedback is held constant. By exceeding the linear working range of the transistor, the feedback reduces, which can be recognised by the different curvature of the envelope, until finally it is just sufficient to make up the losses in the tank circuit. The tank circuit oscillation is then constant in amplitude. Upon switch-off, the amplitude decreases, again in accordance with an e-function. The oscillation fades away completely after a certain time has elapsed.

From the study of the mechanism of the build up to oscillations, it is now known that within the range of the e-function, every instantaneous amplitude at a certain point in time has a distinct relationship to other instantaneous at other points in time. This allows a rough estimate to be made of the initial amplitude: 1.3 μs after switch-on, the amplitude has risen to about 4 V − 52 oscillations have taken place by this time. From this information, it can be deduced that an initial amplitude of 4 μV is required. The active circuit increases this by a factor of 10^6 (120 dB) until the final steady-state oscillation amplitude is achieved. A sampling-gate could sample the RF directly, or it could be done by means of a proportional voltage derived from a rectifying diode. A μV measurement device is thereby available for examination of the switch-on period.

A closer examination of fig. 2 reveals a certain lack of clarity in the leading edge of the envelope which is not present in the trailing edge. This leads one to the conclusion that the initial voltage is not constant – it has noise superimposed upon it! Now it can be surmised, that when the trailing edge decays to zero, it is at that time when the RF-oscillation amplitude falls under that of the noise. If now, the transistor is switched on again but, before the trailing edge has completed its decay function, the new start amplitude is now the residual decay amplitude of the last impulse. A coherent RF wave-train is obtained but this time without the noise in the leading edge.

Examining the build-up to the oscillation even closer, it can be seen from **figure 3** that the time axis has been stretched so far that the switching voltage appears to be quite slow. Now, the details of the leading edge of the RF-oscillation envelope can be seen with more clarity. The RF generator was switched on and delivers a certain initial power. At − 120 dBm, i.e. at the extreme right of the envelope, it can be seen that a strong noise voltage is in evidence. At − 110 dBm the noise is lower and it is finally attenuated until it disappears at all higher start powers (10 dB steps). The start voltage of fig. 2 can now be attributed to a noise voltage of between − 110 and − 120 dBm. This is approximately the thermal noise power of all the loss resistive components of the tank circuit when the transistor is switched off.

The leading edge moves to the left when the start voltage is increased by a factor of 10, assuming that the same units of time are used. **The time taken to attain a certain RF amplitude is a logarithmic function of the start power. The diode circuit of fig. 1 followed by a trigger and a low-pass filter converts the super-**

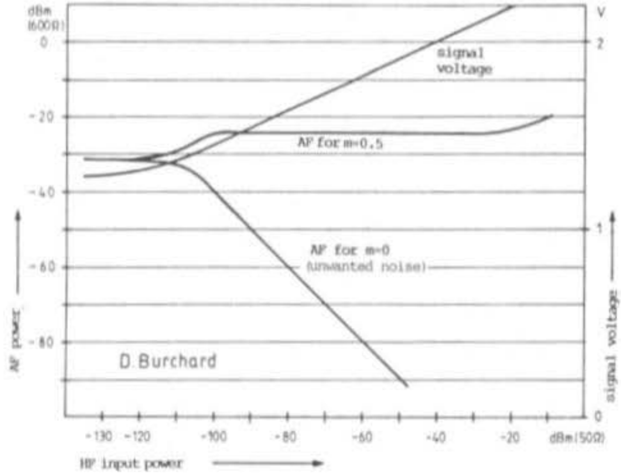

Fig. 4: Audio voltage. Wanted and unwanted signal outputs of the logarithmic detector

regenerative oscillator into a logarithmic demodulator with 80 dB, or more, dynamic range. Its characteristics were measured and presented in **fig. 4**. They are quite remarkable:

1. The circuit amplifies immediately from the internal noise level to a working power, here 5 mW. If an amplifying device of higher power is taken, it can be either 1 W of 10 kW (!), it only requires a little more time.
2. If the amplitude is sampled at a certain time immediately following switch-on but well before the on-set of saturation, one has a linear, highly-sensitive demodulator but which does not possess a barrier point.
3. Measuring the time from switch-on until a certain amplitude has been reached, which must be smaller that the final amplitude, one has a logarithmic demodulator which is capable of working over a dynamic range of many decades.
4. Following the switch-on, the oscillation amplitude quickly increases to a higher level than the noise voltage of the, by now, fully functioning active element. Although it has not been theoretically examined, measurements lead to the conclusion that the circuit has a low noise-figure.

The disadvantages of the super-regenerative feedback cannot be overlooked:

1. The working frequency, the super-regenerative feedback frequency and the selective circuits, both RF and audio, are intimately connected with one-another. Variation of one parameter alters all the others.
2. The incoming signal and the 100 % modulating super-regenerative feedback signal, exist together at the detector input. It is therefore not possible to connect an antenna at this point. When connecting a signal generator, it is debatable whether the instrument's attenuator or its ALC circuits will receive the damage!
3. To ensure stable operation, the super-regenerative feedback stage requires a source which possesses constant characteristics, as the effects of the internal resistance and phase angle in the tank circuit losses are effectively eliminated. This is the same problem that is evident with highly-developed antenna pre-amplifiers.
4. Placing a pre-amplifier before the stage will probably deteriorate both the noise figure and the dynamic range.
5. The demodulated audio frequency can have, as a maximum, the bandwidth of half the

super-regenerative feedback frequency and this is nevertheless only a fraction of the RF bandwidth. Increasing the sensitivity is bought at the price of a decrease in the bandwidth. Filtering the super-regenerative feedback frequency from out of the audio stages does not, however, present any great problems.

In spite of these disadvantages, there are quite a few uses for this type of demodulator: power and field-strength measurements, at a fixed frequency, in the course of antenna and radiation experiments, demodulators for spectrum analyzers and panoramic receivers – also for the shortwave receiver to be described below.

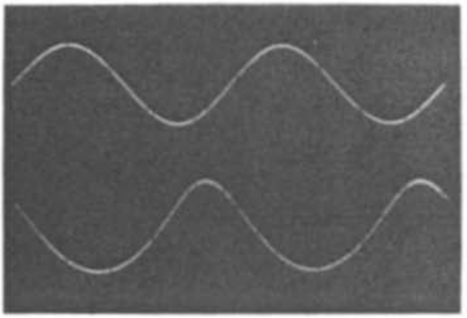

**Fig. 5: Audio distortion in the logarithmic demodulator at m = 0.5
X: 200 μs/div., Y1: 50 mV/div., Y2: 2 V/div.**

3. LOGARITHMIC DEMODULATION

The influence of the rectifier circuit upon the demodulation of AM with regard to linearity, selectivity, and AGC, are hardly ever subjects for discussion. Tüxen (6) has done a lot of work in this area and found considerations which are worthy of a second look.

There is, first of all, the question of the rectifier time constant. The usual diode arrangement has a reservoir capacitor in order to bridge the time between two RF maxima. This is effectively peak rectification. The time constant should be small enough to follow the highest audio frequency. In contrast to this is the anode-demodulator which is fast enough to follow the RF impulses in the modulated waveform. This results, of course, in quite different characteristics. Pure anode demodulation is not really necessary. Owing to the selective circuit before the rectifier, it is not easy to alter the amplitude of the RF. A peak rectifier is fast enough to follow such changes. Under these conditions, a **linear** rectifier delivers distortion-free audio. A simultaneous presence of an unwanted carrier will, however, produce distortion – the frequency difference between wanted and unwanted carrier will appear directly in the output together with their respective modulations. A quantitative analysis is available in (6).

Even worse is the case for **square-law** demodulation: this afflicts all diodes working with very small RF voltages. The audio, even in the absence of spurious signals, is already distorted. The modulation of an unwanted carrier is particularly prominent – a common enough design fault in cheap portable radios. If the radio is turned so that the ferrite antenna is pointed to the optimal direction for the reception of a weak transmission, the s : n ratio may become even worse than before.

It would appear that a detection characteristic which is distorted in the inverse direction, also produces a reverse effect. The **logarithmic** characteristic is capable of suppressing the modulation of the unwanted carrier completely: it will not, however, get rid of the difference tone.

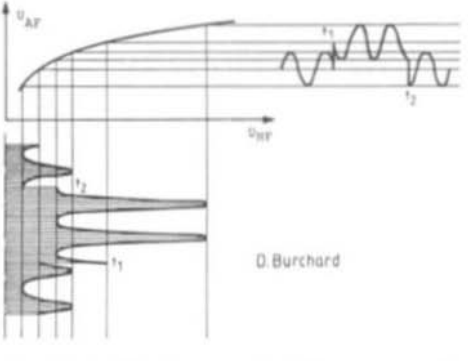

Fig. 6: A logarithmic detector requires no automatic volume control

In addition, the unwanted carrier must be smaller than the wanted carrier – otherwise it will be the wanted carrier that will be the one to be suppressed! This sort of effect is well-known from the interference suppression qualities of Frequency Modulation (FM). This aspect may be seen from the curves of fig. 4 where the similarity to a good FM receiver can be clearly appreciated – the only difference here is the absence of a kink in the characteristic (as is the case in FM detection).

The logarithmic demodulator does, however, distort the detected audio and this can be seen from **fig. 5**. This gives a worse impression than it actually sounds when receiving a speech signal. The oscillogram displays an audio tone which has been demodulated from a carrier having a modulation degree of m = 0.5; the resulting distortion is 12.5 %. The distortion rises to 25 % if the modulation degree is increased to m = 1.0 (full modulation). It will be shown later, that a partial compensation for this distortion is possible using quite simple means. The advantages of logarithmic demodulation are so overwhelming that they cannot be ignored. They allow the construction of receivers with reasonable RF selectivity and yet possess the ability to perfectly suppress unwanted carriers.

A further advantage is that the delivered audio is completely independent of the signal strength of the received signal. This may be surmised from figure 4, but **figure 6** shows clearly that this is an inherent property of the logarithmic characteristic. In time t_1 and t_2, the signal strength rises by some 10 dB but the demodulated audio remains constant in amplitude. Such a receiver has no use for automatic volume regulation circuitry! This avoids completely the difficulties associated with the choice of time constants which is encountered in normal AGC concepts. Logarithmic detection cannot, however, offer any sort of protection for selective fading and loss of carrier.

Figure 6 shows also a form of transmitter modulation which is compensated in such a manner that the demodulated output from the receiver is distortion-free. This highlights the requirement for an exponential modulation characteristic. This is not difficult to arrange and would be suitable for use in a communication system employing the principles now under discussion. If the modulation is so arranged that the peak amplitude is twice that of the quieter passages then the minimum amplitude is half that of the quiet amplitude. This ensures that in the presence of an unwanted transmission, the wanted signal is enhanced to its maximum capability in order to maintain the rectifying action.

The listener of such a receiver will notice immediately that in the absence of signal the output noise level is very loud indeed. This is analogous to the reception by FM without the muting facility. When a signal is tuned in, the noise decreases. Strong signals may be completely free of noise, weaker signals have background noise. Fading alters the intensity of the noise but the output volume remains the same. Naturally, a muting circuit can also be devised for this type of receiver.

Griese & Burchard (2) have employed this transmit and receive principle in a commercial broadcast system.

To be continued

Bring your index up-to-date!

The index of the preceding year is always inserted in the middle of the Spring edition. It can be easily removed to update a comprehensive index.

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Dr. Eng. Jochen Jirmann, DB 1 NV

Coaxial Ceramic Resonators – Interesting Components for the Frequency Range 1 to 2.4 Gigahertz

Coaxial ceramic resonators are able to present very high Q circuits in a compact form for oscillators and filters. This role was fulfilled by the demanding nature of the coaxial resonator in the frequency range 900 to 2400 MHz. This opens up the prospect of more circuit possibilities for 23 cm radio equipment and METEOSAT receivers.

Dielectric resonators fabricated from zircon-tin-titanium ceramics have made large inroads into the field of Gigahertz oscillators in recent years. These DROs (Dielectric Resonator Oscillators) are now standard components in microwave technology. The employment of this form of resonator does, however, become cumbersome below 1 GHz where the dimensions (e.g. 60 mm dia, 20 mm height) offer no real advantage over normal coaxial resonators.

1. CERAMIC COAXIAL RESONATORS

For a good year now, coaxial resonators have been available which are able to offer an extremely compact filter at frequencies in the 1 GHz range. Optically, they resemble a large ferrite bead with a 6 mm diameter, a centre hole of 2.5 mm, and length 5 to 15 mm. They are silvered almost completely – only one end remaining uncoated. The component then forms a coaxial conductor which is short-circuited at one end and which is filled with ceramic.

Given that the ceramic has a dielectric constant of 37.8, the characteristic impedance can be calculated to 8.5 Ω and the velocity factor to 6.15. This means that a quarter-wave resonator at 23 cm is only 10 mm long. The Q is in the region of 500, which, for a component of these small dimensions, is a very good figure. Its temperature characteristics at the frequency of resonance are under 3 ppm/K.

In order to obtain a technical appraisal of these devices, the author obtained the following stock items from Siemens (1): – Resonators type B69500 – S6025 – A115 and B69500 – S6025 – A150 for 1150 MHZ and 1500 MHz respectively – these lying closest to the 23 cm band. For an item price of DM 15 for 10 or more resonators, there was nothing to prevent a comprehensive trial of these interesting components.

Before the description of resonators in various circuit configurations is begun, there will now be a few basic considerations linking the resonators with the circuits into which it will be integrated.

As only the "hot" end of the resonator is accessible, the HF energy can only be tapped with capacitive or inductive coupling. Since capacitive coupling involves capacitances of only a fraction of a picofarad, it is mostly the case that inductive coupling using a coil is more convenient.

The resonator tuning can be effected by means of a **high quality** trimmer capacitor connected between the inner and outer conducting surfaces – the tuning rate being ca. – 100 MHz/pF. Should a more detailed explanation of this calculation be required, it is provided in the appendix. The resonator can be pulled upwards in frequency with a hair-pin inductor inserted between inner and outer conductors, but the Q is drastically reduced.

2. A 1 GHz SELECTIVE AMPLIFIER

The test amplifier used a standard form, a BFR 90 circuit in which the stripline resonators were replaced with ceramic resonators.

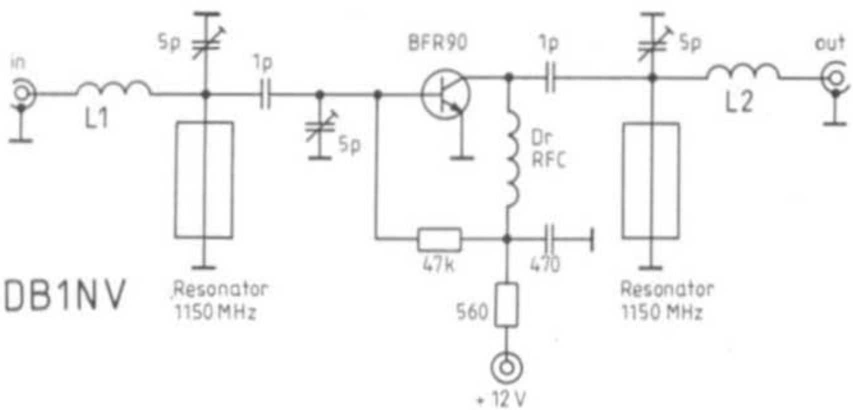
The input signal is coupled in via L1 (**fig. 1**) to the first resonator. This resonator is tuned with a 5 pF teflon capacitor. This solution is not an ideal one as a 3 pF glass-tubular trimmer would yield much finer tuning.

The BFR 90 transistor is fed via an adjustable capacitive divider (1 pF fixed and a teflon 5 pF variable). The collector side also used capacitive (output) coupling whilst the load is again matched inductively to the output resonator. The complimentary use of both inductive and capacitive coupling results in a symmetrical tuning response. The output resonator is also tuned with a 5 pF trimmer.

The following test results were obtained for the circuit working at 1050 MHz using a 1.15 GHz resonator: –

Gain G = 10 dB
Bandwidth B = 25 MHz (between 3 dB points)
Selectivity A_s = 25 dB at ± 50 MHz
40 dB at + 100 MHz
35 dB at – 100 MHz

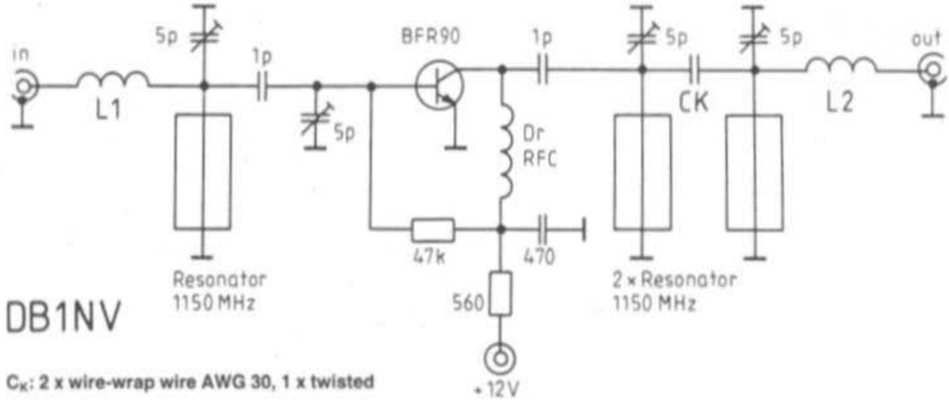
If more selectivity is required, a further resonator can be added to the output circuit. The circuit then becomes that shown in **fig. 2**. The two



L1, L2: 3 turns 0.5 CuL on 3 mm former, self-supporting

RFC: 5 turns 0.5 CuL on 3 mm former, self-supporting

Fig. 1: Selective amplifier using ceramic resonators



C_K: 2 x wire-wrap wire AWG 30, 1 x twisted

Fig. 2: Using a bandpass filter to improve amplifier selectivity

resonators are coupled together using a "fraction of a picofarad" capacitor formed by two conductors in close proximity. As expected, the gain was somewhat reduced but the selectivity much improved. The following test results were obtained: –

Gain G = 8.5 dB
Bandwidth B = 12 MHz (between 3 dB points)
Selectivity A_s = 30 dB at ± 25 MHz
45 dB at ± 50 MHz
65 dB at ± 100 MHz

The photograph of **fig. 3** shows the experimental construction of the amplifier circuit. The cylindrical form of the resonators can be clearly seen as well as the teflon tuning trimmer.

The author also tried to substitute the BFR 90 with a CF 300 but this proved problematical. Perhaps a reader might find a practical solution of how to match this high-impedance GaAs-FET in such a circuit!

3. A 1.3 GHz OSCILLATOR

The possibilities of using ceramic resonators as frequency determining elements was experimented with using the circuit in **fig. 4**. A 1.5 GHz ceramic resonator is tuned by a 3 pF glass-tubular trimmer and coupled via a 1.5 pF capacitor to the collector of a BFR 90 connected in the common-base mode. In the usual manner, the feedback utilises the transistor's internal capacity, the emitter trimmer adjusts the feedback phase. The RF energy is extracted by means of a coupling loop in close proximity to the "hot" end of the resonator. The following data was obtained from this circuit: –

Output power P_o = 1...10 mW according to coupling
Tuning range f = 1200...1380 MHz
Frequency drift Δf = < 1 MHz with supply voltage change of 9 - 15 V
< 100 kHz with an ambient temperature change of 20° - 60° C.

Fig. 3 (right) is a photograph of the constructional details of this experimental oscillator. A narrow-range tunable VCO for a 23 cm synthesizer can be made using a loosely coupled varicap and which possesses good characteristics. The high Q and the low short-term drift ensures low oscillator phase noise.

The author hopes that this article will prompt the generation of a new breed of 23 and 13 cm synthesizer. At the same time perhaps, the first transceiver for these bands will appear.

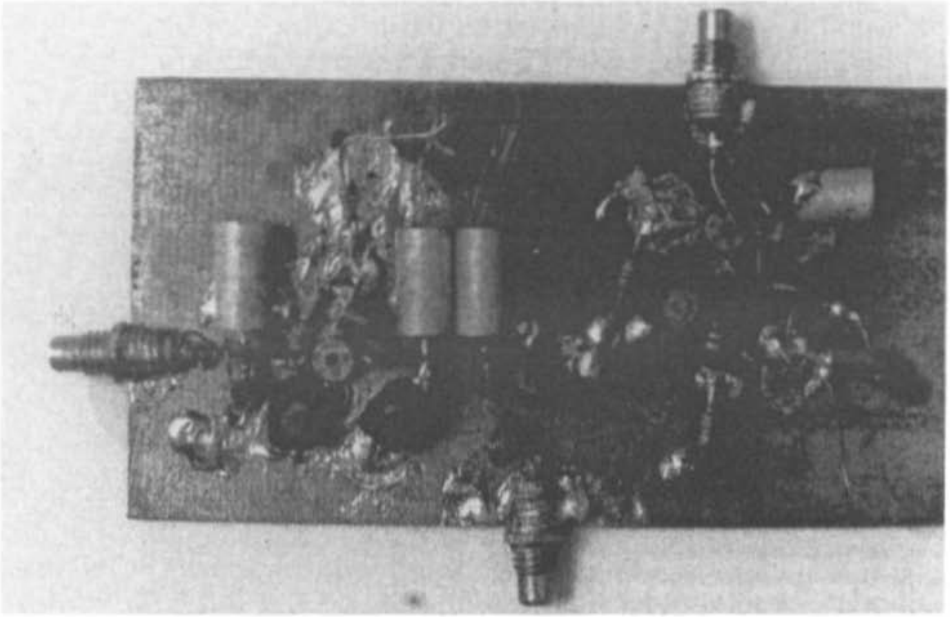


Fig. 3: "Rats-nest" construction of bandpass filter amplifier (left) and oscillator circuit (right)

4. APPENDIX

In order that the quarter-wave resonator may be driven below its "natural" frequency of resonance, an external tuning capacitor is required. To find an optimum value for the capacitor, the following derivation is offered: –

The characteristic impedance of a coaxial line filled with a dielectric of relative permittivity ε_r is given in (2) page 2520.

$$Z_o = \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \frac{D}{d}$$

where D is the external diameter and d the internal diameter.

The velocity factor is $V_K = \frac{1}{\sqrt{\varepsilon_r}}$

When the line is short-circuited at its end, the input impedance according to (2) page 2328 is

$$X_L = j\,Z_o \tan \frac{2\pi l}{\lambda}$$

where l is the mechanical length of the line and λ the wavelength in the material i.e.

$$\lambda = \lambda_o \cdot V_K = \frac{\lambda o}{\sqrt{\varepsilon_r}}$$

The input impedance

$$X_L = j\,\frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \frac{D}{d} \cdot \tan \frac{2\pi l \sqrt{\varepsilon_r}}{\lambda_o}$$

It is well known that at the end of a short-circuited coaxial line having a wavelength of under $\lambda/4$, that the input impedance of a length between $\lambda/4$ and $\lambda/2$ has a capacitive input impedance. In order to use an external capacitor to tune the

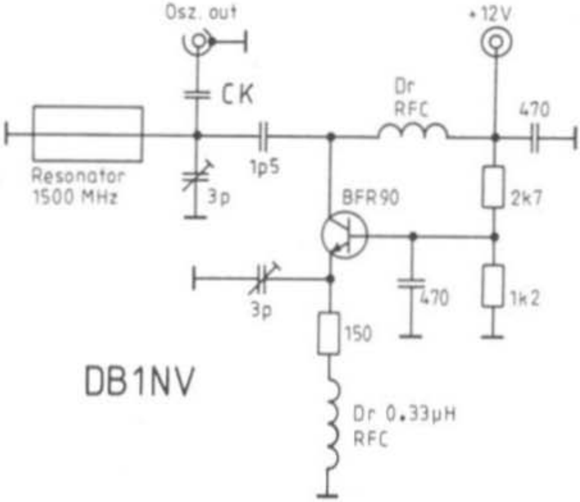


Fig. 4:
Oscillator using ceramic resonators
C_K: see text

line to an exact $\lambda/4$, it must be calculated to equal the make-up impedance but with opposite sign: –

$$X_C = -j\,\frac{1}{2\pi f C}$$

hence it follows

$$C = -j\,\frac{1}{2\pi f X_C}$$

but $X_L = -X_C$ and

$$C = j\,\frac{1}{2\pi f X_L}$$

The required tuning capacity can be calculated as follows: –

$$C = \frac{\lambda_o}{2\pi c}\cdot\frac{\sqrt{\varepsilon_r}}{60}\cdot\frac{1}{\ln\frac{D}{d}}\cdot\cot\frac{2\pi l\sqrt{\varepsilon_r}}{\lambda_o}$$

where C = velocity of light and $f = c/\lambda$.

Substituting the practical data concerning the 1.5 GHz resonator, $\varepsilon_r = 37.8$, D = 6 mm, d = 2.5 mm, l = 8.8 mm, the tuning capacitor for working at 1296 MHz is calculated as follows: –

$$C = \frac{23\ \text{cm}}{2\pi\cdot 3\cdot 10^{10}\ \text{cm/s}}\cdot\frac{\sqrt{37{,}8}}{60}\cdot\frac{1}{\ln\frac{6}{2.5}}$$

$$\cdot\cot\frac{2\pi\cdot 0{,}88\ \text{cm}\sqrt{\varepsilon_r}}{23\ \text{cm}} = 1.34\ \text{pF}$$

Using a similar reckoning, a capacitor of

$$C = 4.35\ \text{pF}$$

would be required to tune the same resonator to 1152 MHz.

5. REFERENCES

(1) Druckschrift der Firma Siemens: Keramikresonatoren für Mikrowellenanwendungen, Lieferprogramm 1987

(2) D. Nührmann: Das große Werkbuch Elektronik, Franzis-Verlag, 5. Auflage 1989

Philipp Prinz, DL 2 AM

41-Element Yagi for the 13 cm Band

The increased activity on the 13 cm band prompted the author to describe a 13 cm antenna which he constructed according to the calculations and principles promulgated by DL6WU (1). A similar antenna for the 23 cm band, but with 44 elements, had already been constructed and with very satisfactory results.

The antenna consists of a boom, a plate reflector and a radiating dipole fitted with a balun and N-socket, 39 directors, an overhead support boom for rigidity and mast-clamp. **Figures 1 and 2** provide a complete overview.

The boom utilises aluminium, 10 mm square cross-section, 1.5 mm tubing. The directors are made from 2 mm welding rods (hard).

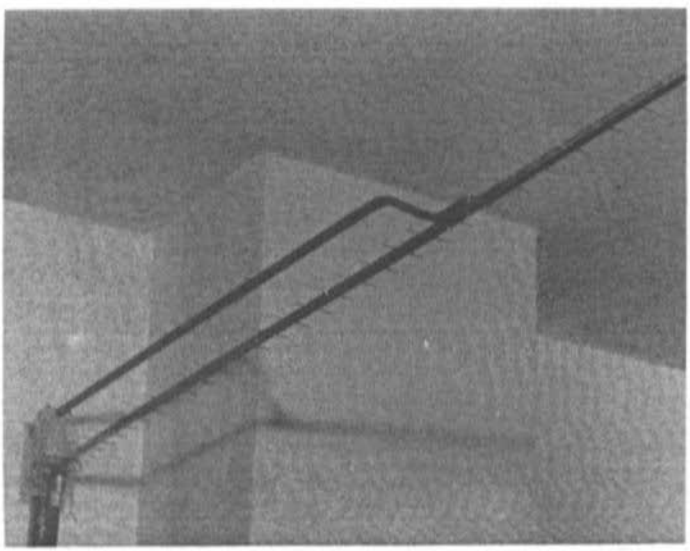


Fig. 1: Don't use this 13 cm-band antenna indoors!

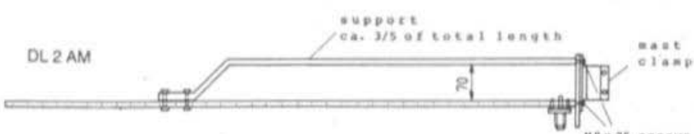


Fig. 2: Side elevation of the 41-element Yagi for the 13 cm band

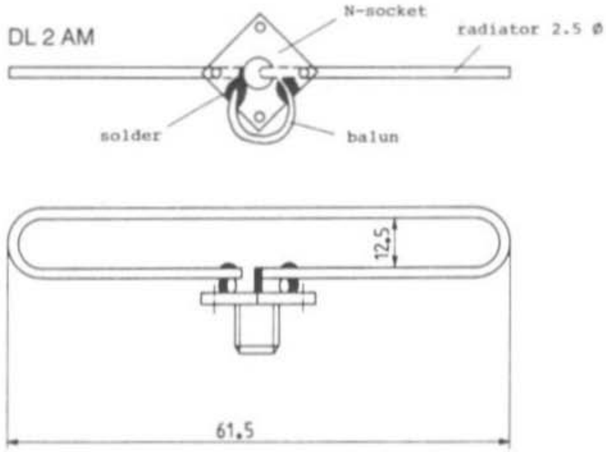


Fig. 3: The radiator/balun assembly

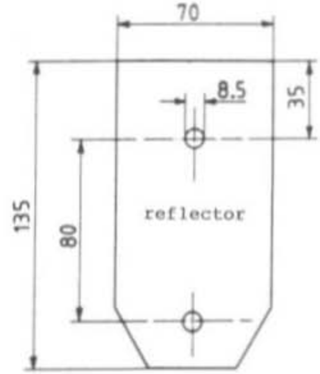


Fig. 4: The reflector

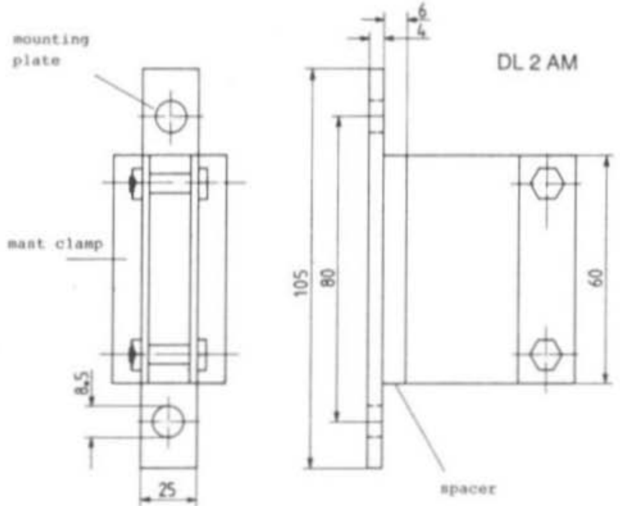


Fig. 5:
Mast clamps with mounting plate

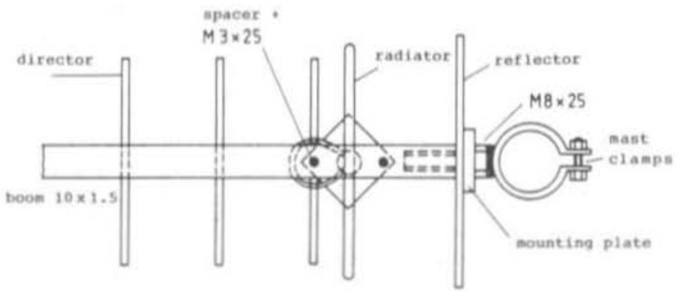


Fig. 6:
Reflector/radiator, a couple of directors and the fixing arrangements

The radiator comprises a bent 2.5 mm copper wire which must be either lacquered or silvered. The balun transformer is also fashioned from 50 Ω cable copper screening mantle – a thin version is preferred with about 3.5 mm dia. The PTFE dielectric has a velocity factor of 0.7 so that the cable mantle is 44 mm long, including the frayed ends in preparation for soldering (**see fig. 3**).

The reflector is made from aluminium gauze plate of about 2 mm thickness with a maximum hole diameter of 6 mm (**see fig. 4**).

The 2 mm steel mast clamp and the welded-on mounting plate (25 x 4) with 6 x 10 x 60 spacer are to be seen in **figures 5 and 6**. They project the entire antenna forward and perpendicular to the plane of the mast.

The directors are cut according to the measurements set down in **table 1** and filed down to the exact length. The ends are then ground to a conical taper.

The holes for the directors are measured and marked as exactly as possible with a steel rule, in accordance with **table 1**. The holes are then drilled, using a drill-press, with a ca. 1.5 mm twist-drill. They are then drilled with a 1.9 mm drill. The holes are then reamered out so that the director rods can pass through with a force fit. The main point to watch is that they are fixed centrally in the boom. The rods may be driven in easier if a smear of grease is used. If any of the holes are inadvertently reamered out too large so that the director fits too loosely, then they must be flanged.

The balun using, semi-rigid cable, must be very carefully bent in order that kinks are avoided. It is measured, marked and cut to size. It is then bent over a 12.5 mm twist-drill shaft. The two ends of the mantle are brought together on the fixing plate of the N-socket and soldered into place. The two inners are soldered to the radiator terminals. (The bending of the radiator is carried out in the same fashion – prior to the soldering)

The completed radiator/balun assembly can be tested in isolation by means of a suitable reflectometer. The VSWR or the return loss should be very good.

Element	Length/mm	Distance	Position/ mm
Reflector	66/66	–	–
Radiator	61,6/12,5	28	28,0
Dir. 1	61,5	10	38,0
Dir. 2	60,7	23,5	61,5
Dir. 3	59,9	28	89,5
Dir. 4	59,2	32,5	122,0
Dir. 5	58,5	36,4	158,4
Dir. 6	57,9	39,2	197,6
Dir. 7	57,4	40,8	238,4
Dir. 8	56,9	42,5	280,9
Dir. 9	56,6	44,8	325,7
Dir. 10	56,2	46,5	372,2
Dir. 11	56,0	48,1	420,3
Dir. 12	55,7	50,4	470,7
Dir. 13	55,3	51,5	522,2
Dir. 14	55,2	51,5	573,7
Dir. 15	54,9	51,5	625,2
Dir. 16	54,7	51,5	676,7
Dir. 17	54,6	51,5	728,2
Dir. 18	54,4	51,5	779,7
Dir. 19	54,2	51,5	831,2
Dir. 20	54,0	51,5	882,7
Dir. 21	53,8	51,5	934,2
Dir. 22	53,6	51,5	985,7
Dir. 23	53,4	51,5	1037,2
Dir. 24	53,3	51,5	1088,7
Dir. 25	53,3	51,5	1140,2
Dir. 26	53,2	51,5	1191,7
Dir. 27	53,0	51,5	1243,2
Dir. 28	52,9	51,5	1294,7
Dir. 29	52,8	51,5	1346,7
Dir. 30	52,6	51,5	1397,7
Dir. 31	52,5	51,5	1449,2
Dir. 32	52,4	51,5	1500,7
Dir. 33	52,3	51,5	1552,2
Dir. 34	52,1	51,5	1603,7
Dir. 35	52,0	51,5	1655,2
Dir. 36	52,0	51,5	1706,7
Dir. 37	51,8	51,5	1758,2
Dir. 38	51,8	51,5	1809,7
Dir. 39	51,7	51,5	1861,2

Table 1:
All the element spacing dimensions of DL2AM's 13 cm-band Yagi

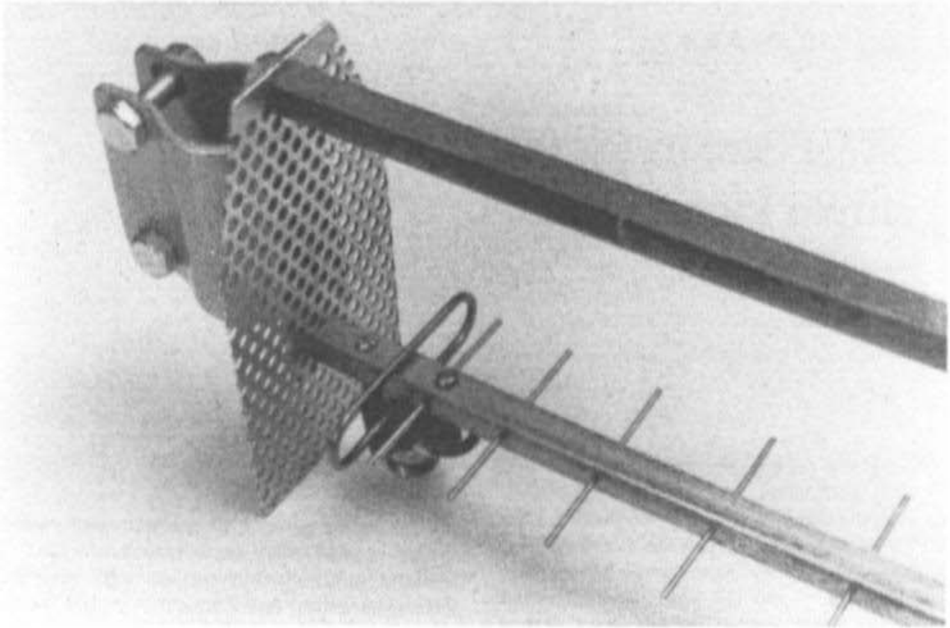


Fig. 7: The most important part of the antenna

The N-socket is secured to the boom with two M3 x 25 which also pass through 10 mm spacers.

The mast clamp shown in **fig. 5** is made from 2 mm tin-plate and is bent according to the mast dimensions. It was found that a length of tin-plate of 60 mm was suitable for a mast of 30 mm diameter. It is welded on to a mounting plate having a length of 105 mm and provided with two 8.5 mm dia. holes. A spacer must be inserted between the clamp and the back-plate in order that the M8 x 25 mm fixing screw heads do not foul the mast thus preventing it from being fed through the clamp. It is, of course, possible to use a discarded television mast clamp.

The end of the boom, and also that of its support, is drilled and then tapped with a M8 thread to a depth of 25 mm. The reflector, complete with mounting plate and clamp, is then screwed on to the boom and its support. **Figure 7** shows the arrangement.

Altogether, the assembly is really easy to complete. It is finished with a sprayed-on coat of plastic paint.

The VSWR proved to be so good that the validity of the measurement and the functioning of the Bird 43 thro-line meter were both needlessly brought into question.

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Joop Kuijntjes, PA 2 JOK

Control Circuit for the METEOSAT Multiple Picture Store

Following the construction of the METEOSAT receiving installation with the multiple picture store (1), the problem still remaining was how to get the desired pictures into the numbers 5 to 10 memories. As the pictures are transmitted according to a definite plan (dissemination schedule), some sort of timing circuit seems to be required. The first thoughts centred around the timer described in (2). This is driven from the 2400 Hz oscillator signal of the FM/AM converter (3). Since this timer must be continually reset when another picture is about to be stored, another solution had to be found. Another consideration had to be, that only the correct picture had to find its way into the store in order that a weather sequence film is not ruined. This could occur when either the desired programmed picture, for some reason or other, does not arrive, or when it does arrive it is not followed by the next in sequence.

The following circuits to be described were arrived at after many experiments and modifications and the prototype was realised on a perforated stripboard. It would probably have looked completely different if it had been developed from a circuit idea and set straight away upon a printed circuit board. For this reason the circuit should be regarded, above all, as a basis upon which to develop a final one. Therefore, even parts of it can be used with confidence.

They offer the following possibilities:

- Indication of the memory in use by means of a 7-segment display.

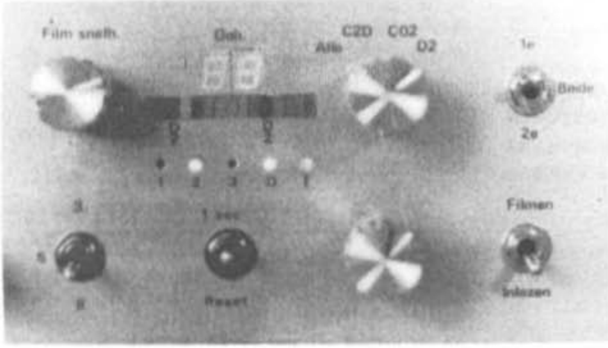

Fig. 1: Control and indicator elements of the author's equipment

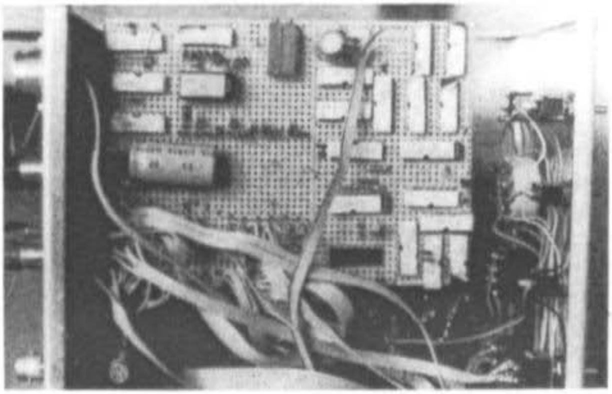

Fig. 2: Perforated strip board and wiring of the control and indicating elements

* Automatic switching between infra-red, water-vapour or visual pictures, as these picture types have differing contrasts.
* The possibility of a choice for reading-in the films from four picture types, namely: D2, or C03, or D2/C2D, or from all pictures. Also the possibility of false pictures appearing in the film is inhibited. When using D2/C2D, the picture is stored in D2 (infra-red) at night and in the day-time, C2D (visible) pictures are stored.

The new dissemination schedule, effective 19th June 1989, could also be taken partially into account: accordingly, fig. 9 has been modified together with the table. The C2D pictures are not sent often enough (at the present time) and therefore this part of the circuit is superfluous.

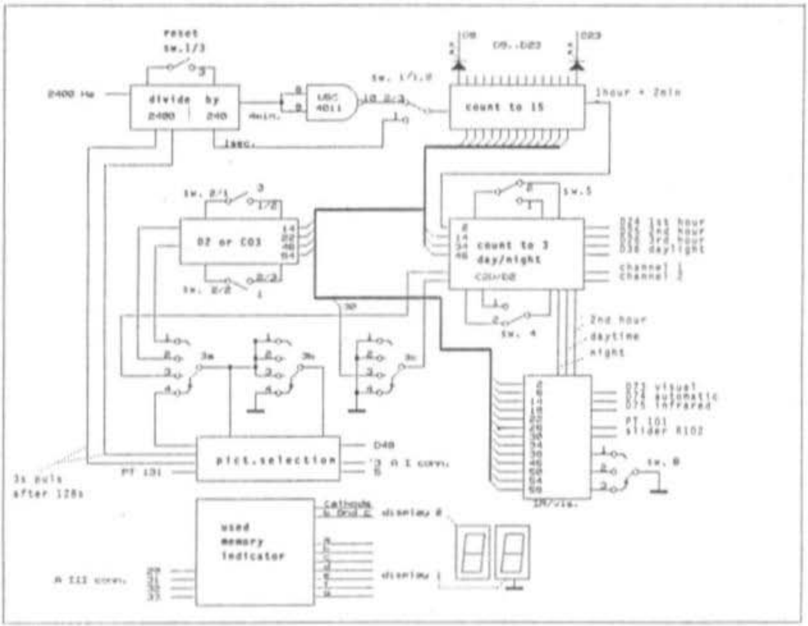

Fig. 3: Block diagram

Switch	Position	Function	
1	1	Sync. period counter with second pulse	
1	2	Normal position (4-min. pulse after period counter)	
1	3	Resets the 4-min. counter	
2	1	First picture	This switch is only
2	2	Both pictures	operative when switch 3 is in
2	3	Second picture	pos. 1 or pos. 2.
3	1	D2 pictures	
3	2	C03 pictures (C02 version shown in photo)	
3	3	D2 und C2D pictures	
3	4	All pictures	
4	1	METEOSAT receive chan. 2	
4	2	Auto-switching chan. 1/chan. 2	
5	1	Summer pictures	
5	2	Winter pictures	
6	1	Pictures in the visible range (VIS) and water-vapour (WV)	
6	2	Auto-switching VIS/IR	
6	3	Pictures in the infra-red range (IR)	

Switches 1, 2 and 6 are toggle switches with an active centre position

Table 1

1. REVIEW

The photographs (**figs. 1 and 2**) of the unit were taken and processed by the author. They show that only a handful of integrated circuits was employed for this project together with a perforated strip board, a container, an indicator and a couple of switches. As the circuit schematic was broken down into six parts for convenience, the block diagram of fig. 3 serves to show how they are pieced together. The switches shown in **fig. 3** have the functions indicated in **table 1** above.

2. CIRCUIT DETAILS

METEOSAT 1 Circuit (Fig. 4)

This part of the circuit consists of the dividers 240 and 2400 together with the digital read-out of the picture memory. The dividers use two CD 4040s and they deliver the second pulses and the 4-minute impulses. The 4-minute counter is reset from switch 1/3. This switch is combined with the one which sets the 15 x 4.minute counter (see METEOSAT 2). The outputs and division factors from the CD 4040 used here are:

Q 5 = 16
Q 6 = 32
Q 7 = 64
Q 8 = 128
Q 9 = 256
Q12 = 2048

The digital read-out consists of a full adder, CD 4560 and a display driver CD 4543. This is actually an LCD driver but it also functions as an LED driver.

The CD 4560 receives, at its A inputs, the BCD code of the pin connector A III, pins 31, 32 and 33. Pin 29 goes high when the second memory card is being read-out. This signal is also present at inputs B1 and B3 where 5 is added to the BCD code of the A inputs. The display 2 (fig. 3) is

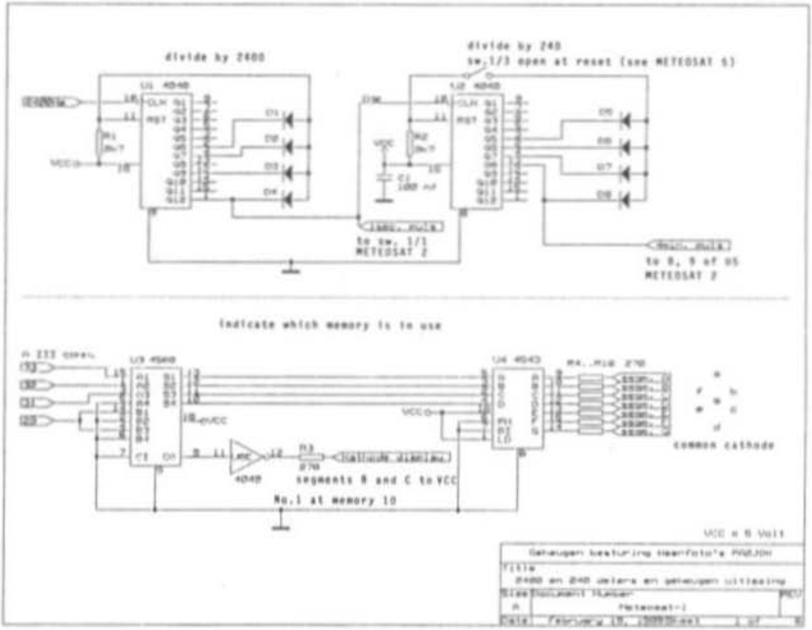

Fig. 4: METEOSAT 1: Divide by 240 and 2400

controlled via the carry-out of the CD 4560 of which only the segments B and C are used. The value of R3 is dependent upon the type of display in use. It serves to control the brightness of display 1. As a low signal is able to deliver more current, an inverter (U8E) is used in order to relieve the burden on the CD 4560.

METEOSAT 2 Circuit (Fig. 5)

The periods four minutes to one hour are counted here. The first segment begins about 2 mins after the full hour. The beginning of a 4-minute period lies between the stop and the start tone from METEOSAT – it has to be set by switch 1/3 (fig. 3). Th counter itself is set via switch 1/1 and 1/2: if switch 1/2 is closed, the counter receives a pulse every 4 mins – this is the normal condition. If, however, switch 1/1 is closed, a 1st train of pulses is received. Switch S1 must be a type having three positions.

The LEDs D11 and D19 have a different colour to the rest of them in order to signal the reception of D2 pictures. Again, inverters are used in order to relieve the load on the outputs of the CD 4017 which also must drive other circuits. As D18 is unable to be driven, to the same brilliance as the other LEDs, by the current supplied by U5D on its own, U8D and U8C are added in parallel. Alternatively, D18 could be taken via its own resistance to the plus-rail instead of using R11.

METEOSAT 3 (Fig. 6)

This part of the circuit carries out the selection between the first picture and/or the second picture. When switch 2/1 is closed, the first picture (D2 or C03) is stored. Switch 2/2 does the same for the second picture. The switch is identical to the one in the METEOSAT 2 part of the circuit. As fig. 6 shows, both contacts can

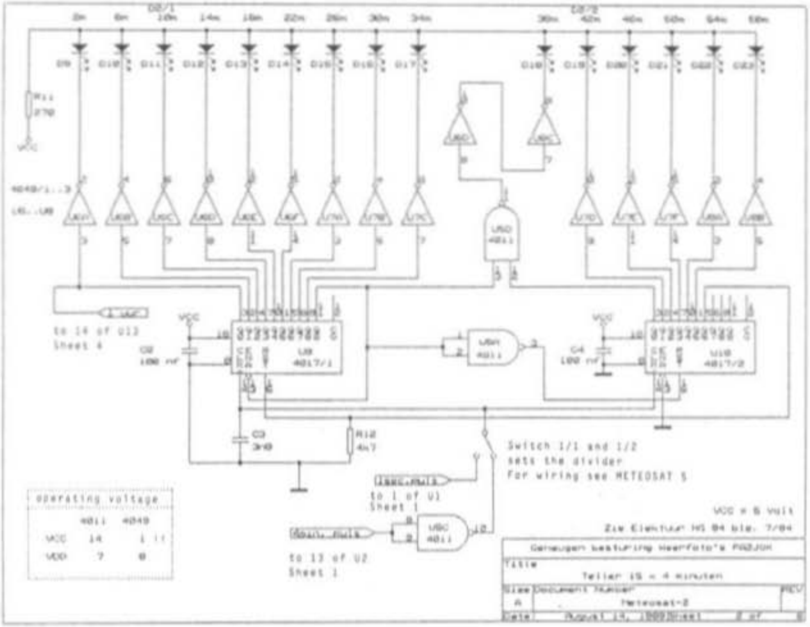

Fig. 5: METEOSAT 2: Divide by 15 x 4 minutes

also be closed, this results in both the first **and** the second picture being stored simultaneously.

METEOSAT 4 (Fig. 7)

This is the hour counter for the selection of the C2D picture. This visible picture, in the format of the D2 picture, is transmitted in METEOSAT channel 1, only every 3 hours (in summer altogether 4 times daily). It appears to be irregularly disposed in the dissemination schedule in channel 2 but it does occur more often. It can then be seen that between changing channels, a C2D picture is received every 1 ½ hours. By these means, it is possible to read-in a film comprising D2 pictures by night and C2D pictures by day.

At the output 3 of the CD 4017/1 appears an hourly pulse which drives the hour counter. The LEDs D24...D26 indicate the hours. D24 indicates the GMT hours 0, 3, 6, 9, 12, 15, 18 and 21 – it is easy to see that they are all divisible by 3.

Whilst between hours 0, 3, 18 and 21 the D2 pictures are read-in, in fact, every 1 ½ hours, whereby in the first hour, the transmission of the 42nd minute, and in the third hour, the transmission of the 10th minute is taken. In all three dissemination slots, a picture is read-in, either a D2 or a C2D.

During hours 6, 9, 12 and 15, C2D pictures are read-in: At the 6th hour, it is possible to read-in a D2 picture via switch 5 (summer/winter).

During the first hour, all pictures are overwritten until the C2D from minute 58 has been stored. Following the stop tone of this picture, LED D24 extinguishes and D25 illuminates. At the same time, an H signal is received at both inputs of U15D, one from an hour counter, the other from the 2 min output of the CD 4017/1. This H signal

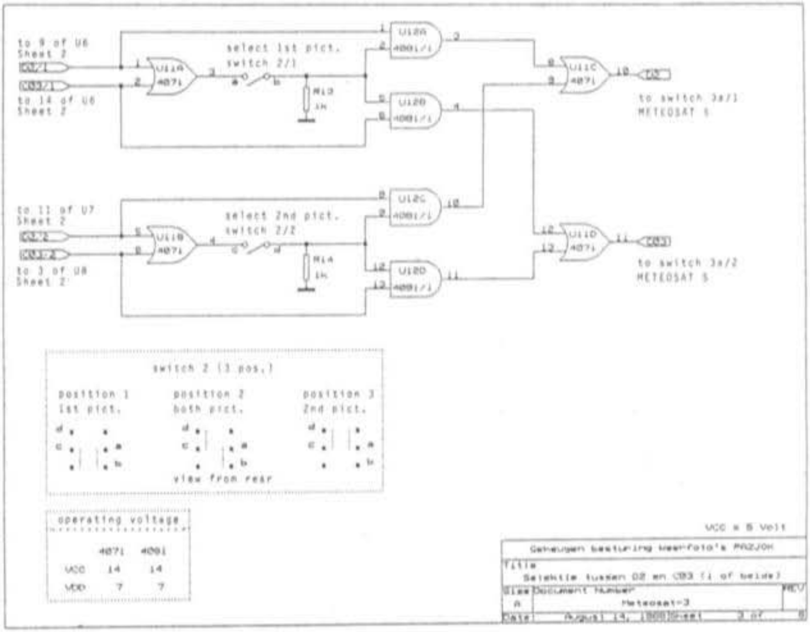

Fig. 6: METEOSAT 3: D2 and C03 selection

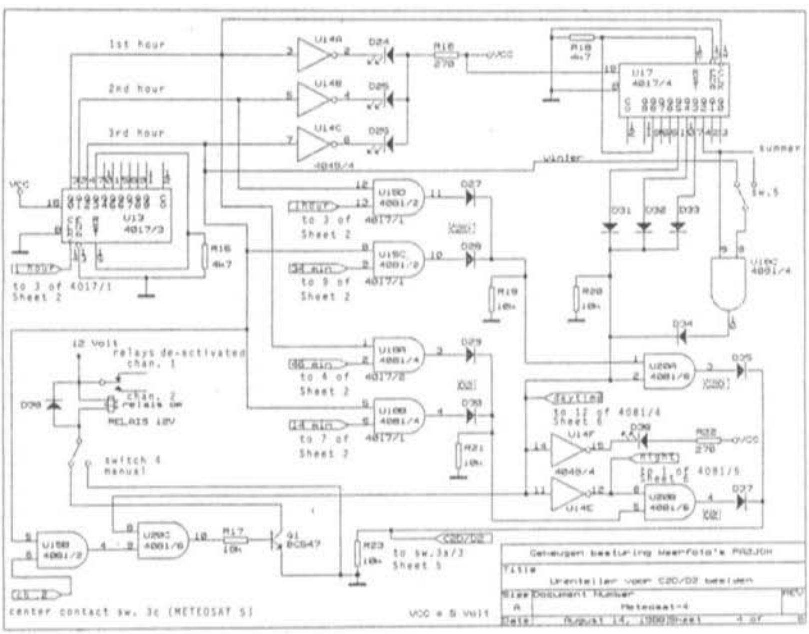

Fig. 7: METEOSAT 4: Hour divider for C2D/D2 pictures

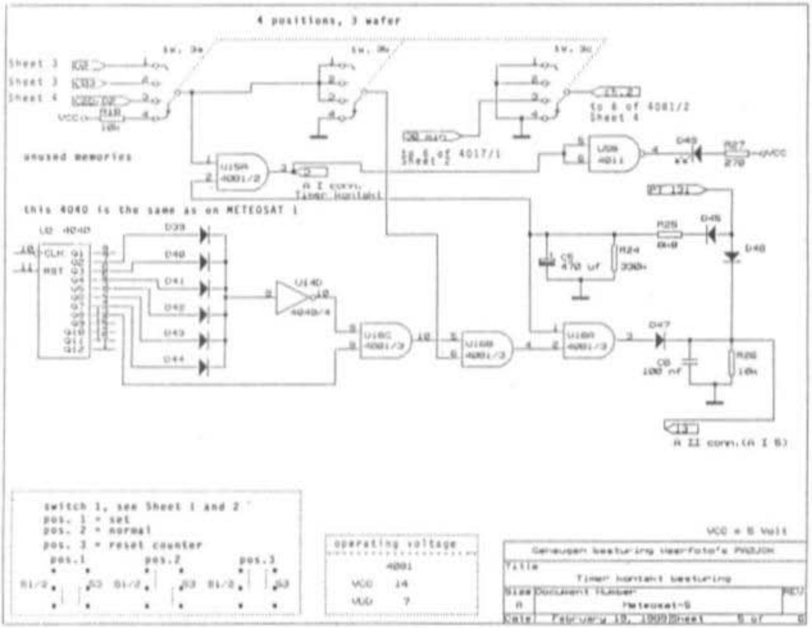

Fig. 8: METEOSAT 5: Timer-contact control

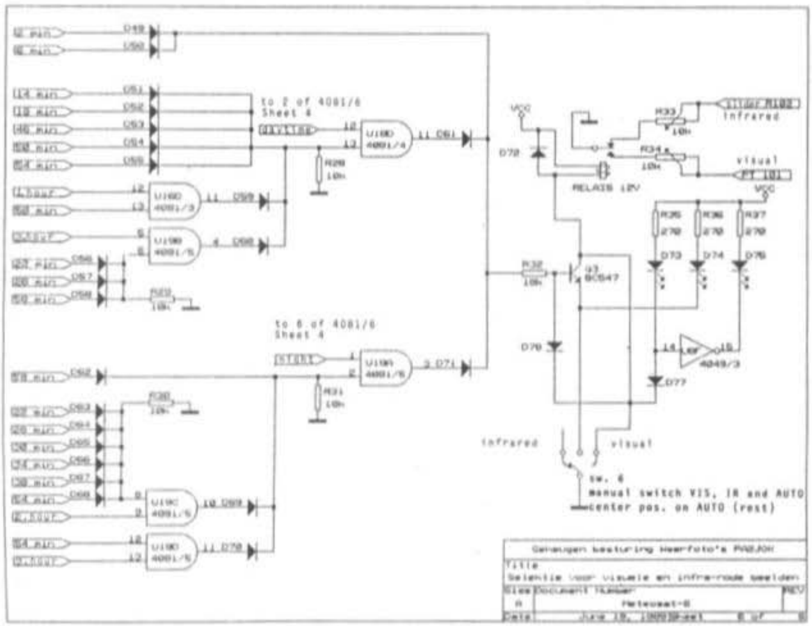

Fig. 9: METEOSAT 6: VIS and IR-picture selection

reads-in the following picture into the next memory (see METEOSAT 5).

LED D26 is illuminated during the third hour and when 30 minutes have elapsed, LED D16 goes on (METEOSAT 2) and the C2D transmission on channel 2 arrives. The signal is taken via switch 3C to one of the inputs of U15B. At the latter, also stands an H signal for the third hour. During daylight hours only, 20C and Q1 are changed over to channel 2. This can also be accomplished manually with switch 4.

After 4 minutes, the relay drops out and an H signal is taken via gate U15C to the input of C2D. This directs the following picture into the next memory. The next hour pulse from the CD 4017/3 input 14 starts the whole sequence all over again from the beginning.

Gates U20A and U20B take care of the D2 and C2D (day/night) transmissions. The combined C2D/D2 output is taken to the switch 3 a/3 (METEOSAT 5).

METEOSAT 5 (Fig. 8)

Diodes D39...D44, the inverter U14D and the AND gate U16C, all form a circuit to provide an H signal at output 10 after a period of 128 seconds. Capacitor C5 is charged by an H signal from Pt 131, according to VHF COMMUNICATIONS, this point can be found at the connection A III, pin 13 – in this circuit, it follows a diode. At pin 13 of this connection, the start signal appears. The latter is then read into the current memory. Capacitor C5 is discharged via R24 in a greater time than the 128 seconds that it took to charge (adjust R24). Pin 3 of the connection A1 is the timer contact which, when high, switches in the next memory.

Switch 3 selects either D2, C03, C2D/D2 or all pictures in a film sequence. In the film mode, after a picture has been read-in, an H signal should appear at the timer contact.

In switch positions 1 and 2, the H signal at the D2 and C03 inputs is offered to the gates U15A and U16B. When this picture has elapsed and C5 is still charged, a H signal is present at U15A, input 2 and the timer contact goes high. If the desired picture has not been received, it could be that C5 is in the discharged state and the timer contact does not go high, thus the next memory cannot be employed. If, after a programmed picture, no further picture arrives, then U16C with its output 10 going high after 128 seconds, initiates an artificial start signal via U16B and U16A.

In switch position 3, the same sequence occurs, then the 30 minute signal is fed to input 6 at AND gate U15B (see description METEOSAT 4).

In switch position 4, all pictures to be received are fed to the memories. Because switch 3b is

	Day			Night		
Minutes	1st	2st	3st	1st	2st	3st
2	x	x	x	x	x	x
6	x	x	x	x	x	x
10						
14	x	x	x			
18	x	x	x			
22			x		x	
26			x		x	
30					x	
34					x	
38					x	
42						
46	x	x	x			
50	x	x	x			
54	x	x	x		x	x
58	x		x	x	x	x

Night		Day				Night	
1st hour = 0	3	6	9	12	15	18	21
2nd hour = 1	4	7	10	13	16	19	22
3rd hour = 2	5	8	11	14	17	20	23

Hours in GMT

Table 2

at ground potential, the circuit for the artificial start signal is now not able to function.

METEOSAT 6 (Fig. 9)

This part of the circuit provides the automatic switch-over between VIS- and IR pictures. The diodes here function as OR gates. With switch 6, IR, VIS or AUTOM can be selected. This is displayed by LEDs D73...D75.

With the two 10 kΩ preset potentiometers, the contrast for the two types of pictures can be adjusted. The diode D77 is required in order to prevent that the supply voltage reaches the input of U8F when a 12 V relay is in use.

Table 2 shows at what times the relay switches to VIS or WV pictures.

My thanks are extended to Chris, PA Ø CVG, for his help in circuit design.

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Martin Althaus, DF 9 DA

Compact METEOSAT Converter

This latest design is characterized by its use of modern, cost-effective components and a well-proven stripline printed circuit board. Only one crystal-controlled frequency multiplier was employed, which of course means two IFs for both METEOSAT channels. These measures enable a very compact converter to be realised in a tin-plate housing of only 111 mm x 74 mm x 30 mm. BNC-sockets are used for both input and output. It is scarcely the type of converter that one associates with earlier generations (1), (2), and will certainly inspire the same form of construction (robust and compact) for other applications such as portable and mobile operation.

The newly developed receiver matches the converter and possesses both required receiving channels of 134.0 and 137.5 MHz, as well as a third channel of 137.62 MHz which enables both APT frequencies of the NOAA weather satellites to be received directly. It has the same compact dimensions and will be described in the next edition.

For experienced UHF-amateurs using the minimum of test-equipment (frequency counter, multimeter) the construction will present no problems. The converter can – if the drilled PCB, the punched housing and suitable components are purchased – be completed in two to three hours. Less experienced amateurs should have, at least test equipment available and also take the time to study the following detailed construction and alignment details and to carry them out step by step. The article was written for this type of constructor.

1. TECHNICAL DATA

* Input frequencies: 1691.0 and 1694.5 MHz
* Intermediate frequencies: 137.5 and 134.0 MHz
* F = 1.7 dB (typical); G = 26 dB (typical)
* Supply: 12 to 14 VDC, circa 80 mA
* Supply lead-in via the IF cable or feed-through capacitors
* Provision for an external pre-amplifier
* RF-connectors: BNC-sockets; 50 Ω

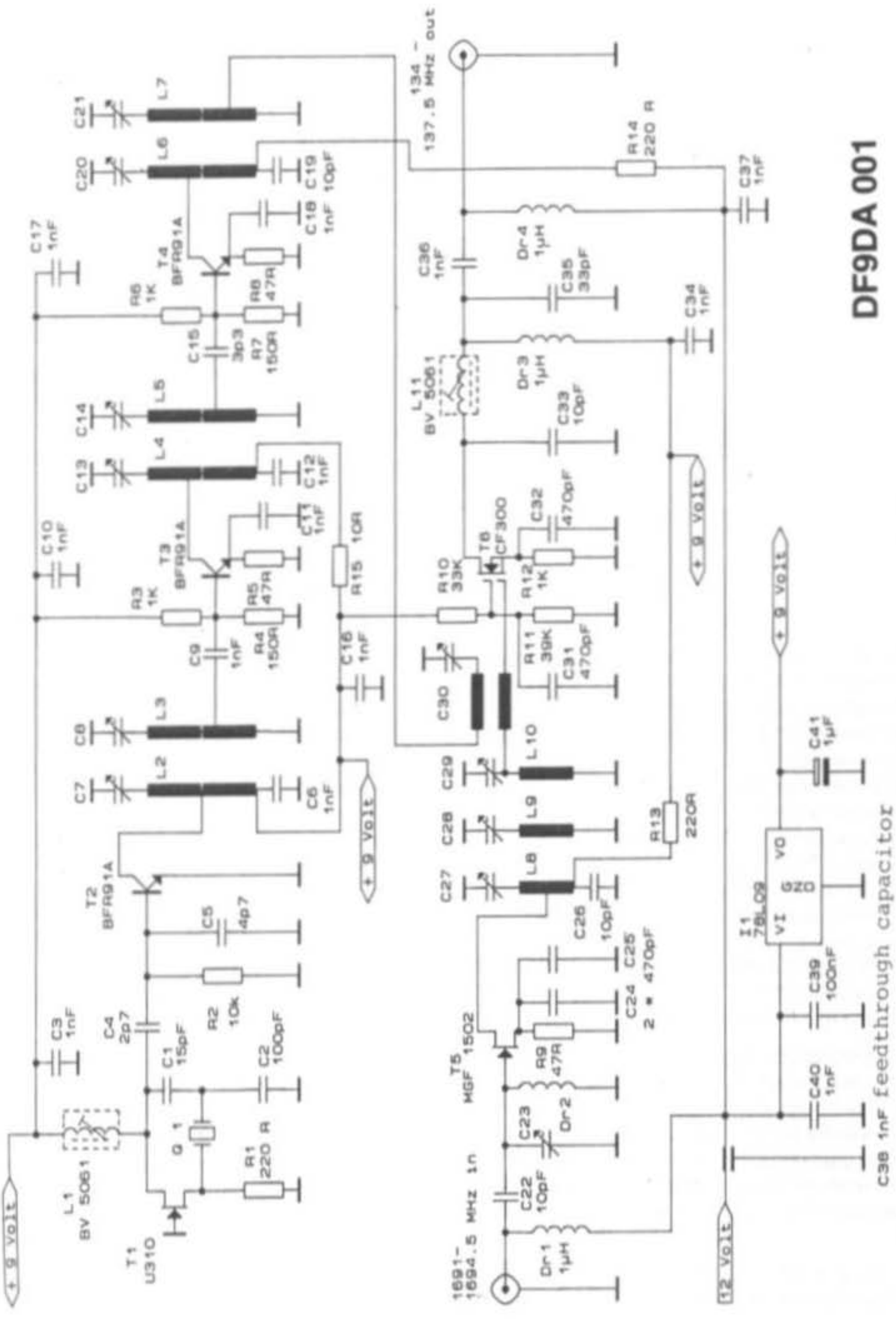


Fig. 1: The complete circuit schematic of the DF9DA 001 converter

2. CIRCUIT DETAILS

Figure 1 shows the complete circuit schematic of the converter with the designation DF9DA 001. The frequency processing derived from a crystal oscillator is to be seen at the top of the diagram together with the x16-frequency multiplier. The middle left of the diagram carries the input amplifier and mixer and below it the small 9 V stabilizer.

The input stage of the converter (T 1) is fitted with a GaAs-FET MITSUBISHI type MGF 1502 but any of the MGF-series types can be used without qualification. The following 3-stage filter with etched striplines L 8, L 9, and L 10 suppresses the image-frequency at approx. 1420 MHz by more than 20 dB. The cost-effective GaAs-FET tetrode CF 300 was able to be used in the mixer owing to the low-noise, high-gain, first stage. The injection frequency is taken via a stripline from the last multiplier circuit L 7 and coupled into gate 1 of the mixer. This circuit results in a mixer gain of 11 dB.

The IF filter in the drain circuit with C 33, L 13, C 35 is arranged as a low-pass filter in order to suppress any coupled-in SHF energy. Choke Dr 4 lies in the DC-supply path when the supply is routed via the IF-output cable. Similarly, Dr 1 has been provided in the supply to an optional external pre-amplifier via the connecting cable. The pre-amplifier is only required if a cable of more than 1 dB loss is used between the METEOSAT antenna and the converter.

The injection frequency is derived from a 97.3125 MHz crystal with a JFET (T 1) – a U 310 – in a well-known low-noise circuit. The tuned circuit with L 1, C 1, C 2 is tuned to the crystal frequency. The stage using T 2 multiplies the frequency x 4 and T 3, T 4 each double the frequency. Between the stages are 2-circuit bandfilters with tunable

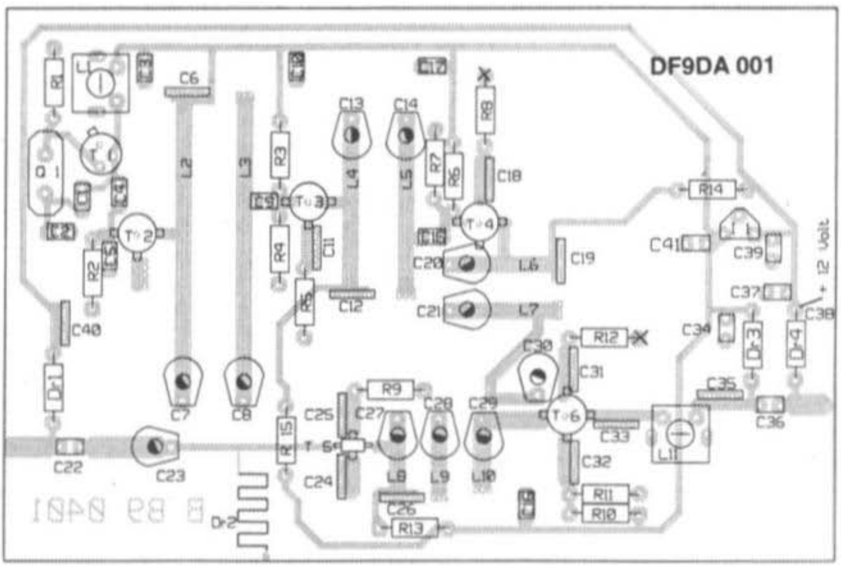


Fig. 2: Component layout plan for the through-contacted glass-fibre PCB DF9DA 001

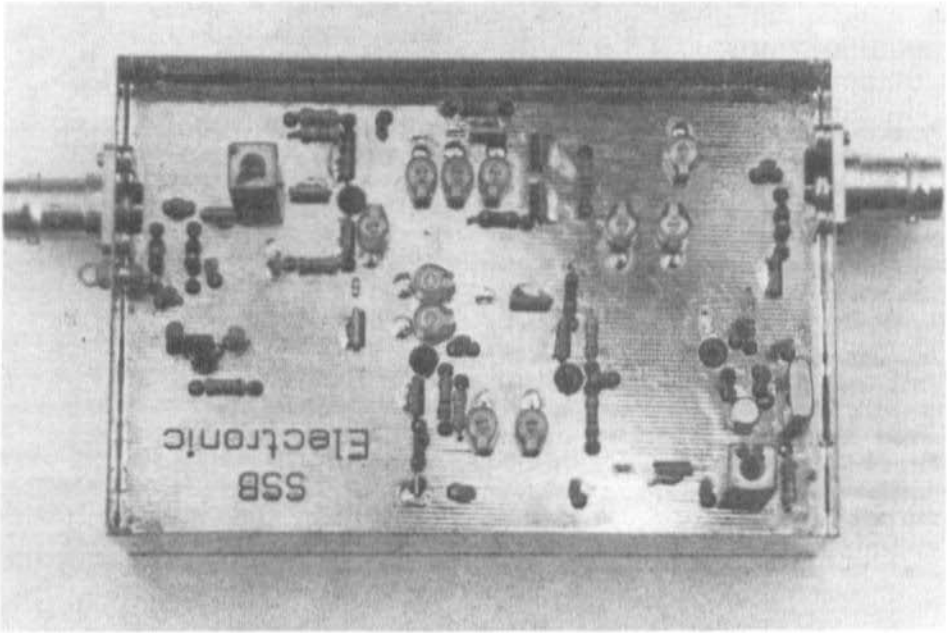


Fig. 3: The METEOSAT converter seen from the component side

micro-striplines inserted so that at L 7's tap a clean signal at 16 times the crystal frequency is available.

3. CONSTRUCTION

The converter is able – as already mentioned – to be completed by an experienced constructor in two to three hours, providing that all components are to hand. With a correct construction by following the detailed instructions, the converter will function immediately following the alignment. If a 500 MHz frequency counter, a multimeter, a spectrum analyzer, a microwave signal generator and a noise-factor measuring set are available, the alignment can be carried out in a expeditious and accurate manner. The converter can, however, be aligned with the aid of the METEOSAT signal coming direct from the antenna

The component layout plan for the 111 mm x 72 mm through-contacted glass-fibre PCB can be seen in **fig. 2**. The step-by-step procedure is now described exactly.

3.1.

The housing components are placed in a cover and soldered together. The BNC-connector sockets must have the excess PTFE-isolation removed so that the centre pins lie directly and fully on the appropriate PCB-track – when they are eventually assembled. Now, the BNC-sockets are screwed on to the box – with the flange on the outside.

3.2.

Place the board into the tin-plate frame and align it horizontally within the housing. The correct mounting of the board is ensured by the recesses in the housing. The board is provisionally spot-soldered at two points to the housing. After ensuring once more that the board is horizontal, it is soldered all the way around its perimeter to the frame. On the B side of the board, the etched choke DR 2 is soldered to the housing. **Figure 3** is a little advanced as the board is shown as being already loaded with components.

3.3.
Solder in all the chip capacitors after checking carefully their values. Half of the chip capacitors are placed into the appropriate slots on the board and soldered on both sides. The slots should be checked on the B-side to ensure that there are no short-circuits to ground – remove any burrs with a scalpel. The soldering iron temperature should not exceed 300° C otherwise the chip capacitors could disintegrate.

3.4.
Now the decoupling capacitors should be soldered in. They are marked in the layout plan without a value given. Afterwards, the rest of the capacitors are soldered in according to the layout plan. Care should be taken to select the correct value and also that tantalum capacitors are inserted in accordance with their polarity.

3.5.
The flat connectors of the trimmer capacitors are then bent at 90°, shortened, located and then soldered to the ground plane. The trimmers C7 and C8 are SKY trimmers (black), all others being green.

3.6.
After inserting the resistors, the crystal, the oscillator transistor and both NEOSID-filters are to be placed on the component side. The connection sequence of these components can be seen in **fig. 4**. The unconnected third leg of the filter is carefully withdrawn from the coil former. The U 310-oscillator transistor's housing tab is soldered to the ground plane. The resistors R8 and R12, as opposed to the other resistors, are not connected with both connections going through PCB but the ground connection is taken directly to the component side ground. The correct side is marked in the component layout plan with a cross.

3.7.
Finally, the multiplier transistors and the two GaAs-FETs are inserted (**fig. 5**). The transistors T2, T3 and T4 must have their base and collector leads shortened to a length of 2 mm. The transistors are inserted into their respective holes and soldered to the PCB conductor tracks. Note: do not mix-up the base and collector connector leads! When handling the GaAs-FETs, care must be exercised to minimise static damage by using a grounded soldering-iron tip and also grounding the housing. Discharge the constructor's body before handling these devices by touching a grounded object (the housing). The drain leads of transistors T5 and T6, each have a ferrite bead slipped on to them before the FETs are soldered in.

As the converter was intended originally to be supplied remotely, the connecting leads of the feed-through capacitor C38 must be carefully bent downwards and soldered to choke Dr 1 on the component side of the board.

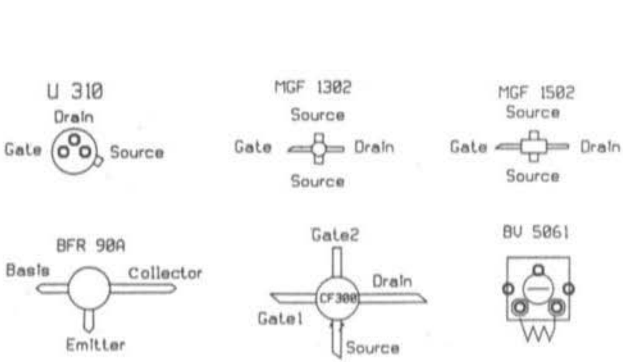
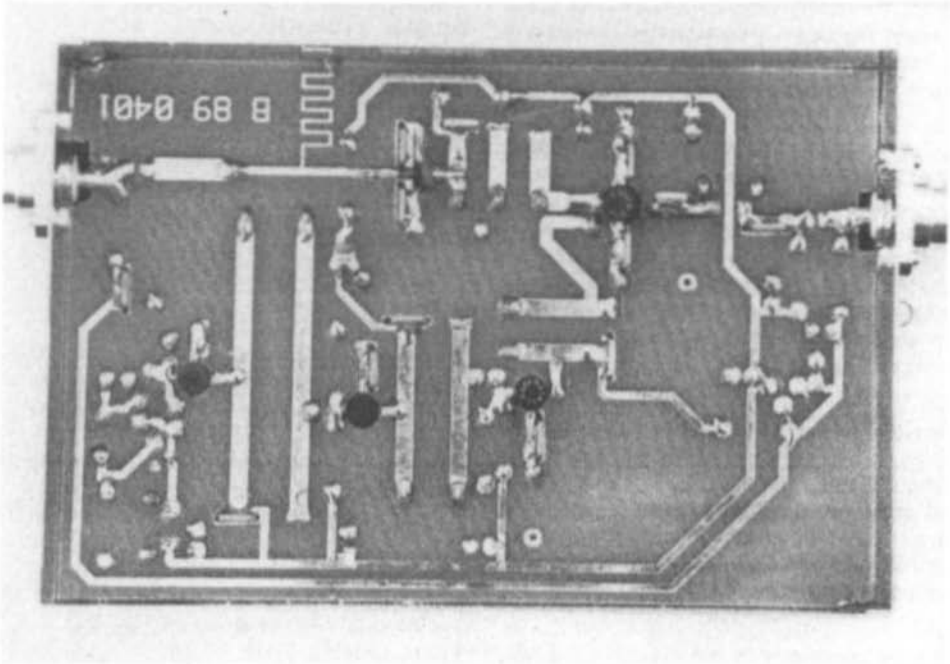


Fig. 4:
Components seen from above



4. ALIGNMENT

All trimmers must be installed according to the component layout plan; the final positions of the rotors lie in the vicinity of the position indicated. If, when tuning the multiplier, large divergencies occur, it can be assumed that an incorrect alignment has taken place.

4.1.
Apply the supply potentials to the feed-through capacitors and check the current consumption of the module – it must lie between 50 and 60 mA. The voltage-regulator output should measure 9 V. The drain potential of the MGF 1502 should be 5 V and the gate-2 potential of the CF 300 is ca. 4.5 V. Small deviations from these voltages can be ignored but large differences could indicate either a defective component or one wrongly installed.

4.2.
Measure the voltage at the source of T1 and turn the coil core to its extreme out position, and then slowly in again. When the voltage drops by about 0.15 V, the oscillator is oscillating. The oscillatory condition can, of course, be checked by using a frequency counter. It should be adjusted to a nominal frequency of 97.3125 MHz.

4.3.
The alignment of the following tuned circuits must be carried out intuitively. First check the emitter voltage of T3 – it measures ca. 0.4 V when the oscillator is not working and rises to 0.7 V when the oscillator is working and assuming an optimal alignment of the filter. During the alignment, it must be observed that with C7 and C8 a clearly-defined maximum of the emitter voltage should be obtained. If this is not the case, it is likely that the circuit has been tuned to the third harmonic instead of the fourth.

4.4.
Measure the emitter voltage of T4, and tune the bandpass filter with C13 and C14 for a maximum – about 0.7 V.

4.5.
Using a sufficiently strong METEOSAT signal, or that from a signal generator, the rest of the alignment is carried out. Assuming that the trimmers C20 to C30 have been turned to the start positions, shown in the component layout plan, it should be possible to hear a signal in a connected receiver. If not, C30 should be carefully mis-tuned until it is. In order to obtain the maximum gain from the converter, the alignment is commenced from C20 and C21. Then, the 3-stage filter is aligned by tuning C27, C28 and C29. The adjustment of C28 is very critical and must be carried out very carefully. Finally, with the tuning of the mixer with C23 and L13, the alignment of the converter is completed.

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Dr. Robert Dorner, DD 5 IK

4-Channel 140 MHz Oscilloscope Part 1: Salient Circuit Details

The oscilloscope to be described was developed by the author over a period of several years. It completely fulfils the demands which can be imposed upon a modern, wideband oscilloscope in the upper price bracket. Its construction may be readily undertaken by the experienced amateur constructor or by the professional with the benefit of a small outlay in terms of cost. The design has taken into account the possibilities of further extensions and options which may be required and was therefore constructed using the plug-in module concept. For the mechanical construction and the circuitry only readily obtainable standard parts have been specified.

1. FUNCTIONS REVIEW

1.1. Short-Form Data

General

- CRT: 8 x 10 cm, internal grid illumination, 16.5 kV
- Overlapping ranges for vertical (amplitude) and time base
- Adjustments for brilliance, focus, astigmatism and trace rotation
- Trace-find button and band-limit (20 MHz) switch
- Calibration voltage output 0.2 V_{pp}, 1 kHz
- AUTO-range function (hands-off operation)
- Outputs for trigger and time base 1
- Universal socket with provision for switched external X/Y/Z presentation
- Power consumption: 50 W
- Dimensions (without handles): 210 x 300 x 400 mm
- Weight: approx. 9.5 kg

Vertical Section

- Bandwidth 140 MHz, 4 chan. (max), chop/alt., differential inputs
- DC range: approx. 100 x scale div.
- Sensitivity: 10 mV to 5 V/div. in 1-2-5 steps plus variable
- Input: AC-DC-50 Ω
- Y delay: 40 ns

Triggering

- Source: chans 1 to 4, alternate, mains, ext.
- Coupling: AC-DC-LF-HF
- Switch: AUTO/NORM
- Level: pos/neg. flank

Horizontal section and time base

- Limit frequency in X/Y operation: 1 MHz
- Operational mode horizontal: TB1, mix, search, TB2, X/Y
- Delay setting: 10-turn potentiometer
- Variable hold-off
- Time base ranges: 10 ns to 0.5 s/div. in 1-2-5 steps plus variable
- X shift: coarse and fine

1.2. General Constructional Considerations

The mechanical construction is both simple and functional. The upper half is occupied in the middle by the cathode ray tube (CRT). To the left of this are located general operating elements (brilliance, ON/OFF etc.) and below the elements for trigger adjustment and to the right, those for the time base. The lower half contains the pre-amplifier plug-in units whereby in a simple manner, new plug-in arrangements are possible e.g. (electro-meter amplifier, clamp-ammeter amplifiers etc.) – all units having a common pin-out to feed in potentials and signals to the main frame. Situated under the plug-in bay are switches for the beam selection and X position.

This clear, uncluttered front-panel layout is made possible because the oscilloscope's main frame only contains controls for operation and for low-frequency signals. The use of more switches, possessing a fewer number of wafers, can be made to be more versatile and give greater operational clarity than the use of fewer expensive multi-wafer units. Special constructions, such as concentric time base switches fitted with locking arrangements, have been excluded for both technical and financial reasons. The range switching for Y amplifiers and time bases is carried out by switches via a micro-controller. Theoretically, all the other functions could have been controlled by these means, using menu inputs, but it has been found that a minimum

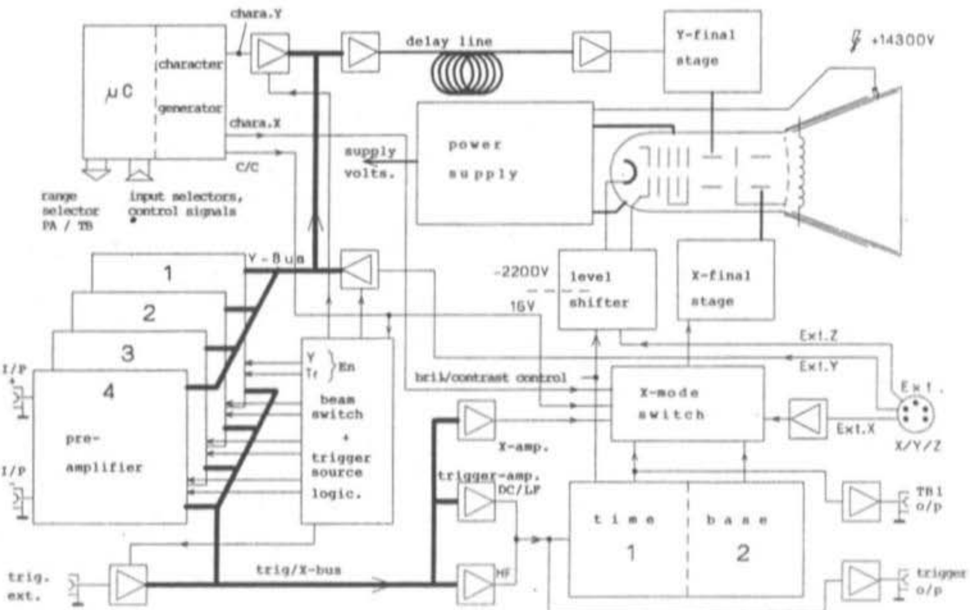


Fig. 1.1.: Complete constructional block-diagram

of mechanical function controls, with the appropriate inscriptions, give the best operational panel layout.

1.3. Circuit Overview

Figure 1.1. gives the block diagram. In the pre-amplifiers, the separation between the trigger and Y-signal circuits is carried out. The block "beam switch + trigger-source logic" determines which channel and at what time it is switched to either the trigger or the Y bus. In addition, control signals for other amplifiers are generated here and which also operate over one of the busses in the same manner as the selection of external trigger or the external Y input.

The multiplexed Y signal is taken from the bus via a driver stage to the delay line and the final stage. The trigger signal, also multiplexed in "ALT" operation, is taken to the trigger- and X amplifier and is only active in the X/Y mode.

The trigger amplifier possesses two parallel paths, one for DC and LF and one for HF. By activating the various paths, the trigger coupling is switched over. The trigger signal is taken to the time base and via an output stage to the back panel. Here, the signal may be intercepted for test purposes; a frequency counter, etc.

The time base comprises two similarly constructed units, the first is activated by the trigger signal and the second by attaining a certain output voltage at time base 1. By selecting the output magnitude of time bases 1 and 2, or an arithmetic coupling, various sweep modes are formed in the X-mode switching. The time bases generate also the output signals for the trace flyback blanking and the brilliance control during segment highlighting. The time bases are fed to a level shifter and then on to the CRT cathode. The latter is biased to − 2200 V in accordance with the Wehnelt system.

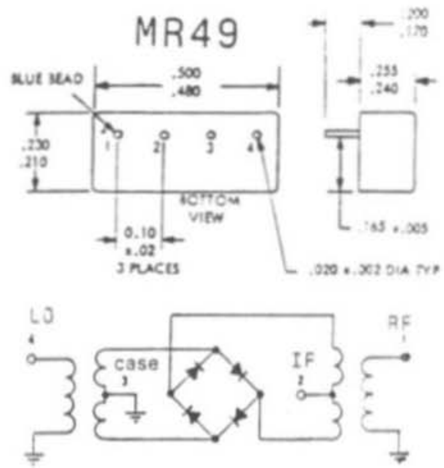
The X-mode switching also carries the X/Y selection control and the EXT-X facilities. The appropriate signal is processed and taken to the CRT via the X-driver amplifier.

The range selection is dependent upon the input selectors and the overlapping of the ranges on the upper and lower edges of the screen. It is controlled by the micro controller which possesses a coupled character generator. The characters are split into individual bits which carry both X and Y information. During about 5 % of the sweep time, these signals lie directly at the X or Y amplifiers. The switching is accomplished by the W/C (Write/Character) signal. During the character overlap, the flyback blanking is rendered inactive in order to avoid interference to the trace sweep. The power supply generates three different groups of voltages: The low-potential supplies for amplifiers, micro controllers and drive amplifiers, the supply for the cathode-Wehnelt system at about − 2200 Volt and finally the HV of 14 kV for the cathode ray tube. Interference-sensitive circuit points are fed from a conventional series controlled regulated power supply, the others (especially the driver amplifiers) obtain their power from a switched-mode power supply (SMPS).

The following chapters will introduce individual circuits and explain their operation. Simplifications and generalizations will be made which will enable an easier appreciation of the circuit operation. The diagrams, therefore cannot claim to represent the actual operational circuit.

To be continued

BRIEFLY SPEAKING...



Ring mixers

We have discovered a new, reasonably priced and quick supply for ring mixers. The firm, "Tele Tech", is represented in Germany by "mikro lambda", Munich (089-152050). The Tele Tech ring mixers are, for all intents and purposes, identical with those from Mini Circuits but retail for only half the price – at least in DL! For example, a 2 GHz mixer MR49 costs DM 78.00.

DB 6 NT/DB 1 NV

Miniature crystal oscillators

The well-established oscillator TQO 52 which has been offered by TELE QUARZ for years in TO8 housing, are now obtainable in TO5 packaging. This integrated oscillator is suitable for the supply of TTL circuits with a stable source of clock pulses or basic working frequencies. As opposed to conventional solutions, the following advantages are offered: –

Owing to its small construction, it offers a markedly lower profile and board space requirement. The reduced weight allows the module to be soldered directly onto the circuit board at any desired attitude.

The TQO 52 in the TO5 housing is available in four models: without divider, with two, with four and with eight dividers.

Frequency range:
500 kHz - 50 MHz
Temperature range:
– 55° C to + 125° C

Those oscillators which incorporate a start-stop function fulfil, for mechanical demands (shock, vibration), the IEC-norm (68 part-2 long-term test).

New Microwave Beacon

England has a new microwave beacon GB3OHM. (GB3OHM) IO92AJ 3456.9 GHz, 10 watts, ERP: omni. Reports to G6KOA QTHR: 458-1941 Birmingham, England.

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MATERIAL PRICE LIST OF EQUIPMENT

DF9DA	**Compact METEOSAT Converter**	**Art.No.**	**Ed. 1/1990**
DF9DA 001	kit with all components	6510	DM 395.00
	ready-to-operate module	3029	DM 575.00
DF9DA 002	**The matching weather-satellite receiver**		**Ed. 2/1990**
DF9DA 002	kit with all components, with 1 crystal for 137.5 MHz	6511	DM 445.00
Receive crystal for 134.0 MHz (METEOSAT-channel 2)		6512	DM 34.00
Receive crystal for 137.62 MHz (NOAA-9 and 11)		6513	DM 34.00
Ready-to-operate module (with 3 crystals)		3310	DM 890.00
YT3MV	**DSP Computer for Radio Amateur Applications**		**Ed. 2/1988, 1 + 2/1989**
Set of PCBs,	**progr. EPROM with authentic documentation** Contains: PCBs YT3MV 003 (bus), 004, 005, 006 (4 x), 007, 008 and 009; a programmed EPROM (operating system with Compiler, Editor), a set of copies of all diagrams and component lay-out plans (in A4), the operating-system manual.	6004	DM 599.00
YT3MV 010	**Interface for KR-5600 Rotators** PCB, programmed EPROM, together with A4 copies of circuit schematics and component location plans	6003	**Ed. 1+2/1989** DM 100.00
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