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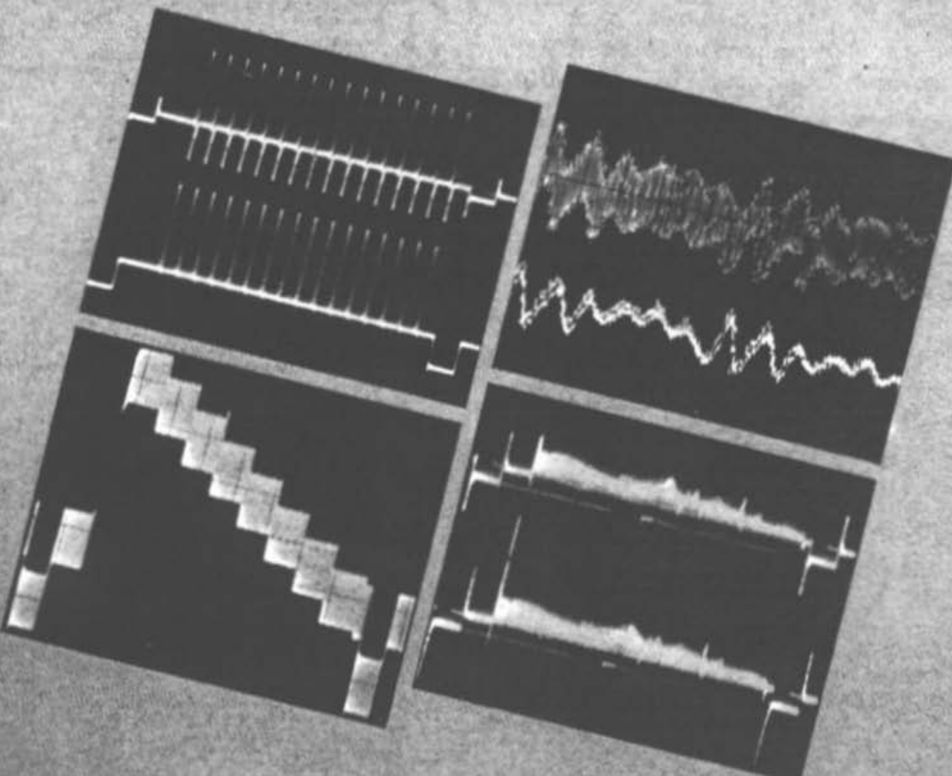
A Publication
for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 22 Summer 2/1990 DM 7.50



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VHF communications

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Volume No. 22 · Summer · Edition 2/1990

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

Publishers: TERRY BITTAN OHG

Editors: Corrie Bittan
Colin J. Brock (Assistant)

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

Advertising manager: Corrie Bittan

VHF COMMUNICATIONS The international edition of the German publication UKW-BERICHTE is a quarterly amateur radio magazine especially catering for the VHF / UHF / SHF technology. It is published in Spring, Summer, Autumn and Winter. The 1990 subscription price is DM 27.00 or national equivalent per year. Individual copies are available at DM 7.50 or equivalent each. Subscriptions, orders of individual copies, purchase of PC-boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative, or – if not possible – directly to the publishers.

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

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ISSN 0177-7505

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Andrew Bell, GW 4 JJW

10 GHz Varactor Tuned Gunn Oscillator

The Gunn oscillator to be described is tunable over about 100 MHz using surplus 2 pF varactor diodes, or 60 MHz if 4 pF diodes are used. The cavity is essentially a 3/4 wavelength type shortened to compensate for the presence of the varactor and Gunn diodes. Three such oscillators have been made by the author without problems.

1. CONSTRUCTION

The oscillator was constructed from a short length of brass WG16. First a brass plate is made according to **figure 1**, but the holes are not drilled at this stage. Although 2 mm thickness has been specified, a thicker piece may be better. This plate is high-temperature silver-soldered (780° C) to the WG16 taking care that the centre of the plate is aligned with the centre of the waveguide. The two 4BA holes are now drilled right through the plate and both faces of the waveguide. It is important that these two holes are correctly aligned and perpendicular to the waveguide. The holes are then enlarged to 8.5 mm on the side opposite to the plate and finally reamed out so that the brass tube is a tight fit. The two brass tubes, see **figure 2**, are pushed into the waveguide so that they are flush with the inside wall and then carefully aligned. They are then soldered into the waveguide using a lower temperature, 617° C, silver solder. Finally with wet tissue or cloth in the waveguide the brass cross member, shown in **figure 3**, is silver-soldered between the two brass tubes. This will connect to the screens of the cables leading to the diodes.

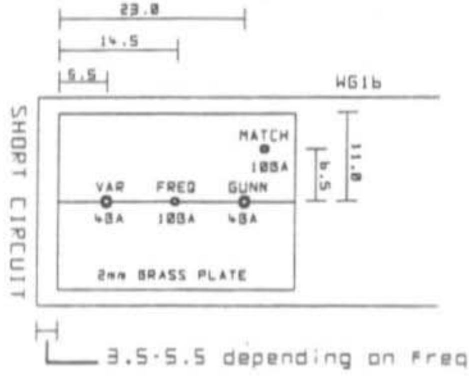


Fig. 1: Cavity details

The cross member is drilled and tapped 4BA, or as required. The 4BA holes in the brass plate are then tapped and the 10BA ones drilled and tapped. Finally the four 10BA holes in the tops of the tubes are drilled and tapped. These four bolts hold the wood in place. Wood has been used to absorb any RF that manages to bypass the chokes. The threads should all be tight – especially in the brass plate. In the case of one of the prototypes this was not so and 10BA brass nuts were soldered onto the brass plate for extra support – (**see Photographs**).

In the prototype the chokes were cut to size on a lathe, but these could probably be constructed by soldering different diameters of brass end-to-end. **Figure 4** gives the choke details. The diameters are not particularly important, but the larger the step in diameters the better. The outer diameter will probably be determined by the inner diameter of the tube and the thickness of 1.5 turns of Selotape (or similar). The inner (smaller) diameter should be as small as possible whilst still maintaining an acceptable mechanical strength. One and a half turns of Selotape are then wound around the outer (larger) diameter sections and excess trimmed off with a sharp knife. The chokes with Selotape should be a close fit in the tubes.

A length of wood is turned, or sand-papered, down to the correct size to fit inside the brass tubes and a 2 mm hole is drilled through the centre of the wood to take the end of the choke.

Holes are then drilled in the bottom of two 4BA brass bolts, **figure 5**, to take the Gunn and varactor diodes. A lock nut is screwed well on to each bolt which is then screwed into the cavity. The diodes are placed in the holes in the bolts using tweezers. Once the short circuit is soldered in place it is obviously better to place the varactor diode first. The choke is then dropped down the tube to engage the diode. The bolts are adjusted such that the wide diameters of the chokes are flush with the inside wall of the waveguide. Finally the four 10BA bolts in the tops of the brass tubes are tightened into the wood to fix the chokes in position.

Coaxial lines are connected between the tops of the chokes and a suitable PSU or TX. The brass short circuit, see **figure 6**, is either held with tight rubber bands or temporarily lead-soldered into place over the end of the wave-guide. The frequency range of the oscillator is now determined. It can be corrected upwards by

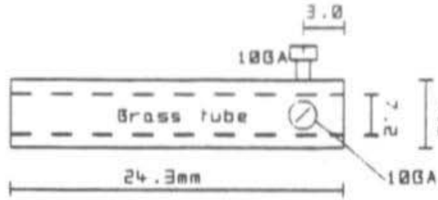


Fig. 2: Preparation of choke tubes

Fig. 3: Cross member

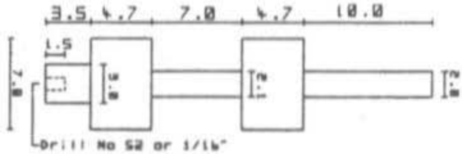


Fig. 4: Choke details

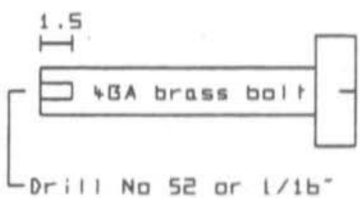
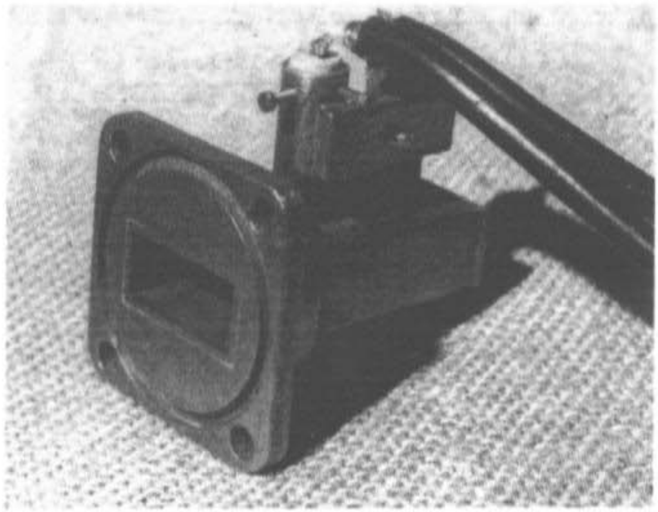
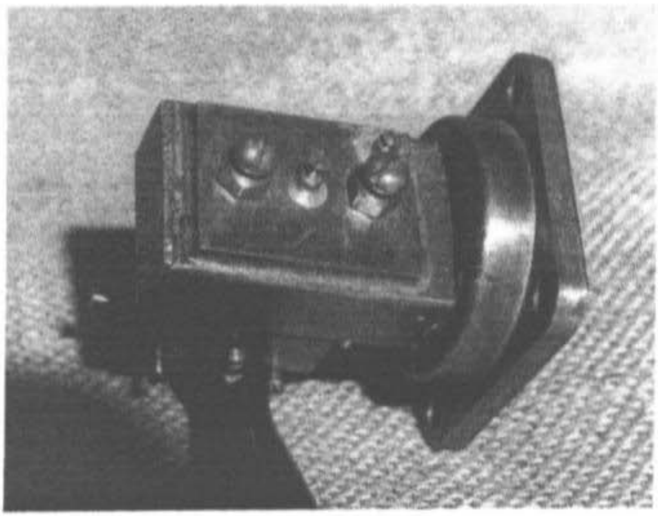


Fig. 5: Preparation of brass bolts



Photograph 1:
Varactor oscillator (top view)



Photograph 2:
Varactor oscillator (bottom view)

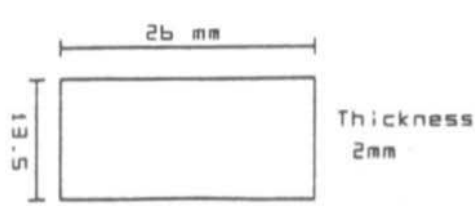


Fig. 6: Short circuit

removing the short circuit and filing a small amount of brass off the end of the waveguide. (Note that the frequency can be lowered about 50 MHz by using a varactor diode of larger capacitance, say 4 pF instead of 2 pF.)

When the frequency has been set to the required range, the short circuit is permanently lead-soldered into place allowing as little solder to get inside the waveguide as possible.

2. TYPICAL RESULTS

The following tables show the degree of frequency shift obtained with 2 pF and 4 pF varactor diodes and the 10BA tuning screw. The readings were obtained by repeatedly screwing the frequency adjusting bolt by one turn and measuring the minimum and maximum frequency obtainable using the varactor. The total amount of frequency shift available from the varactor is shown in the penultimate column and the actual frequency shift caused by the tuning screw approximately one turn is shown in the right hand column. No adjustment was made to the matching screw while taking measurements. The power measurement is only relative, it is in fact the rectified current through a SIM2 diode on the output of a wavemeter coupled to the cavity via an attenuator.

2 pF Varactor diode

Varactor = 0.16 V		Varactor = 8.8 V				
Freq.	Power	Freq.	Power	Turns	Varactor shift	Screw shift MHz/turn
10.265	2.8	10.365	2.4	0	100	–
10.261	3	10.361	2.4	1	100	4
10.251	3	10.345	2.6	2	94	16
10.231	3.8	10.331	2.6	3	100	14
10.208	5	10.305	2.6	4	97	26
10.181	Unstable	10.278	3.0	5	97	17
10.065	3.4	10.258	1.5	6	77	20
9.985	2.8	10.168	6	7	103	Unstable
		10.085	3.5	8	100	83

Table 1:
The degree of frequency shift obtained with 2 pF varactor diodes

4 pF Varactor diode

Varactor = 0.16 V		Varactor = 8.8 V				
Freq.	Power	Freq.	Power	Turns	Varactor shift	Screw shift MHz/turn
10.235	4.4	10.295	3.4	0	60	–
10.231	5	10.291	3.4	1	60	4
10.325	5.1	10.278	4.8	2	53	13
10.208	6	10.265	5	3	57	13
10.181	6.5	10.241	4.8	4	60	60
10.155	5.5	10.211	5.3	5	56	30
10.108	4	10.165	6	6	57	46
10.038	3.5	10.098	4.1	7	60	67
9.961	3.0	10.015	3.4	8	54	83
9.851	5	9.908	3.8	9	57	57
9.718	3.5	9.771	6.8	10	53	137
9.541	4	9.598	3.6	11	57	173
9.341	4.6	9.381	6.8	12	40	217

Table 2:
The degree of frequency shift obtained with 4 pF varactor diodes

Dipl. Eng. Detlef Burchard, POB 14426, Nairobi, Kenya

Shortwave Reception – Based on the Thirties Principles Part 2 (Conclusion)

4. A PRACTICAL SW RECEIVER EMBODYING THESE PRINCIPLES

In order to put these theories into practice, the circuit of **fig. 7** was developed and built on "Minimount" (**fig. 8**) experimental board using good RF lay-out practice.

This article is not a constructional guidance as the construction is in no way critical compared with other projects in the GHz-region – but others may put the priorities differently. He may, for example, prefer to have as good a noise figure as is technically possible or to have a continuously tunable input selection. Developments from Günther (3) and Kestler (4) can be combined in this concept. An IM-free dynamic range of at least 80 dB was aimed for. This can – as calculated previously – result in a usable signal range of 54 dB. The de-sensitizing and blocking effects are minimal in this region. For them to take effect, the input signal level must be some 20 dB higher. The IF and image suppression should also be better than 80 dB.

The input is provided with a high-ratio transformer. It is suitable for antennas having a 50 Ω impedance and also for the previously mentioned ground-plane CB antenna with five to ten metres of feeder coaxial cable. The input circuit response has a band selection influenced by the series-tuned circuit to the base and the 8-pole filter behind the transistor and is shown in **fig. 9** for the two cases of antennas mentioned.

The primary inductance of the input transformer has, together with the cable capacitance, a resonance at about 4 MHz when the CB antenna is in use. This results in a rising response with frequency which is matched to the atmospheric conditions. With the input terminated with a real impedance of 50 Ω the response is, of course, sensibly flat over the bandpass. The rather sharp upper-frequency flank of the response is, due to the employment of high-Q inductors, for all the tuning elements.

The Tschebyscheff 8-pole low-pass filter also serves as a 150 Ω to 50 Ω transformer and has a Cauer end stage. Theoretically, the in-band ripple characteristic should be within 2.5 dB. The Cauer end stage gives the response an attenuation of

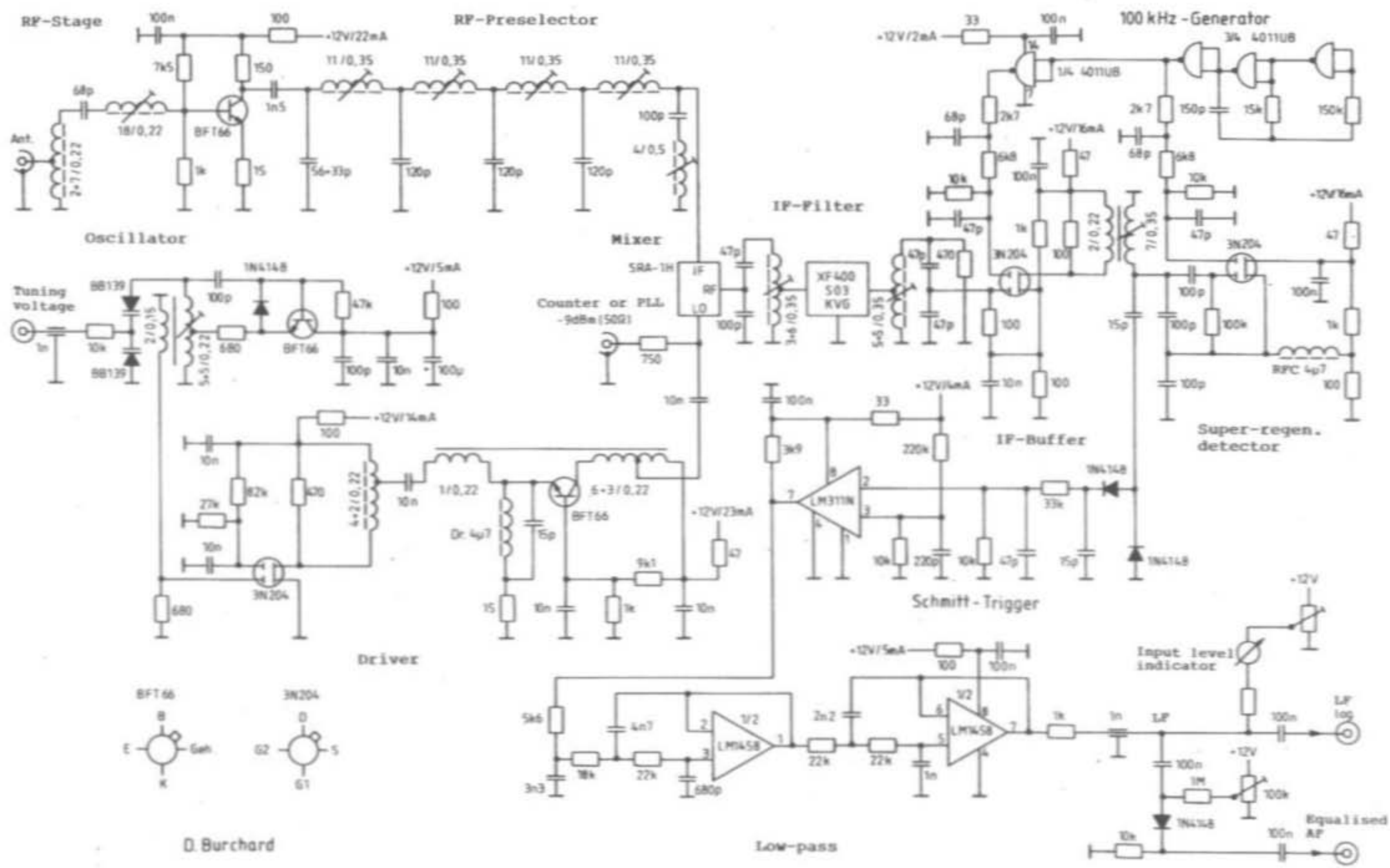


Fig. 7: The complete circuit of the shortwave receiver. All variable inductors are: Vogt-coil kits 51404 00000. All RFCs are marked Dr. All other inductors are Siemens double-hole beads B62152-A0007-X001

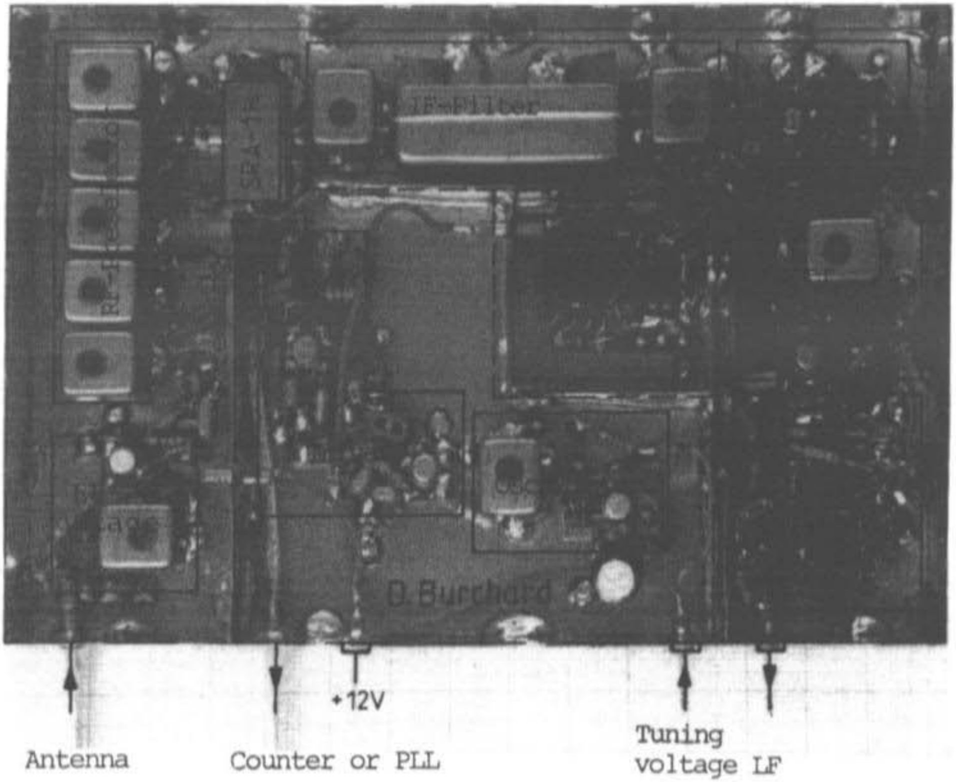


Fig. 8: Bread-board construction of the receiver using "Minimount"

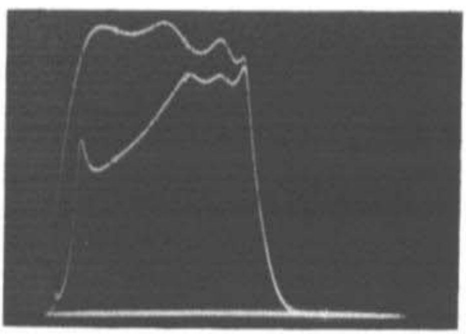


Fig. 9:
Preselector response in the frequency range 0 - 50 MHz
X: 5 MHz/cm, Y1: 5 dB/cm (50 Ω source). Y2: 5 dB/cm and terminated in the antenna (see text)

some 90 dB at 40 MHz. Oscillator signals in the range 43 MHz to 68 MHz are attenuated by 75 dB to 88 dB. Image-frequencies in the range 83 MHz to 108 MHz are attenuated by 87 dB to 103 dB. These high-order measurement cannot be undertaken with a normal wideband sweeper, a selective level measurement must be employed.

The choice of a bi-polar device for the input stage may well be surprising, however, on account of the current feedback in the emitter circuit the stage works sufficiently linear. Every kind of feedback, as for example in integrated-circuit RF-amplifiers, lowers the isolation between output and input thus allowing a greater level of oscillator signal to be radiated from the antenna. Here, an unfavourable −82 dBm (**fig. 10**) was measured, which represents a 26 dB deterioration on the oscillator radiation permitted in commercial practice. The oscillator is tuned by means of varicap diodes over a range of 45 MHz to 68 MHz with a 2 - 10 V tuning voltage. It is thereby suitable for the control by a PLL which can work from a + 12 V supply. The oscillator must be extremely low-noise, possessing a very good filtering of the tuning voltage. A side-band noise of −77 dB was measured in an 8 kHz slot, 10 kH removed from the oscillator signal. That appears to be very little, but when it is expressed in terms of unitary bandwidth it comes out at −116 dB/Hz. Taking again the 8 kHz measurement slot and noting the manner with which the noise rises at a greater rate towards the carrier, one obtains the same sort of values as in (4) even though the frequencies lie considerably higher. An improvement would, nevertheless, be desirable, especially if a higher dynamic range is required.

The oscillator output power level is always within ± 0.3 dB over the tuning range thanks to the high negative feedback in the transistor as well as a constant load. A level regulation is therefore not necessary, as the following two-stage driver is broad-banded and its response can be neclected. Its first stage serves as a buffer for returning signals and also to establish a defined source resistance for the second stage. This forms, theoretically, a transformer negative feedback and has therefore a good efficiency. It has also the property to possess a 50 Ω internal impedance when it is terminated in a 50 Ω source impedance. The latter is provided by the first stage. The provision of such an internal impedance is therefore important since the LO-port of the mixer is highly non-linear. It is not possible, therefore, to arrive at the driving power, not even by a voltage measurement at the LO-port. It must be measured into an external 50 Ω load termination. Without having a 50 Ω internal impedance, the mixer is not correctly operated.

The mixer uses an MCL ring mixer type SRA-1H and is driven with the IF and the RF-ports interchanged so that the input frequency range starts at zero. That can be useful if the medium-wave band is also desired. A diplexer behind the mixer did not appear to confer any advantage but purists may want it included anyway. The importance of good additional stop-band rejection characteristics for the IF stage can be satisfied with LC-circuits both before and behind the IF filter apart from their use as matching elements. The KVG filter has a mid-band frequency of 40.000 MHz and ± 3.6 kHz bandwidth (−3 dB) and uses third-harmonic oscillations. It could have been used in the basic mode or on other harmonics (like any other crystal filter). It was chosen because it was available and had the required characteristics. The adjacent channel selectivity amounts to some 60 dB. This is sufficient to ensure that the conditions for an infinite demodulation selection, irrespective of the power of the interference in the presence of a weak wanted signal (within the specified dynamic range), are provided.

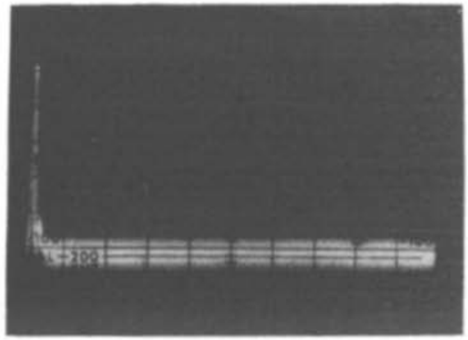


Fig. 10: Spurious spectrum at the antenna input. Left is the zero mark. X: 0 - 200 MHz, Y: 10 dB/cm, the centre line = −102 dBm

Between the IF stage and the detector there is a buffer stage having no amplification. Its purpose is to ensure that the high oscillator input power does not reach as far as the filter, the detector stage is also presented with a constant source impedance which is continually being switched on and off.

From the discussion about the detector feedback, it will be recalled, that after the provision of the start energy, no further requirement for RF energy is necessary. The buffer can therefore be switched off and is switched on again when the detector tuned circuit is in a decremental

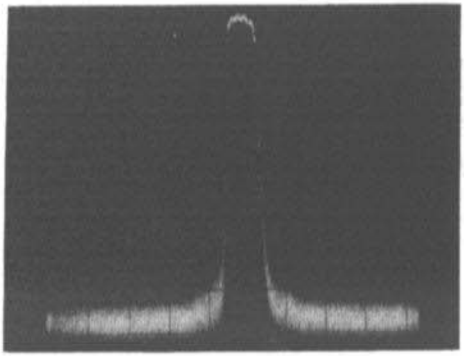


Fig. 11: Selectivity curve and noise pedestal
X: 10 kHz/cm, Y: 10 dB/cm

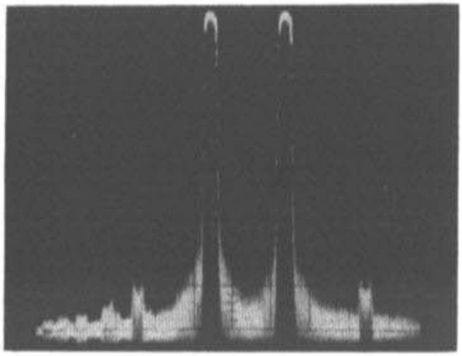


Fig. 12: IM products of two −25 dBm signals
X: 20 kHz/cm, Y: 10 dB/cm

condition. This is very advantageous under the conditions of high wanted signal levels. The tuned circuit oscillates at its free resonant frequency which can be quite different from that of the driven frequency. If the driven frequency has sufficient power, beat undulations can be audible in the output, otherwise its working is in accordance with the characteristics discussed.

The oscillating frequency is generated in an C-MOS gate-circuit astable multivibrator, two free gates serve as decoupling and phase reversal. In this case, the use of un-buffered gates are recommended. Their flank slopes are smaller and have no tendency to HF oscillations at the switch-over point. In order that the control of the detector and the buffer stage is carried out in a spurious free manner, the slopes must be decreased by a two-stage RC network. At this point, purists could use either an active or a passive low-pass filter of a higher order.

The dual-MOSFETs used were measured previously and then a working point with a high linearity and symmetrical limits were determined. It is given by the following: U_{G1S} = 0 V; U_{G2S} = + 2.5 V; I_D = 12 mA; S = 14 mS. The amplification is reduced by 60 dB (switched off), when U_{G2S} = −1 V. The source is raised to + 1 V when installed in the circuit. Such a stage is able to process an input signal of 1.8 V_{pp} linearly. The same type of transistor from another manufacturer or from another series can exhibit large deviations from this characteristic. To obtain optimal results it will be worth-while to carry out this selection measurement procedure. A circuit for serial production would have to be designed upon quite different lines. It is, however, to be understood that the devices which were installed in the circuit, were employed merely because they were available. There must be several dozen similar types which would do the job just as well. The important thing to watch is that the amplification at the working frequency is high enough and that their noise-figure is not unduly high.

The Schmitt trigger and the low-pass filter following the detector stage were realized with integrated circuits. Since they are standard circuits they do not require any explanation. Behind the low-pass filter the rectified voltage can be removed and indicated on an instrument. This gives an exact logarithmic display of the signal voltage. The audio is decoupled via a capacitor which must match the sound balance. The indicated value results in a limit frequency of 300 Hz at a 4.7 kΩ load.

Finally, the pre-tested diode is connected, this carries out a partial compensation of the logarithmic demodulation. Partly because of the low-pass, all low frequencies above 2 kHz contain no harmonics. They are also compensated by the diode. A compromise must be found which is most easily accomplished by listening to the sound output. If the process of compensation was carried out before the low-pass, these demodulation products would be regenerated by the non-linearity – which the inherent properties of the logarithmic rectification have eliminated when under the influence of a high-level interference signal!

The low-pass is dimensioned in order that at the limit frequency of the IF frequency it exhibits a small rise in its characteristic. Altogether, an audio range of up to 3.8 kHz +0/−3 dB result when the receiver is on the correct tuning point for the incoming signal thus giving an extraordinarily good speech intelligibility.

4.1. Measured Characteristics

The measurement of the receiver's characteristics turned out to be no simple matter in the absence of a laboratory full of test instruments. The spectrum analyzer trace of **fig. 11** is the best indication of the selectivity performance which could be presented. The Y-range spans 80 dB and it can be seen immediately, that the signal generator is responsible for the production of more noise than the receiver's local oscillator. The latter was caused to wobulate by varying DC (from a battery) as the tuning voltage and a small saw-tooth voltage from a function generator. Following that, the crystal oscillators were tested: the reference frequency of 10 MHz taken from an electronic frequency counter proved to be contaminated with divider products and increased the noise floor up to a level of −55 dB, the reference frequency from the signal generator synchronizer (Marconi TF2171) also caused the noise floor to increase to −76 dB. The trace was obtained in the presence of a noise floor of −80 dB in the absence of signal i.e. at the extreme limit of indication. A comparative estimate of the oscillator side-band noise of −77 dB could then be carried out.

The spectrogram of **fig. 12** was obtained by feeding the receiver with two equal-level signals, the lower frequency (left) was derived from a Marconi TF2015 signal generator and the other from a home constructed generator. It may be seen that the noise pedestals are much higher than in figure 11, the noise floor is lower but there are IM products in evidence. The third-order products are the ones which are closest

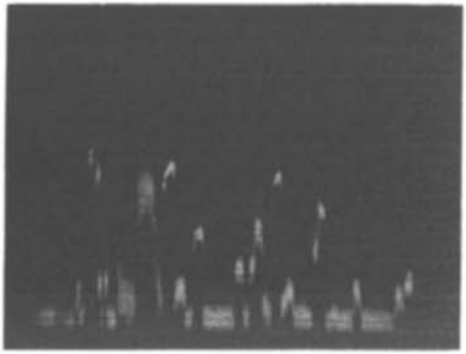


Fig. 13: Occupancy of the 19 m band on 14.3.1989 at 0945 GMT
X: 50 kHz/cm, F_{mid} = 15.70 MHz, Y: 10 dB/cm

to the two carriers and at a level of −72 dBc. The two input signal powers were −25 dBm and therefore the IM_3 intercept point can be calculated to be + 11 dBm at the receiver input.

The specification aim of 80 dB IM-free range and −109 dB internal noise was therefore, not quite achieved. The point here though, is to demonstrate the benefits of logarithmic detection in the super-regenerative demodulator. When a large dynamic range is aimed for, then better signal generators are required in order that this quantity can be measured. In this respect, signal generators from the 1960s using valves, are the better instruments for this purpose. They deliver a purer output signal of some 20 dB better than the latest "high-tech rubbish".

4.2. Other Uses

Just think now, of an optimal spectrum-analy filter curve having a form factor (60 dB to 3 dB bandwidth ratio) of 4.4 instead of the rectangular selectivity curve shown in fig. 11. This type of selectivity is able to be wobbulated at a faster rate in applications such as in panoramic adaptors or spectrum analyzers. It is well known that the logarithmic detector in such equipment demands a good deal of complex design work. Here, it is done with a single transistor circuit.

The spectrogram of **figure 13** repesents a segment of the 19 m band. Somewhat right of the centre can be seen two transmissions having a separation of only 5 kHz. These could be received, thanks to the inherent separation qualities of this form of detector but it must be said that luckily, the two signals were just about equally strong. The 5 kHz spurious tone could be attenuated by 30 dB using the audio filter and it was hardly discernable. The author is now working on a suitable PLL for this project. If there is enough interest displayed, an article will appear on the subject.

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Martin Althaus, DF 9 DA

Compact Weather-Satellite FM Receiver

The receiver to be described has three crystal-controlled channels at 134.0, 137.5 and 137.62 MHz. This enables the receiver to work the APT transmissions of the polar-orbiting NOAA weather satellites using the last two frequencies. By using an additional converter – described in (1) – the reception of all WEFAX transmissions of the geo-stationary weather satellites METEOSAT, GOES, GMS is possible. They use the frequencies 1691.0 MHz and an additional METEOSAT frequency at 1694.5 MHz. The compact converter translates these frequencies to 134.0 MHz and 137.5 MHz.

The compact receiver makes an extensive use of integrated circuits which contributes not only to its compact form but also to its reliability of replication. As far as the technical specifications are concerned, the low-noise input stage and the S-meter output indicate a dynamic range of greater than 60 dB.

1. LEADING CIRCUIT DETAILS

The basic receiver for a microwave converter, such as the METEOSAT converter, does not need to be particularly sensitive. As this receiver, however, is employed for direct reception of the APT transmissions the input stage is fitted with a low-noise, dual-gate MOSFET. The circuit schematic is shown in **fig. 1**.

It may be seen that the first stage is provided with a two-stage pre-selector before the active device and a three-stage filter following it. Capacitive impedance transformers are used both on the antenna side and at the mixer.

The integrated-circuit mixer is an active type, the Signetics (Valvo) NE 612, which is characterized by a high conversion gain and a low component count necessary to complete the circuit. The mixer translates the input signal to an IF of 10.7 MHz. The IF is symmetrically coupled out and matched to a ceramic filter via a tuned circuit. This 30 kHz bandwidth filter largely takes care of the selectivity requirements of the receiver. A further tuned circuit is used for output matching and also assists in the selectivity. The IF amplifier also uses an integrated circuit, the Signetics (Valvo) NE 614. This contains a two-stage amplifier which allows the insertion of a further ceramic filter which reduces the effects of wide-band noise. The NE 614 also contains a demodulator, a muting circuit as well as an output for an S-meter indication (max. 5 V in 100 kΩ.). The latter is buffered by an operational amplifier in order that a moving-coil voltmeter may be directly

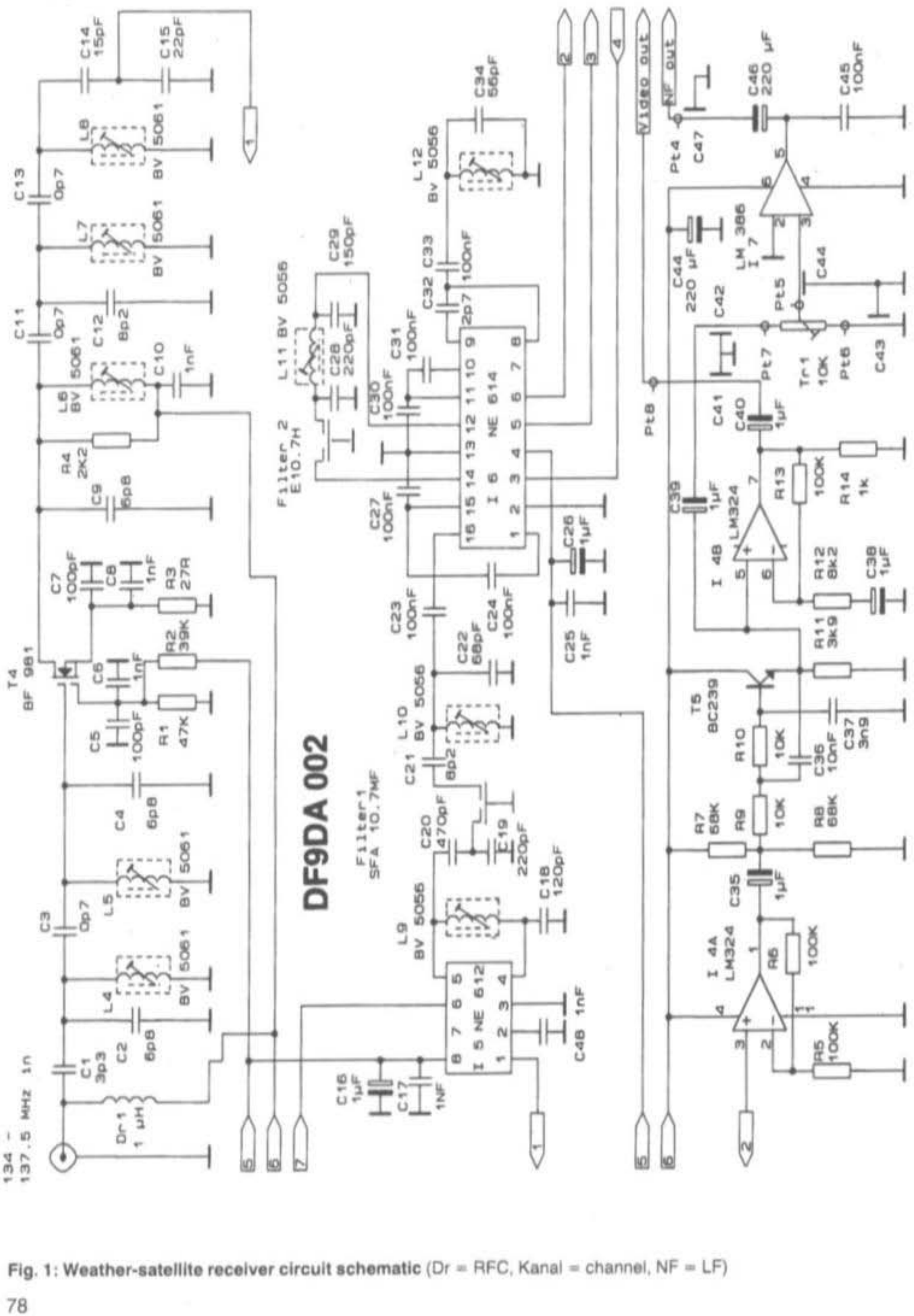
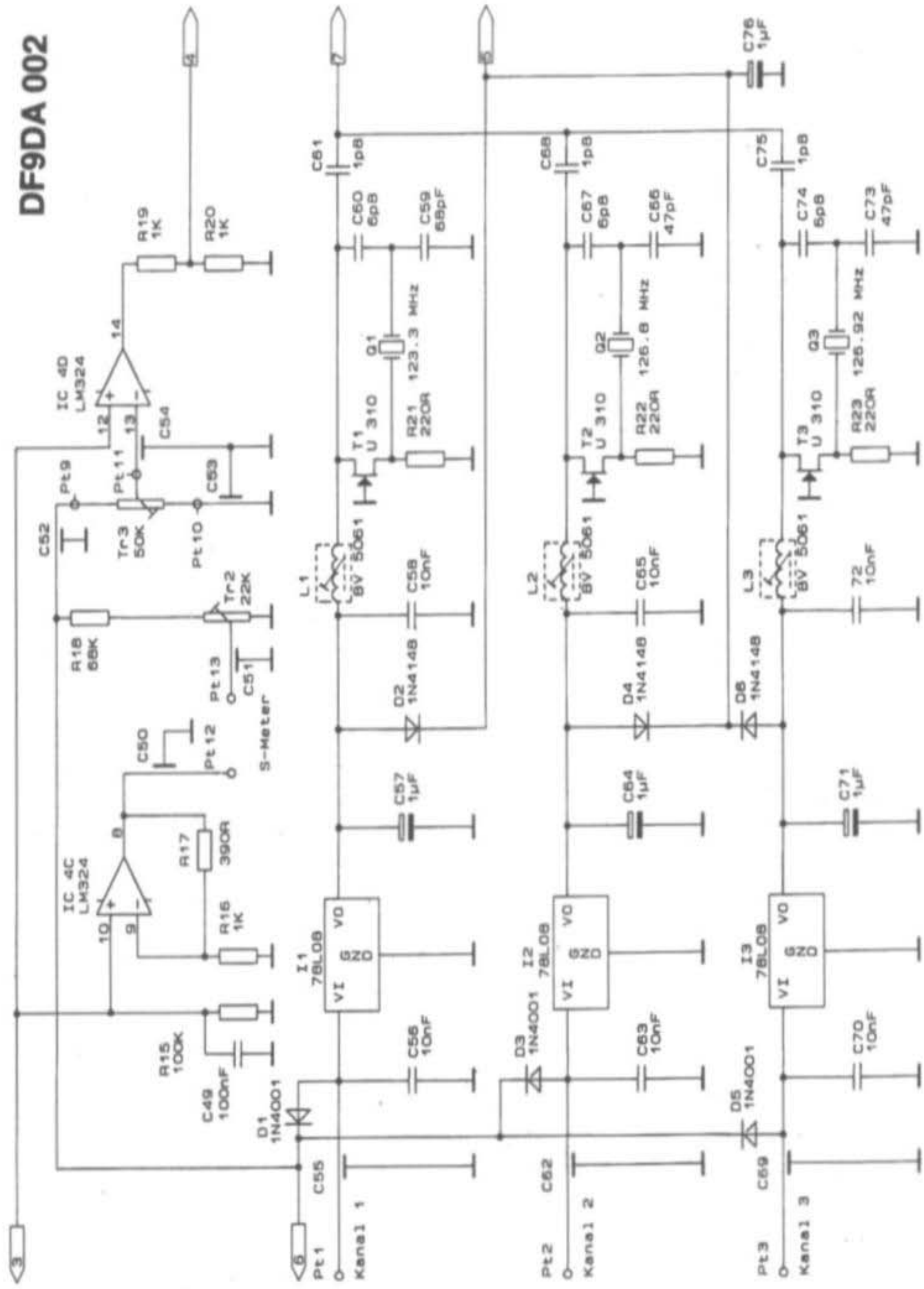


Fig. 1: Weather-satellite receiver circuit schematic (Dr = RFC, Kanal = channel, NF = LF)


DF9DA 002
IC 4D LM324
IC 4C LM324
R19 1K
R20 1K
R15 100K
R16 1K
R17 390R
R18 68K
Tr2 22K
Tr3 50K
C49 100nF
C50
C51
C52
C53
C54
Pt9
Pt10
Pt11
Pt12
Pt13
S-Meter
C61 1p8
C60 6p8
C59 68pF
Q1 123.3 MHz
T1 U 310
R21 220R
L1 BV 5061
C58 10nF
D2 1N4148
C57 1µF
I1 78L08
C56 10nF
D1 1N4001
C55
Pt1
Kanal 1
C68 1p8
C67 6p8
C66 47pF
Q2 126.8 MHz
T2 U 310
R22 220R
L2 BV 5061
C65 10nF
D4 1N4148
C64 1µF
I2 78L08
D3 1N4001
C63 10nF
C62
Pt2
Kanal 2
C76 1µF
C75 1p8
C74 6p8
C73 47pF
Q3 126.92 MHz
T3 U 310
R23 220R
L3 BV 5061
72 10nF
D6 1N4148
C71 1µF
I3 78L08
C70 10nF
D5 1N4001
C69
Pt3
Kanal 3

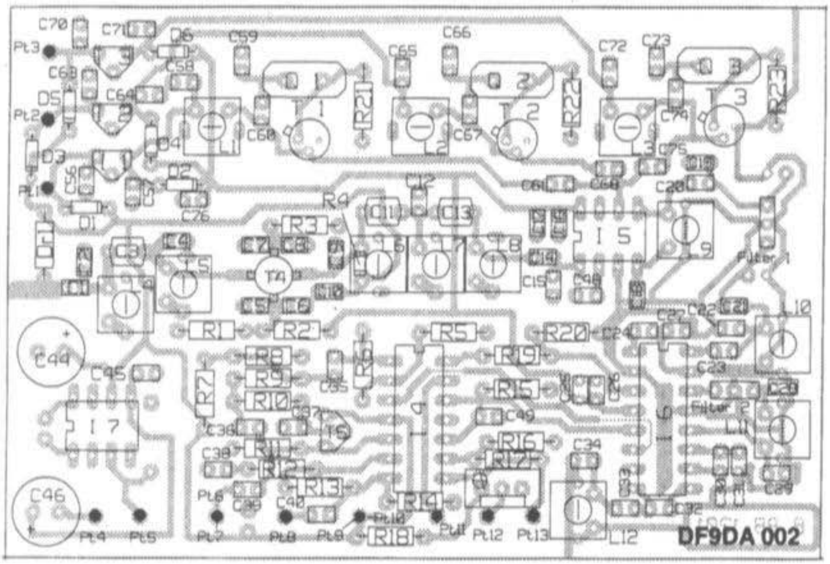


Fig. 2: Double-sided printed circuit board of dimensions 111 cm x 74 cm

connected to this output. The demodulated signal is amplified in another op-amp. and is then taken to an active low-pass filter (T5). A final stage of amplification follows in which the video output is boosted to an amplitude of 1.5 V_{pp}. A small audio IC power amplifier, the LM 386, takes care of the sound side of things. The 4th op-amp. in the IC, LM 324, is used for the muting circuit.

The three oscillators in the lower part of the schematic work with crystals on their 5th or 7th overtones together with a tuned circuit on the frequency of operation. The same low-noise circuit is used as that in the converter oscillator, with a barrier-FET U310. The RF outputs are connected together with the mixer oscillator input (pin 6), the channel selection is effected by applying the supply potential (12 V to 15 V) to one of the three inputs. The supply voltage is available via decoupling diodes as is the stabilised 8 V, both minus 0.7 V diode drop, for eventual use in the other receiver parts.

2. CONSTRUCTION

The complete circuit is realized on a circuit board of only 111 mm x 74 mm dimensions (**fig. 2**). It is double-coated copper foil and has the designation DF9DA 002. The component side remains free of any etching, is through-plated and serves as a ground-plane. There is a little mechanical work to be carried out on the complete module, as can be seen in **fig. 3** – the completed receiver unit.

The first thing to do is to put the two halves of the frame together into a cover and solder them together – but not to the cover! The inner conductor of the BNC socket is shortened to a length of 3 mm and the socket is then inserted from the outside into a hole in the frame and secured with M2.5 mm screws from the inside.

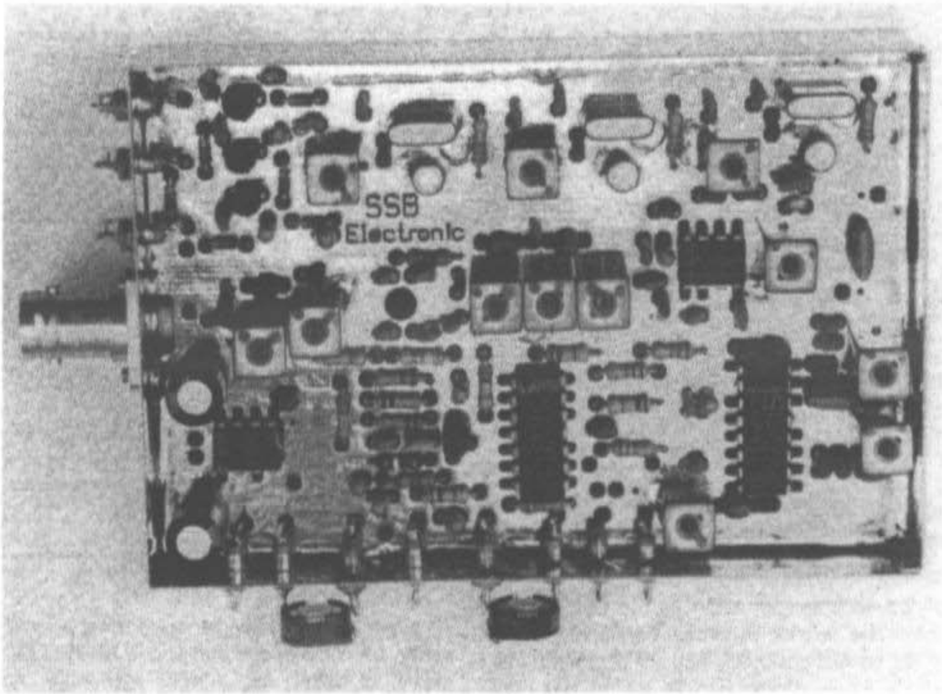


Fig. 3: An operational receiver module

Now the PCB with the etched side is so positioned that it lies in contact with the BNC socket, the ground-plane side is turned towards the feed-through capacitors. After aligning the PCB to be exactly in the plane of the frame it is soldered, on the ground-plane side, to the walls of the frame.

Now the whole board can be equipped beginning with diodes D1 to D6 – watch the polarities! As the PCB is through-contacted, not all the components need to be soldered to the ground-plane side as well.

Then install all the resistors and capacitors according to the component layout plan and solder them in.

The experience of aligning many samples of this project has shown the requirement for a 0.3 pF capacitor to be wired across capacitor C3 (0.7 pF). This is located on, and soldered to, the track side of the board.

When installing the Neosid filter, the colour-code markings should be carefully noted. The surplus-to-requirement pins (one pin on the BV5061 and three on the BV5056) should be carefully removed with side-cutters. **Figure 4** shows the connection array.

Soldering the semi-conductors is effected with the usual measures associated with the risk of static damage – earth the iron tip and the metal work. Slip a ferrite bead over the drain lead of the BF 981 before installation (**fig. 5**) and solder the gates of the U310s directly to the ground-plane. Installing the ICs should present few problems as their orientation is clearly marked in the component layout plan. No IC sockets should be used!

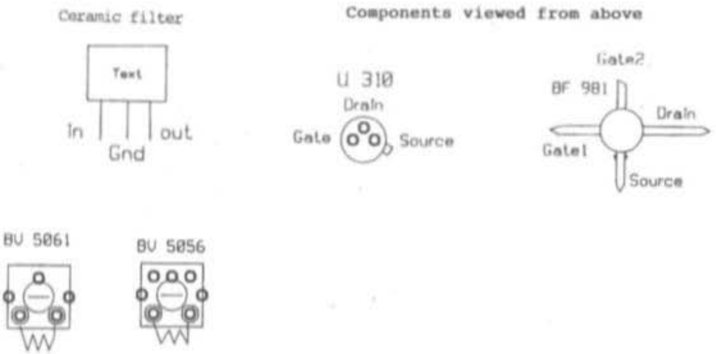


Fig. 4: Component lead identification

If a converter, or a pre-amplifier is to be remotely supplied from the receiver via the coaxial cable, the choke RFC 1 is inserted.

The following points are to be settled for the rest of the construction: solder in the crystals and watch the ceramic filters that they are soldered in the correct positions, they are different bandwidths F1 = 30 kHz and F2 = 110 kHz. The damping resistor R4 is located and soldered on the board track-side in the drain circuit of the BF 981. The 2.2 kΩ chip resistor is soldered in parallel with the inductor L6. That completes the loading of the printed circuit board.

Now the feed-through capacitors are soldered in – use a larger soldering bit. These capacitors are shown in the circuit diagram without any values. The supply connections each have a solder tag for the ground connection under them – bend them out a little before soldering the connector/lug assembly into position.

The volume and squelch presets are soldered on to the appropriate feed-throughs. The connecting wire (except to C70 and C75) must be shortened – the PCB connections are made with wire cuttings from resistors etc. The feed-through connections to C43 and C51 (at the presets, via C42 or C52) are bent over and soldered to the housing.

With this, the construction of the receiver is completed and it should be ready for an easy alignment!

3. ALIGNMENT

An exact alignment can be made using a multimeter, an oscilloscope and strong METEOSAT signal, or better still, a signal generator. An important item is a correct type of plastic trimming tool. It must be used in order that the Neosid filter cores are not destroyed.

Before the alignment is commenced, a few spot checks and adjustments should be made at the following points:

Squelch preset TR 3 is completely turned ACW. Volume preset TR 1 is tuned fully CW. The following voltages are to be measured at either channels 1, 2 or 3.

According to the channel selection, the appropriate voltage regulator should be delivering 8 V at the output. The other two regulators should have no voltages at the output. This applies also to the source resistors of the oscillator transistors: the voltage at these resistors should be approximately 1.7 V when the oscillator is working. All other voltages listed below, are to be measured independently of the channel selection.

If the voltage at the drain and at I4 is not present, either diode D4, D5 or D6 has been wrongly installed or there is a short-circuit.

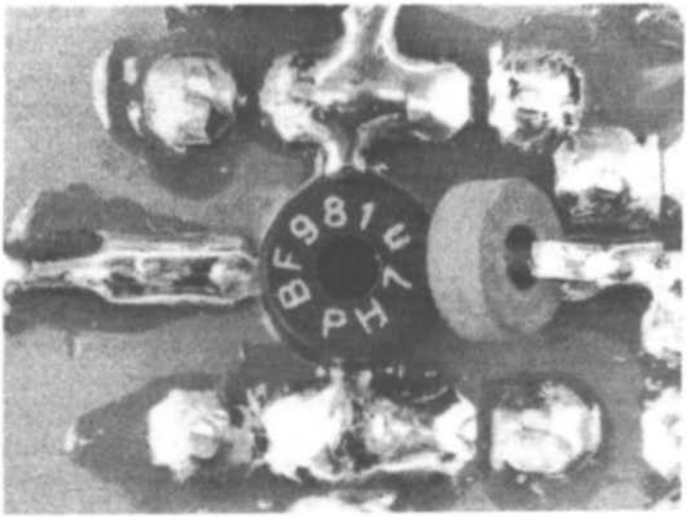


Fig. 5:
The Pre-amplifier stage T4 showing the ferrite bead over the drain connection

Test point	Voltage
T4 drain	$U_b - 0.7$ V
T4 gate 2	4 V
I 5/pin 8	7.3 V
I 6/pin 4	7.3 V
I 6/pin 3	approx. 6 V
I 4/pin 4	$U_b - 0.7$ V

If the measured voltages are in accordance with the values given in the table, the alignment can proceed.

First of all the pre-selector cores L4, L5 are turned out two turns, and L6, L7, L8 are turned 1½ turns. The oscilloscope is then connected to the video output Pt 8 (Y sweep: 0.5 V/cm). Now the filter cores L9 to L12 are turned roughly to their centre points. The oscilloscope should then start to display a noise voltage. If only a limited noise voltage can be observed, the cores must be turned again until the noise voltage is 0.5 V max.

Upon switch-on of the oscillators, adopt the following procedure: Apply the supply potential to the appropriate channel. With a counter probe at pin 6 of I 5, the oscillator frequency is adjusted with the oscillator inductor core. The source voltage will sink from 1.7 V to about 1.55 V. The functioning oscillator should now result in an increase in the noise at the video output.

If now, an FM signal is fed into the antenna terminals from the signal generator, or from METEOSAT, the demodulated signal should be displayed on the oscilloscope – if only a distorted version. If this is not the case, the IF-tuned circuits must be re-aligned.

The correct position of the demodulator inductor core (L 12) can be determined in the following manner: The noise disappears when the core is turned fully in. If the core is now turned slowly out again, the noise increases. The core is turned out until the noise starts to decrease again. Two points of maximum will be observed. Now the core is returned to the first maximum. The fine tuning of the IF filter L9 to L 11 and the demodulator inductor L 12 requires a very fine touch.

In the course of the alignment, the voltage at the S-meter output can also be mesured. Owing to the high-dynamic range of the S meter (60 dB), this is readily available. It must be ascertained that, even without an input signal, a voltage of 1.5 V approx. can be measured when a converter is added. Without the converter this voltage is 0.5 V approx. This output voltage is caused by the high pre-amplification. In order to back off any moving-coil instrument to zero, the voltage divider R 18/Tr 2 has been provided. If this facility is to be utilised, the feed-through capacitor C 51 is not connected to the chassis but to the PCB track leading to the trimmer.

If the alignment of the IF filter and demodulator circuits do not result in an improvement of the audio signal, try re-aligning the pre-amplifier. A correct alignment will result in the signals from both METEOSAT channels to have the same amount of noise behind them. This completes the alignment of the receiver.

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Compact METEOSAT Converter
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E. Chicken MBE, BSc MSc CEng FIEE, G 3 BIK

Stacked Loop-Yagi Antenna for METEOSAT Reception

1. INTRODUCTION

For those interested in receiving the weak 1.69 GHz signals from the METEOSAT Weather Observation satellite in geostationary orbit some 36000 km above the equator, the requirement for an antenna of suitable gain and directivity poses a problem.

It is not uncommon to read articles on the subject in which reference is made to the use of a parabolic dish antenna of 1.0 - 1.5 m diameter, which to say the least, is costly and unwieldy.

The antenna described here was designed and constructed to overcome the difficulties of the parabolic antenna, whilst providing sufficient gain to fully silence the FM receiver, and with a beamwidth adequate to minimise the possibility of interference from off-beam signals.

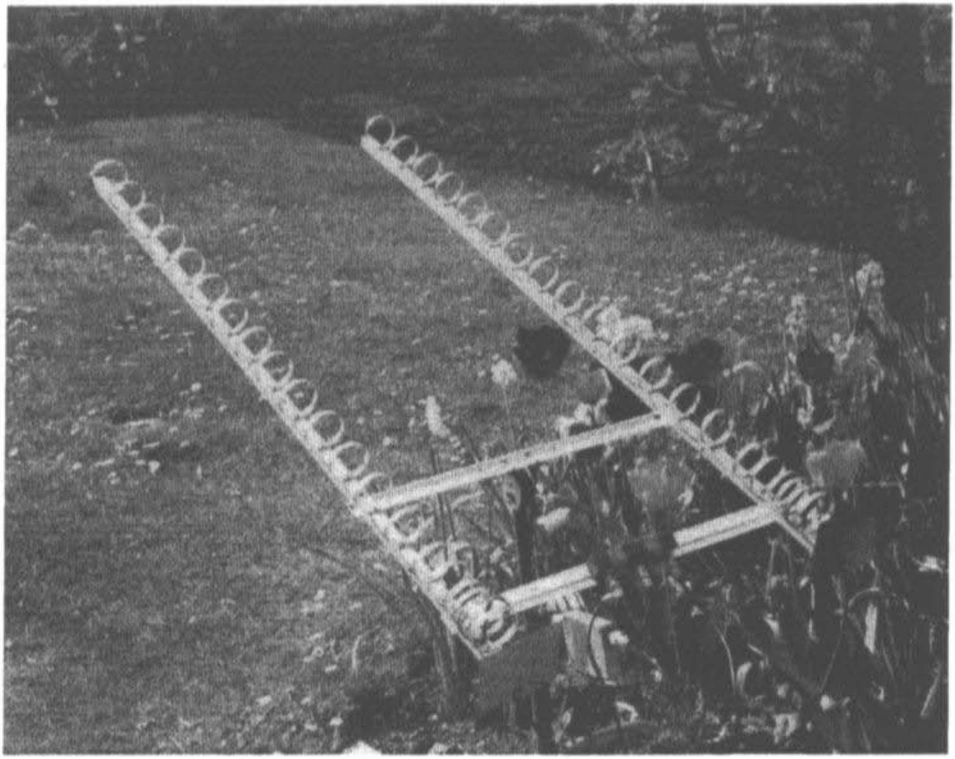
It was also intended to be suitable for indoor use by being light-weight and not unduly space-consuming, because the author is fortunate in having the use of an upstairs room with a window which gives a reasonably clear view to the METEOSAT horizon. The mechanical design allows the array to be tilted to the vertical plane when not in use, so that it can be stored out of the way in a hitherto unused space between the side of a bookcase and the wall.

No great claim is made for its all weather outdoor suitability, but it has on many a summer's day during its ten years of useful service been carried out into the garden to obtain a clearer view of the satellite's location, to overcome the severe absorption of the signal by an oak-tree when it is in full leaf. On such occasions it has withstood the wetness of summer storms without adverse performance, and its mechanical stability was adequately enhanced by the placement of two bricks upon its base. **See photograph.**

A wing nut provides convenient and stable adjustment of the required tilt angle, typically 25 - 30 degrees above the horizon.

The antenna array comprises two 21-element loop-Yagis, horizontally stacked with their signals combined for optimum gain and directivity.

Two correctly stacked identical Yagi antennae can offer a power gain of circa 3 dB over an individual Yagi.



Photograph: The author's stacked loop-Yagi

Whilst the two loop-Yagi antennae are stacked here to provide a realistic signal margin, each on its own is capable of yielding a reasonably usable signal.

Combining of the signals from the two Yagis is achieved by a simple-to-construct dual quarter-wave coaxial line transformer or its micro-strip-line counterpart, both of which are fully described in the text.

Design philosophy for loop-Yagi antennae is well documented by others (1, 2), and several articles have been published with constructional details for antennae of that type for the 23 cm and 13 cm amateur bands.

The paper (1) "Optimum Design of Yagi Array of Loops" by Shen and Raffoul was used as the basis for this design, and the final dimensions were cross-checked by linear-scaling from the published values (2) for a 23 cm antenna.

METEOSAT transmission frequencies in use at present are 1691.0 MHz and 1694.5 MHz, so this antenna is dimensioned on the mid-frequency of 1692.75 MHz, the wavelength of which is 177.23 mm.

2. CONSTRUCTIONAL DETAILS

Two identical loop Yagis are required, each with a total of 21 elements consisting of a reflector plate at the rear, a reflector loop, a loop dipole

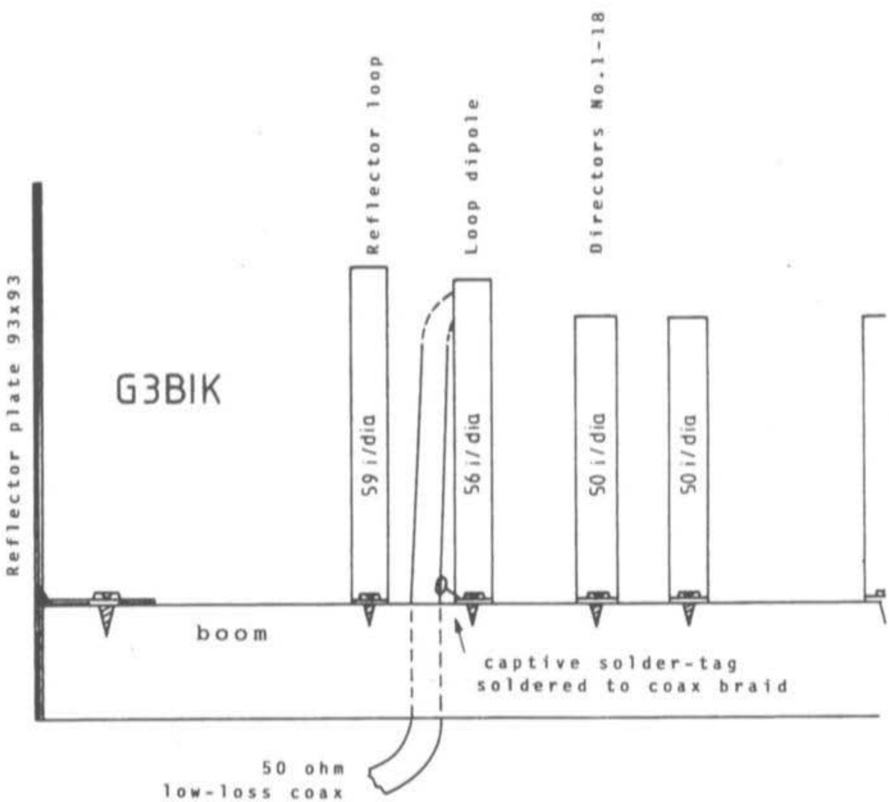


Fig. 1: Element fixing detail

of circumference equal to one wavelength, and 18 director loops **Figure 1**

The exact number of director loops is not critical, the quantity 18 having been adopted because they just happened to conveniently fit onto the available 1.2 m length of aluminium boom. A few more or less would not significantly affect the performance of the finished antenna.

2.1. Boom

For the boom, the ideal material would be a 1.2 m length of 20 mm square or round section aluminium tubing, such as is used for commercial TV antennae. Alternatively, a 20 mm diameter thin-walled copper pipe or thick-walled PVC pipe or conduit could be used.

Using a sharp pointed object, a line is scribed along the length of each boom, running exactly parallel to the sides to avoid spiralling.

Referring now to **table 1** and **fig. 2**, the element locations are marked at the distances shown, by very carefully measuring along the axial line on the boom from its rearmost end.

To avoid cumulative errors of meaurement, it is better to take the dimensions from table 1 and progressively measure from the end of the boom,

mm	Element	mm	Element
0	Reflector plate	459	Director loop no. 8
10	Reflector plate fixing tab	523	Director loop no. 9
59	Reflector loop	591	Director loop no. 10
77	Dipole loop	659	Director loop no. 11
99	Director loop no. 1	727	Director loop no. 12
115	Director loop no. 2	795	Director loop no. 13
149	Director loop no. 3	863	Director loop no. 14
183	Director loop no. 4	931	Director loop no. 15
251	Director loop no. 5	999	Director loop no. 16
319	Director loop no. 6	1067	Director loop no. 17
387	Director loop no. 7	1135	Director loop no. 18

Table 1: Element locations from rear end of the boom

rather than to measure and mark out each individual spacing as shown in **figure 2**.

With the aid of a centre pop, preferably a spring-loaded version, an indent is carefully made at each of the marked locations, then a 1 mm-diameter hole is drilled at each indent. The elements are not yet to be affixed.

2.2. Cross-Bracing

Two cross bracers and one diagonal bracer are used to provide mechanical rigidity for the finished array, with a support/tilt bracket to be mounted centrally on the underside of the forward cross bracer. (Material for the bracers can be either metal or hardwood.) If aluminium tubing is

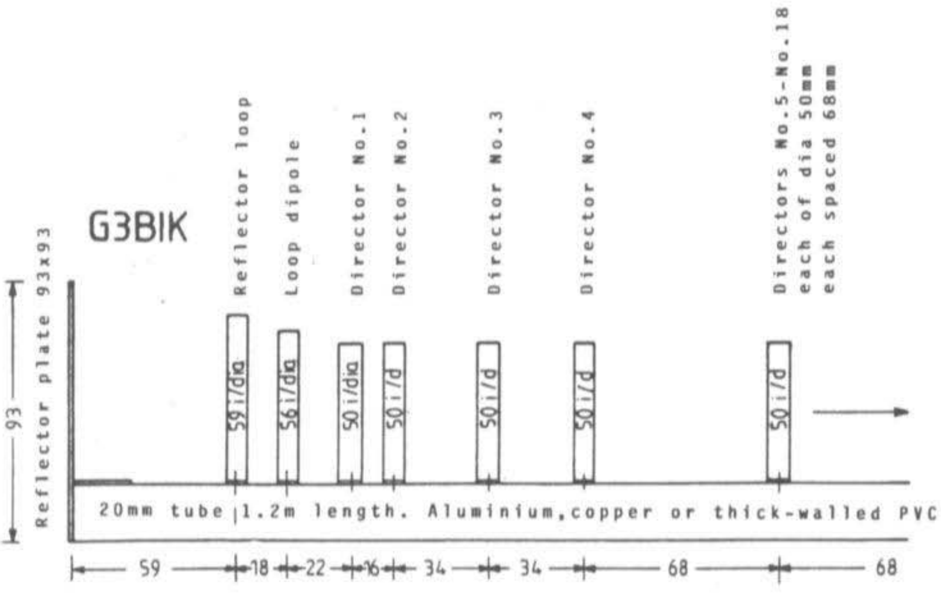


Fig. 2: Element spacings

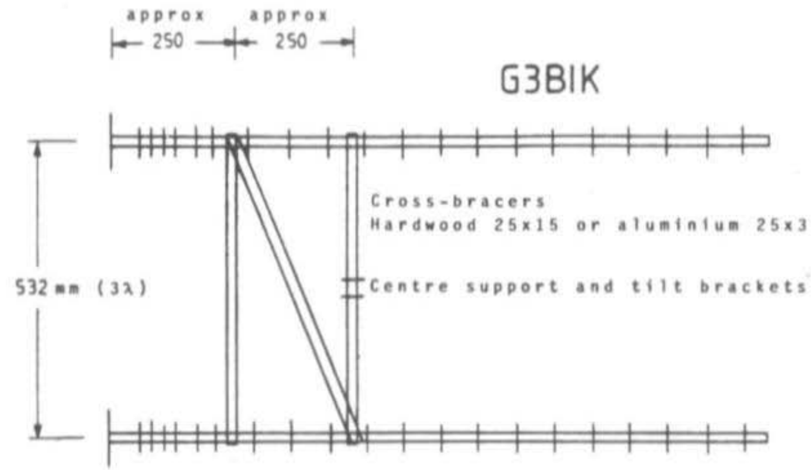


Fig. 3: Plan view

used for the booms, it might be convenient to use it also for the bracers, in which case the ends would need to be flattened for ease of fixing.

Bracers for the prototype were offcuts from aluminium thresh-plates of the type used to protect floor carpets where they meet at a doorway.

Although the cross bracers are shown at locations 250 mm and 500 mm from the rear end of the boom, some slight repositioning may be necessary in practice to avoid conflict with the installed elements.

The boom is marked at the chosen locations for the two cross bracers, and a 3.5 mm hole is drilled vertically through the boom at each marked location to accomodate the fixing bolts for the bracers.

Each of the two bracers is then marked to locate its fixing holes at a separation distance of 532 mm (= 3 wavelengths), and a 3.5 mm hole is drilled through the bracer at each marked location.

It is important that the hole-spacing is exactly equal on the two bracers, to ensure correct parallel stacking of the two Yagis, but the 532 mm separation distance between the Yagis is not critical to a few millimetres.

The two cross bracers are now bolted to the undersides of the two booms, using M3.5 x 30 mm bolts and nuts plus shakeproof washers.

After confirming an exact rectangular relationship between the booms and bracers, the diagonal bracer is marked and drilled at each end with hole-spacing to match the measured diagonal, then it is fitted to the booms using two of the now existing cross bracer retaining bolts, as shown in **figures 3 and 4**.

2.3. Tilt Bracket

The two right-angled brackets, as detailed in **fig. 5**, are made from a 100 mm x 100 mm piece of 12 swg aluminium plate similar to that to be used later for the reflector plate.

They are now to be mounted on the underside of the forward cross bracer as indicated in fig. 4, one on each side of the centre line to leave a space between them of 20 mm to accomodate the support tube.

2.4. Hole for Coaxial Cable

A hole is required to feed the coaxial cable vertically through each boom, such that it will

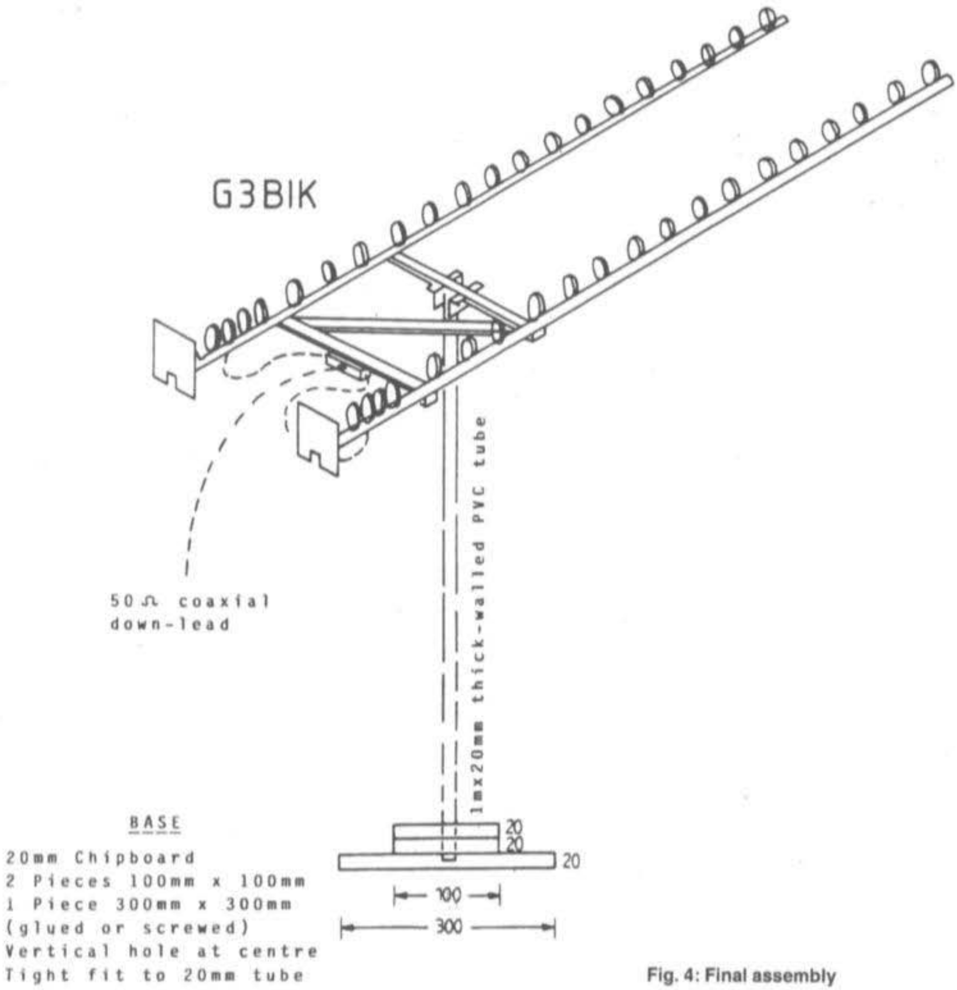


Fig. 4: Final assembly

emerge from the upper side of the boom immediately to the rear of the dipole loop.

The dipole is the second loop forward of the rear plate reflector, and it is located at 77 mm from the plate end of the boom. The boom is marked at 70 mm from the rear, or 7 mm to the rear of the dipole fixing hole, and a 6 mm hole is drilled vertically through the boom at that mark to accomodate later the coaxial cable.

2.5. Reflector Plate

Thin aluminium plate of about 12 swg is preferred, of dimensions 93 mm x 93 mm square, with two 20 mm vertical cuts in the lower edge which allow a 20 mm x 20 mm tab to be bent upwards at right angles to the plate (**fig. 6 refers**). This tab will be used later to fasten the plate to the rear end of the boom.

2.6. Reflector and Director Loops

With the exception of the dipole, all of the loops are formed from 7 mm x 1 mm aluminium strip, of the type used by electricians for clipping cables into place. Rolls of this PVC-covered material are available from electrical stockists, and the PVC is easily removed with a pocket knife. (Some 7 metres should suffice for both Yagis.)

The required lengths of strip, as detailed in table 2, are cut from the roll of aluminium strip, and are then formed into sensibly circular loops with the aid of a round-section bottle or similar container of outside diameter equal to the inner diameter required for the loop.

For convenience, all 36 director loops are of 50 mm inside diameter, and 59 mm for the two reflectors.

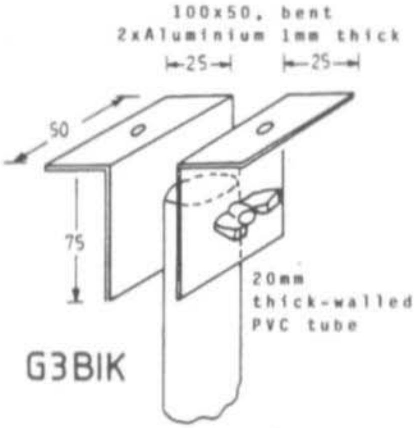


Fig. 5: Centre support and tilt detail for stacked Yagi

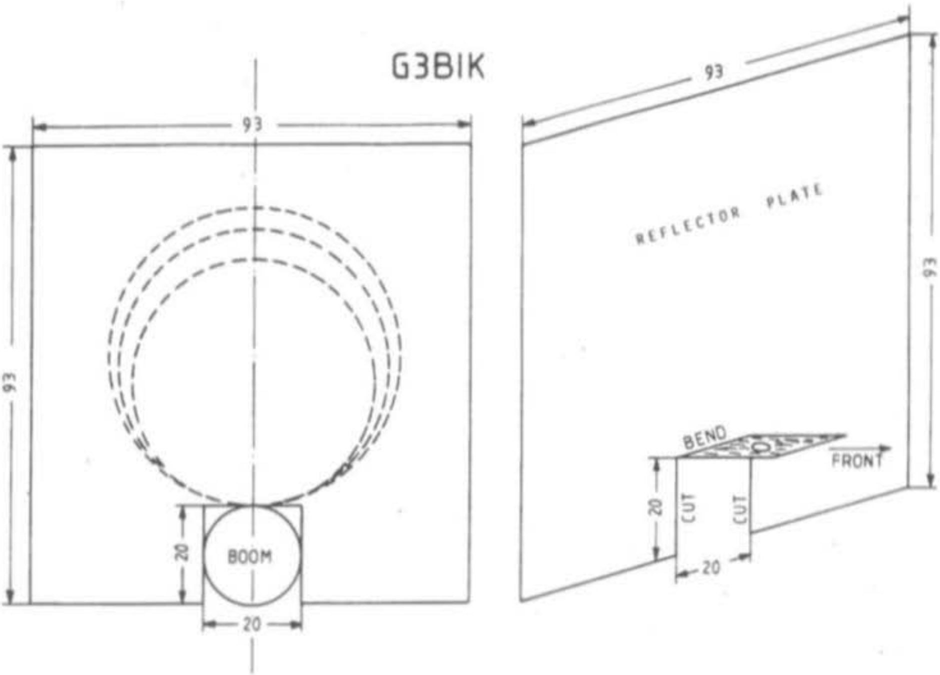


Fig. 6: Element alignment

Qty	Element	Cut length	Hole spacing	Inside diameter	Material
2	Reflector loop	195	185.4	59	7 x 1 aluminium strip
36	Director loop	167	157	50	7 x 1 aluminium strip
2	Dipole loop	177.2	------	56	7 x 1 copper or brass strip 1 hole at centre of each
2	Reflector plate	93 x 93			12 swg aluminium

Table 2: Element dimensions (in mm)

It would make good sense therefore, to search the house for medicine bottles etc of suitable diameter, before commencing the loop-shaping operation.

The three diameters required are 59 mm, 56 mm and 50 mm, for reflector, dipole and director respectively.

Reference to **table 2** will show that two length dimensions are given for each of the reflector and director elements. This is because the finished diameter of each element is related to the design frequency of the antenna, and the diameter is determined by the distance between the two fixing holes, one at each end of the strip.

This important distance between the holes is given in column 4 of table 2.

The length dimension given in column 3 of table 2 refers to the cut length of the strip before the holes are drilled, to allow an overlap of about 5 mm at each end of the strip to bring the two fixing holes into coincidence after the loop has been formed.

Each strip is carefully centre-popped at both ends to give the exact spacing required between holes, and a 2.5 mm hole is then drilled at both locations.

Thus, when the strip is formed into a loop of the correct internal diameter, the ends of the strip should overlap by some 5 mm, and the two holes should exactly coincide.

Ultimately, each reflector and director loop will be affixed to its boom by means of a small self-tapping screw, through the two overlapping holes.

2.7. Dipole Loop Element

This is of 7 mm x 1 mm brass or copper strip rather than aluminium, because it needs to accept the soldering to be described later.

The dipole loop is formed and affixed in a manner which is distinctly different from that of the reflector and director.

Column 3 of table 2 gives the cut length of the dipole. This length is also the length of the circumference of the finished loop, which is exactly one wavelength at the METEOSAT frequency. Unlike the reflector and director elements no overlap is required, because the dipole loop is affixed via a hole located at the exact centre of its strip length rather than at the ends of the strip.

The dipole loop when installed later will have its two free ends adjacent but unconnected so forming a small gap at the top centre of the loop, for subsequent connection of the coaxial feeder cable.

3. INSTALLING THE ELEMENTS

Each element is to be secured into place by one No. 4 x 6.5 mm pan-head self-tapping screw.

Beginning with the reflector plate, it is affixed by a self-tapper through the 20 mm square horizontal tab, so that the plate is vertical at what is to be rear end of the boom.

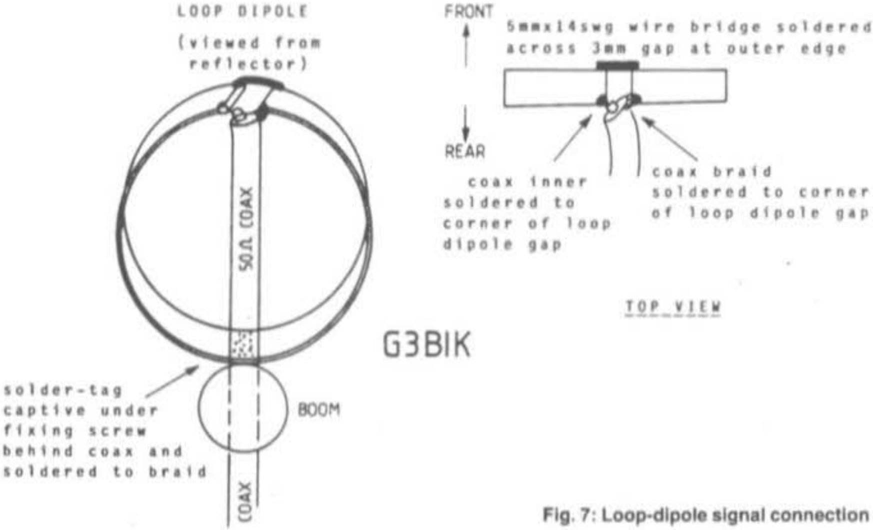


Fig. 7: Loop-dipole signal connection

The reflector loop is next installed, with a screw through the two holes previously drilled at the ends of the strip, to form a 59 mm inside-diameter loop in a plane vertical to the boom, and at right angles across it as indicated in figures 1 and 2.

Next in the forward line is the dipole loop, which is secured into position by one screw through its central hole, to provide a 56 mm inside-diameter loop parallel to the reflector loop, but with a 3 - 5 mm gap on the top of the loop opposite its fixing screw. **Figure 7** refers.

A 3.5 mm solder tag is fitted over the dipole fixing self-tapper before it is entered, with the tongue of the tag bent upwards and to the rear.

All of the 50 mm inside-diameter director loops are now installed progressively towards the forward end of the boom, in the same manner as was used for the reflector loop.

It may be noticed that fixing the elements directly onto the upper-side of the boom as described, means that the loops are not truly aligned through their centres along the axis of the boom, as illustrated in **figure 6**. In practice, this slight misalignment does not noticeably degrade the performance of the completed array, but if so desired, the inclusion of a nut of thickness 1.5 mm between the dipole and the boom, and similarly a nut of thickness 3.5 mm between each director and the boom, would provide an acceptable degree alignment.

3.1. Coaxial Cable Installation

The choice of coaxial cable presents a conflict of interests, because the microwave frequencies used by METEOSAT demand the use of coaxial cable with the lowest possible attenuation factor, but such cables tend to be of diameter greater than 10 mm which is somewhat bulky for this application.

A realistic compromise is adopted by the use of readily available URM43, 50 Ω coaxial cable of about 6 mm external diameter, and its length kept to a minimum.

Two exactly equal 500 mm lengths of the chosen coaxial cable are cut, and a 50 Ω BNC free-plug is carefully fitted to one end of each cable.

The actual length of the cable sections is not critical, but it is important that they be of identical length.

Before trimming the free ends, one cable is fed through the 6 mm hole from the underside of each boom, leaving enough free length protruding above the booms to allow easy access for trimming the cables.

The outer PVC covering is carefully trimmed back by 6 mm and the braid teased out until it fans neatly to one side, then the internal insulation is cut back by about 3 mm to expose 3 mm of the inner conductor.

Using the minimum of heat, the exposed braid and centre conductor of each cable are tinned with a soldering iron and resin-cored solder.

Before progressing the cable installation further, the free ends are temporarily fastened back away from the dipole loops, and the four corners of the gap at the top centre of each dipole loop, are individually tinned with the soldering iron.

A 5 mm length of 12 or 14 swg tinned copper wire is now soldered between the two corners of the 3 mm gap, on the forward face at the top centre of each dipole, to form an electrical bridge which, while mechanically stabilising the gap, also plays an important role in the impedance matching of antenna to coaxial feeder.

Each cable is now gently eased back through the hole until its previously trimmed free end is stationed immediately adjacent to the two rearmost corners of the gap in the dipole loop.

Again using the minimum of heat, the centre conductor is soldered to one of the two rear corners of the gap, and the braid to the other.

3.2. Balun

There is one school of thought which proposes that if a balanced antenna, such as the loop Yagi, is used in conjunction with an unbalanced feeder, such as coaxial cable, then its directivity pattern may be adversely distorted and the complete antenna system may be undesirably prone to interference from extraneous signals.

If that hypothesis were valid, the inclusion of a balance-to-unbalance or balun device between balanced antenna and unbalanced coaxial feeder, should overcome such a problem.

An effective balun can be produced quite simply by electrically connecting the outer braid of the coaxial feeder cable to ground potential, at a distance of one quarter wavelength from the point at which the braid is connected to the balanced antenna.

On the loop-Yagi antenna now being considered, a low-impedance location on each loop element is electrically connected to ground potential via its fixing screw to the metallic boom.

It would therefore be feasible to produce a balun simply by electrically connecting the braid of the coaxial cable to the mechanical fixing point of the dipole loop, but unfortunately, the diameter of the particular loop is 56 mm while the quarter wavelength of braid required for the balun is 44.3 mm.

Again there is a conflict of interests. By squeezing the dipole loop downwards slightly from the top, the braid length could be adjusted until the 44.3 mm point lies opposite the solder tag at the base of the loop, but in so doing the geometrical symmetry of the Yagi antenna is compromised.

In practice, making the electrical connection to the braid to form a balun on the prototype made no discernible difference to the received signal, which confirms the author's considerable experience with commercially produced folded-dipole Yagis for radiocommunications at 1.5 GHz and TV antennae at half that frequency, none of which used a balun.

The decision on whether or not to form a balun by soldering the tag to the braid becomes one for the experimentalist.

Each individual Yagi antenna is now complete and functional, but their coaxial feed cables must be combined in a manner such that the two received signals are truly additive, and the characteristic impedance of the conjoined cables matches that of the 50 Ω down-feeder cable.

Two alternative designs are given for the signal combiner.

a: Signal Combining Unit, Trough-line Version

Assuming the impedance at the feed point of a loop-Yagi antenna to be 50 Ω, and that it is correctly translated to the distal end of the 500 mm length of 50 Ω coaxial feeder, the physical act of connecting in parallel the two 50 Ω cables would in effect reduce the characteristic impedance at the point of connection to 25 Ω.

That would present a problem of matching to the 50 Ω input impedance of most modern communications receiving equipment, so the 25 Ω must be transformed to 50 Ω before the received signal is fed to the 50 Ω coaxial cable which connects ultimately to the preamplifier of the receiving system.

To obtain an impedance of 50 Ω at the junction of two coaxial cables in parallel, would require that the impedance of each of the two was 100 Ω, whereas the impedance of the individual 500 mm sections of coaxial cable coming from the Yagis is 50 Ω.

This means therefore, that the 50 Ω of each cable must first be transformed into 100 Ω before the two are connected in parallel, to reobtain a 50 Ω impedance at their point of connection.

Such impedance transformation may be conveniently effected by means of a quarter-wave transformer, which is simply a short section of coaxial-line one quarter of a wavelength long at the frequency concerned, connected in series with the coaxial cable whose impedance is to be transformed.

A quarter-wave transformer section is required in each of the two Yagi coaxial feeder tails, and for mechanical convenience is located at the BNC-plug end of the tail.

The characteristic impedance $Z_{\lambda/4}$ of the transformation line section, is different from the impedances at either end of the transformation, and is given by the expression:–

$Z_{\lambda/4} = \sqrt{Za \times Zb}$

where Za is the impedance of the Yagi's coaxial tail = 50 Ω

and Zb is the impedance required after transformation = 100 Ω

So, $Z_{\lambda/4} = \sqrt{50 \times 100} = \sqrt{5000} = 70.7\ \Omega$

For mechanical convenience, an air-spaced coaxial line is used for the quarter-wave transformer, the dimensions for which are derived from the expression:

$Z = 138 \log_{10} (D/d)$ for an air-spaced coaxial line of circular section

where D is the internal diameter of the outer line and d is the external diameter of the inner line.

From the constructor's point of view it is even more convenient to use coaxial line of square outer section in preference to circular for this particular application.

A square of side dimension S has an equivalent circle of diameter D = (1.128 x S) which converts the expression for Z to:–

$Z = 138 \log_{10} (1.128\ S/d)$

for an air-spaced coaxial line of square section
where S = side of square
and d = diameter of inner line
and $Z = Z_{\lambda/4} = 70.7\ \Omega$
now for convenience let
d = 6.35 mm = (0.25 inch)

So $Z = 70.7\ \Omega = 138 \log_{10} (1.128\ S/6.35)$

from which:–

$S_{mm} = (6.35/1.128) \times \text{antilog}_{10} (70.7/138) =$
= **18.3 mm** inside square

Therefore, the dimensions for a square-sided air-spaced coaxial line quarter-wave transformer suitable for converting the 50 Ω Yagi coaxial tail to 100 Ω, are:–

Trough of square section inner dimension:

	S = 18.3 mm
Inner line of diameter	d = 6.35 mm
Inner line length	L = 44.3 mm

The requirement is of course for two such quarter-wave transformer assemblies, one for each Yagi coaxial feeder tail, and their 100 Ω ends are to be electrically connected in parallel.

Figure 8 illustrates the mechanical arrangement by which the parallel connection is achieved, by effectively joining the two quarter-wave assemblies end to end, to produce what seems like a half-wave section of air-spaced 70.7 Ω line, with one signal feeding into each end, and

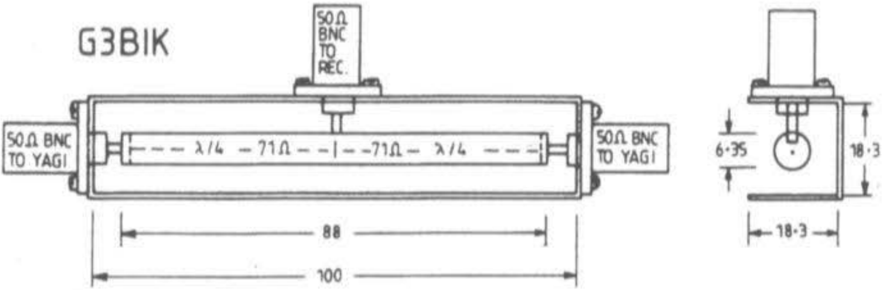


Fig. 8: Signal combiner dual trough-line 71 Ω λ/4 transformer

the combined output signal taken from the centre of the composite line.

The signal-combining assembly is made from light-gauge metal, cut and shaped to form an open-sided trough of internal depth 18.3 mm, and internal length 100 mm, with a 6.35 mm x 88.0 mm brass or copper tube centrally located within the trough. It should be noted that the theoretical 88.6 mm has been reduced slightly to allow for end effects.

Copper or brass would allow the corner edges of the trough to be soldered for mechanical strength, but aluminium would serve equally well given that fold-over tabs were provided at the end faces to accomodate small fixing screws.

Whether or not a lid is provided for the open-sided trough is a matter of personal choice. In either case it makes but little practical difference to the finished characteristic impedance (3).

The trough is now drilled to accomodate the three BNC flanged sockets, which are to be installed later.

A 1.5 mm hole is drilled radially through and at the centre of the 6.35 mm x 88.0 mm copper or brass tube which is to be used as the inner line.

Both ends of the tubing are to be partially sealed with a small brass nut of diameter chosen to make it a tight fit into the open end of the tube, leaving a small axial hole at the centre of each end face.

The two nuts are soldered into position, and their central holes drilled, if necessary, to provide a 1.5 mm clearance for the centre pin of each BNC end socket.

The central socket is now installed and the inner-line tubing offered to it such that the centre pin of the socket engages with the radial hole at the mid point of the line.

At each end of the trough, a 50 Ω BNC flanged socket is affixed into place with its centre pin entered into the hole at the end of the inner-line, after which all three socket pins are soldered to the line.

b: Signal Combining Unit, Microstripline Version

This design for a signal combiner is simpler to construct than the trough-line version described earlier, but is equally effective.

It is based upon the use of microstripline technology and 1.6 mm low-cost glass-fibre double-clad copper circuit board, which behaves sensibly enough even at METEOSAT frequencies.

The complex design formulae (4) and calculations are deliberately omitted from this text, but the design philosophy and calculated dimensions are included.

As with the previous version, a quarter-wave transformer principle is employed to produce the required impedance transformation, albeit in a different manner.

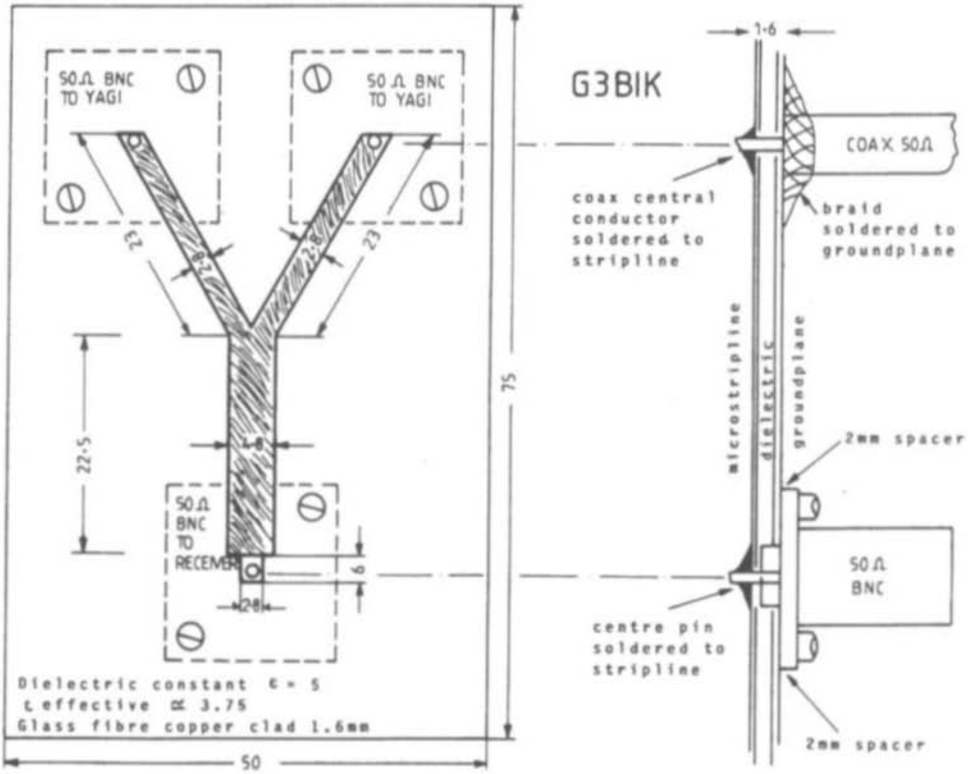


Fig. 9: Microstripline signal combiner

In this design, the distal ends of the two Yagi 50 Ω coaxial feeder tails are connected directly in parallel, to present an impedance of 25 Ω at their junction.

To match that 25 Ω to the 50 Ω coaxial main down-lead, a quarter-wave transformer section of microstripline is connected in series between the 25 Ω junction and the 50 Ω coaxial down-feeder.

The characteristic impedance required for the quarter-wave transformer microstripline section, is given by:–

$$Z_{\lambda/4} = \sqrt{Z_j \times Z_m}$$

where Z_j = Impedance at junction = 25 Ω and Z_m = Impedance of main line = 50 Ω

$$Z_{\lambda/4} = \sqrt{25 \times 50\ \Omega} = \mathbf{35.35\ \Omega}$$

Figure 9 illustrates the finished arrangement for the microstripline circuit which is to be etched onto one side of the double-clad copper board, leaving the copper on the opposite side of the board intact, to serve as the ground plane which forms an essential part of microstripline circuits.

The circuit comprises two short sections of 50 Ω stripline electrically connected together at one end, and that junction connected to a stripline section of 35.35 Ω characteristic impedance.

The two 50 Ω sections connect to the Yagi coaxial feeder tails, and the 50 Ω main down-feeder cable connects to the free end of the 35 Ω section.

Actually, the physical length of the two 50 Ω sections of microstripline is not critical although they must be of equal length, but care must be taken to ensure correctness of the width of all three sections of stripline, because it is the width which determines the characteristic impedance.

Of critical importance are the width and length of the 35 Ω microstripline section. The width, as already noted, not only determines the characteristic impedance of the section, but it is that impedance value which determines the impedance-transformation property of the quarter-wave transformer.

A physical peculiarity of microstriplines is that their electrical length, when expressed as fractional wavelength, differs from the theoretical free-space wavelength because the velocity of the microwave signal, as it travels along the stripline, is modified by the physical constants of the material.

The effect of this particular glass-fibre material is that the quarter-wave length required for the 35 Ω section, is reduced from the theoretical value of 44.3 mm to the much shorter 22.5 mm.

For convenience of attachment for the 50 Ω mainline, a very short section of 50 Ω microstripline has been included at its point of connection to the 35 Ω section.

Figure 9 also shows how the three BNC plugs and sockets may be omitted and the coaxial cables soldered directly to the microstripline signal-combiner printed circuit board, but from the experimental aspect it is more convenient to retain the plugs and sockets.

4. CONCLUSION

The signal cables from the two loop Yagis are now plugged into the end sockets of the signal-combiner, and the combined output signal is available from the middle socket.

Finally, the signal-combiner unit is strapped centrally to the rear-most cross bracer by means of plastic ratchet-type cable ties, then the support rod is bolted to the tilt brackets and its lower end inserted tightly into the base stand.

The antenna is now ready for use.

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Foundations of Microstrip Circuit Design

Günther Borchert, DF 5 FC

Universal Synthesizer for Frequencies up to and above 1000 MHz Part 1

The following article describes a frequency synthesizer circuit whose hardware must be complimented by appropriate software in order that it can be conditioned to fulfil a variety of functions. The highest frequency, for the time being, is about 1 GHz but the further development of the pre-scaler section can lead to the extension of this limit to higher realms. Owing to the employment of the most modern integrated technology, the performance can be achieved with the minimum of power consumption thus making it very suitable for portable work. Besides the digital portion, two differing oscillator circuits will be considered which are optimised for the frequency in use.

1. CIRCUIT CONCEPT

1.1. General

The basic idea of a PLL synthesizer oscillator is to produce a multitude of frequencies with a defined frequency separation by means of a main oscillator and a reference oscillator, utilising digital techniques to achieve this aim. The theoretical advantage with this method is that no mixer or filters are necessary in the signal path and therefore the production of undesired spurious mixing products is avoided.

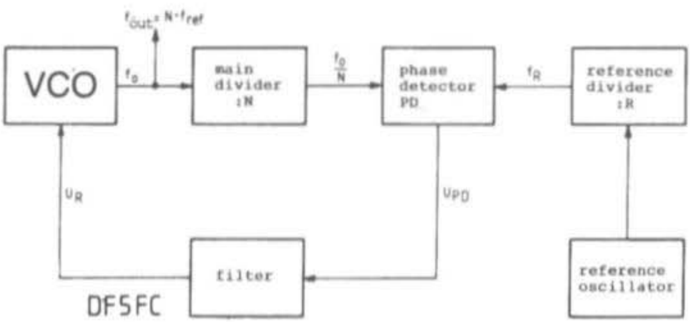


Fig. 1: Basic block schematic of a PLL synthesizer

The output frequency of the Voltage Controlled Oscillator (VCO) is brought to the reference frequency f_r via a programmable divider N as shown in **fig. 1**. This reference frequency represents the frequency separation of all juxtaposition signals which form the output "comb" of frequencies – or channels. The divider R divides down the output frequency of a crystal-controlled reference oscillator to the value of that possessed by the separation (step) frequency – this is again, the reference frequency f_{ref}. The crystal oscillator determines the frequency stability of all the frequencies available in the output. The two derived frequencies are then compared in a frequency and phase detector. This delivers an output pulse train which varies in accordance to which of the two comparison frequencies is the highest. This pulse train is then filtered and the integrated output, which forms the tuning voltage, then controls the VCO. The circuit thereby causes the two inputs to the phase detector to lock together and maintains them in this condition by correcting the VCO as required. When now, the divider N is required to shift to a one higher increment in frequency, the frequency at the PD at that instant, is lower than that demanded. The VCO frequency is then raised until the original conditions of frequency balance are achieved. The output frequency then moves up to the new channel. The whole process is effectively the same, but reversed, when the value of N is required to be smaller.

The technical limits of this process are set by the programmable divider. Because of their internal transit times, they have a relatively low, higher frequency limit. With ECL chips, this limit lies around 60 to 80 MHz and with modern C-MOS ICs, 15 to 20 MHz can be achieved. The direct use of any ICs for this purpose is not possible at 145 MHz. The limit for highly integrated dividers in both MOS or in TTL is not very much higher. For all higher frequency applications the VCO frequency must be brought down to the frequency of the programmable divider. There are three techniques for doing this:

- Mixing
- PLL with a fixed pre-scaler
- PLL with a switchable pre-scaler (swallow or modulus divider)

The first two techniques will now be considered briefly together with their relative advantages and disadvantages. The third will be described in some detail.

1.2. The Mixing Synthesizer

The block diagram of **fig. 2** differs from fig. 1 owing to the addition of a mixer and a local oscillator. The local oscillator works at a frequency which is 5 to 10 MHz lower than the VCO, depending on the type of divider in use. It is crystal controlled and could also contain a multiplier.

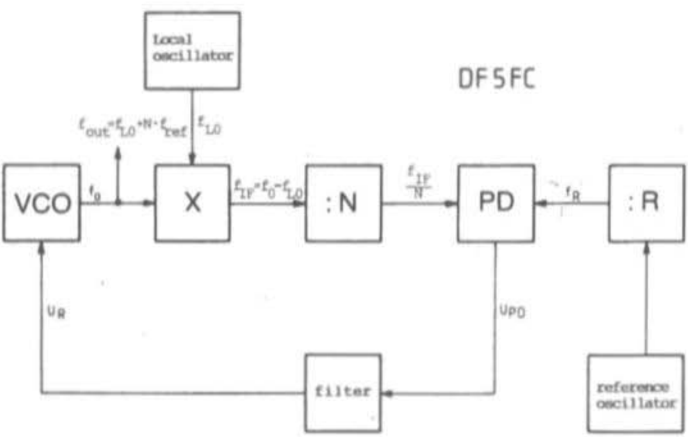


Fig. 2: Synthesizer with mixer and local oscillator

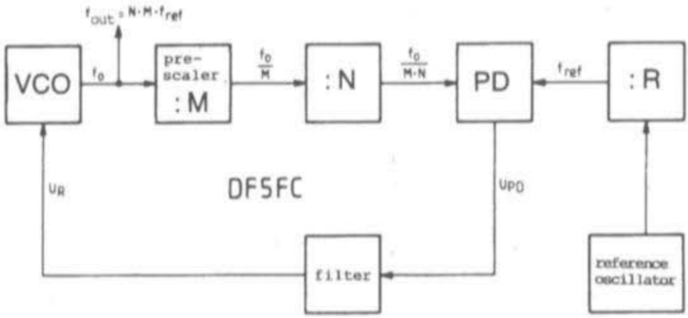


Fig. 3: Synthesizer with a fixed pre-scaler

The advantage of this circuit is that, despite a small step frequency, a signal with a relatively low phase-noise can be obtained. This, of course, presumes that all oscillators are constructed in accordance with good practice and that the rest of the circuit is very carefully dimensioned. Furthermore, a quasi-continuous tuning can be realised, if the local oscillator can also be tuned over a small range. A large proportion of modern radio equipment, having continuous tuning, works using this principle. It does require a certain complexity in the tuning circuits but this presents no difficulties for modern micro-processors.

The disadvantage of this circuit is the complexity of the circuitry which is required. Looking at circuits which were published some time ago in VHF COMMUNICATIONS, for example "Süd-wind" by DJ8IL in (1) or the 2-metre synthesizer by DC1QW in (2), it can be seen that as well as the actual digital part, which did not need alignment, the local oscillator multiplier and mixer stages all were present. This sort of circuit technique is always prone to the risk of spurious outputs and false locking of the PLL.

The use of the mixer limits the tuning range of the circuit. Theoretically, band segments are received whose range extends from zero Herz up to the maximum working frequency of the dividers. When a broadband equipment is required, several down-conversions (with their oscillators) are necessary. Another problem occurs when the local oscillator is brought too near the working frequency inasmuch that the PLL locks to the image frequency.

This would mean that for a universal synthesizer, a different mixer crystal must be employed for each application. This also entails a different dimensioning of the multiplier for each case. This is a much too high a price to pay in terms of complexity and on these grounds the technique was not considered further. Another disadvantage is that the LO is normally very close to the output frequency. If the whole unit is not very carefully constructed, a large component of the oscillator power appears in the output. Filtering out the unwanted LO power is difficult owing to the close proximity of the wanted signal.

1.3. Synthesizer with Fixed Pre-Scaler

This circuit obviates the use of local oscillators as the diagram of **fig. 3** shows. The circuit complexity remains within bounds and the actual circuit has no alignment stages to bother with. The overwhelming disadvantage lies in the use of the fixed pre-scaler. The inherent problem is that the size of the incremental steps cannot be constant as they are determined by N * M where M is the dividing factor of the pre-scaler. In order to obtain the required constant steps in frequency, the reference frequency must also be divided by M in sympathy. But this means that the reference frequency must drop which has a deleterious effect upon the phase loop (see filter calculation in appendix). Owing to the low reference frequency, the loop filter must also be designed for a low limit frequency which entails a lengthening of the time required for the lock to occur. This is unacceptably long when, in operation, large frequency spans are switched and the control

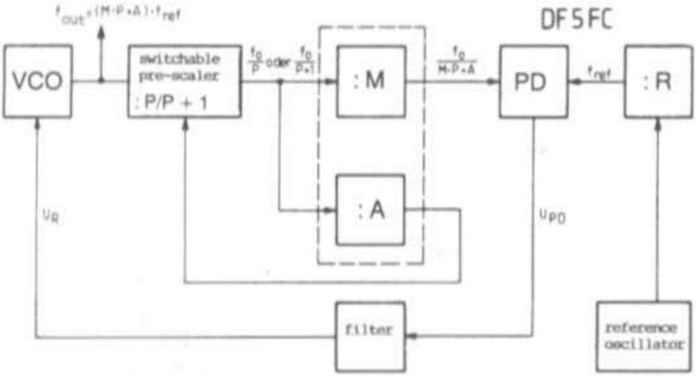


Fig. 4: Synthesizer with a dual-modulus pre-scaler

voltage is hunting towards the selected frequency. It is tolerable only for fixed channel equipment.

A low-frequency loop filter also tends to inhibit the property of the loop to phase out spurious signals. Also, the reference frequency tends to appear in the loop control voltage which causes a spectral comb of frequencies to occur in the output of the VCO. This problem is, of course, inherent in all synthesizers but it is exacerbated when the loop frequency is low. Circuits of this type have been described by Helpert (3) in his "Gartenzwerge" transceiver article.

This form of circuit is very suitable when the required step frequency is to be very large as, for example, in a converter local oscillator. In a 23 cm transverter, the local oscillator could have frequency incremental steps of 2 MHz. This would entail a reference frequency of 31.25 kHz – even if a divider scaling factor of 64 is used. The technique is also suitable for employment as a local oscillator for a down-converter synthesizer, but the frequency planning must be given due consideration (4). The output signal contains no spurs as only one oscillator is working. It is not suitable, however, for the planned universal synthesizer as the reference frequency must sink to an unacceptably low value.

1.4. Synthesizer with a Variable Pre-Scaler (Dual Modulus Divider)

In this type of circuit (**fig. 4**), a pre-scaler having a variable divider precedes the "N"-divider. The division factor can be switched over by a control logic from a programmable divider. This has the effect that the intended step increments are obtained despite the pre-scaling and without changing the reference frequency. The loop is easier to arrange and does not carry the disad-

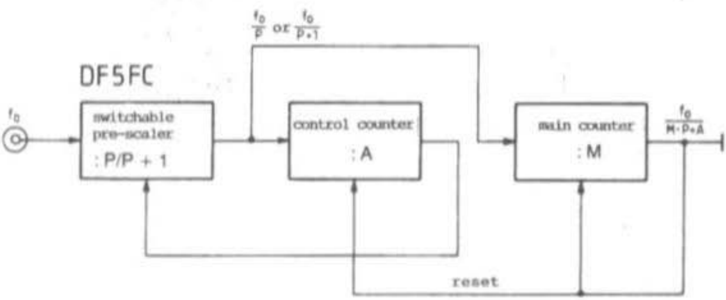


Fig. 5: Principle of dual-modulus pre-scaler

vantages of a fixed pre-scaler. Simply programming the main divider allows a very large frequency range to be spanned, making this circuit variant seems a suitable candidate for the universal synthesizer. Because of the absence of an LO/mixer there are no tuning points. Continuous tuning, within certain limits, is possible by making the reference frequency tunable.

At first sight, this type of frequency processing seems rather complicate. If the complete divider block is closely examined and the sequences taken separately, there is nothing terribly difficult about it. The diagram of **fig. 5** will help to explain its function.

The following conditions are assumed to begin the functional explanation:

1. The pre-scaler divides by factors of P and P + 1, switchable via the modulus control.
2. The dividers A and M are down-wards counters which, together with a reset pulse, are set to the pre-determined value, both of which are **not allowed** to underflow zero.
3. The A-counter is **always** set to a smaller number than the M-counter which means that is **always** the first one to reach zero.

It is assumed that both counters have just been set to their pre-determined value and that the counting cycle has started afresh. It is now necessary to consider how many VCO oscillations are necessary in order to obtain this condition again. This number will then correspond to the scaling factor of the circuit.

To start with, the pre-scaler is switched to P + 1. Since the A and the M counters have been switched in parallel, each time after P + 1 oscillations of the VCO by an impulse from the pre-scaler, the counter condition is stepped one lower. If, for example, the A counter has been set to 7, it has reached zero after 7 * (P + 1) (or in general after A * (P + 1) oscillations, with A as the pre-determined value. Because they are connected in parallel, the M counter has also decreased by a count of 7 (or in general, by A), its actual value is now M-7 (M-A). The A counter, upon reaching zero, initiates a signal which switches the pre-scaler over from P + 1 to P. The A counter locks itself so that it now works only in conjunction with the M counter. Now, each time after P oscillations of the VCO it is decreased by 1 until it also reaches zero i.e. (M-A) * P. The output of the counter controls, on the one hand, the phase discriminator, and, on the other hand, the reset logic. Upon the M counter reaching the count of zero, both counters are reset back to their original condition and the whole cycle starts all over again.

The total number of oscillations which were undertaken by the VCO in order to run through this sequence of events, is a consequence of the sum of the A counter A * (P + 1) and the **remainder** sum of the M counter (M-A) * P, that means:

$$\begin{aligned} N_{tot} &= A\,(P+1) + (M\text{-}A) * P \\ &= A * P + A + M * P - A * P \\ &= A + M * P \end{aligned} \qquad (1)$$

In order to clarify this, an actual numerical example is considered.

P = 40, i.e. P + 1 = 41, (determined by the type of pre-scaler).

The M counter is pre-adjusted to the figure 362 (M = 362).

The A counter is pre-adjusted to the figure 20, (A = 20).

This means: that the VCO signal is divided 20 times by a factor of 41 and then 342 times (M-A) by 40. This results in a total divisor of:

$$N_{tot} = 20 + 40 * 362 = 14500$$

This count is appropriate to a 2 metre transmitter which is tunable in 10 kHz increments:

$$N_{tot} * f_{ref} = f_{VCO}$$

When considering this procedure, the following limitations must be mentioned. The most important one is that imposed by the pre-scaler. This has a lower limit, beneath which, frequencies are completely unobtainable. If the formula of (1) is resolved, the result will be obtained for when P = 40 (typical value for these pre-scalers):

$$N = 40 * M + A$$

Now it may be imagined that M is a decade value (better, a multiplier by 40) and A is in units (A runs from 0 to 39). Under the conditions initially mentioned, that M must always be greater than A, results in the fact that M must also be at least equal to 40. When M is smaller, the reset condition is attained (counter A and M are connected in parallel, see fig. 5), before P + 1 is switched over to P. In this manner, the units presented by the A counter are never reached and the process goes no further. The minimum divisor of M must therefore be:

$$N_{min.} = 40 * 40 = 1600 \text{ or in general } N_{min.} = (P)^2 \qquad (2)$$

This condition is important if the pre-scaler has a large P, because high frequencies must then be processed. The Plessey NJ8820 used here (to be considered later in more detail) has a specified worst-case working frequency of 10.6 MHz. This value should not be exceeded bacause it causes interference to the internal pulse sequencing (the internal gate transit-times limit the scaling, depending upon the adjusted numerical value of the divisor). This results in frequencies being generated other than those which have been pre-determined (e.g. some channels more or less). For a 1.3 GHz VCO a pre-scaler with a divisor of, for example 128/129 should be chosen. The maximum input frequency to the PLL chip is then approximately 10.2 MHz which is just about admissible. The minimum divisor is therefore:

$$N_{min.} = (128)^2 = 16384$$

For a required frequency incremental step of say 10 kHz the lower limit would be: 16384 * 10 kHz = 163.84 MHz. This arrangement would not be very suitable for a 2-metre radio. A further limitation is occasioned by the PLL chip itself. The maximum value for M is here 1023 and for A is 127 (for NJ8820). The dividing process treated here would result in the evaluation of the division factor for a frequency in the following manner:

$$\text{Maximum working frequency/step} = N_{max.} \qquad (3)$$

$$\text{Value for M counter: } M = N_{max.}/P \qquad (4)$$

From this result the whole number before the comma is directed to the M counter. The value after the comma is again multiplied by P, the resulting whole figure is the value for the A counter.

For example:
Maximum working frequency, 1300 MHz
Channel step, 10 kHz:

$$1300 \text{ MHz}/10 \text{ kHz} = 130000 = N_{max.}$$

$$M = 130000/128 = 1015.625$$
$$M = 1015$$

$$A = 0.625 * 128 = 80$$

This example makes it clear that the limit for M is rather extended. If, using this calculation, the upper limit is reached, then another step must be chosen in order that $N_{max.}$ is reduced; e.g. 25 kHz – or another scaling factor may be required.

The disadvantage of this PLL technique is that modulation of the loop control-voltage with the switching frequency of the pre-scaler takes place. Because of the constantly changing pre-scaler divisor, the pulse-length at the input of the phase detector also changes. This superimposed frequency is very low making it difficult to be filtered out in the loop. It manifests itself as an audible rumble in SSB receivers. This indicates that this narrow-band concept for the synthesizer is only usable within certain limits. Owing to the use of the double phase-detector and by careful construction, this feed-through can be kept to low limits. In commercial practice (signal generators up to 1000 MHz from almost all well-known manufacturers), this effect is minimised by a judicious amount of a counter voltage. This is possible since the superimposed frequency is derived directly from the division factor and can therefore be used for compensation. Th required complexity is extensive and would be much too much for any home-brewed amateur equipment.

To be continued

Günter Sattler, DJ 4 LB

A Universal Sound-Vision Unit for FM-ATV Transmitters

The unit to be described displays characteristics which justify the claim that it is truly universal.

The composite signal comprises the processed video signals and the frequency-modulated sound carrier. The so-called base-band signal is available at a high level (4 V_{pp}) at the 75 Ω coaxial output. As any video signal at this impedance, it can be taken via any length of coaxial cable (**figs. 1a and 1b**). This module does not have to be in the vicinity of the modulated GHz oscillators (or Gunnplexers) and it can, moreover, feed two ATV transmitters simultaneously (**fig. 1c**).

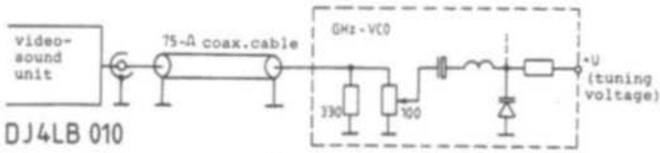


Fig. 1a: VCO level adjustment

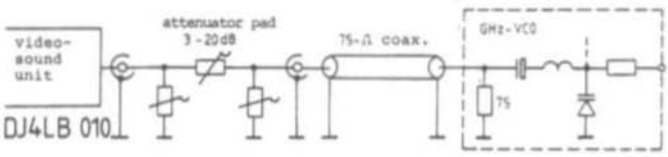


Fig. 1b: Level adjustment using a 75 Ω pad at the video/sound output

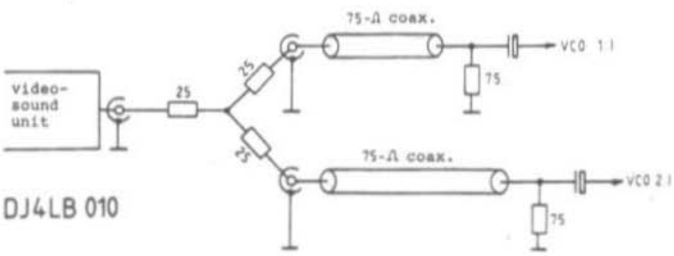


Fig. 1c: Using two VCOs at a video/sound output

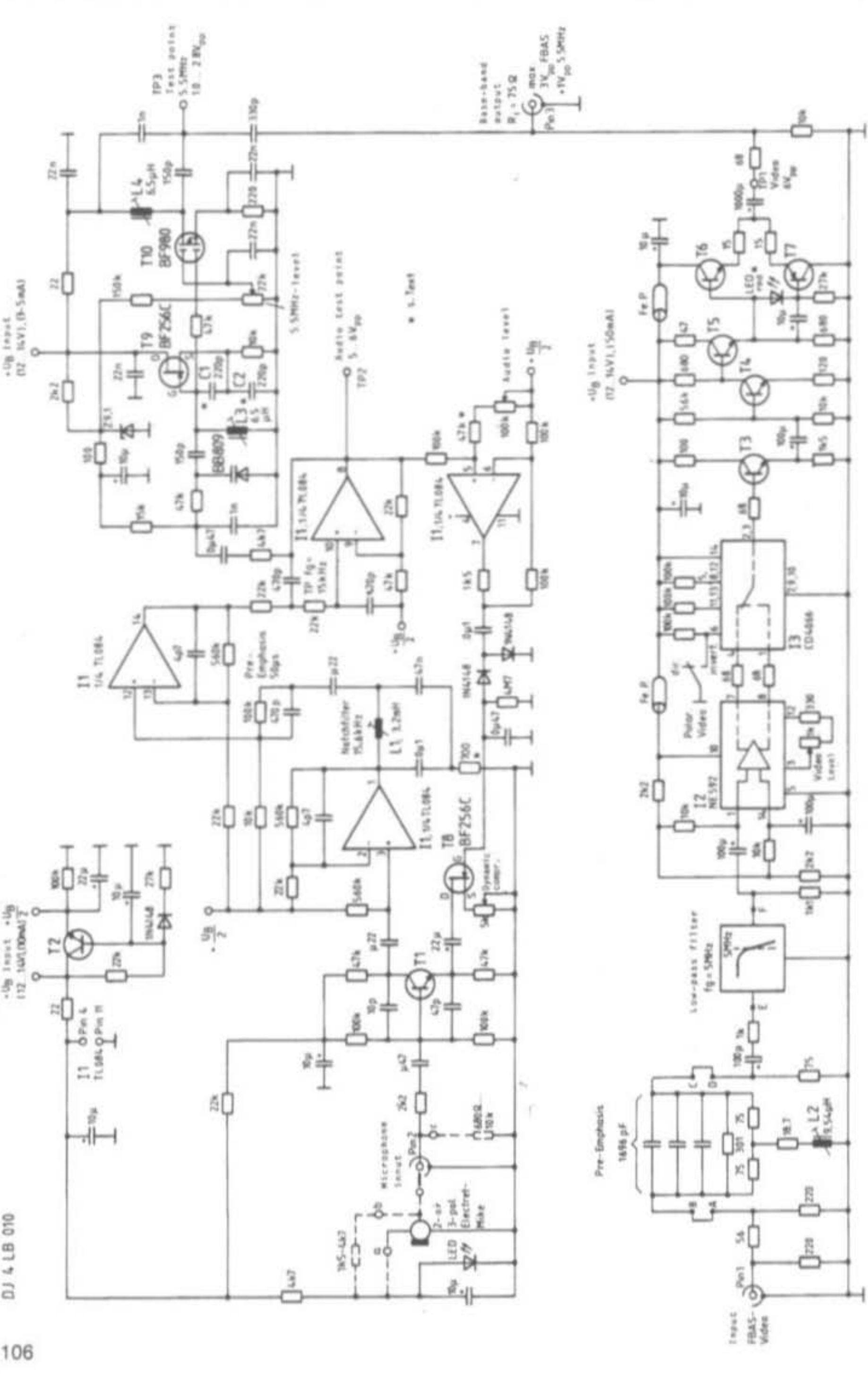


Fig. 2: Complete circuit schematic of the FM-ATV transmitter universal sound/vision unit

The polarity of the processed video signal is reversible. This facility may be necessary under three circumstances as explained in (1).

The sound-carrier circuit is suitable for all frequencies which are used in both amateur- and satellite television (approx. 5.5 to 7.5 MHz).

This unit is self-contained inasmuch that all the components necessary for the base-band processing, including a supply voltage for electret microphones, and a video block filter are incorporated. Thus, no additional modules should be necessary.

The relatively high sophistication not only ensures a high sound and vision quality but also serves to limit the modulated spectrum to that which is necessary compatible with the fidelity offered.

1. CIRCUIT DESCRIPTION

The complete circuit of the universal sound-vision unit may be seen in **fig. 2**.

1.1. Video Processing

1.1.1. Video Pre-Emphasis

A 6 dB pad precedes the pre-emphasis network in the signal path from pin 1. The PE-network attenuates signals lower than 100 kHz by some 14 dB (2) whilst allowing signals of greater than approximately 3 MHz to pass without attenuation. Both BAS and FBAS signals contain relatively large components at frequencies over 3 MHz which remain largely unaffected by the PE-network but the form of the composite signal has been clearly changed (**see fig. 3a**). It will be seen that the video test-signal has encroached deep into the synchronizing pulse area and exceeds the amplitude of the sync-pulse. This would preclude satisfactory synchronization in any monitor! The signal relative amplitudes must therefore be restored in the demodulator with a de-emphasis network (2). If the latter were to be experimentally applied to the output of the video-amplifier at pin 3, the original signal form is then restored, as in **fig. 3b**.

1.1.2. Video Low-Pass Filter

Experience has shown that many amateur video signals are transmitted with components well in excess of 5 MHz. This is not normally to increase the resolution (exception: character-generators etc) but the result of unnecessarily abrupt sync- and blanking pulses. Video signals components which are higher in frequency than

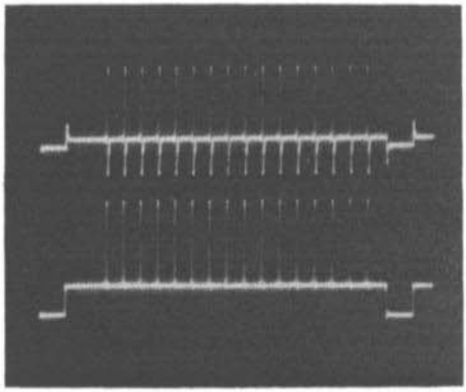


Fig. 3a: Line frequency: before pre-emphasis (below), after pre-emphasis (above) using same unit scale

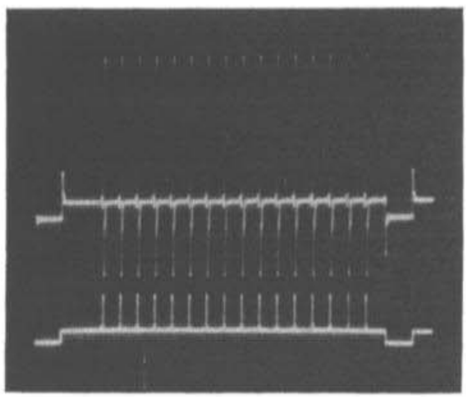


Fig. 3b: Line Frequency: before de-emphasis (above), after de-emphasis (below) using same unit scale

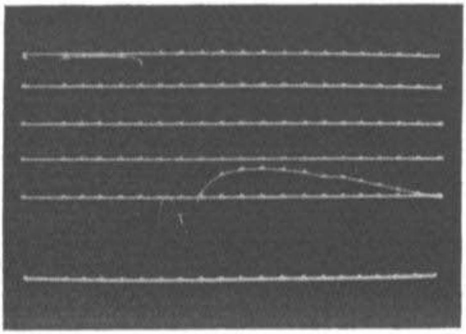


Fig. 4: 0 - 20 MHz 1218 block filter

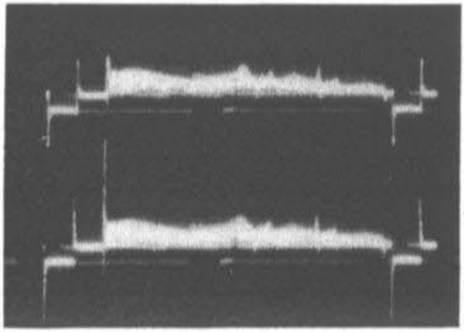


Fig. 5: One line of B/W video: before the video block filter (below), after the block filter (above)

5 MHz cannot be displayed in a normal PAL-norm monitor and therefore it is not necessary for them to be transmitted in the first place. They can cause modulation to the sound carrier which further increases the total transmitted bandwidth in the range 5.5 to 7.5 MHz.

These unwanted high video components are removed by the low-pass filter which has a 5 MHz limit frequency. The LPF is located in the video path shortly before it is combined with the sound carrier. A suitable filter here would be the pre-aligned "video block-filters" which are available in a miniature LC-technology. These may be obtained from Componex Düsseldorf under the designation 5 VFQ 1218 (or 1919).

Fig. 4 shows the video-output frequency response with the filter fitted between pin 1 and pin 3 (without pre-emphasis). Both filter types attenuate all frequencies above 5.5 MHz by more than 30 dB. The filter 1919 is characterised by a wider video bandwidth making it suitable for character and computer transmission but the "1218" has a higher attenuation in the vicinity of the sound carrier between 5.5 and 8 MHz. These small differences do not, in practice, appear to make a great deal of difference even with a variety of video sources, so it is quite in order to select on the basis of what is available.

Fig. 5 demonstrates the effect of the described video block-filter on a BAS signal. It can be seen that the high-frequency components of both the sync and the spurious needle pulses have been suppressed at the end of the front black porch. The original value of the voltage has been halved and thereby the sender deviation – for the same video quality!

1.1.3. Video Amplifier

The integrated circuit I 2 (NE 592) delivers at its output (7 and 8) two equal amplitude video signals but of opposite phase to each other. They are taken on to I 3 4 x CMOS switch (CD 4066). Two of these CMOS-switches form the actual video change-over. The opposite polarity control voltage switches the other "contacts" in the chip to earth.

The drive to the complementary output stage T6/T7, is a conventional x 4-transistor stage (T4) and is necessary because the undistorted output of I 2 is only 3 V_{pp}.

1.2. Audio Processing

1.2.1. Audio Amplifier

Figures 6a, b and c show the circuit details of the connections of the various microphone types to the audio amplifier. These input stages are fitted with a single low-noise transistor as standard ICs, even FET input types such as the TL 074/084, exhibit 10 dB more noise. A careful component layout and arrangement of the board conductor tracks avoid any inter-action with the video circuits sharing the common PCB and serve to prevent the ingress of the video voltage at 15 kHz into the sensitive microphone amplifiers.

1.2.1.1. Audio Notch-Filter

TV receivers and monitors radiate their line frequencies not only in an electrical fashion but also audibly. It is hardly avoidable that any microphone in the vicinity will pick up this 15 kHz tone and feed it, together with the speech frequencies, through the audio chain. **Fig. 7 above**, shows the AF voltage corresponding to the spoken "A" and which is being modulated by the line frequency. The consequence of this, is a 15 kHz spaced spectral comb on the sound carrier and/or an automatic overloading of the audio amplifiers which results in a lower deviation for the speech signals. This may be avoided, as shown in fig. 7, by the notch-filter comprising L1 tuned to 15.6 kHz. The effects of this filter may be seen in the **lower trace of fig. 7.**

1.2.1.2. Audio Pre-Emphasis

A standard 50 μs pre-emphasis network has been included in the audio signal path. Its function is described in detail in reference (1).

1.2.1.3. Audio Low-Pass

The noise spectrum of the input stage does not end at the AF limit frequency. In addition, the presence of the audio pre-emphasis network boosts the noise at 30 kHz by some 6 dB relative to that at 15 kHz. Therefore, in order to limit the audio bandwidth, the last stage of the AF amplifier is fitted with a Butterworth characteristic, low-pass filter with a 15 kHz limit frequency.

1.2.1.4. Automatic Level Control and Dynamic Compression

A fast-acting control circuit influences the amplification of transistor T1. This has the effect of reducing input amplitude signal variations of

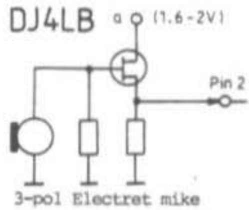


Fig. 6a: 3-pole electret microphone connections

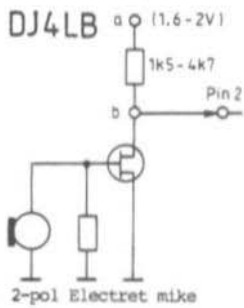


Fig. 6b: 2-pole electret microphone connections

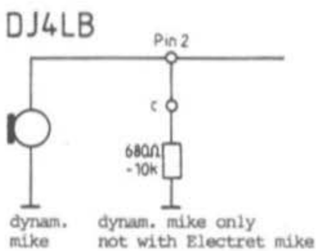


Fig. 6c: Dynamic microphone connections

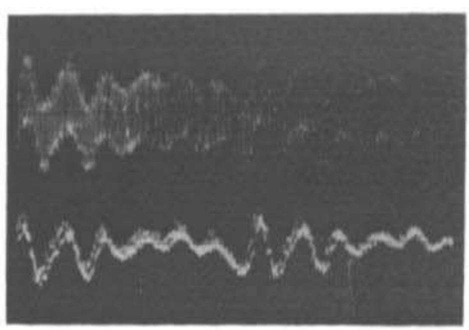


Fig. 7: A spoken "A" before the notch-filter (above), after the notch-filter (below)

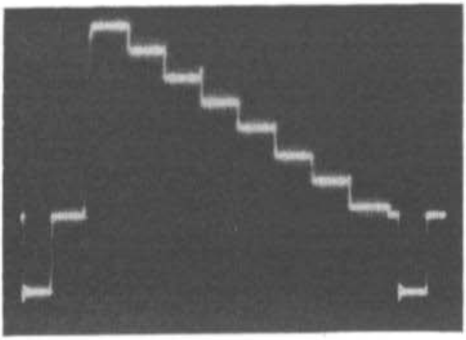


Fig. 8: Line of grey steps without sound carrier

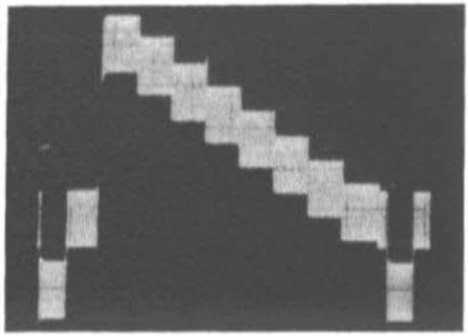


Fig. 9: Line of grey steps with sound-carrier component (−13 dB)

between 1 mV_{pp} and 70 mV_{pp} to an almost constant output voltage of 5.5 V_{pp} at TP 2. This serves to give a defined frequency deviation for the TV-sound carrier.

1.2.2. Sound Carrier Generation

The spectral purity of the sound carrier demanded in (1) can be attained by a combination of circuit measures:

A very small current (0.3 mA) through the oscillator transistor T9.

Galvanic coupling from the hot end of the low-harmonic oscillator tuned-circuit to gate 1 of T10. It will be apparent, that because of this, the effect upon the oscillator frequency when it is being aligned, either by varying the tuning voltage or by de-tuning the tuned-circuit L4 core, will be some 30-times smaller than would be the case if the coupling had been taken directly from the source via a coupling capacitor in the normal manner.

Frequency-independent coupling to the 75 Ω output via a capacitive divider instead of the more usual 10 pF capacitor direct from the hot end, also favours the suppression of harmonics.

At the specified deviation of ± 50 kHz, extensive tests have determined that at a mean DC-operating point on the varicap of 9 V, only the very smallest of distortion is measurable.

1.3. Video-Sound Carrier Coupling

The sound-carrier voltage is taken capacitively from the 75 Ω output, pin 3. This has the advantage over a common output for both video and sound carriers that the sound carrier is not affected by excessive video over-modulation.

Fig. 8 shows a test picture taken from one of the output sockets which has no audio component. **Fig. 9** shows a sound carrier which has a −13 dB (approx. 23 % of the video amplitude) of video modulation.

2. CONSTRUCTION

2.1. Printed Circuit Board

For the reasons given in 1.2.1., the module is best constructed using the copper-foil, single-sided, printed circuit board designation DJ4LB 010 (**fig. 10**). It will be found convenient to install the proprietary coil L2 after adjusting it first to an inductance of 9.5 μH.

If the sound-carrier oscillator is required to work on a frequency other than that of 5.5 MHz, the source resistance of T9 must be changed to achieve the same RF level. The appropriate resistances are given in **table 1**. It may also be

Frequency	$R_{source\ T9}$	AF voltage at TP 2 for ± 50 kHz deviation
5.5 MHz	10 kΩ	5.4 V_{pp}
6.0 MHz	9.1 kΩ	5.0 V_{pp}
6.5 MHz	8.2 kΩ	4.6 V_{pp}
7.0 MHz	6.8 kΩ	4.2 V_{pp}
7.5 MHz	5.1 kΩ	4.0 V_{pp}

Table 1

necessary to change the value of the 47 kΩ resistor in series with the "audio level" set pot'meter in order to bring the specified deviation into the range of this pot'meter.

2.2. Component Selection

T1: BC 550 C, BC 413 C (low-noise NPN)
T2...T6: BC 547 B or equiv. (NPN)
T7: BC 557 B or equiv. (PNP)
T8, T9: BF 256 C or BF 245 C (JFET)
T10: BF 980 or equiv. (DG-MOSFET)

I 1: TL 084 or TL 074
I 2: NE 592
I 3: CD 4066

1 red LED, 5 or 3 mm dia, U_F = 1.6 V
1 LED, any
3 universal diodes 1 N 4148 or equiv.
1 Z diode 9.1 V
1 varicap BB 809 (Siemens)

L1: Inductor approx. 3.2 mH (Neosid-Type 5902)
L2...L4: Inductor approx. 6 - 10 µH (Neosid 5800)
2 ferrite beads 5 mm long

Ceramic Capacitors with RM = 2.5:
2 x 4 p 7
1 x 10 p
1 x 47 p
2 x 1 n (RM = 2.5 and 5)

Styroflex Capacitors:
2 x 150 p
1 x 220 p
1 x 330 p
3 x 470 p
1 x 1696 pF (3 x 560 p)

Mica (TK = 0):
C2: 220 pF

MKH-Caps. (Siemens), RM = 7.5 and 10:
4 x 22 nF
1 x 47 nF
2 x 100 nF
2 x 220 nF
3 x 470 nF

Elco, 16 V, upright, RM = 2.5:
8 x 10 µF
2 x 22 µF
4 x 100 µF (or RM = 5)

Elco, 16 V, horizontal, max. 16 mm dia., max. 30 mm long: 1 x 1000 µF

Preset pot'meters, horizontal, RM 10/5:
each 1 x 1 k, 5 k, 25 k, 100 k

The following 4 resistors are to be found in the E96 series:
18.7 Ω, 75 Ω (3 x), 301 Ω, 698 Ω.

Note:
RM = lead spacing (mm),
Elco = electrolytic capacitor

The temperature drift of the sound-carrier oscillator depends largely upon the type of tuned-circuit capacity used. With a temperature change of $\Delta t = +1°$ C the following frequency changes are to be expected using the indicated types of capacitor for C1 and C2:

$\Delta f \approx -250$ Hz (styroflex)
$\Delta f \approx +350$ (mica)
$\Delta f \approx +100$ when C1 is styroflex and C2 is mica

Once the proximity effects of other components have been adjusted for, the sound carrier will work in a stable fashion even without enclosure in a metal housing.

2.3. Housing

The Euroformat (100 x 160) board mounted with 5 mm clearance above the housing floor, will fit into a proprietary tin-plate box which has a height of 30 mm – see **fig. 11**.

When wiring the connector sockets, the following points are to be observed: The microphone-cable screening mantle should not make contact

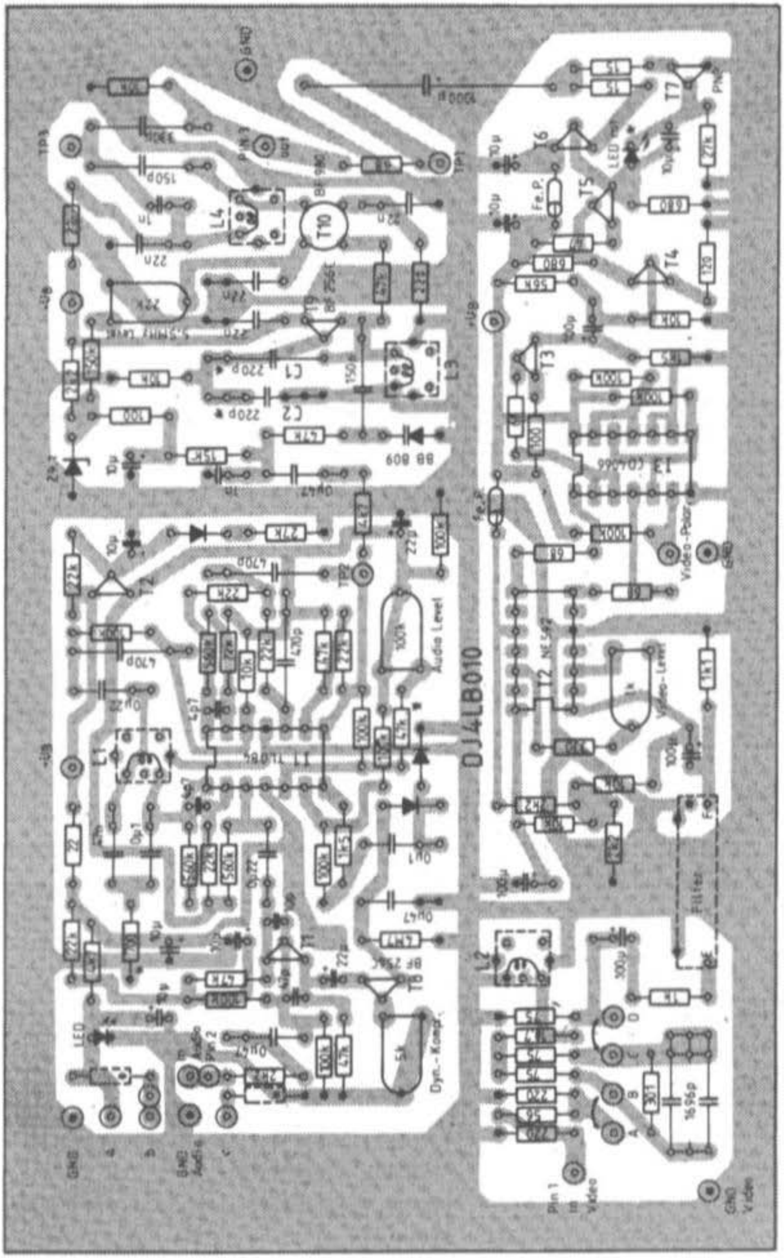


Fig. 10: Component layout plan of PCB DJ4LB 010

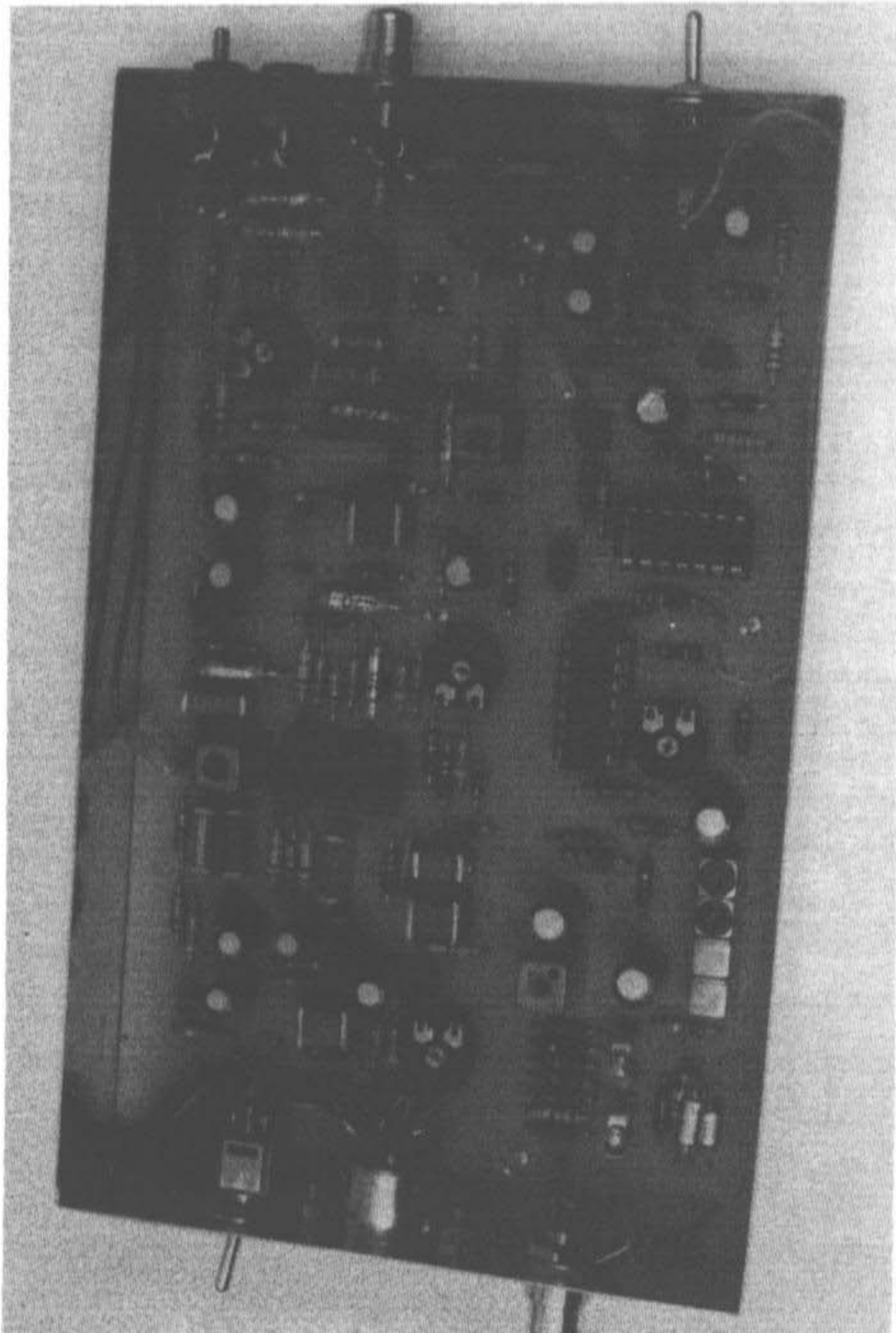


Fig. 11: A completed sample of the FM-ATV transmitter sound/vision unit

with the metal of the housing. It must be taken directly to the PCB point marked "GND Audio". This applies also to the video input where the cable mantle must be taken to the point marked "GND Video". This is simplified owing to the use of DIN "diode sockets" for both audio and video inputs. Cynch and BNC-sockets, if required, must also be in the isolated versions.

3. COMMISSIONING AND ALIGNMENT

Take supply potentials to the three parts of the circuit and check the current consumption. Video amplifier: when shorting the LED between T6 and T7, the current consumption should fall to some 5 mA – the final stage quiescent current. If a higher value is indicated, try another type of LED (older, dark red variety) or perhaps, three silicon diodes in series.

The carrier oscillator is adjusted to the nominal frequency by means of L3. L4 is adjusted for a maximum RF voltage at TP 3. Do not forget the wire or plug bridges A - B and C - D!

Operation without video pre-emphasis:
bridge A - D
Operation without video block filter:
bridge E - F
When all pots are set to their mid-range, the unit is ready for operation.

3.1. Fine-Tuning

With pot "video level", adjust to obtain 3 V_{pp} at the output pin 3 whilst using the largest normal video input signal. Or, simply adjust until no signs of over-modulation are apparent on the monitor screen.

The audio voltages at TP 2 are given in table 1. The adjustment of the AF level can be carried out without instruments by listening tests from the TV set.

The 5.5 MHz level can be changed during the course of a TV contact but the sound-carrier frequency does not drift by more than 250 Hz.

The pot. "dynamic compression" is turned only as far that the noise in speech-pause periods is not audible – use headphones connected to TP 2 to check this.

3.2. Notch-Filter

To adjust the notch-filter, feed in a 15.625 kHz signal to pin 2 preferably via a microphone which is standing close to a TV receiver. Adjust L1 for a minimum 15 kHz signal at TP 2. Using closely toleranced capacitors for the 0.1 µF and the 47 nF, the resistor at the output I 1/pin 1 will be 698 Ω – otherwise, put a small pre-set in series and use it in conjunction with L1 to minimize the 15 kHz. Measure the resistance of the pre-set and replace it with a fixed resistor of equal value. Then fine-adjust L1.

4. REFERENCES

(1) Sattler, G.: Baugruppen für frequenzmodulierte Amateurfunk-Fernsehsender (FM-ATV-Sender) in den GHz-Bereichen. Scriptum der Vorträge 32. Weinheimer UKW-Tagung (1987) S. 181 - 190

(2) Grimm, J., DJ 6 PI: Frequency Modulated Amateur Television (ATV) VHF COMMUNICATIONS, Vol. 18, Ed. 3/1986, P. 165 - 176

Andrew Bell, GW 4 JJW

WG20 Dish Mount

This article describes a method of mounting small 24 GHz parabolic dishes on a photographic tripod, when they are fed from the rear with WG20 or a smaller waveguide. Typical dish sizes are in the region 45 cms to 60 cms diameter.

The major problem with WG20 is that it is not really strong enough to support the weight and wind loading of a dish. To overcome this problem the author used a length of WG16 which structurally supports the dish on the tripod. Two square WG16 flanges and two purpose made adapters are used in conjunction with the WG16 to make the WG20 focusing and support mechanism for the dish.

Figure 1 shows a photograph of the completed unit. The dish has been drilled to take the four fixing bolts of a WG16 flange. Between the

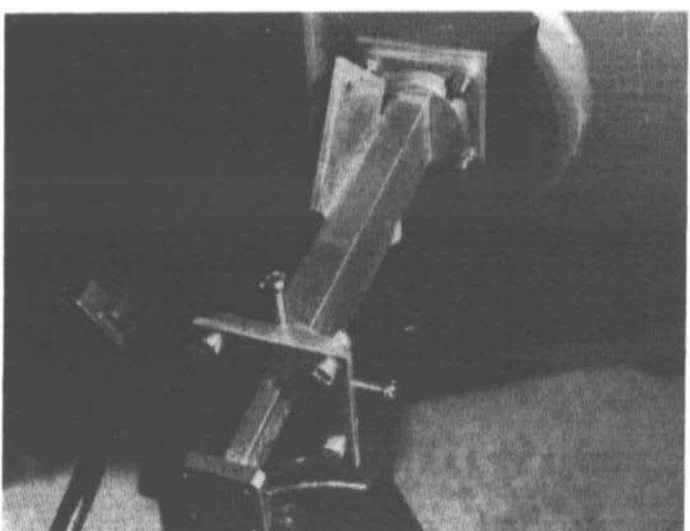


Fig. 1: The completed WG20 dish mount

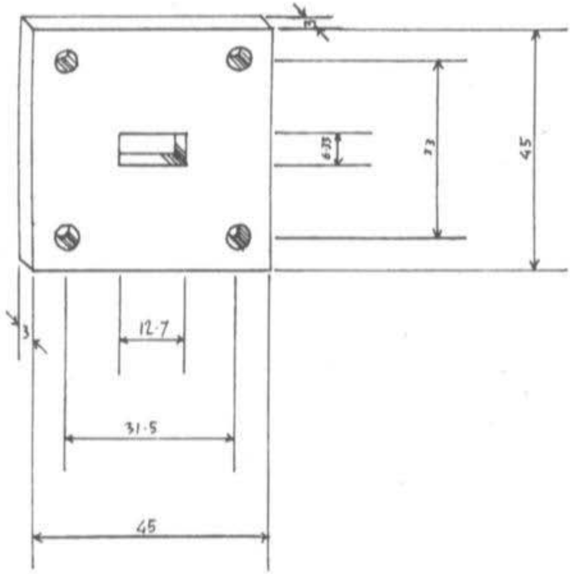


Fig. 2: WG16/20 adapter plate

dish and this flange is mounted a WG16/20 adapter. A short length of WG16 connects the "dish mounting" flange with the tripod mounting brackets and the rear WG16 flange and its WG16/20 adapter. Squares of brass were used for the tripod mounting brackets, but this was only because the author had these to hand and in fact orthogonally mounted rectangular brackets would probably be more suitable. Two brackets were used to allow the waveguide to be mounted either horizontally or vertically. (It was found that the tripod's own mechanism for turning a

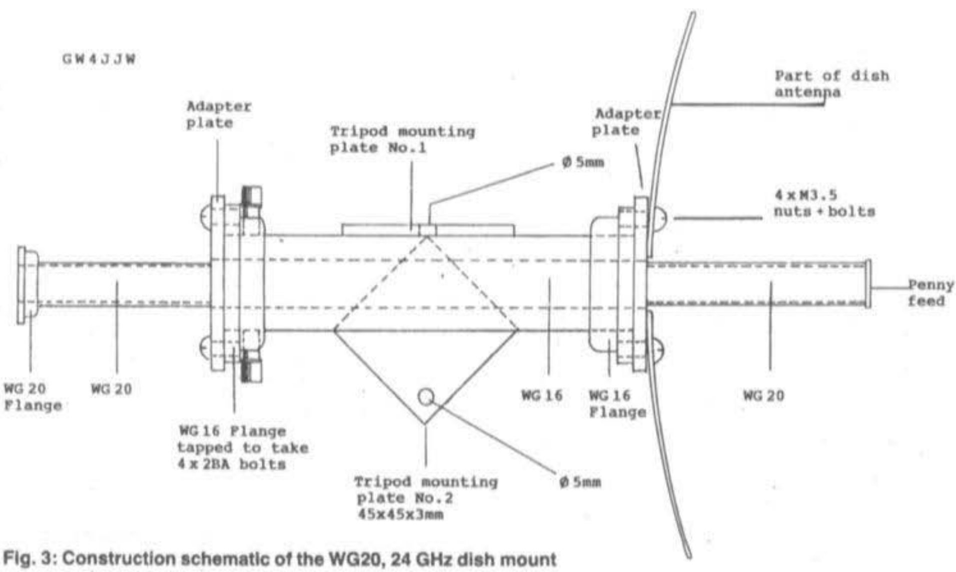


Fig. 3: Construction schematic of the WG20, 24 GHz dish mount

camera on its side was not satisfactory when supporting the weight involved.) The two brackets were silver-soldered to the WG16, though lead solder is probably strong enough.

The rear WG16 flange has been drilled and tapped radially with four 2BA or M4.5 bolts. These bolts, when **finger tight,** grip the WG20. Ensure that the ends of these bolts are smooth before using them to grip the WG20. The WG 16 must also be drilled such that these bolts pass straight through its wall with some clearance. Before lead-soldering this WG16 flange to the WG16, zinc-plated bolts were screwed into these holes to avoid them filling with solder.

Figures 2 and 3 give material details. It is most important that the WG16 flange is mounted centrally on the axis of the dish. It is also very important that the WG16/20 adapters are made accurately so that the WG20 is mounted in the exact centre of the WG16. If this is not so the radiator will appear off centre in the dish. In practice a small amount of adjustment can be made by replacing the M4 bolts with M3.5, or even M3, and adjusting the position of the two plates before tightening the bolts. A very small adjustment can be made by selecting the order in which the WG20 fixing bolts are tightened. **These bolts should only be tightened finger tight to avoid distorting the waveguide.**

John Winsor, G Ø JXU

A Magnetic Loop Antenna for 2 Metres

The described antenna was intended for mounting on a car using a magnetic mount. However, it may be used in other applications where a compact 2 metre antenna is required.

1. INTRODUCTION

The antenna as seen in **figure 1**, is constructed from standard copper water-pipe and plumbing fixtures available from most hardware stores. The antenna is omni-directional and horizontally polarised. If used in vertical polarisation, the polar diagram should be figure-of-eight.

The antenna functions on the principle of the magnetic loop, which is essentially a series-tuned circuit in which the inductor is a single-turn loop. A high voltage is developed across the capacitive gap, hence a voltage gradient exists around the periphery of the loop.

For the antenna to work well it must generate a high current and voltage, hence the Q will be

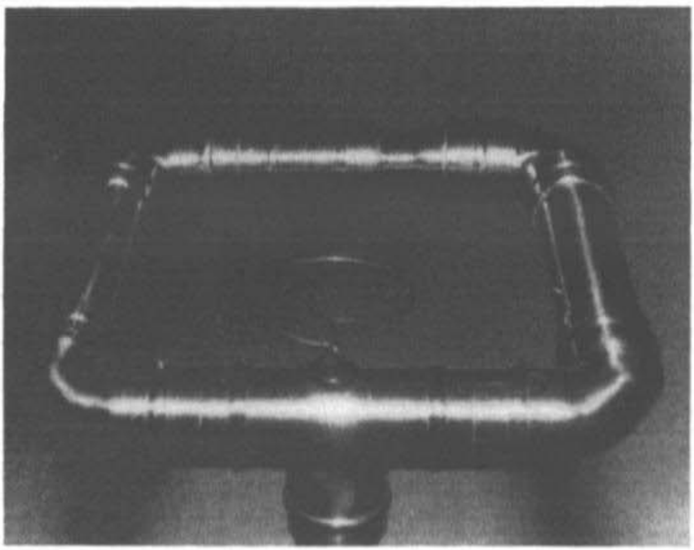


Fig. 1: Magnetic loop antenna for 2 metres

high and the bandwidth correspondingly narrow (about 200 kHz). The resonant frequency is intended to be 144.300 MHz which is the s.s.b. calling frequency.

To make the Q high the losses must be kept very low, and for this reason the antenna is constructed from copper water-pipe. Sturdy construction is also necessary to provide stability for the resonant frequency. The antenna is made to the minimum size possible using 22 millimetre pipe. It is believed, however, that the size could not be increased significantly because the capacitance required for resonance would have to be impracticably small. The constructor is recommended, therefore, to keep strictly to the dimensions supplied in **figure 2** and to construct the antenna as accurately as possible.

It is emphasised that the antenna is experimental and is **not** intended for use in rain, because the presence of water around the capacitive gap (**fig. 3**) would cause de-tuning. The writer believes that a plastic "radome" around the antenna would resolve this problem, and the constructor is invited to use his ingenuity.

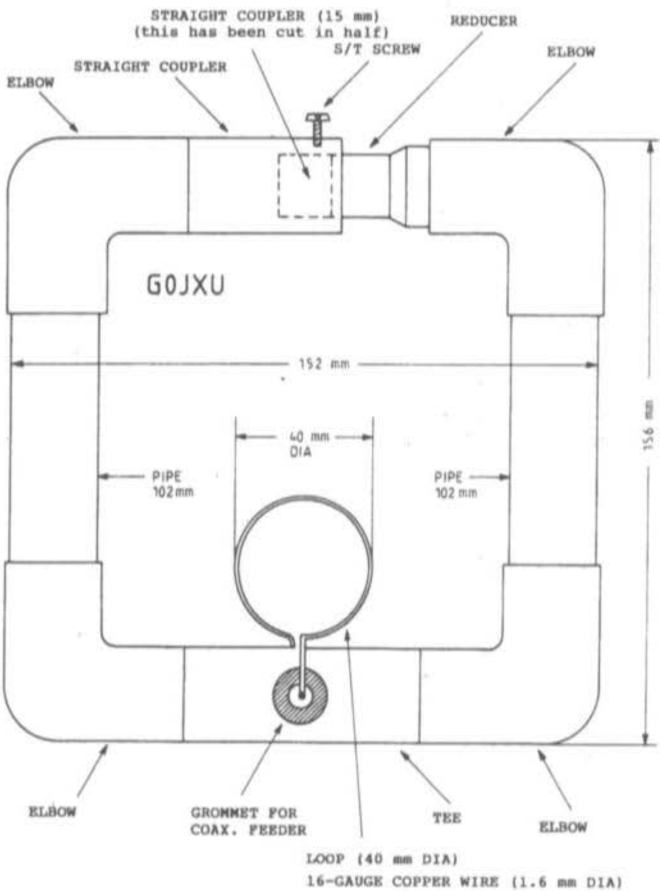


Fig. 2:
Dimensions of the magnetic loop antenna

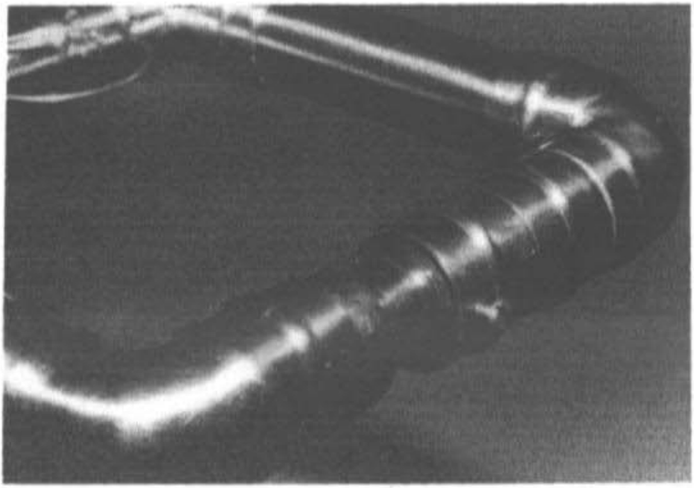


Fig. 3:
Capacitive gap

2. CONSTRUCTION

2.1. Parts List

Item Description	Qty
1 Elbow (22 mm)	4
2 Tee (22 mm)	1
3 Straight coupler (15 mm) (It is necessary to cut this in half)	1
4 Straight coupler (22 mm)	1
5 Reducer (22/15 mm)	2
6 Copper water-pipe, 15 mm diameter, 27 mm long	1
7 Copper water-pipe, 22 mm diameter, 36 mm long	3
8 Copper water-pipe, 22 mm diameter, 102 mm long	3
9 Plastic straight coupler (22 mm), used for insulating the antenna from the vertical supporting pipe.	1

(Items 1 to 5 are of a type which contains an internal deposition of solder.)

A solder tag is required to secure the braid of the coaxial cable and one end of the inductive loop.

2.2. General Remarks

It may be necessary to develop a little skill in plumbing. Soldering is best performed with a butane blowtorch. Apply flux beforehand. Be careful not to overheat the joints or the solder will drain out. Soldering may best be performed with the joint vertical. While heating watch the underside of the joint for the first appearance of solder, and remove the heat when this is seen. Joints already soldered may be wrapped in a wet rag to reduce the chance of accidentally de-soldering them while soldering an adjacent joint.

2.3. Construction, Step by Step

1. Cut the pipe to the specified lengths, and clean the surfaces with steel wool.
2. Drill holes in the tee for the feeder grommet, and for the screw which will secure the solder tag for connecting the feeder braid.
3. Assemble the antenna using the elbows and the tee, but on the capacitive gap side of the square, substitute a 102 mm long pipe in place of the capacitive gap components.
4. Solder the joints, but do not solder those holding the substitute pipe.
5. Make two saw cuts in the substitute pipe, and

remove and discard the pieces. Then assemble the capacitive gap components in their final location. Do not solder the 22/15 mm reducer to the elbow at this time.

6. In order to make fine adjustments to the resonant frequency, a very small self-tapping screw must be screwed into the 22 mm straight coupler, close to the end. The screw should protrude about halfway across the airgap to the 22/15 mm reducer. Use a screw without a sharp point at its end, or otherwise file off the point.

7. The antenna is supported by a 15 mm vertical pipe soldered to another 22/15 mm reducer, which is connected to the tee at right-angles to the plane of the loop. The vertical pipe may be attached to a magnetic mount using a suitable plumbing fixture with the antenna at about one quarter-wave above the car roof. Originally it was found that the 15 mm supporting pipe was slightly "hot" to r.f., and in order to avoid this effect a short plastic tube is used to insulate the loop from the vertical pipe (although this is not essential). The plastic tube is, in fact, a 22 mm straight coupler intended for waste water, also obtained from a hardware store.

8. The coaxial cable is coupled to the aerial by an inductive loop of 40 mm diameter (see diagram). Use 16 gauge copper wire, soldered to the same tag to which the coaxial cable braid is connected (**fig. 4**).

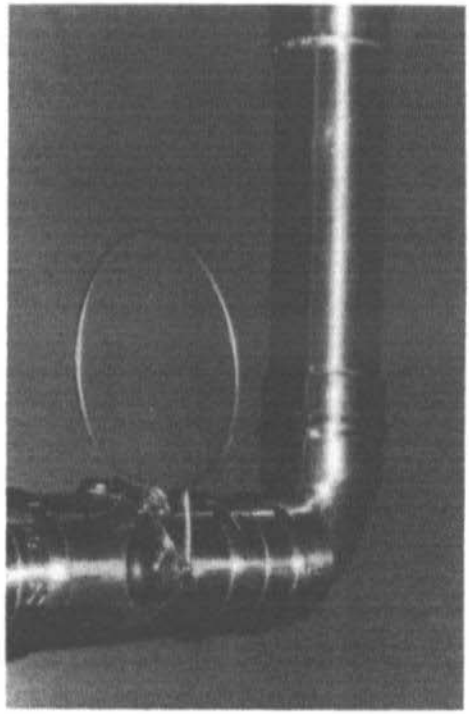


Fig. 4: Coupling from feeder

3. ADJUSTMENT

Initially it is convenient to adjust the resonant frequency while receiving a weak signal close to 144.3 MHz. The 22/15 mm reducer should be positioned to optimise the received signal strength. If the received signal cannot be tuned in this way, then some change to the dimensions of the loop would be necessary.

The final adjustment must be made while transmitting. Connect a VSWR meter between the transmitter and antenna. Be very careful, because output transistors can easily be destroyed by operating the transmitter into a gross mismatch, so use as little power as possible when making the adjustment. One way to generate low power (in the s.s.b. mode) is to whistle very softly into the microphone. The sensitivity control of the VSWR meter should be turned-up high.

With the VSWR meter indicating reflected power, the impedance match is optimised by sliding the 22/15 mm reducer. Finally, secure the 22/15 mm reducer to the elbow by soldering. Do this very gently and without further moving it. (This may be difficult because flux cannot be applied for this operation, so make sure that the inner and outer surfaces are cleaned with steel wool beforehand).

The closeness of the aerial to its correct resonant frequency may be tested by moving a finger near to the capacitive gap, whereupon the VSWR meter should indicate a change in the reflected power. If necessary, adjust the resonant frequency with the self-tapping screw already inserted in the 22 mm straight coupler.

Another means of tuning the aerial if close to the correct frequency is to distort the structure very slightly, hence either increasing or reducing the capacitive gap.

4. PERFORMANCE

The performance may best be assessed in comparison with a simple dipole. Although objective tests have yet to be performed, initial experience suggests that the magnetic loop antenna performs at least as well as a dipole. The chief advantages are its compact size and approximately omni-directional radiation pattern in horizontal polarisation. In vertical polarisation the radiation pattern should be figure-of-eight, in which unwanted signals may be nulled-out by rotating the antenna. This would, of course, be a disadvantage in mobile applications.

The main disadvantage is the narrow bandwidth (about 200 kHz as measured between the frequencies at which the VSWR = 1.5/1) which is an inevitable consequence of its low-loss construction and high Q. This bandwidth would present no problem to most c.w. or s.s.b. operators, but devotees of f.m. seem to require larger areas of the spectrum.

5. CONCLUSIONS

The writer believes that much remains to be discovered about the principles of the magnetic loop antenna. The theory of its operation seems to be rather obscure, and experiments are likely to yield surprising results. Especially, the conclusions found in articles describing h.f. loops should not always be accepted uncritically.

Caution – Possible Hazard to Health

Because of the compact size of the antenna it may seem convenient to use it in close proximity to the operator, especially when first tuning up. The possible harmful effects of electromagnetic radiation on the body are the subject of international concern and controversy. The constructor is therefore advised to use the antenna with reasonable caution, especially to use as little power as possible when tuning and to avoid unnecessary exposure at all times. **In particular, a hand should never be placed inside the loop with power applied.**

Matjaž Vidmar, YT 3 MV

Amateur-Radio Applications of the Fast Fourier Transform Part 1

This article is more than just a description of the latest software for the DSP Computer, as interested readers will soon find out. Nevertheless, it is also a continuation of the series of articles about DSP Techniques and written by the same author.

1. INTRODUCTION

The Fast Fourier Transform (FFT) is a very efficient numerical algorithm to compute a discrete Fourier transform. The FFT algorithm has found many different applications and it is not limited only to digital computers and digital signal processing. For example, a completely analog implementation of the FFT algorithm can be made with a number of 3 dB directional couplers and delay lines to feed an antenna array.

Obviously the FFT algorithm is widely used in digital signal processing too. In this article a particular DSP application of the FFT algorithm is described and discussed in detail: FFT spectrum analysis. Since the FFT algorithm can be implemented on almost any digital computer and a FFT spectrum analyzer only requires a little additional hardware, FFT spectrum analysis is very convenient for many practical applications.

In order to understand the operation of a FFT spectrum analyzer and its advantages and drawbacks when compared with other spectrum analysis techniques, a short description of the various Fourier transforms and the operation of the FFT algorithm is included. Building a practically working spectrum analyzer around the bare FFT algorithm is also discussed. No mathematical proofs are given just to keep the discussion as simple as possible. The interested reader can find the former in almost any book describing digital signal processing.

Although FFT spectrum analysis is presently limited to the audio frequency range, at least for amateur resources, it has many amateur-radio applications ranging from weak signal detection to modulation analysis. A few typical amateur applications are discussed later in this article, including spectrum function plots and intensity spectrograms obtained from real signals using a FFT spectrum analyzer.

Finally, a FFT spectrum-analyzer program for the DSP computer, described in UKW-BERICHTE/VHF COMMUNICATIONS, is presented, including a short description of its commands and performance limits imposed by the hardware. All the practical examples shown in this article were obtained with this software: most figures are just hardcopy prints of the computer screen on a laser printer.

2. THE FOURIER TRANSFORM AND THE DISCRETE FOURIER TRANSFORM

The Fourier transform is a mathematical operation that computes a new function $F(\omega)$ from an original function $f(t)$, as shown in **fig. 2.1.** Both functions have real arguments t and ω while their values $f(t)$ and $F(\omega)$ are in general complex numbers. $F(\omega)$ is called the Fourier transform of a given function $f(t)$. The Fourier transform has some interesting properties including a relatively simple inverse operation: computing $f(t)$ back from a given $F(\omega)$ is very similar to the Fourier transform itself.

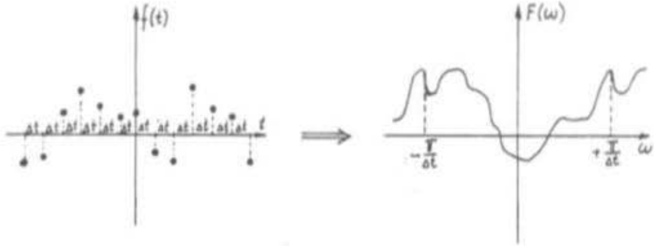
The Fourier transform is frequently used in physics to compute the frequency spectrum of a time-dependent physical quantity. In a physical problem t represents time and $f(t)$ represents a physical quantity (force, displacement, pressure, electrical current or voltage) as a function of time. In all physical problems $f(t)$ is a real function. The new variable ω represents the frequency and the function $F(\omega)$ is the frequency spectrum of the physical quantity observed. The frequency spectrum is a complex function of a real variable ω. The absolute value of $F(\omega)$ represents the magnitude of a given spectral component and the argument of $F(\omega)$ represents the phase relative to the (chosen) time origin. Since $f(t)$ is always a real function one does not really need to compute $F(\omega)$ for negative frequencies: $F(-\omega)$ is simply the complex-conjugate of $F(\omega)$ for a real $f(t)$.

While the Fourier transform is a powerful analytical tool to perform theoretical computations it has at least two constraints which could never be fulfilled in practice: infinite bandwidth and infinite frequency resolution (which implies an infinite observation time), regardles of the method, analog or digital, used to perform the Fourier transform.

A real-world, finite-bandwidth signal can be sampled without loosing any information provided that the sampling frequency is high enough, at least twice the signal bandwidth. The Fourier

$$F(\omega) = \int_{-\infty}^{+\infty} f(t)\, e^{j\omega t}\, dt$$

Fig. 2.1.: Definition of the Fourier Transform



$$F(\omega) = \sum_{n=-\infty}^{+\infty} f(n\Delta t)\, e^{j\omega n\Delta t}$$

Fig. 2.2.: Fourier Transform of a sampled signal

transform of a sampled signal is shown in **fig. 2.2.**. The integral is replaced with a sum, the bounds of the sum are, however, still infinite. An important difference should be noticed in plot of F(ω): the spectrum of a sampled signal is a periodic function and its period is inversely proportional to the sampling period. It is therefore sufficient to compute (either in an analog or in a digital way) just one period of F(ω), usually between $-\pi/\Delta t$ and $+\pi/\Delta t$.

The (original) Fourier transform integral (or sum) has infinite bounds: the integration (or summation) should be performed from minus infinity to plus infinity. Of course no real-world signal will ever last that long! It is therefore completely sufficient to compute the integral or the sum only over the time interval when the signal exists. Integrating over a finite amount of time limits the frequency resolution, which is inversely proportional to the integration time. This implies that the frequency spectrum could also be represented by discrete samples in place of a continuous function.

The procedure that computes a finite number of spectral lines from a finite number of signal samples is called the Discrete Fourier Transform (DFT) and is shown in **fig. 2.3.** To properly describe the signal spectrum one needs at least the same number of frequency samples as there are input signal samples. Of course, to obtain meaningful results, the frequency interval (spectral line spacing) has to be chosen in a close

$$F(m\Delta\omega) = \sum_{n=-N/2}^{+N/2} f(n\Delta t)\, e^{jm\Delta\omega n\Delta t}$$

Fig. 2.3.: Definition of the Discrete Fourier Transform

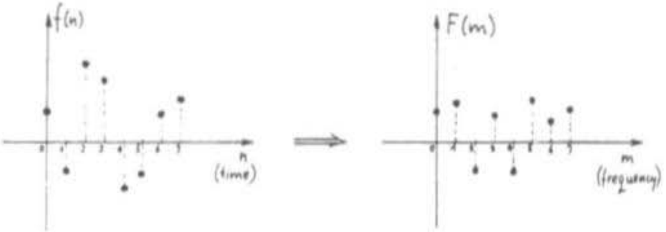


Fig. 2.4.: Normalizing time and frequency units

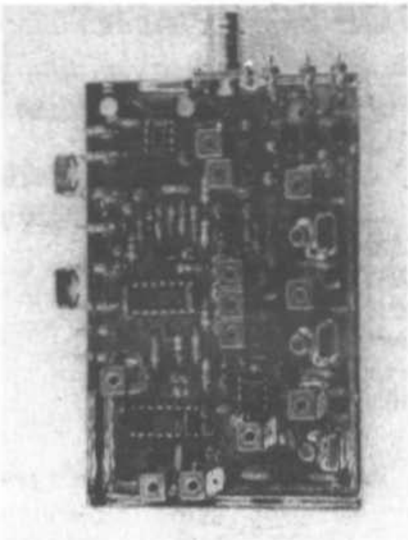
$$F(m) = \sum_{n=0}^{N-1} f(n)\, e^{j\frac{2\pi}{N}nm}$$

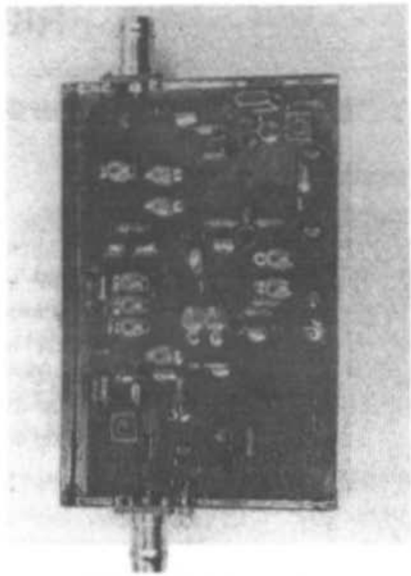
relationship to the time interval (signal sampling step). Similar constraints also apply to a completely analog Fourier transform (analog spectrum analyzer).

In order to simplify computations, both time and frequency units are usually normalized to 1 as shown in **fig. 2.4.** Time t now only takes the values of 0, 1, 2,... (N − 1) and frequency ω also takes the values of 0, 1, 2,... (N − 1), if the discrete Fourier transform produces N spectral lines from N signal samples. As a result of this normalisation the constant 2 ∗π/N appears in the complex exponent function. This constant is chosen such that the resulting spectral lines cover exactly one period of the periodic signal spectrum.

The DFT can be computed on any general-purpose computer. Its operation is similar to N bandpass FIR filters, each filter having N stages and tuned to its own frequency. The DFT is not computationally efficient, since the number of computations required increases with N^2. If the DFT is computed on a real signal (real f(t)) with N signal samples, then the result only includes N/2 spectral lines ranging from zero to half the sampling frequency. The remaining N/2 spectral lines are simply complex conjugates that can be computed in a much simpler way once the first N/2 spectral lines are known, reducing the total number of computations required to about $N^2/2$.

Will be continued





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