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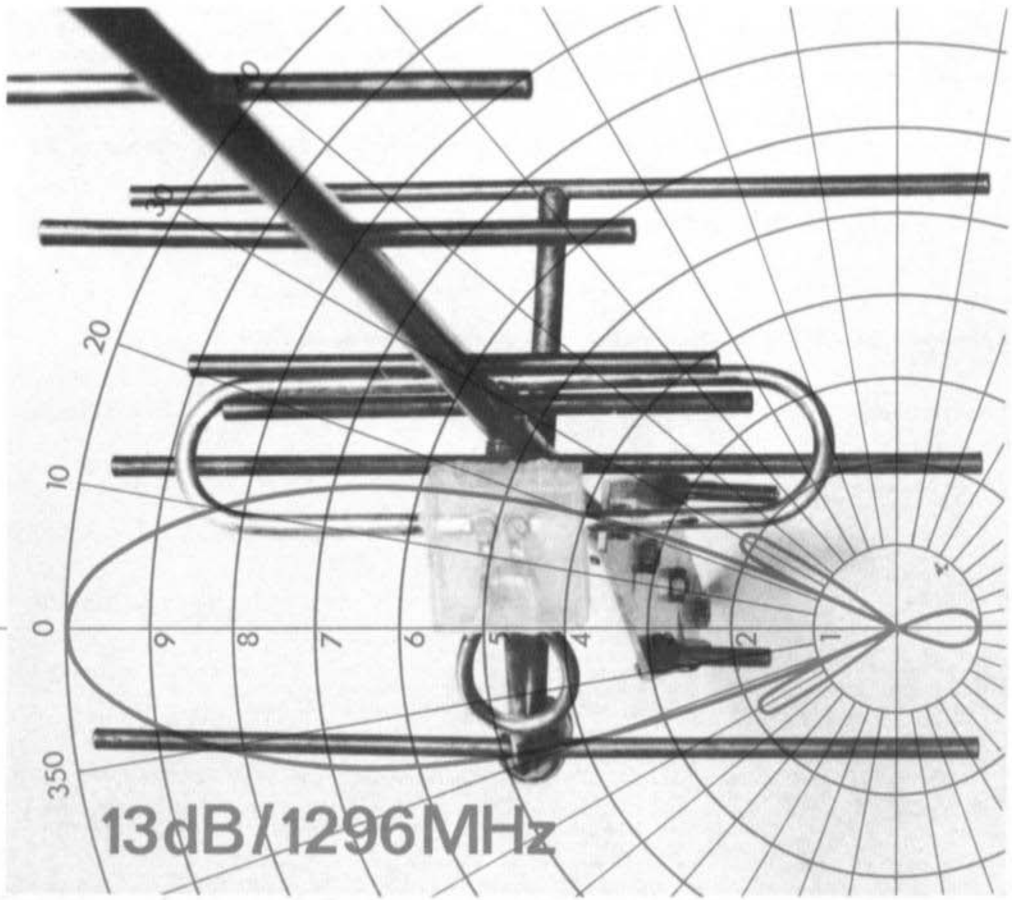

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AN SSB-EXCITER WITH RF-CLIPPER

by J. Kestler, DK 1 OF

A detailed description discussing the various methods of increasing the mean modulation level of voice transmissions was discussed in detail in (1). It was determined that the RF-clipper was superior with respect to speech quality to the other methods such as dynamic compressor, AF-clipper etc. However, the circuitry is considerably more extensive with RF-clippers. This is not too important with SSB transmitters since one of the two crystal filters is already available.

An SSB-exciter is to be described that is equipped with such an RF-clipper and whose degree of limiting is continuously variable. In order to be able to use this module for other modes, it has been provided with a product detector to demodulate the clipped RF-signal so that it is possible to feed it to an AM or FM modulator. Of course, it is then also possible for the module to be connected to any existing equipment at AF-level.

1. OPERATION

The operation of a RF-clipper is so well-known that it need not be discussed in detail here. As can be seen in the block diagram given in Figure 1, the circuit consists of a crystal-controlled oscillator whose output signal is mixed with the AF-signal in a ring modulator. The subsequent crystal filter suppresses the unwanted sideband and the resulting SSB-signal is fed via a variable amplifier stage to the RF-clipper where its maximum amplitude is limited. Of course, harmonics are generated here as with all limiting processes but do not cause any interference since they are at multiples of the carrier frequency (e.g. 18, 27, 36 MHz etc.). This means that they can be easily suppressed using simple resonant circuits. This is the great advantage of RF-clippers over AF-clippers. With the latter, the harmonics of the lower frequencies of the clipped AF-signal fall into the passband range causing a deterioration of the readability and speech quality.

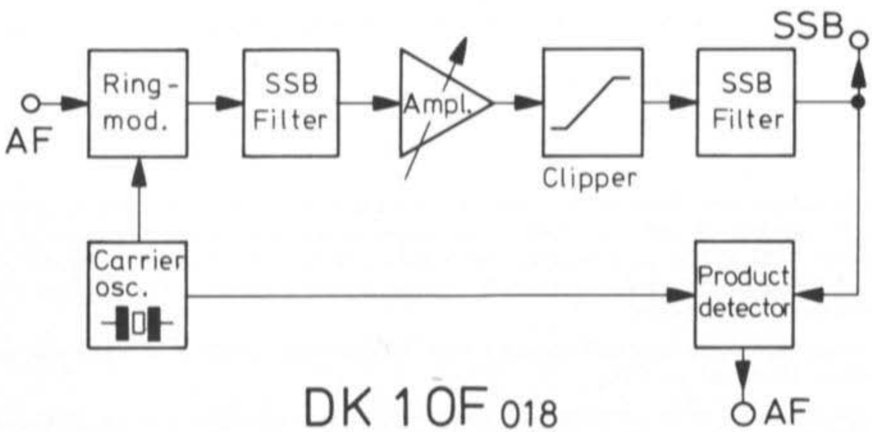

Fig. 1: Block diagram of the SSB-exciter with RF-clipper

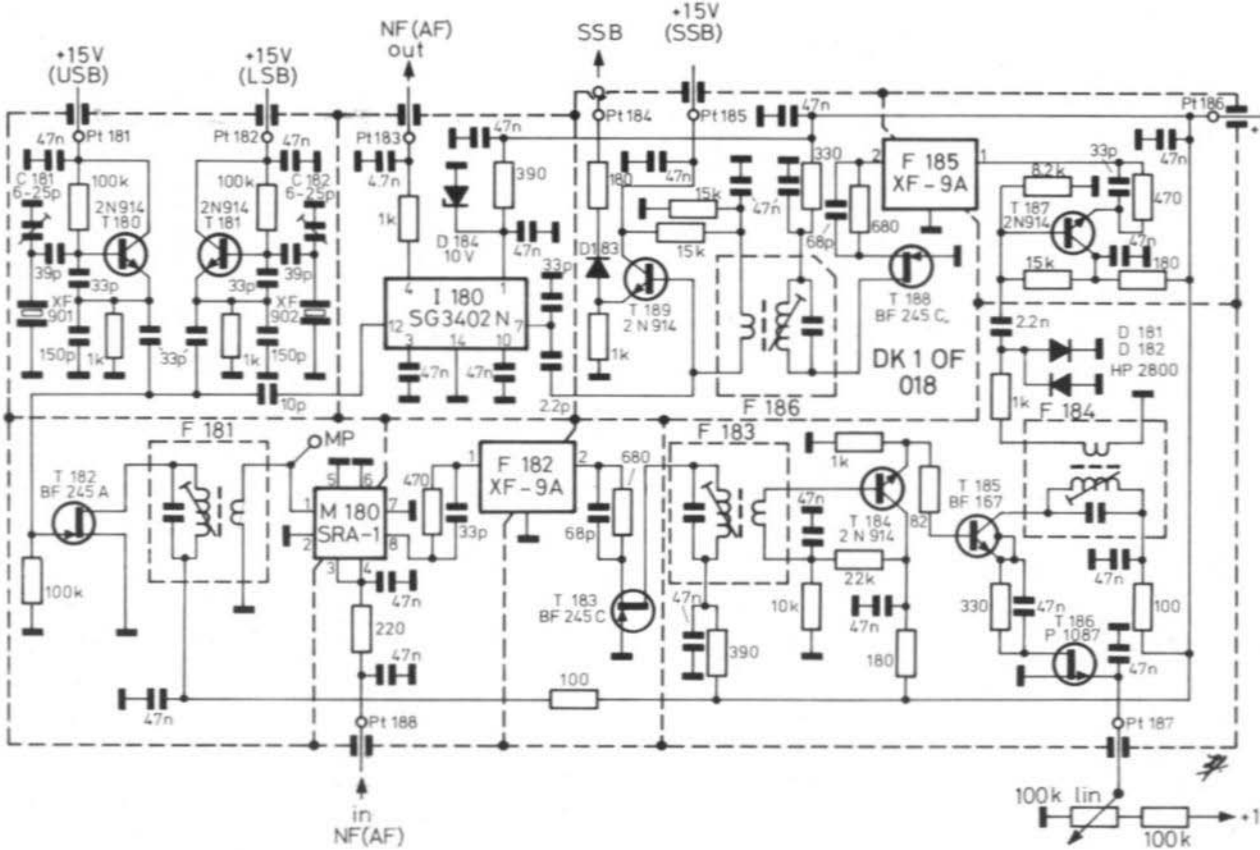

Fig. 2 : Circuit diagram of the SSB-exciter for 9 MHz or 10.7 MHz with RF-clipper

Conversion products from the individual AF frequencies deteriorate the signal far more than the harmonics since they are far nearer to the passband. This is shown in the following example:

The AF-signal is assumed to comprise two AF tones of 300 and 900 Hz. At a carrier frequency of exactly 9 MHz, these two AF-signals will be transposed to an upper sideband signal of 9000.3 and 9000.9 kHz, and the harmonics will be 18,000.6 and 18.001,8 as well as 27,000.9 and 27,002.7 kHz. Fourth and higher-order harmonics are to be neglected in this example. The following conversion products are formed from these harmonics.

27000.9 - 18001.8 = 8999.1 kHz
18000.6 - 9000.9 = 8999.7 kHz
18001.8 - 9000.3 = 9001.5 kHz
27002.7 - 18000.6 = 9002.1 kHz

It will be seen that the above frequencies consist of difference frequencies between odd and even harmonics. It is, of course, possible to considerably suppress the formation of even-order harmonics when an exactly balanced limiter is used (1), however, these components can also be generated by conversion between the odd harmonics and the fundamental frequency.

For this reason, it is necessary for a second SSB-filter to be provided directly after the limiter that is able to suppress these unwanted conversion products. In this manner, it is possible for a clean, narrow-band SSB-signal to be generated even when high clipping levels are used.

As can be seen in Figure 1, a product detector is provided which allows the clipped SSB-signal to be re-converted to the AF-level with the aid of the original carrier frequency.

2. CIRCUIT DESCRIPTION

The circuit diagram of the SSB-exciter is given in Figure 2. It consists of a crystal oscillator for generating the carrier frequency for the upper and lower sideband which comprises transistors T 180 and T 181. The selection of the sideband is made by switching the operating voltage between Pt 181 and Pt 182. Trimmers are provided in parallel with the crystals that allow the exact adjustment of the carrier frequency to the slopes of the crystal filter.

The carrier signal is amplified in transistor T 182 and then fed via the matching filter F 181 to the balanced mixer M 180 where it is mixed with the AF-signal from Pt 188. The filter link (220 Ω, 2 x 47 nF) ensures that no RF is fed to the AF-amplifier. For simplicity and reliability, a completed Schottky-ring modulator was used which means that no alignment is required. The carrier suppression was measured on five such mixers. It was found that a carrier suppression of 40 to 50 dB was exhibited. This is further suppressed on the slope of the two crystal filters by a further 30 dB. This means that a carrier-suppression of over 50 dB is provided even at the highest clipping level.

The first crystal filter (F 182) is matched to the 50 Ω output of the mixer with the aid of the RC-link of 470 Ω/33 pF. The same matching circuit is also used between the filter and the first RF-amplifier stage (T 183). As can be seen, the drain-current of the FET, which operates in a common-gate circuit flows via the secondary winding of the crystal filter. In spite of this, the filter is not endangered even with a direct short-circuit of the FET since the maximum possible direct current is limited to less than 15 mA with the aid of the 680 Ω and 390 Ω resistors. The SSB-signal is then passed via the resonant circuit F 183 to the emitter-follower T 184, which feeds the control stage T 185 at low-impedance. A variable TV IF-transistor BF 167 has been selected for T 185. The gain-adjustment is made with the aid of emitter feedback. The FET T 186 is used as a voltage-controlled resistor; in order to ensure that this transistor can be controlled with a positive control voltage (Pt 187), it is necessary to use a P-channel FET. If a negative voltage (approx. -7 V) is available, it is possible for the transistor to be replaced with a BF 245 C. A 5 kΩ potentiometer can be used instead of transistor T 186. However, since RF-voltage is present, it is necessary for it to be mounted within the screened module.

The actual RF limiter follows the matching circuit F 184. It comprises two anti-phase diodes D 181 and D 182. Extensive measurements have shown that even "super-fast" conventional diodes are not usable since they possess charge storage effects during the current flow-phase which are then released during the blocking phase. This means that they operate as capacitors at high limiting levels. Similar effects were found when using various circuits comprising differential amplifiers (CA 3028); the discharge times of these were also partially too long even at the relatively low operating frequency of 9 MHz. The best results were provided with Schottky diodes (HP 2800); using these diodes, it was possible for the energy of the unwanted voice peaks to be immediately dissipated into heat and not distributed over the following oscillations. It is not necessary to exactly pair these diodes, since even with an unbalance of 10% of the forward voltage, even harmonics will not cause any interference effects. However, both diodes should be from the same packing.

The RF-clipper is followed by an emitter follower (T 187) which then drives the second crystal filter (F 185). In spite of its relatively low ultimate suppression of approximately 45 dB, a crystal filter type XF-9 A is completely sufficient since the conversion products are already approximately 40 dB under the required signal level. This means that the intermodulation distortion of the subsequent power amplifier will be most certainly higher.

The second crystal filter is followed by a further buffer amplifier equipped with the FET T 188. This stage also operates in a common gate circuit and drives the output stage T 189 and the product detector I 180 via the resonant circuit F 186. Transistor T 189 operates in a common collector circuit. The operating voltage of this transistor is provided separately via connection Pt 185, so that it can be switched off in the AM and FM mode. The SSB signal is fed via the decoupling diode D 183 to the RF output Pt 184, where approximately 2 V (peak-to-peak) is available at an impedance of 500 Ω.

A cheap active ring mixer is used as product detector. The carrier signal is fed via the 10 pF capacitor to pin 12 of the integrated circuit; the sideband signal is fed from the secondary winding of filter F 186 via a voltage divider (2.2 pF/33 pF) to pin 7. The operating voltage (pin 1) is stabilized with the aid of a zener diode (D 184). The demodulated AF signal is fed via a filter link of 1 kΩ/4.7 nF and is available at connection Pt 183.

2.1. COMPONENTS

T 180, T 181, T 184, T 187, T 189: 2 N 2222 or 2 N 914
T 182: BF 245 A
T 183, T 188: BF 245 C
T 185: BF 167
T 186: W 1087, P 1087 (Siliconix)

D 181, D 182: HP 2800 or similar Schottky diodes
D 183: BAW 76 or 1 N 4148
D 184: ZF 10 or similar 10 V zener diode

M 180: SRA-1 or IE-500 (MCL)

F 181, F 183, F 184, F 186: FFM-2 10.7 MHz IF circuits
F 182, F 185: XF-9A with sideband crystals XF 901 and 902

I 180: SG 3402 N (Slicon General).

2.2. CONSTRUCTION

The component locations and PC-board DK 1 OF 018 are shown in Figure 3. The PC-board is single-coated and has the dimensions 125 mm x 80 mm. As can be seen in the photograph given in Figure 4,the module is split into seven chambers with the aid of screening panels. The total height of the module amounts to 30 mm. All connections with the exception of Pt 184 are provided with feed-through capacitors of 2 nF or more. The inputs and outputs of the two crystal filters are connected with the aid of short wire bridges to the required points of the PC-board.

3. CONNECTION AND ALIGNMENT

Firstly, the operation of the crystal oscillators is checked. This is achieved by connecting an operating voltage of +15 V to Pt 181 and Pt 186. A tube-voltmeter (or a test circuit comprising germanium diode, bypass capacitor and multimeter) is used to measure the RF voltage at the inputs of the balanced mixer (measuring-point "MP"). Filter F 181 is now aligned for maximum voltage reading. A value of approximately 1 V_{rms} (or approx. 1.3 V DC) should be achieved. If a frequency counter is available, the adjustment trimmers can be adjusted to the required frequency:

Pos. op. voltage to	Alignment on	Frequency at MP
Pt 181	C 181	8998.5 kHz
Pt 182	C 182	9001.5 kHz

An AF voltage of approximately 1 V (peak-to-peak) at a frequency of approximately 1 kHz and an impedance of approximately 300 Ω is now fed via a 10 µF capacitor to Pt 188. The clipper is adjusted to maximum (0 V at Pt 187) and the tube voltmeter (VTVM) connected to Pt 184 (terminated with a 500 Ω resistor to ground). Connection Pt 185 should now be connected to the operating voltage and circuits F 183, F 184 and F 186 aligned in this order for maximum

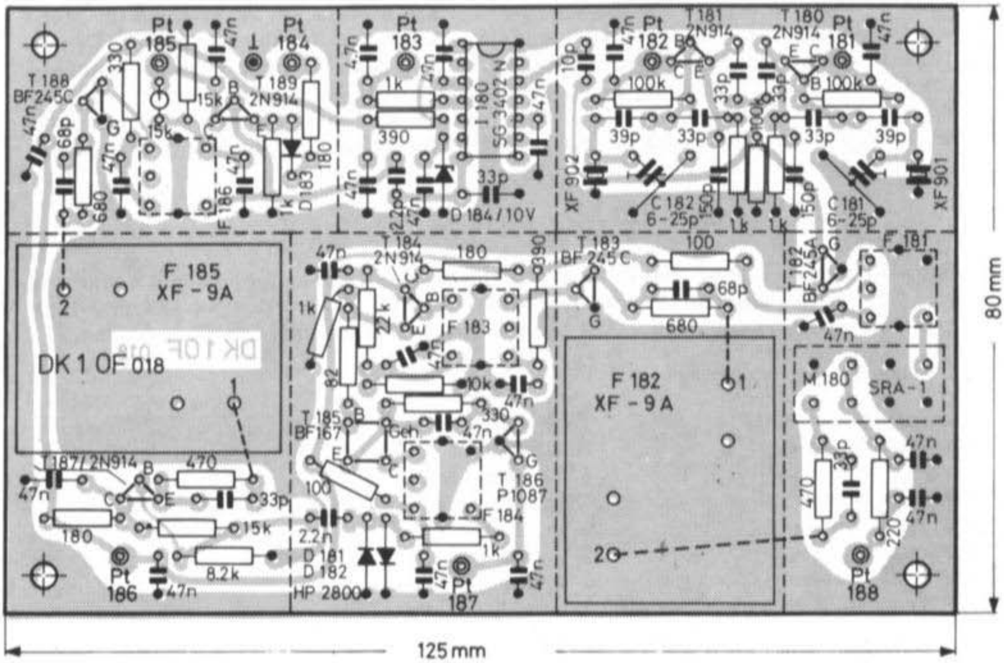

Fig. 3 : Component locations on PC-board DK 1 OF 018 (SSB-exciter)

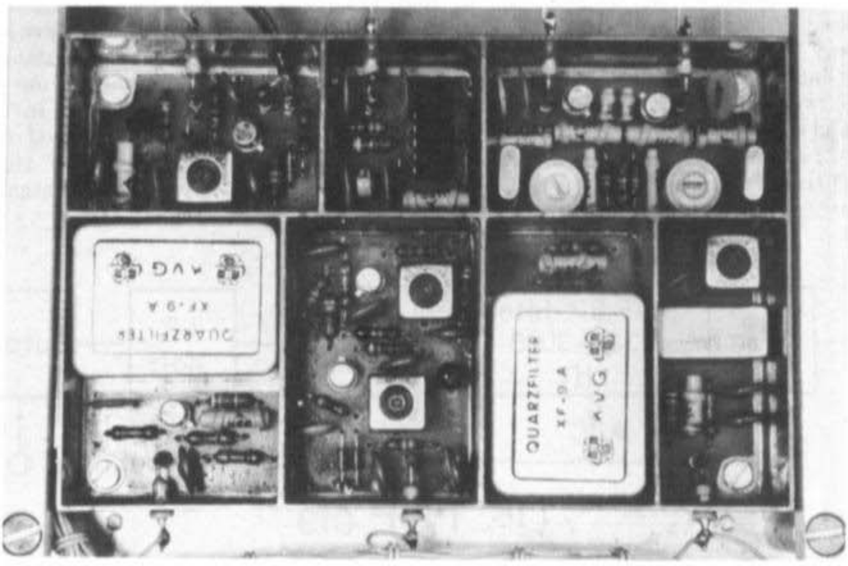

Fig. 4 : Photograph of the author's prototype of the SSB-exciter DK 1 OF 018

output voltage. It will be necessary for the AF voltage to be reduced in steps so that the limiter just does not actuate. At an AF input voltage of 150 mV (peak-to-peak), an RF output voltage of 2 V (peak-to-peak), corresponding to 0.7 V_{rms} should be available across 500 Ω. With the clipper level reduced, (approx. +7 V at Pt 187), this output voltage will be available with approximately 1.5 V (peak-to-peak) at the AF input. The adjustment range of the clipper therefore amounts to:

$$20 \log \frac{1.5\ \mathrm{V}}{150\ \mathrm{mV}} = 20\ \mathrm{dB}$$

The demodulated signal should be audible on a high-impedance headset connected to Pt 183 via a capacitor of approximately 0.1 μF. Connection Pt 183 provides a voltage of approximately 1 V (peak-to-peak) under non-load conditions.

4. MEASURED VALUES

Operating voltage:	15 V (12 V - 18 V)
Carrier frequency:	USB: 8998.5 kHz
	LSB: 9001.5 kHz
Carrier suppression:	Clipper min. >70 dB (Clipper max.) >50 dB
AF input voltage:	max. 1.5 V into 300 Ω
RF output voltage:	2 V into 500 Ω
AF output voltage:	1 V into 100 kΩ
AF frequency response (-6 dB):	300 to 2700 kHz

All voltage values are peak-to-peak.

5. MICROPHONE AMPLIFIER

In order to utilize the advantages of the RF clipper to the full, it is necessary for the microphone amplifier to satisfy certain demands. Only those AF frequencies should be fed to the exciter that are required for the voice transmission. Virtually none of the SSB exciters that are in operation at present have any special selectivity at AF level. The result of this is that the balanced mixer can be overdriven by frequency spectra that are far above and below the voice-frequency range. These frequencies are, of course, attenuated in the sideband filter, but not the conversion products that fall into the required frequency range. For this reason, it is extremely advisable for a steep AF filter to be provided before each modulator that suppresses all unwanted AF signals so that no overloading can take place.

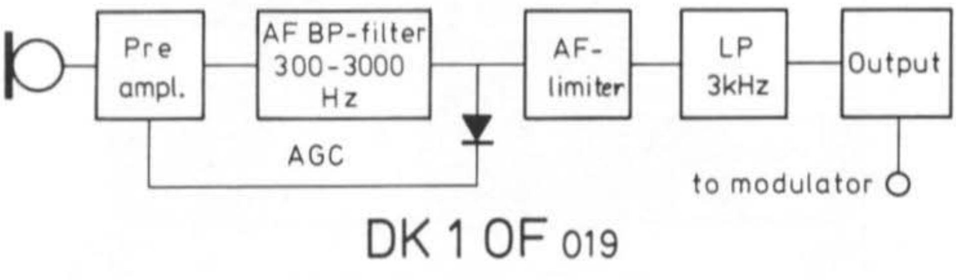

Fig. 5 : Block diagram of the microphone amplifier

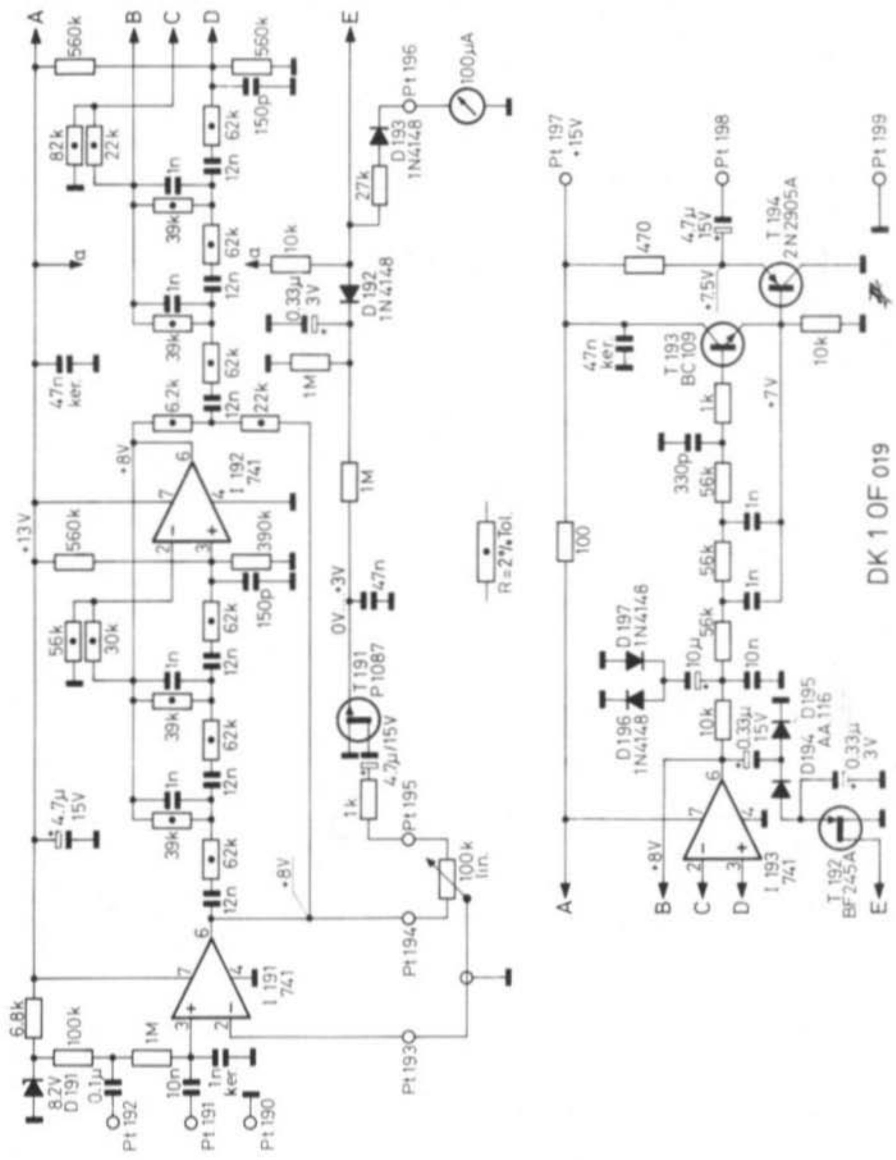

Fig. 6 : Circuit diagram of the modulation amplifier with dynamic compressor, limiter and filters

Furthermore it must be ensured that all limiting exclusively takes place between the two crystal filters at RF level and not, for instance, in the balance mixer. If the latter were the case, harmonics of the AF voltage would result which will then fall within the passband range. It is therefore advisable for a level-control circuit to be provided which is able to protect the mixer against overloading.

5.1. CIRCUIT OF THE AF AMPLIFIER

A block diagram of the microphone amplifier is given in Figure 5. The controlled input stage is followed by an AF bandpass filter which limits the AF range to the required 300 Hz to 3 kHz. Various descriptions recommend a combination of lowpass filter and pre-emphasis. However, a bandpass filter is preferable since the natural frequency characteristic of the voice remains intact. An active AF bandpass filter is hardly more extensive than a corresponding low-pass filter. As is to be described later, it only requires the same number of active components.

The next stage consists of an AF limiter. This stage is designed so that its threshold is only reached during transients of the AGC voltage of the input stage. This ensures that the modulator is not overdriven before the AGC can control the signal. Overshoot effects from the steep AF filter are also suppressed. The tailored AF signal is finally passed via an active low-pass filter with a cut-off frequency of 3 kHz. Any harmonics generated by limiting the overshoot peaks are therefore suppressed. The output stage must provide a relatively high output current in order to drive the balanced mixer.

The complete circuit diagram is given in Figure 6. Input Pt 191 is designed for crystal microphones. Input Pt 192 is a further input for connection of a tape recorder or AF generator (CW) and has an input impedance of approximately 100 kΩ. An operational amplifier I 191 is used as preamplifier. Its operating point is stabilized with the aid of the zener diode D 191. The manual gain-control is made with the aid of an external 100 kΩ potentiometer. The automatic gain-control drives the P-channel FET T 191 to higher impedance with the aid of a positive gate voltage and thus reduces the gain of I 191.

The subsequent circuit comprising amplifiers I 192 and I 193 forms a two-stage active bandpass filter. They are virtually formed by combining a high-pass and a low-pass filter, as described by DJ 4 BG (2), as shown in Figure 7. It will be seen that only three capacitors and three resistors are required per stage in order to form a bandpass filter instead of a low-pass filter. Figure 8 shows the selectivity curve of the first stage and Figure 9 the passband curve of the complete filter. These diagrams were made with an AF sweep-generator and XY-recorder; the X-axis (frequency) possesses a logarithmic scale, whereas the Y-axis (amplitude) is linear.

The output voltage of I 193 is rectified in a voltage-doubler circuit (D 194, D 195) and blocks the FET T 192 according to the drive level (this transistor is fully open under no-signal conditions). The increasing drain voltage under blocking conditions charges the 0.33 μF capacitor via diode D 192 which then drives the previously mentioned control transistor T 191. A short rise-time (30 ms) is achieved in this manner as well as a long fall-time constant (0.3 s). A meter of 100 μA can be connected to Pt 196 for monitoring the drive. The

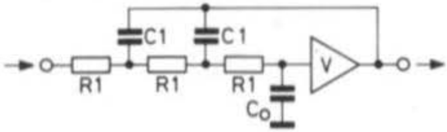

Fig. 7a: Active low-pass filter

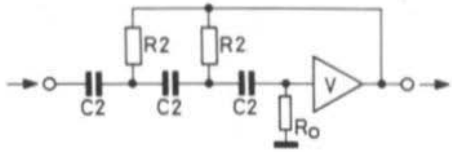

Fig. 7b: Active high-pass filter

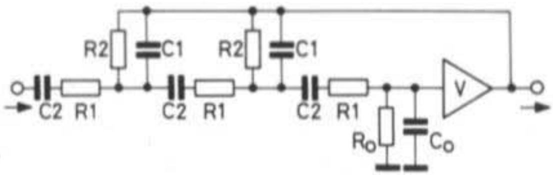

Fig. 7c: Active bandpass filter

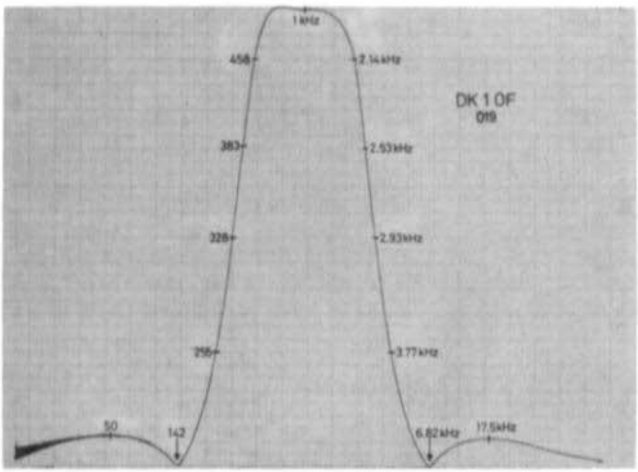

Fig. 8: Selectivity curve of the first stage of the modulator

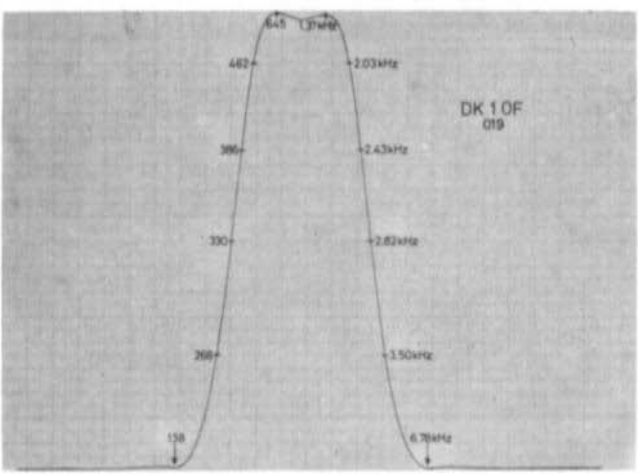

Fig. 9: Passband curve of the whole AF filter

threshold voltage of diode D 193 ensures that the saturation voltage of T 192 does not falsify the zero point of the meter.

The AF-limiter consists of diodes D 196 and D 197. Since the subsequent part of the circuit exhibits a voltage gain of 1, the maximum output voltage of the module is limited to 1.4 V peak-to-peak (twice the threshold voltage of the silicon diodes). The low-pass filter comprising T 193 as active component is also based on details given in (2), and has a cut-off frequency of 3 kHz. The output stages equipped with transistor T 194 in a common collector circuit. The collector current amounts to approximately 16 mA and the output impedance is approximately 30 Ω.

5.2. COMPONENTS

I 191, I 192, I 193: TBA 221 B (Siemens) or LM 741 CM

T 191: P 1087, W 1087 (Siliconix) or similar P-FET
T 192: BF 245 A
T 193: BC 109, BC 413 or similar (B > 200) AF transistors
T 194: BC 160, 2 N 2905 A or similar PNP-transistors (TO-5)

D 191: ZF 8.2 or similar 8.2 V-zener diode
D 192, D 193, D 196, D 197: 1 N 4148 or similar silicon diodes
D 194, D 195: AAY 25, AA 116 or similar germanium diodes

All resistors that are marked with a dot in the circuit diagram should have a tolerance of maximum ± 2%. Siemens MKM-series have been found very suitable for the 12 nF capacitors. The 1 nF capacitors in the active filters are styroflex-types. Ceramic-disc capacitors are not advisable due to their large tolerance ranges (values of e.g. +100/-50% have been measured) and due to microphonic effects.

5.3. CONSTRUCTION AND ALIGNMENT

The printed circuit board DK 1 OF 019 has been developed for accommodating the microphone amplifier. The dimensions are 125 mm x 50 mm and it is single-coated. The component locations on this board are given in Figure 10. As will be seen, two ground connections are provided: Pt 199 is connected to the ground of the transmitter and Pt 190 is only connected to the screening of the micro-

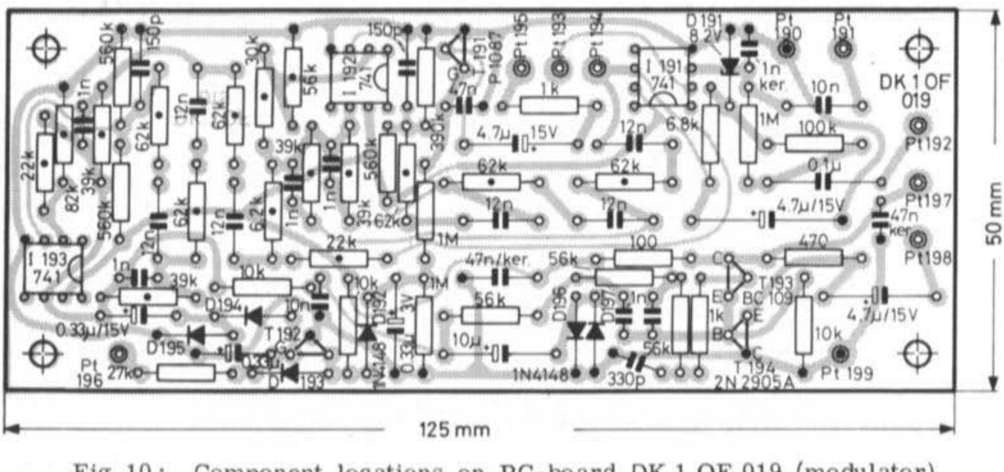

Fig. 10: Component locations on PC-board DK 1 OF 019 (modulator)

phone cable which should only be grounded at this point. It is possible in this manner to ensure that no hum voltage is injected, which is very important when operating with dynamic compressors and clippers due to the increase of the mean modulation level.

Of course, a well-stabilized and filtered operating voltage is required for operating the module since no filter links are provided on the board for this purpose. Any hum or hum-noise voltages on the power supply lines could be directly injected into the signal path via the voltage dividers and the non-inverting inputs of the operational amplifiers.

The operating voltage of +15 V is connected to Pt 197, after which the DC voltage values given in Figure 6 should be checked. Deviations of ± 2 V are permissible and will be the result of component tolerances. If an AF-generator and VTVM (or calibrated oscilloscope) are available, it is possible for the selectivity curve to be measured. Figure 9 shows the ideal curve which will probably not be available before alignment. Mainly, the two humps of the passband curve will differ in amplitude (approx. ± 3 dB). This is due to the inavoidable tolerances of the frequencydeterminingR and C components and will hardly cause any undesirable effects during practical operation. The curve can be optimized by altering the voltage divider at the inverting inputs of I 192 (56 kΩ/30 kΩ) and I 193 (82 kΩ/22 kΩ). This is achieved easier by providing trimmer resistors and determining the required resistance values experimentally, after which they can be replaced by fixed resistors of the nearest standard value.

The dynamic compressor can also be checked easily. At full gain (Pt 193 to Pt 195), the output voltage at Pt 198 should only alter by 6 dB at input voltage levels between 5 mV and 0.5 V at Pt 191.

Finally it should be noted that the described AF amplifier is by no means only suitable for use together with the RF-clipper. The advantages of the AF-selectivity and level control makes it very suitable for installation in existing shortwave and VHF equipment by placing it in the microphone cable.

5.4. MEASURED VALUES

Operating voltage:	+ 10 to 18 V
Voltage gain:	Min. 6 dB, max. 46 dB (1 kHz)
Input impedance:	500 kΩ
6 dB bandwidth:	330 to 2800 Hz
3 dB bandwidth:	390 to 2400 Hz
Slope of the filter:	24 dB/octave
Control range of the compressor:	40 dB
Max. output voltage:	1.4 peak-to-peak into 300 Ω
Output impedance:	approx. 30 Ω

6. PRACTICAL EXPERIENCE WITH THE RF-CLIPPER

Although measures to increase the mean modulation level can be very effective one cannot expect wonders. If one's signal cannot be heard without clipper by the partner station, no reliable QSO will be possible when the clipper is in operation. However, when the partner station reports that the signal can be heard but not read, then the clipper should be able to increase the readability to from four to five.

Extensive tests have shown that the improvement of the signal-to-noise ratio corresponds to an increase of output power by approximately 6 dB.

Experiments with a higher degree of clipping than 20 dB brought no improvement in readability even with distant stations. The modulation was then said to be unnatural.

When the clipping level of the described exciter is at maximum (20 dB compression), the plate current meter of the output stage remains at maximum. This means that all stages of the transmitter must be designed for continuous operation at full load.

7. REFERENCES

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VHF COMMUNICATIONS 1 (1969), Edition 4, Pages 218-235

ACTIVE BANDPASS FILTERS USING RC COMPONENTS EASY TO CALCULATE AND REPRODUCEABLE THEORETICAL PART

by D. E. Schmitzer, DJ 4 BG

1. INTRODUCTION

Active filters, mainly low-pass and high-pass filters as well as combinations to form bandpass filters have been discussed in detail in the past (see references (1), (2), (3)). This article is to describe an active bandpass filter which is of interest for a large number of applications. It is easy to calculate and is very reproduceable if a few simple rules are followed. The basic details for this were given in (4), and have been accepted as being correct. The derivation of these formulas is not of interest for practical applications. This article pays special attention to simplification possibilities and shows how these formulas can be best used.

The circuit diagram given in Figure 1 is used. The operational amplifier and the demands placed upon it are discussed in further detail in Section 7. The frequency-determining parts of this circuit are resistors R 1, R 2, R 3 and capacitors C 1 and C 2.

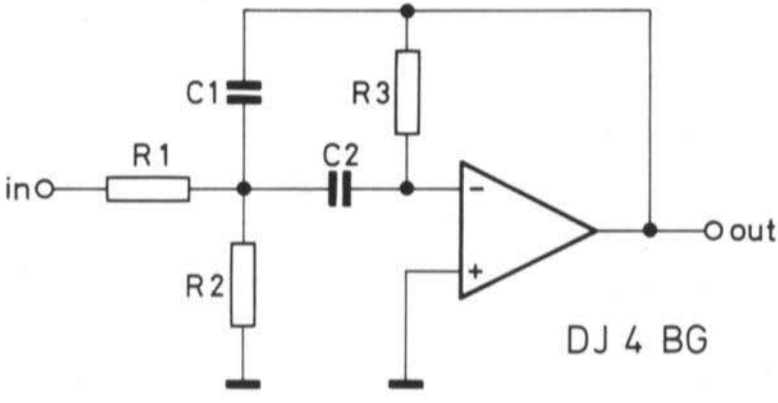

Fig. 1: Circuit of an active bandpass-filter (without DC-paths)

The parallel connection of R 1 and R 2 is designated R_p, and can be calculated according to the well-known formula:

$$R_p = \frac{R\,1 \times R\,2}{R\,1 + R\,2} \qquad (1)$$

In addition to this, the known relationships:

$$\omega_o = 2\pi \times f_o = 6.28 \times f_o \quad (2) \qquad \text{and} \qquad Q = \frac{f_o}{b} \qquad (3)$$

are required where f_o = resonant frequency (required centre frequency of the filter) and b the 3 dB bandwidth. If the two -3 dB corner frequencies are given instead of the centre frequency, the centre frequency can be calculated as the geometric mean value between these frequencies, e.g. $f_o = \sqrt{f\,1 \times f\,2}$ (4).

2. REQUIREMENTS

It is necessary to determine which characteristics are required from the filter before calculation can be commenced. Required are: The centre frequency of the filter (f_o), the Q or the bandwidth b; where the Q can be calculated according to formula (3). In addition to this, it is necessary to know the required gain G of the filter. Normally, the gain is given as being factor 1 which means that the output voltage is equal to the input voltage. However, it is advisable for some applications that a portion of the overall gain is provided by the filter so that this need not be obtained elsewhere.

After these demands (f_o, Q and G) have been established, it is necessary to determine which of the two possible calculations should be used with the aid of a simple preliminary process: This is done by calculating the value $\sqrt{G/2}$. If the required Q is greater than this value, the formulas given in Section 3 will be valid; if, however Q is less than $\sqrt{G/2}$, the formulas given in Section 4 should be used. It should be noted that a transition takes place in which both types of calculation are valid. This will be explained in Section 5.

3. FILTER CALCULATION WITH A $Q \geq \sqrt{G/2}$

The two capacitors C 1 and C 2 should be readily-available, standard values which can be based on the following order of magnitude:

f_o in the order:	1 Hz	10 Hz	100 Hz	1 kHz	10 kHz
C 1, C 2 approx.	10 μF	1 μF	0.1 μF	10 nF	1 nF

The value of the capacitance and the centre frequency result in a value of approximately 10^{-5} when multiplied with another. It is now possible using the known capacitor values and the previously determined requirements made on the filter to establish the required values of resistors R 1 to R 3 according to the following formulas:

$$R\,1 = \frac{Q}{G \times \omega_o \times C\,1} \qquad (5)$$

$$R\,3 = \frac{Q\,(C\,1 + C\,2)}{C\,1 \times C\,2 \times \omega_o} \qquad (8a)$$

or

$$R_p = \frac{1}{Q \times (C\,1 + C\,2)\,\omega_o} \qquad (6)$$

$$R\,3 = G \times R\,1\,\left(1 + \frac{C\,1}{C\,2}\right) \qquad (8b)$$

$$R\,2 = \frac{R\,1 \times Rp}{R\,1 - Rp} \qquad (7)$$

It is far simpler when the same value is selected for C 1 and C 2, e.g. when C 2 = C 2 = C. This then results in:

$$R\,1 = \frac{Q}{G \times \omega_o \times C} \qquad (9)$$

$$Rp = \frac{1}{2 \times Q \times C \times \omega_o} \qquad (10)$$

R 2 should be calculated according to formula (7).

$$R\,3 = \frac{2\,Q}{\omega_o \times C} \quad \text{or simplified} \quad R\,3 = G \times R\,1 \times 2 \qquad (11)$$

It is even possible to further simplify these already simplified equations if the gain is assumed to be G = 1. At this gain value equation (9) will be simplified to:

$$R\,1 = \frac{Q}{\omega_o \times C} \qquad (12)$$

and equation (11) will then be R 3 = 2 x R 1 (13).

Whereas R 2 can be calculated according to equation (7).

4. FILTER CALCULATION WITH $Q \leq \sqrt{G/2}$

If the required Q is less than the value $\sqrt{G/2}$, it is not possible for the calculation to be commenced with the capacitors but is necessary for R 1 and R 2 to be determined (e.g. with 10 kΩ). R 3, C 1, and C 2 should now be calculated according to the following equations:

$$R\,3 = \frac{G \times R\,1}{1 - \frac{Q^2 \times Rp}{G \times R\,1}} \qquad (14)$$

$$C\,2 = \frac{G/Q \times R\,1}{Rp \times R\,3 \times \omega_o} \qquad (15) \qquad = \qquad \frac{G \times R\,1}{Q \times Rp \times R\,3 \times \omega_o} \qquad (15a)$$

$$C\,1 = \frac{Q}{G \times R\,1 \times \omega_o} \qquad (16)$$

A special simplification is possible with this method of calculation: It is possible for R 2 to be calculated as ∞ e.g. deleted. In this case R 1 will be equal to Rp. This allows the equations to be simplified in the following manner:

$$R\,3 = \frac{G \times R\,1}{1 - \frac{Q^2}{G}} \qquad (17) \qquad\qquad C\,2 = \frac{G}{Q \times R\,3 \times \omega_o} \qquad (18)$$

however, the formula for calculating C 1 is not modified (16).

A further simplification is also possible here if the gain is assumed to be G = 1. The following equations are then valid:

$$R\,3 = \frac{R\,1}{1 - Q^2} \qquad (19) \qquad\qquad C\,2 = \frac{1}{Q \times R\,3 \times \omega_o} \qquad (20)$$

Equation (16) remains still valid for C 1.

5. TRANSITION BETWEEN THE TWO METHODS OF CALCULATION

If equation (9) is divided by equation (10) the following will result:

$$\frac{R\,1}{Rp} = \frac{2\,Q^2}{G} \quad \text{which means} \quad Q = \sqrt{\frac{G}{2} \times \frac{R\,1}{Rp}} \qquad (21)$$

Since Rp is the parallel connection of R 1 and R 2, Rp can never be higher than R 1 with positive resistances. Under the limit conditions of R 2 = ∞, it is only possible for Rp to be equal to R 1. In this case the following is valid from equation 21:

$$Q = \sqrt{\frac{G}{2}} \qquad (22)$$

If Q is to be less than this value, the calculated values for R 2 will become negative and can no longer be realized. This is why the preliminary calculation given in Section 2 is necessary.

$$Q \geq \sqrt{G/2} \qquad \text{and} \qquad Q \leq \sqrt{G/2}$$

If equation (16) is observed, which is valid for the second case, it will be seen that such a transition is present which is as follows:

Q^2/G may not be greater than 1, since the result of this equation will either be infinite or negative. The limit is therefore at:

$$Q^2/G = 1 \text{ or transposed: } Q = \sqrt{G} \qquad (22)$$

This means that both methods of calculation are possible between the limits $Q = \sqrt{G}$ and $Q = \sqrt{G/2}$; it should be remembered that the calculation according to Section 4. allows one resistor (R 2) to be deleted. On the other hand, the calculation according to Section 3 allows the calculation to be made with available capacitance values, which is often more important than the saving of one resistor.

To summarize, it is advisable to calculate in the following manner:

Use equations given in Section 3 with $Q = \sqrt{G/2}$
Use equations given in Section 4 with $Q = \sqrt{G}$

6. IMPEDANCE TRANSFORMATION

Generally speaking, values result from the equations that do not represent standard values. It is true that resistors are available with tolerances of better than 1%, however, these are not readily available to amateurs. This means that it is necessary to obtain the required resistance values by parallel and series connection, or to select the required value from a large number of large-tolerance resistors using the appropriate measuring equipment.

When using the calculation given in Section 3, it is possible for standard capacitance values to be used where C 1 = C 2 is usually selected. However, the resistors must be selected as above. If the calculation given in Section 4 is used, it is possible for the calculation to be based on the available resistors, however, usually very unsuitable capacitance values result.

Of course, it is also possible for the capacitors to be connected in series or parallel to obtain the required value. An impedance transformation can often be of assistance in this problem. In this case, all resistors can be increased by a factor "f", whereas the capacitance values will be divided by the same factor. With a little luck it will be possible to obtain standard values for the capacitors (at least for one of the two) and for another standard value to be used for the resistors since the latter are available in smaller steps than the capacitors. If this process is used correctly, it will have no effect on the characteristics of the filter. This factor "f" must be used for all frequency-determining components of the filter, e.g.:

$$R\,1' = f \times R\,1; \quad R\,2' = f \times R\,2; \quad R\,3' = f \times R\,3 \text{ and}$$
$$C\,1' = C\,1/f \text{ and } C\,2' = C\,2/f.$$

Example: C 1 has been calculated as being 53 nF, however, a capacitor of 47 nF is only available as the next standard value. This means that the calculated value for C 2 should be then reduced by the factor 47/53 whereas R 1 to R 3 should be increased by the factor 53/47 in order to maintain the required Q and centre frequency.

7. OPERATIONAL AMPLIFIER AND APPLICATION LIMITS

A prerequisite of the operation of the circuit is that the no-load gain of the operational amplifier is consideralby higher than the intrinsic gain of this circuit. This amounts to $2 \times Q^2$ (see ref. (5)) which indicates that the characteristics of the operational amplifier used are important especially when using high-Q filters and at higher frequencies. If, for instance, a Q of 10 is to be achieved, this means that the intrinsic gain of the circuit will be 200. In order to ensure that the errors are really negligible, it is necessary for the gain of the operational amplifier to be one hundred times higher, e.g. 20 000. This represents a value that conventional inexpensive operational amplifiers of the 741 series only achieve at very low frequencies (see minimum values given in the data sheets). It is, of course, an illusion to assume that the typical values are higher than the minimum values, since the cheapest types available on the market very often do not achieve the minimum values given on the data sheets. This means that special attention must be paid to the characteristics of the operational amplifier when high-Q filters are to be constructed. The deterioration of its no-load gain at 1 kHz is considerable when compared with the gain at very low frequencies. This means that higher-quality operational amplifiers must be used for filters with a higher Q, for example: Series SG (and LM) 218/318, or 218 A/318 A, or an operational amplifier that can be compensated externally or possesses wideband characteristics.

Another point must be considered: The input impedance of the amplifier. This must be considerably higher than the impedance of the external components which is in the order of Rp.

The input impedance of the 741 series of amplifiers is in the order of 500 kΩ. As a rule, the value Rp should be at least ten times, preferably 100 times lower: In this case less than 50 kΩ, preferably less than 5 kΩ. With a ratio of only 1/10 between Rp and R_{in}, measurable deviations will exist between the calculation and actual measurements, whereas these will be negligible at a ratio of 1/100. According to the demands and the application, the ratio between Rp and R_{in} should be selected so that the most favourable component values are achieved and that the calculated characteristics of the filter are sufficiently realized. Concrete designs are to be given in Part II of this article which offers several proven circuits. If unfavourable component values result from a calculation, it is possible for them to be brought to a more suitable order of magnitude using impedance transformation.

If the intrinsic noise of the operational amplifier could interfere with the amplification of weaker signals, it should be noted that the signal is firstly attenuated by the voltage divider comprising R 1/R 2 before the signal is selectively amplified in the filter. In this case, it is better to design the circuit so that no voltage divider is present at the input, e.g. when R 2 is deleted. This is achieved when a gain of $G = 2 \times Q^2$ is permissible (transposition of equation 22) and when a voltage divider is provided after the filter if this gain is not of importance at low signal levels.

Part II of this article is to describe several proven circuits as well as to describe an universal PC-board on which they can be accommodated. This will be brought in one of the next editions of VHF COMMUNICATIONS.

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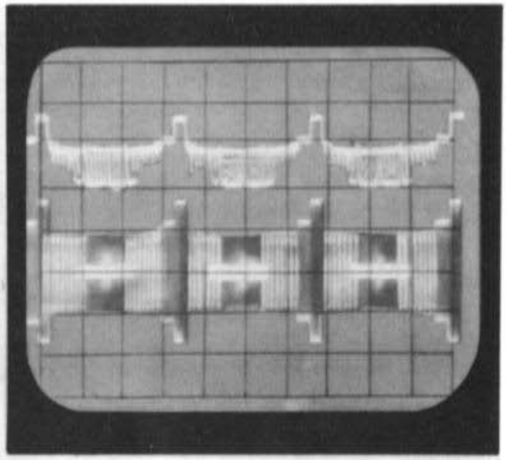


ANTENNA NOTEBOOK

FURTHER DETAILS ON CIRCULAR POLARISATION

by T. Bittan, G 3 JVQ / DJ Ø BQ

1. FURTHER EXPERIENCE WITH CIRCULAR POLARISATION

The author received quite a lot of feedback from readers regarding their experiences with circular polarisation and crossed Yagi antennas. As was to be expected, the results were very dependent on the locations, and confirmed the authors statement that: The poorer the VHF-UHF location, the greater will be the advantages of circular polarisation. It was, however, found that the crossed Yagi should be switched in all but the poorest location until circular polarisation is in general use, if the versatility of such antennas is to be used to the full, and if the most favourable polarisation mode is to be selected.

It has also been confirmed that it is very advisable to use non-conductive (glass-fibre) masts in order to maintain the circularity of the circular polarisation and not cause a reduction of the vertical component. This is also true for all vertically polarised Yagi antennas. In a large number of cases it was found that amateurs had not mounted their antennas with sufficient spacing to other antennas, or metal structures. This is very important with all Yagi type antennas, which are considerably affected by ground effects. A Yagi antenna requires a minimum spacing from ground or other conductive structures which is in the order of 3 m for a 144 MHz 10 element Yagi to 1 m for the smallest antennas.

The experience gained on 70 cm using circular polarisation has shown that although it is advisable to switch polarisations on 2 m, this brings hardly any advantage on 70 cm due to the far higher component of reflected signals that are present at 432 MHz. For this reason, the author would like to recommend that 70 cm crossed Yagis should only be switched between circular clockwise and anticlockwise, and not to linear polarisation.

2. FURTHER DEVELOPMENT OF SWITCHING UNITS FOR CIRCULAR POLARISATION

2.1. OPTIMIZING THE SWITCHING UNIT GIVEN IN (1)

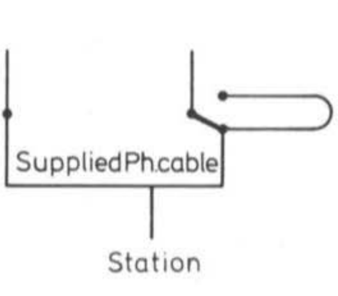


Fig. 1 : Circuit for selecting circular clockwise and anticlockwise polarisation

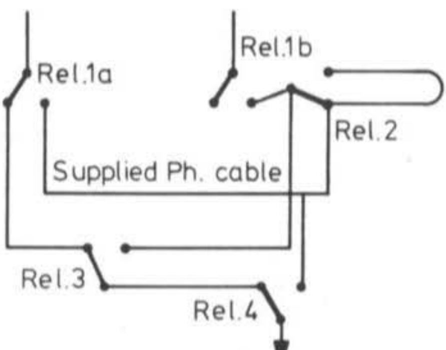


Fig. 2 : Circuit for selecting four polarisation modes

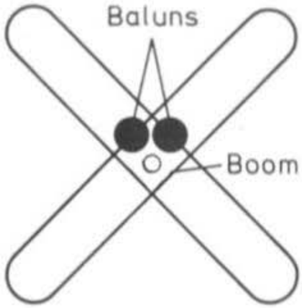


Fig. 3 : Correct mounting of the antenna to obtain given six polarisation modes

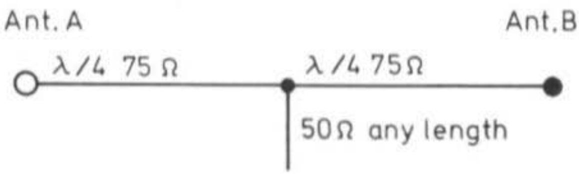


Fig. 4 : Phasing for vertical polarisation

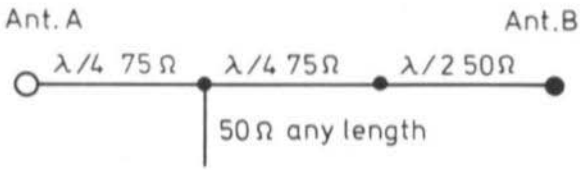


Fig. 5 : Phasing for horizontal polarisation

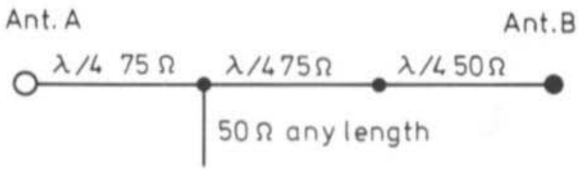


Fig. 6 : Phasing for clockwise circular polarisation

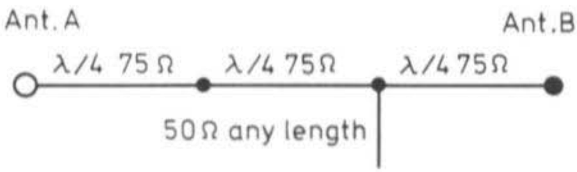


Fig. 7 : Phasing for anticlockwise circular polarisation

A basic switching unit for selection of four polarisation modes was described in (1). This switching circuit has been found to be very satisfactory in operation. One small modification could be made to really optimize the unit, which was not noticed until designing such a switching unit for higher frequencies. The switching of the additional $\lambda/2$ line for selecting anticlockwise circular polarisation should not be as shown in Figures 1 and 2 of (1) but as given in Figure 1. The difference is very small in practice, but the modified circuit does provide a more reliable phase condition.

2.2. A NEW SWITCHING UNIT FOR A MAXIMUM OF SIX POLARISATION MODES

Due to the difficulties in obtaining suitable relays and the non-availability of glass fibre masts at a reasonable price, the author continued to experiment with various methods of phasing in order to minimize these difficulties and reduce costs. Several different circuits were used such as λ and 1.5 λ hybrid rings with partial success. The final switching unit is relatively inexpensive and is very suitable for switching the polarisation in the shack using two equilength feeders to the antennas (and compensating for the stagger or spacing between the two individual dipoles). The author recommends that the following circuit should be used when the switching is made at the operating position whereas the circuit given in Figures 1 and 2 should be used for switching the polarisation at the masttop, or under the roof.

The main advantage of the new switching unit is that the crossed Yagi antenna can be mounted diagonally in the form of an "X" and still provide vertical and horizontal linear polarisation. The circuit is to be described with the crossed Yagi mounted in this position, but will be valid for other mountings assumed that the two baluns face away from another and that the antenna is mounted as shown in Figure 3, with the baluns mounted diagonally above the boom. If the antenna is mounted differently, this will lead to a different polarisation, but this is not serious since it is relatively easy to establish whether the linear polarisation is vertical, horizontal or diagonal after completing the switching unit. The six polarisation modes that can be selected are: Circular clockwise and anticlockwise as well as linear vertical, horizontal, diagonal to the left (135°) and to the right (45°). The latter are rather a biproduct which need not be used in practice.

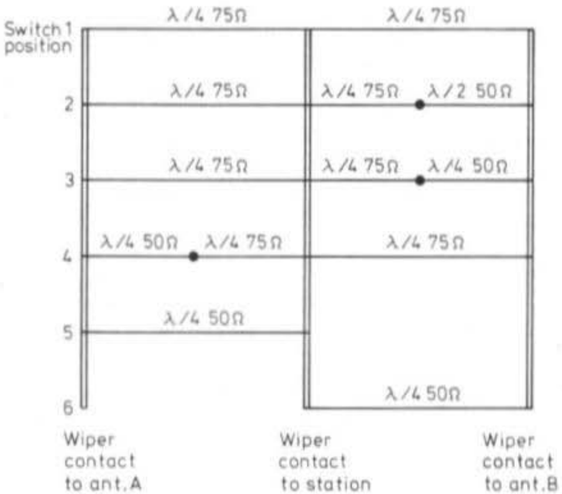
The required phasing lines or cables for the various polarisations are given in Figures 4 to 7. The two diagonal modes are obtained by selecting the required individual antenna without phasing it to the other antenna.

2.2.1. CONSTRUCTION OF THE SWITCHING UNIT

Solid dielectric coaxial cables should be used for the phasing cables due to the large variations in the velocity factor of semi-airspaced and foam dielectric cables. RG-58/U is suitable for the 50 Ω cables and RG-59/U used for the 75 Ω sections.

The electrical $\lambda/4$ and $\lambda/2$ lengths should be best measured with the aid of a dip meter in conjunction with a 2 m receiver. The best way of obtaining the required lengths is to cut off more than the required $\lambda/4$ or $\lambda/2$ times velocity factor and obtain the actual electrical length as follows: Make a small loop at one end of the $\lambda/4$ cable leaving the other end open. The loop should be as small as possible in order not to falsify the frequency indication; a loop of approximately 1 cm in diameter should be sufficient. Adjust the dip meter to find the dip. If the dip is below the 2 m band, it will be necessary to cut off some of the cable carefully until the dip occurs within the amateur band. If the same cable is used, the same mechanical length can be used for all the phasing lines of this impedance. Although the electrical $\lambda/2$ length can be measured in the same manner with a short circuit at the far end, the author prefers the following method due to the fact that it is not necessary to keep making the short circuit after cutting: The $\lambda/2$ length is found by reducing the frequency of the dip meter by half which means that the length is once again $\lambda/4$ at half the frequency and a dip will occur with an open far end. The harmonic of the dip meter frequency will still be heard on the 2 m receiver.

Until now it is only the required phasing cables that have been discussed. This has been made mainly to allow those readers that only require one or more specific polarisations to be able to construct the phasing cables so that they can be connected to both antennas directly without switching. This forms, of course, the cheapest method of changing from one polarisation to another. However, the majority of readers will wish to switch the polarisation with the aid of a rotary switch. This can be achieved using a six position, three wafer switch. If the two diagonal, linear polarisation modes are not required, then a four position, three wafer switch will be sufficient. The interconnection is very simply made by connection the cable from the station to the central wafer, and the antenna feeders (equi-length plus stagger difference) to the outer two wafers as shown in Figure 8. It should be mentioned that this switching unit is only suitable for crossed Yagis having a feedpoint impedance of 50 Ω. However, a different circuit for other impedance could be brought for other antennas in one of the next editions of VHF COMMUNICATIONS if required.



Polarisation modes (only when antenna mounted as in Fig. 3)

Position 1: Vertical
Position 2: Horizontal
Position 3: Circular, clockwise
Position 4: Circular, anti-clockwise
Position 5: Diagonal, 45°
Position 6: Diagonal, 135°

Fig. 8: Interconnection of the three-wafer switch

Due to the interest in such a switching unit, VHF COMMUNICATIONS will be bringing out a ready-to-operate switching unit to allow the selection of all polarisation modes which will be available at a reasonable price.

It is recommended that a ceramic wafer switch be used in order to keep dielectric losses at a minimum. Also attention should be paid that connection lengths are kept to a minimum. The use of cables having a diameter of 5 mm to 6 mm represents a good compromise between accommodating the cable within the box, and the power handling capability of the switching unit. The author suggests the use of RG-58/U and RG-59/U. However, thin cable RG-174/U or similar PTFE (Teflon) cables can be used for low power applications, and RG-213/U and RG-11/U for higher power applications (in the kW range).

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MEASUREMENTS ON A QUADRUPLE QUAD ANTENNA FOR THE 2 m BAND

by G. Schwarzbeck, DL 1 BU

A so-called quadruple quad antenna was described in Edition 2/1971 of VHF COMMUNICATIONS. This antenna is extremely light and suitable for portable operation. The author was able to make a number of measurements on this antenna. Unfortunately, not all measurements could be made due to lack of time. However, diagrams are to be given showing the gain, front-to-back ratio and matching relationships as a function of the measuring frequency. The results are then to be briefly evaluated.

Since the original description is nearly four years old, the original article in (1) may no longer be available. For this reason, the main construction of the antenna is shown in Figure 1. As can be seen, the antenna consists of a group of four quad antennas stacked one above the other with a similar reflector system.

1. CONSTRUCTION OF THE QUADRUPLE QUAD ANTENNA

The practical construction of this antenna is similar to a rope-ladder, where the horizontal elements are made from thin brass tubes of approximately 4 mm diameter and 520 mm in length (which differs slightly from the original description). Stranded wire of 2 mm in diameter is used for the vertical elements. The spacing between the horizontal bars is 500 mm. This results in four quad loops stacked one above the other. The reflector system is built up in a similar manner with a spacing of 520 mm from the main quad loops. The same type of construction is used with 560 mm tubes with a vertical spacing of 500 mm. The upper, centre and lower tube of the radiator and reflector are joined in the centre with the aid of 520 mm long brass tubes which can be connected to the mast.

Since the antenna can be easily packed by removing the metal bars from the mast and folding the vertical, stranded wire elements, it is extremely suitable for portable operation. It only weighs several hundred grams and is far more portable than a 10 element or 8 element yagi antenna of similar gain.

The polar diagrams, bandwidth and feedpoint impedance are very similar to that of a colinear group antenna, however, it is far easier to transport.

2. FEEDPOINT AND MATCHING

The centre, horizontal rod of the radiating element system is interrupted at the centre. The impedance at this point is in the order of 200 to 250 Ω. At power levels of up to approximately 30 W, the simplest means of matching to 50 or 75 Ω coaxial cable is to use an UHF core as shown in Figure 1. This is made from two turns of a thin 120 Ω twin lead which is connected in parallel at the low impedance side and in series at the antenna. Such cores can often be obtained from older VHF-FM or TV-receivers.

A resonance-type balun can also be used. This comprises two air-spaced coils wound one above the other having a diameter of approximately 8 mm diameter with two turns at the 50 Ω end and 5 turns for the 200 Ω winding. A trimmer

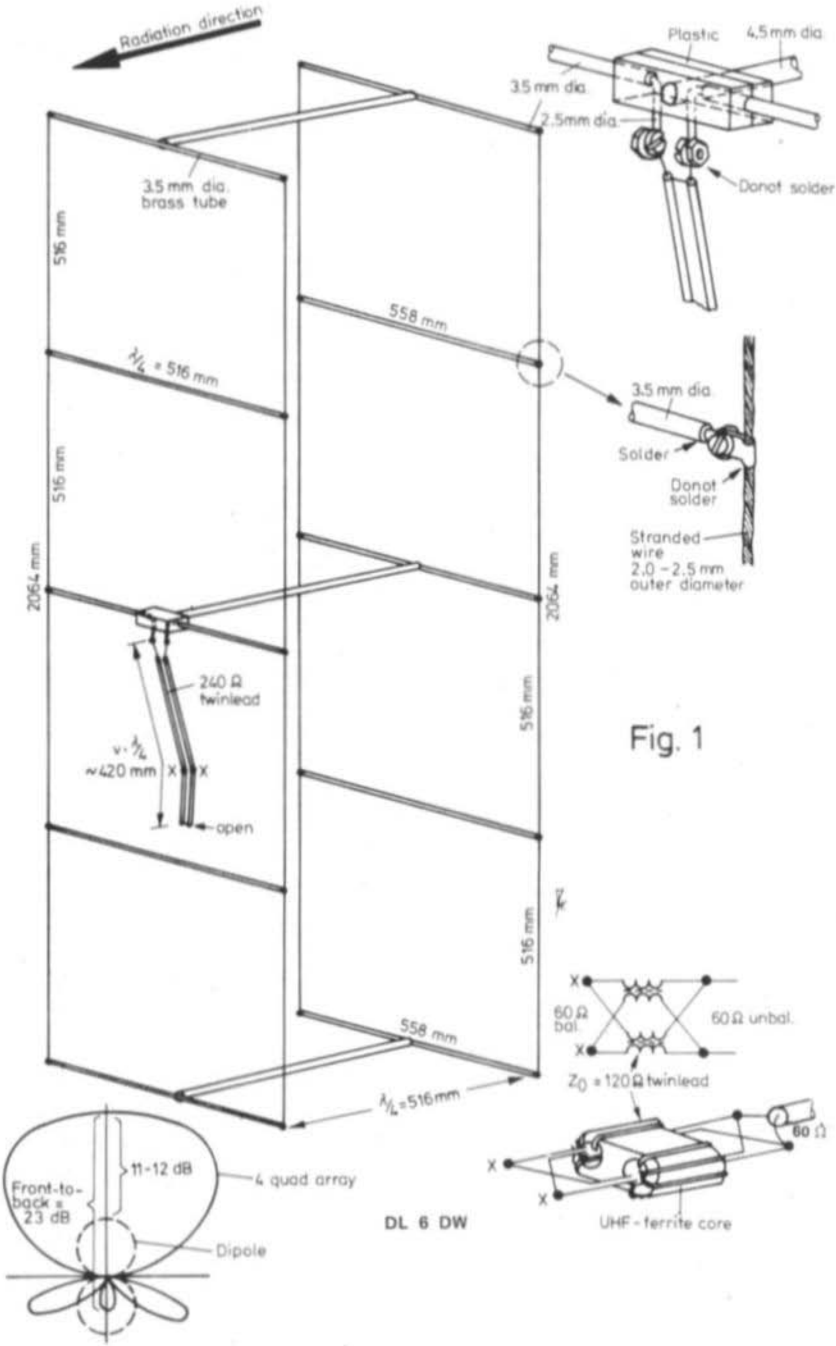

Fig. 1: The quadruple quad antenna

capacitor of 40 pF is connected in series with the low-impedance winding; a 10 pF trimmer is connected in parallel with the 5 turn winding connected to the feedpoint of the antenna. These two trimmers allow the reactive component to be tuned out and the circuit brought to resonance.

For higher power levels, it is possible for a coaxial half-wave balun to be used. This is made from a 80 to 95 cm long piece of 75 Ω coaxial cable (the exact length is dependent on the velocity factor of the actual coaxial cable). The two inner conductors are then connected to the feedpoint of the antenna. The feeder cable is connected so that its inner conductor is connected to one side of the antenna feedpoint and the outer conductor connected to the outer conductor of the coaxial balun loop. All these baluns transform the voltage by a ratio of 2 : 1 and the impedance by 4 : 1.

Due to the low weight of the antenna, readers will often use a thin coaxial cable. If only low power is to be used, it will be possible for a 2 to 3 mm teflon (PTFE) coaxial cable to be used. However, its length should not exceed 5 m in order to keep losses to a reasonable level. If a longer cable is required, the light-weight teflon cable should be kept as short as possible and extended using a thicker cable. Of course, it is not worthwhile paying exaggerated attention to an extremely low standing-wave ratio or exact resonance of the antenna and then to loose the small advantage of the exact matching due to cable losses. As can be seen in the first diagram given in Figure 2, even the large mismatch condition of directly connecting an unbalanced 50 Ω cable to the balanced 200 Ω feed-point of the antenna only causes a loss of 3 dB (half and S-point). The same loss exhibited by an 8 m length of thin teflon cable and perfect matching.

3. MEASURING SYSTEM USED

The actual antenna measurements were made on a flat, reflection-free area. The measurement of gain and front-to-back ratio were measured against a precision halfwave dipole using an automatic attenuation measuring system. The same measuring system was also used for the matching measurements using a professional 50 Ω directional coupler with a constant coupling attenuation. The signal source comprised a tracking generator, which was locked to the momentary frequency of a spectrum analyzer.

By selection of a small analyzer bandwidth, it was possible to virtually suppress all interference effects. A further advantage is the extremely logarithmic level indication, which is given on a linear dB-scale as well as the linear frequency scale. A XY-recorder was connected to the outputs of the spectrum analyzer and calibrated by a crystal-controlled spectrum in the X-direction and by a calibrated attenuator in the Y-plane.

The measurement is made in the form of a four-pole measurement showing the total loss between the input of the transmit antenna and the outputs of the antenna under test. The cable losses and system errors are not present since the same cable is used during the measurement as is used for calibration, and since most measurements are compared to a halfwave dipole. A wideband dipole antenna was used as transmit antenna.

4. MEASURING RESULTS

The first diagram (Fig. 2) shows the system loss as a function of the measuring frequency using a wideband transmit dipole and the quadrupical quad antenna. The antennas were spaced 20 m from another and the mean height was 4 m.

The coupling attenuation between the two antennas was 28 dB within the 2 m band when using the exactly tuned resonance balun; it amounted to 29 dB with the ferrite balun and was 32 dB when directly connecting the 50 Ω coaxial cable to the antenna without balun. This measurement shows that even an extremely high standing-wave ratio of approximately 6 : 1 only results in a loss of half an S-point, which is the same as when using a coaxial cable with 3 dB loss and perfect matching.

The second diagram (Fig. 3) shows the gain of the quadruple quad antenna of approximately 8 dB when compared to the reference dipole. However, although virtually practically dístant conditions with respect to the ratio of magnetic and electrical field components exist at the relatively low measuring distance of 10 wavelengths, no homogenius wave front exists due to the effects of the ground reflection. This results in a lower gain indication than would be present at greater distances and greater antenna heights. This means, that the gain of the quadruple quad antenna is in the order of 10 dB, which is extremely good for such a compact antenna and is only a few dB less than the larger colinear group antennas or long yagis.

The dot-dashed curve in Figure 3 gives the front-to-back ratio as a function of frequency. In contrast to the wideband characteristic of the gain curve (virtually constant from 130 to 160 MHz), the front-to-back ratio is relatively sharp and concentrated to the 2 m amateur band. This means that the design of the antenna is correct. The dashed diagram indicates the very high, and virtually frequency-independent attenuation to the side of the quad group.

Several things can be learned from such diagrams. It is possible, for instance, to establish that a receiver having a spurious reception, say in the 160 MHz band will receive such transmissions with an attenuation of 10 dB from the rear of the antenna and 28 dB from the side. In the aeronautical frequency band (130 MHz), the receiver will be provided with just as much energy from the rear of the antenna as from the front. A yagi antenna is far more selective; the gain usually falls off by at least 40 dB above its operating frequency so that a spurious transmission from an amateur transmitter does not usually cause so much TVI in TV-band III when using a yagi as when using a simple dipole.

The third diagram (Fig. 4) shows the matching as a function of frequency. The directional coupler was calibrated with matching standards so that the standing-wave ratio can be read off directly.

As was to be expected, the best results were obtained using a resonance balun that can be aligned both with respect to impedance and frequency.The resonance, could, of course, be brought completely into the 2 m band by slightly reducing the value of the parallel trimmer. When using the wideband ferrite balun, the resonance frequency is slightly too low which was mainly caused by an inductive component on connecting the balun (connection wires too long with too large a spacing). The dashed line indicates the extreme mismatch condition (SWR

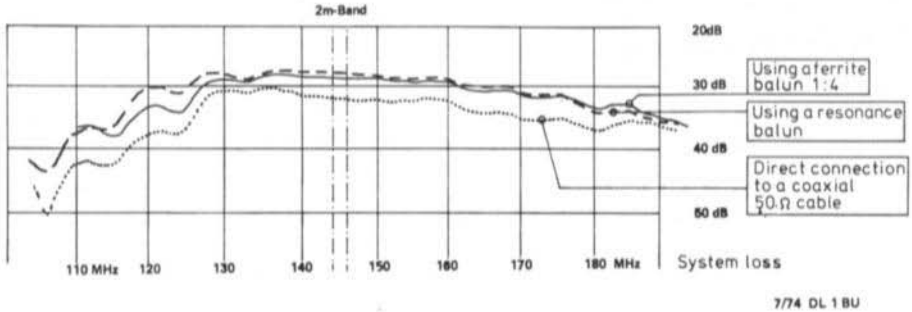

Fig. 2 : The quadruple quad antenna using different types of matching. The diagram shows the system loss between a horizontal-polarized transmit dipole at a height of 3 m and the quad group as described by DL 6 DW at a height of 4 m.

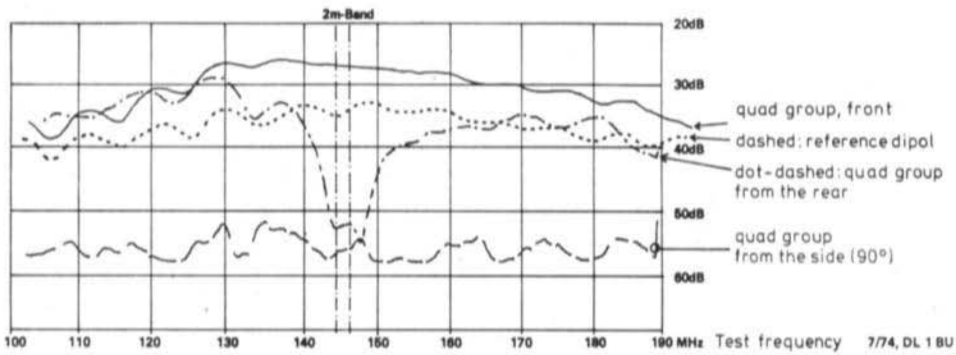

Fig. 3 : Gain and front-to-back ratio of the quad group. System loss measurement between transmit dipole at a height of 3 m and the receive antenna (quad group) at a height of 4 m and a spacing of 20 m. Dashed line: Reference half-wave dipole, horizontally polarized.

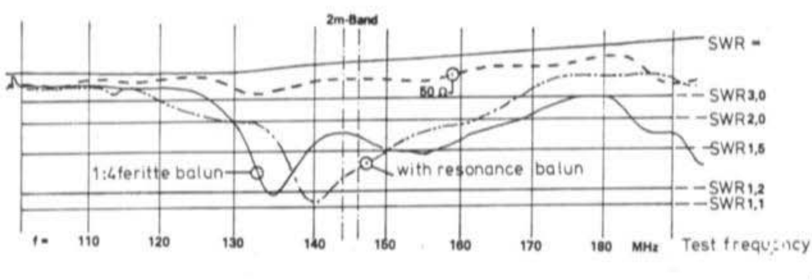

Fig. 4 : Matching diagram of the quad group. The diagram gives the standing wave ratio as a function of frequency with direct connection of an unbalanced 50 Ohm cable (dashed), using a 4:1 ferrite balun (50: 200 Ohm) (continuous line) and using a tuned resonance balun (dot-dashed line).

approx. 6) when directly connecting the 50 Ω cable (300 Ω : 50 Ω). It will be seen that the loss is only in the order of 3 dB.

The radiation diagram was found to be similar to that given in the original article. It will be seen that the antenna possesses a relatively wide horizontal diagram and that the gain is obtained with the aid of the sharp vertical beamwidth. This is often favourable in practice where a large number of stations will not be heard due to the narrow beamwidth of longyagi antennas.

This article was designed to firstly indicate the excellent suitability of this antenna for portable operation due to its low weight and small dismantled dimensions as well as to give general details regarding the measurement of VHF antennas in general.

5. REFERENCES

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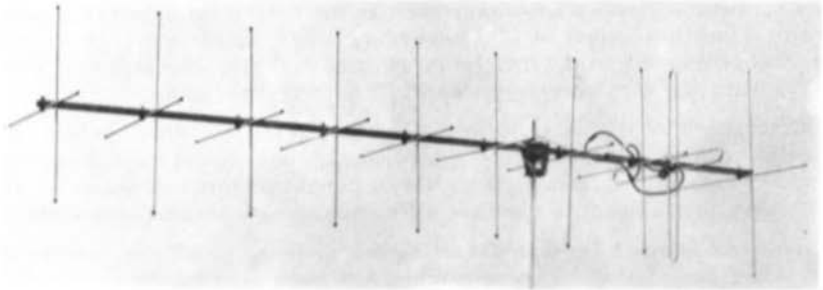


A LONG YAGI ANTENNA FOR 1296 MHz

by R. Lentz, DL 3 WR

A long yagi with a boom length of 4 λ (1 m) is to be described which is suitable for operation on the 24 cm band. The author has constructed several such antennas and has used them for portable operation and for home station use. The development was not made by the author but is based on a design given in (1) which was modified by the author. Since this original article went virtually unnoticed, even for German amateurs, and since no antennas are available on the market for this band it was thought that this antenna would be of interest.

The following data were given for the antenna in the original article:

Gain: approx. 13.5 dB
Front-to-back ratio: approx. 22 dB
Hor. beamwidth: approx. 33-34°
Recommended stacking distance: 385 mm.

A gain of approximately 13 dB has been confirmed during several comparison measurements with other antennas, such as the 6-element colinear antenna with reflector plate described in (2). According to the information given in (3) the horizontal beamwidth would then be in the order of 37°. The standing-wave ratio was 1.5 with the first prototype and 1.2 with the second.

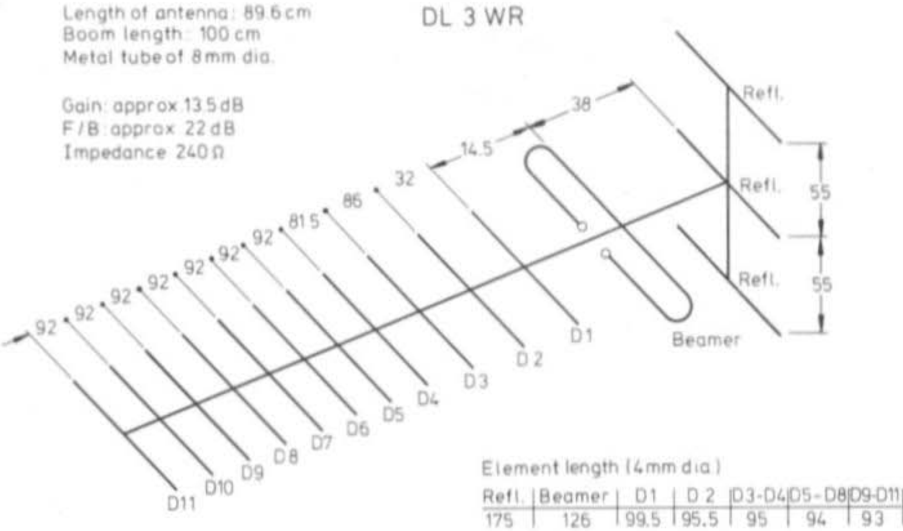
The dimensions of the antenna are given in Figure 1, which is not to scale. The small spacings between the folded dipole and directors D 1 and D 2 are typical for long yagi antennas. It should be noted that these dimensions are only valid when a metal boom is used and all elements have mechanical contact to it.

Steel tube can be used for manufacturing the antenna, which can then be copper-plated after completion. Such an antenna has been used by the author for mountain top portable operation and has stood up to some very hard treatment. For fixed-station operation, the antenna can be made more favourably from brass rod or tubing and lacquered after construction. In the case of the steel version, the elements are hard soldered to the boom, whereas the brass elements can be soldered into place using normal solder.

The most difficult part of the construction is the centre point of the reflector elements. In order to ensure that sufficient material ramains after drilling, the diameters of the tubes is as follows: Boom diameter = 8 mm, vertical rod of the reflector = 6 mm, all elements 4 mm in diameter.

The spacing between the upper and lower parts of the folded dipole amounts to 23 mm; a spacing of 2.5 mm is present at the feed point, where the balun transformer can be soldered or screwed into place after which it should be sealed in a water-tight manner. The folded dipole is fixed into place with the aid of a screw; this means that the folded dipole can be silver-plated, and mounted by sliding it over the various directors. (It would also be possible to modify the dimensions of the folded dipole so that the printed balun DJ 5 XA 002 described in (2) can be used).

It is in these details that the antenna differs from the original description, since hardly any details regarding material thickness and the dimensions of the folded dipole were given. However, the element lengths are given in the original description. Figure 2 shows a photograph of the launching elements of the author's prototype.



Element length (4 mm dia.)

Refl.	Beamer	D1	D 2	D3-D4	D5-D8	D9-D11
175	126	99.5	95.5	95	94	93

Fig. 1: Dimensions of a 13 dB long yagi for the 24 cm band

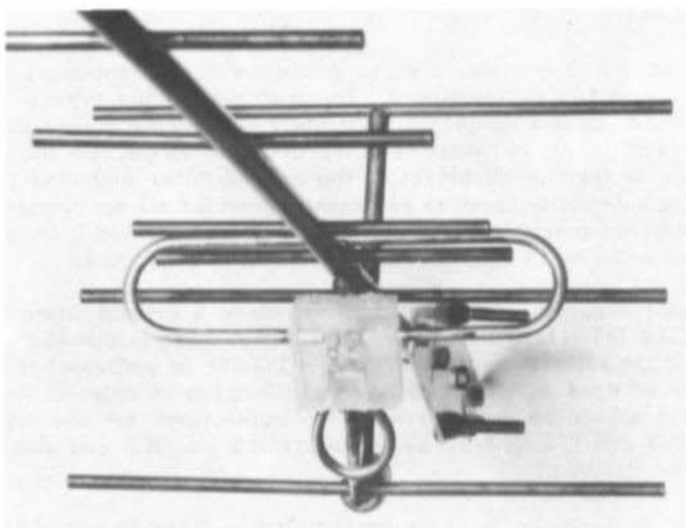


Fig. 2: Launching elements of the long yagi antenna with waterproofed balun

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FUNDKTECHNIK, Band 22 (1967), No. 24, Pages 941-944

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VHF COMMUNICATIONS 6 (1974), Edition 2, Pages 82-84

SSB/CW IF MODULE AND AGC CIRCUIT

by D. E. Schmitzer, DJ 4 BG

A module is to be described that can be used to complement the previously described units of the modular receiver. It consists of an IF-module and AGC circuit. The AGC voltage is obtained from the AF voltage provided by the product detector DJ 4 BG 014.

1. SSB/CW IF MODULE

As was mentioned in (1), an additional IF amplifier will not be necessary if the high sensitivity of the product detector is used to the full, and when no automatic gain control is required. However, an IF amplifier offers three advantages:

a) more IF gain in order to provide the product detector with feedback so that further reserves are available for the automatic gain control.

b) Adding the IF amplifier to the AGC line extends the control range in conjunction with module DJ 4 BG 011.

c) Ensures sufficient drive power for the product detector of module DJ 4 BG 014 even when the amplifier stages of the intermediate frequency chain are fully controlled.

The circuit of the IF-module is given in Figure 1. It contains a few special features which should be mentioned. The matching to the crystal filter using resistor R 2 results in a loss of approximately 6 dB. This means that the total insertion loss of the crystal filter is in the order of 10 dB. On the other hand, the advantage is that no alignment of the crystal filter matching is required. The resulting selectivity curve is sufficiently good for all applications and only when the required measuring equipment is available, would it be possible for the filter curve to be corrected with the aid of C 1 and C 2.

A further note regarding the crystal filter: Since a crystal filter is provided in module DJ 4 BG 011, a further crystal filter at this position is not absolutely necessary. If the selectivity of module DJ 4 BG 011 is sufficient to satisfy the demands placed upon it, the second crystal filter can be deleted. In this case, a wire bridge should be made between the connections for the crystal filter; capacitors C 1 and C 2 as well as resistors R 2 and R 3 can also be deleted (R 2 must be bridged).

If a multi-mode receiver is to be constructed, a filter having a larger bandwidth (FM filter) can be provided in module DJ 4 BG 011 and the required bandwidth for SSB or CW can be provided in module DJ 4 BG 013. In order to satisfy especially high demands in a purely SSB receiver, a crystal filter type XF-9B can be used in module DJ 4 BG 011 and a XF-9A or another XF-9B in module DJ 4 BG 013. In the case of a purely CW receiver, a crystal filter XF-9M should be used in both modules which will result in a bandwidth of approximately 400 Hz and an extraordinary high selectivity.

The first IF stage of the module is equipped with a dual-gate MOSFET and the gain is controlled at low level with the aid of the gate 2 voltage. In contrast to the RF-stage in module DJ 4 BG 011, gate 1 is not controlled. This results in a faster risetime of the control voltage and a larger control range. In this

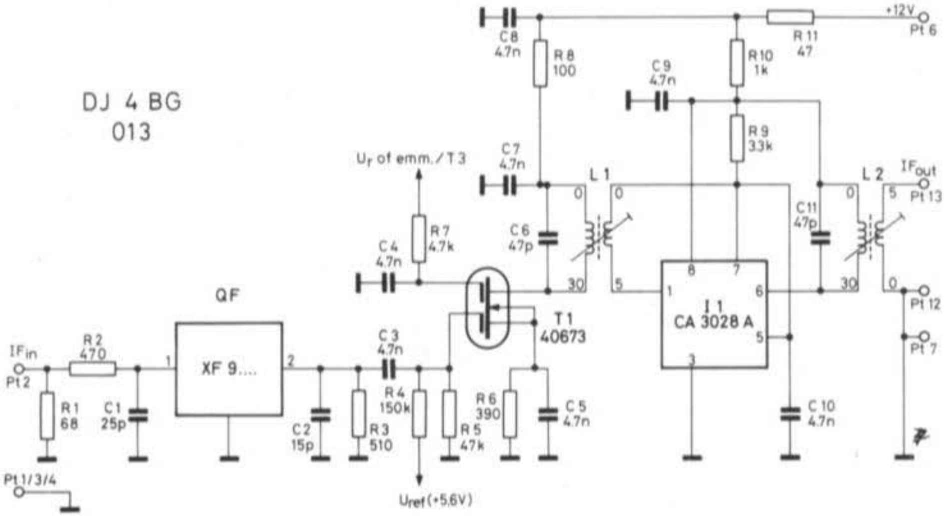


Fig. 1: Circuit of the IF amplifier in module DJ 4 BG 013

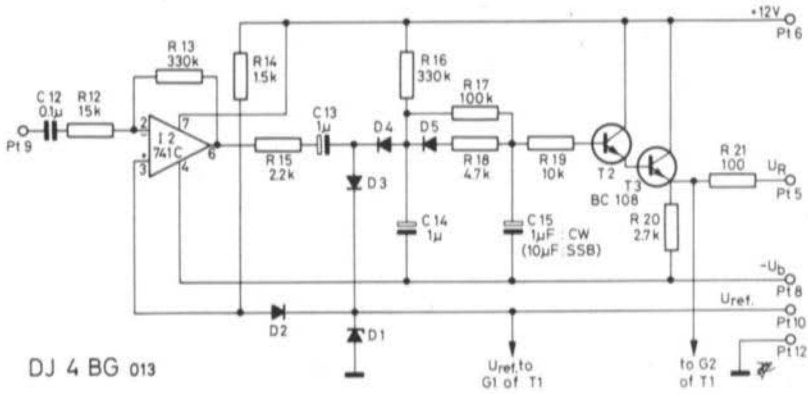


Fig. 2: Control voltage circuit of module DJ 4 BG 013

manner, it is possible for this stage to have a large effect on the control characteristics of the receiver so that the RF-stage of module DJ 4 BG 011 need only be controlled to any great extent when a good signal-to-noise ratio is provided. Since it provides a stable gain with a very low extent of external components, the integrated circuit CA 3028 A is used in the output stage of the IF-module. The amplified IF-signal is finally fed via a coupling winding to the output of the module (Pt 13). The overall gain is in the order of 40 dB including the insertion loss of the crystal filter and its matching.

2. AGC CIRCUIT

The following demands are placed on the automatic gain control circuit:

A quiescent voltage of approximately +5 to +6 V must be provided under no-signal conditions in order to allow the controlled IF stages to operate at full gain. With an AF-signal that is somewhat below the maximum drive of the AF output stage, the control voltage should be in the order of -2 V in order to block the dual gate MOSFETS and reduce their gain to a sufficient degree. It is especially important in the SSB mode that the rise-time of the control voltage is very quick, whereas the fall-time is slow.

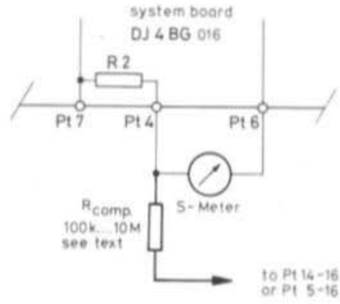
Figure 2 shows the circuit diagram of the module. The AF voltage from the product detector is amplified in the operational amplifier I 2 and rectified in a voltage-doubler circuit comprising diodes D 3 and D 4. The DC voltage obtained in this manner is dependent on the amplitude of the AF voltage. It is filtered and finally fed via the Darlington-stage comprising T 2 and T 3 to Pt 5.

Now a few notes regarding parts of the circuit whose operation will probably not be seen immediately from the circuit diagram: Diode D 2 is used to shift the voltage level of the operational amplifier by approximately +0.7 V. This is done firstly to increase its drive-range somewhat so that its output DC voltage is more centrally positioned between 12 V and -2 V; secondly it ensures that C 13 is not incorrectly poled under no-drive conditions. The voltage across diode D 3 and D 4 is discharged via the base-emitter diodes of transistors T 2 and T 3 so that the quiescent output voltage of the control circuit is practically equal to the reference voltage of D 1. This results in a good temperature stability of this voltage since it is only the low temperature coefficient of diode D 1 that remains effective. Only a small voltage difference in the order of 0.1 V remains between the reference voltage and the control voltage output under quiescent conditions, which can be neglected with a control voltage in the order of more than 6 V. A level-meter (S-meter) that is to be connected between these two voltages should have a full scale deflection of maximum 0.5 mA (dropper resistor approx. 12 kΩ); however, it is more favourable for a 100 μA-meter to be used (dropper resistor approx. 60 kΩ).

The previously mentioned error of approximately 0.1 V can be compensated using a high-value resistor as shown in Figure 3. Since the value of this resistor is dependent on the amplitude of the residual voltage and on the impedance of the meter, no exact values can be given. It is necessary to find the value experimentally which should be in the order of 100 kΩ to several MΩ. According to the polarity of the error voltage, the resistor should be connected to +12 V or -2 V.

Diode D 5 and resistor R 18 provide the required characteristics of the AGC voltage. A fast rise-time of the signal voltage will cause a rapid discharge of C 15 via D 5, R 18, whereas the higher impedance path via R 16 and R 17 is only available for the recharge cycle. Suitable selection of the previously mentioned components allows the most favourable control characteristics to be achieved for any application. The values given in the circuit diagram (Fig. 2) should be most favourable for CW, whereas the filter capacitor C 15 should be increased to 10 μF or 22 μF in the SSB-mode. If the fall-time is still considered to be too short, the value of resistor R 17 can be increased from 100 kΩ to 220 kΩ or more.

Fig. 3 :
Zero-point
adjustment
of the S-meter



3. MEASURED RESULTS

The two following diagrams give the characteristics of the described module, or rather the characteristics of a receiver using this module:

Figure 4 shows the relationship between AF voltage fed to the AGC circuit and the control voltage present at connection Pt 5. It is referred to the reference voltage (Pt 10) and measured at various values of the input resistor R 12. The absolute values are given in the second scale.

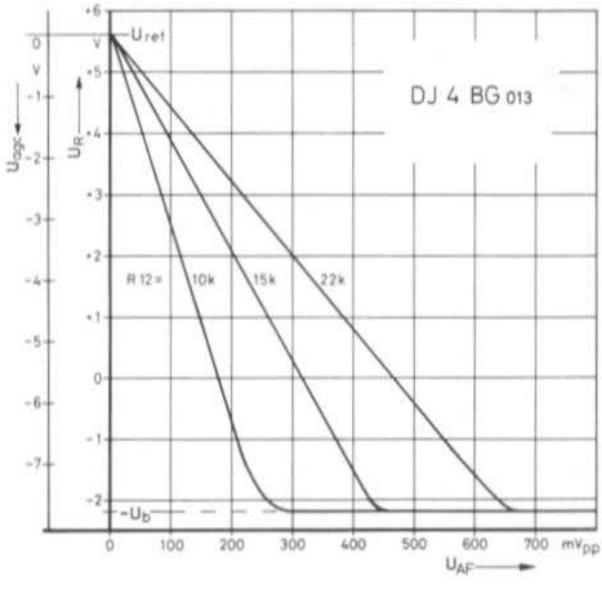


Fig. 4 :
Control voltage
as a function
of the AF-signal
(1 kHz)

Curve 1 of Figure 5 shows the relationship between the IF-input voltage and the AF-signal using a combination of modules DJ 4 BG 013 (IF-module) and DJ 4 BG 014 (product detector). In this case, only one stage of module DJ 4 BG 013 was controlled (T 1). Curve 2 shows the relationship using a combination of modules DJ 4 BG 011, 013 and 014 where the IF voltage from the signal generator was fed to gate 1 of T 3 in module DJ 4 BG 011, and this transistor also controlled. In both cases, the feedback resistor in the product detector (R 4 - 14) was 100 Ω.

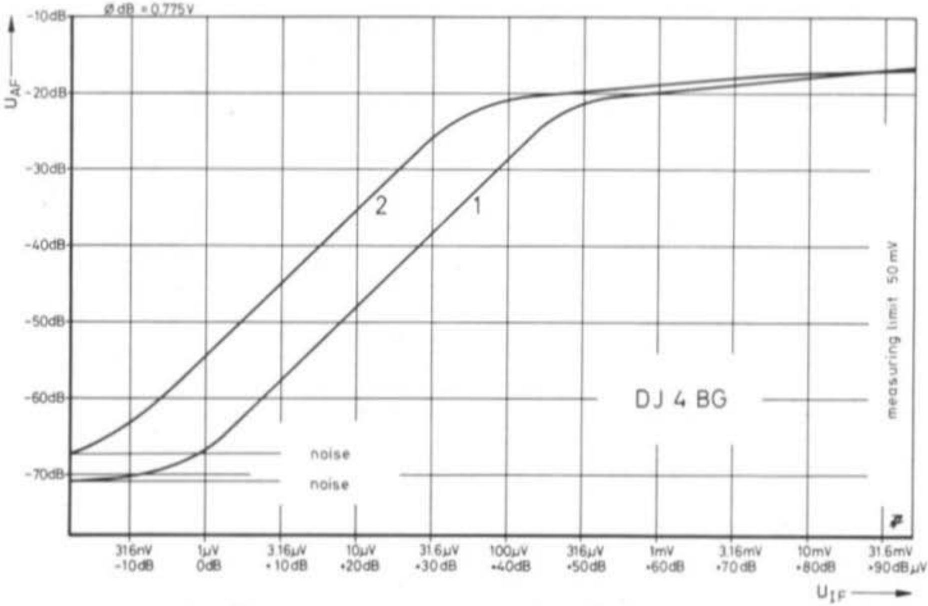


Fig. 5 : AF voltage as a function of the IF voltage of a receiver constructed from TEKO modules

Experiments have shown that this combination provides a sufficiently large overall gain on the "louder" amateur bands such as 80 and 40 m. On the 20, 15, and 10 m amateur bands, it may be necessary for the feedback resistor (R 4) in module DJ 4 BG 014 to be reduced to values in the order of 10 to 33 Ω. If the RF stage is not controlled in a receiver comprising these modules, the control characteristics of the overall receiver will correspond to curve 2 of Figure 5, if the X-axis of this curve is shifted to the amount of the RF and conversion gain after deducting the insertion loss of the crystal filter. If the RF stage in module DJ 4 BG 011 is also connected to the control line, a somewhat flatter curve will result, in other words, the control range will be increased.

4. CONSTRUCTION AND COMPONENT DETAILS

The PC-board DJ 4 BG 013 has been developed for construction of the described module. Figure 6 shows this PC-board and the component locations. With exception of R 11, all resistors are for 10 mm spacing, as are C 12 and D 1 to D 5. In the case of C 3, the connections should be spread to a spacing of 12.5 mm. Capacitors C 1, C 2, C 6 and C 11 are ceramic disc capacitors having a low-temperature coefficient; C 13, C 14 and C 15 are tantalium electrolytics, whereas all capacitors with a value of 4.7 nF are disc types. In other words, the components used are virtually the same as have been used for the other modules and units described in VHF COMMUNICATIONS.

I 1: CA 3028 A or CA 3053 (RCA)
I 2: 741 C (various manufacturers), TBA 221 B (Siemens)

T 1: 40673, 40820 (RCA) or similar dual-gate MOSFETS
T 2, T 3: BC 108, BC 183, BC 413 or any other silicon NPN audio transistors

D 1: 5.6 V zener diode, e.g. BZY 83/C5V6 or BZX 55/C5V6
D 2 - D 5: 1 N 914, 1 N 4148 or similar silicon diodes

L 1, L 2: Resonant circuit 30 turns of 0.2 mm dia. (32 AWG) enamelled copper wire in special coil set. Coupling winding 5 turns.

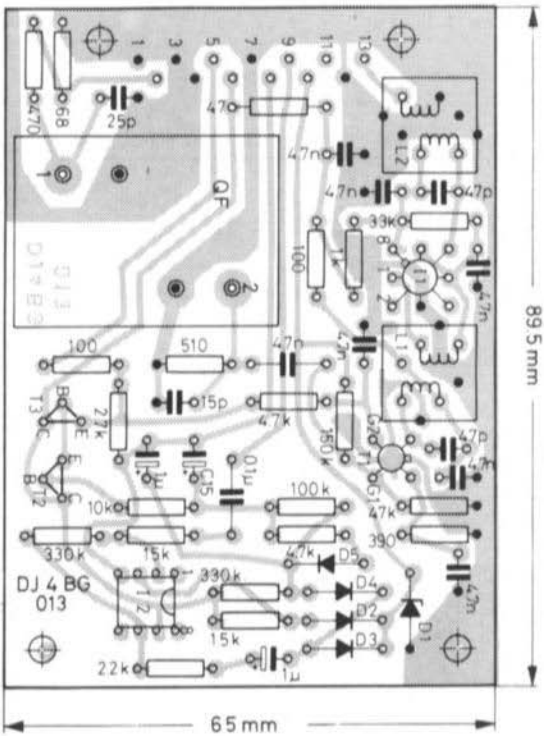


Fig. 6 : Component locations on PC-board DJ 4 BG 013

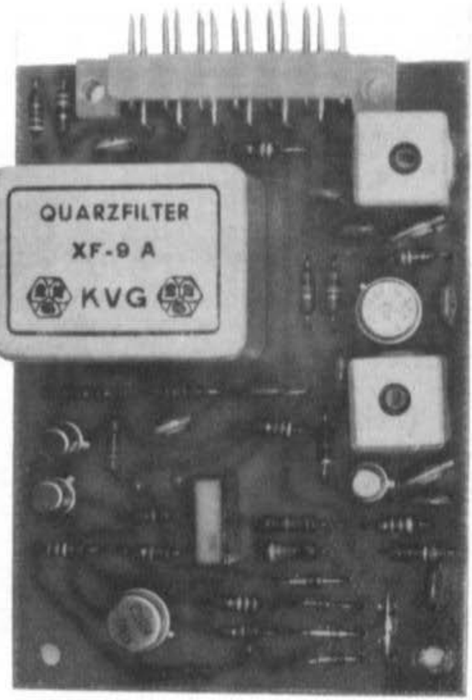


Fig. 7 : Author's prototype of module DJ 4 BG 013

The connections of the operational amplifier are arranged so that the plastic version in an eight-pin dual-inline case can be used.

Crystal filter: According to the required operating mode, XF-9A or XF-9B in the SSB-mode, XF 9 M in the CW-mode or an XF-9A if an XF-9M is provided in module DJ 4 BG 011.

5. REFERENCES

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VHF COMMUNICATIONS 6 (1974), Edition 4, Pages 220-229

USING THE PHASE-LOCKED OSCILLATOR DK 1 OF 011/014 FOR REPEATER/DUPLEX OPERATION WITH A FREQUENCY SPACING OF 1.6 MHz OR 0.6 MHz

by H. Hanserl, OE 5 AN

The phase-locked oscillator described in (1) was originally designed for simplex operation. If repeater operation is required, a switching arrangement can be used to provide the required 1.6 MHz or 0.6 MHz spacing. By correct selection of the crystal frequencies and tuning range of the VFO, it is possible to select these frequency spacings without modifying the phase-locked oscillator. The frequency plan and the required wiring is as follows:

The following table contains all the frequencies used for the European two meter band:

Frequency	Oscillator Freq.	Freq. Crystal	Crystal	VFO Freq. Range
144.0-144.6 MHz	135.0-135.6 MHz	130.0 MHz	65.0	5.0 - 5.6 MHz
144.6-145.2 MHz	135.6-136.2 MHz	130.6 MHz	65.3	" "
145.0-145.6 MHz	136.0-136.6 MHz	131.0 MHz	65.5	" "
145.6-146.2 MHz	136.6-137.2 MHz	131.6 MHz	65.8	" "

Figure 1 shows the wiring and switches required.

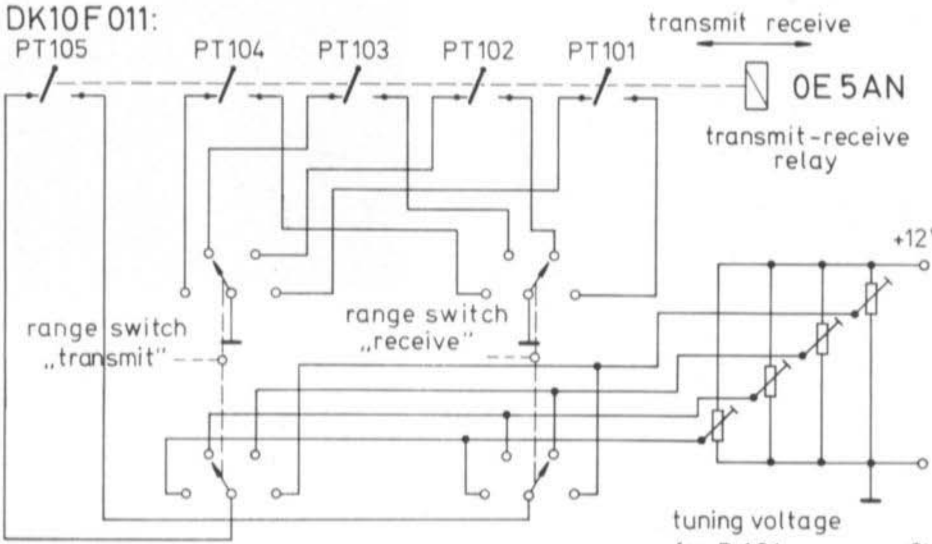

Fig. 1: Required switching for a frequency spacing of 0.6 or 1.6 MHz

This circuit differs from the original in that it possesses three other crystal frequencies and that the tuning range of the VFO has been increased to 600 kHz. In the receive mode, a higher-frequency crystal is used so that a crystal-controlled frequency spacing from the transmit frequency of 0.6 MHz, 1.2 MHz or 1.6 MHz is provided according to the crystal used. It is unfortunate that four scales are required for calibrating the VFO. However, this is also the case with most shortwave transceivers. The scale calibration is as follows:

145.6	145.7	145.8	145.9	146.0	146.1	146.2 MHz
145.0	145.1	145.2	145.3	145.4	145.5	145.6 MHz
144.6	144.7	144.8	144.9	145.0	145.1	145.2 MHz
144.0	144.1	144.2	144.3	144.4	144.5	144.6 MHz
5.0	5.1	5.2	5.3	5.4	5.5	5.6 MHz

Editors: The Frequency range can be extended to the North-American 2 m band by continuing the first table.

REFERENCES:

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VHF COMMUNICATIONS 6 (1974), Edition 2, Pages 114-124.

A LINEAR TRANSPONDER FOR AMATEUR RADIO SATELLITES

by Dr. K. Meinzer, DJ 4 ZC

The task of this article is to summarize the experience gained with the ballon-carried ARTOB and OSCAR satellite transponders and to indicate the state of the art using the 70 cm/2 m OSCAR 7 satellite transponder as an example.

1. SYSTEM CONSIDERATIONS

Transponders are required for telecommunication satellites that receive signals from the ground stations and retransmit these on a different frequency after providing the required amount of gain. Whereas with FM-repeaters only one station can operate at a time, telecommunications and amateur satellites are designed to allow a large number of stations to operate simultaneously via the satellite.

This is achieved by amplifying a complete frequency segment in a linear manner and transposing it (not demodulating it) to a new frequency segment. This allows each user to occupy a small portion of the frequency segment provided by the transponder.

The commercial telecommunication satellites use a different method to allow several ground stations to use the satellite simultaneously: Each user is provided with a time window which means that each of the ground stations is periodically switched to the satellite. Due to the synchronization problems involved in this method, it is not suitable for use for amateur radio applications. On the other hand, linear transponders have never been seriously considered for the commercial satellite services due to the danger of intermodulation products and power distribution. The experience gained by radio amateurs using linear systems has shown, however, that these problems have been exaggerated and can be solved. Of course, several conditions must be fulfilled which were firstly established during the ballon-carried ARTOB program.

The first ballon flights used a linear transponder that received signals in the lower part of the 2 m band and retransmitted them at the upper end of the band. The experience gained during the first few flights indicated that the interference was considerable, and although a large number of stations were heard only relatively few contacts were achieved.

The main problem was that hardly any stations could monitor their own transposed signal when single-band transposers were used. This means that virtually all transmissions were made "blind" and often more power was used than required. This problem was solved subsequently by using two different bands for transmit and receive. The question was then which frequency band should be used in which manner. The following technical considerations were made:

The path loss a between two isotropic antenna is:

$$a\ (\mathrm{dB}) = 22 + 20 \times \log D/\lambda$$

D = distance between the antennas

λ = wavelength

This means, for example, that the path loss is approximately 10 dB greater on the 70 cm band than for 2 m. On the other hand, higher antenna gains are possible on 70 cm with the same size of array than would be possible at 2 m. If this is considered using a parabolic antenna as an example, the following will be valid for the antenna gain g:

$$g\ (\text{dB}) = 7 + 20 \log d/\lambda$$
$$d = \text{diameter of the antenna}$$

This means that the higher gain of the parabolic antenna at the higher frequency is just able to compensate for the higher path loss. Since either omni-directional antennas or antennas with a prescribed beamwidth can be used on the satellite (gain does not increase with frequency), this will mean that the pass loss between the receiver and transmitter will not be frequency-dependent as long as the increase in path loss is compensated for at the ground station. As was mentioned in (1) an increase in gain (on increasing the frequency) will reduce the beamwidth so that such antennas must track the satellite.

If yagi antennas are considered instead of parabolic dishes, it will be found that it is hardly possible to achieve more than 5 dB more gain on 70 cm than on 2 m when using a reasonable amount of antennas and tracking system. This means that the 70 cm path has a disadvantage of approximately 5 dB.

The transmission path from the satellite to earth is usually less favourable due to the limited output power of the satellite. For this reason, the 2 m band was selected for the down path for both the ARTOB ballon-carried transponder and for the 70 cm/2 m transponder of OSCAR 7 (which is to be called H-transponder in the subsequent text).

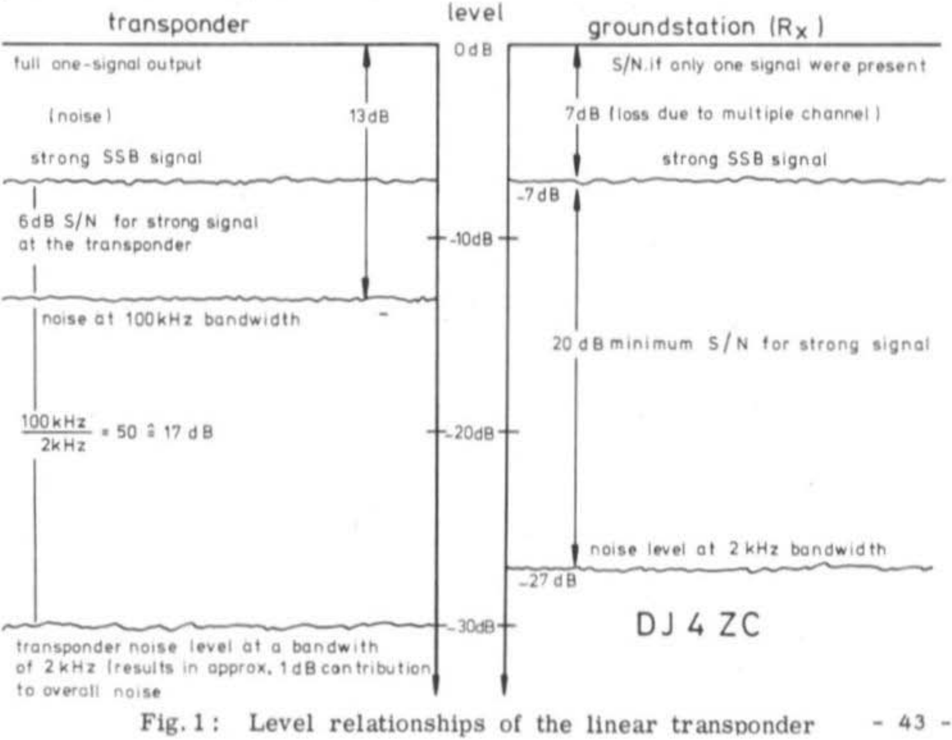


Fig. 1 : Level relationships of the linear transponder

Ballon flights using this concept allowed far better results to be obtained than using a 2 m/2 m transponder. Most stations were able to monitor their own, transposed transmission so that interference between stations was very seldom present. Furthermore, the monitoring of ones own signal allowed true break-in operation in the SSB mode in a similar manner to normal telephone conversation, and the good strength of the monitored transmissions ensured that the transmit power was kept to a minimum since low-power levels did not bring any subjective disadvantages.

2. LEVEL RATIOS

The level ratios present in linear transponders is now to considered (Fig. 1):

An amateur signal is usually considered to be strong when the signal-to-noise ratio is approximately 20 dB. In order to obtain the required psychological effect so that the output power of the ground stations is kept to a minimum, the strongest signals should at least achieve this level. At the same time, signals of 15 dB weaker stations will also be audible at a signal-to-noise ratio of 5 dB. In practice, the output power levels between amateur radio stations are usually less than 15 dB which means that it is possible for virtually all stations to operate via the transponder. Since the transponder must transpose a large number of signals simultaneously, the strongest signals are usually approximately 7 dB weaker than would be the case when the transponder is only used by a single station. This value was measured experimentally during the ARTOB ballon flights. Theoretically speaking, a larger ratio was to be expected. However, the low mean on-time of SSB and CW transmissions together with the inavoidable difference in power level seems to have a compensating effect. This means that the transponder should provide enough output power so that it can provide a total signal-to-noise ratio of 27 dB. The OSCAR 6 satellite was not able to fulfill these demands; this led to the use of higher output powers in the case of several ground stations which then caused interference to other stations and led to the fact that it could only be used by the few powerful stations.

This means that the levels have already been determined for the transponder. If it is considered that the up-path should not provide any considerable component to the overall noise, it is necessary for it to be approximately 3 dB better than the down-path. When referred to the full output power of the transponder the signal-to-noise ratio should be approximately 30 dB at a bandwidth of 2 kHz. The noise power P_n is:

$$P_n = k\,T\,B$$

k = Boltzmann's constant = 1.38×10^{-23} Ws/k
T = noise temperature of the receiver
(typ. 1000 k = 5 dB)
B = bandwidth (Hz)

The noise power is thus proportional to the bandwidth.

If a transponder bandwidth of 100 kHz is used, this will result in a signal-to-noise ratio of 13 dB to be provided in the case of a full transponder band. This will mean that a typically strong signal will only have a signal-to-noise ratio of approximately 6 dB. This shows how important the receiver of the ground station is for improving the signal-to-noise ratio.

Intermodulation products (IM) have a similar structure to noise. Since the internal signal-to-noise ratio of the transponder only amounts to 13 dB, these IM products will only interfere when they approach this order. However, since 25 dB IM rejection can be achieved relatively easily, these distortions can usually be neglected. Of course, such distortion products can be of a virtually line structure under unfavourable conditions so that they will be audible. However, experience gained with the ARTOB and with OSCAR 6 have shown that interference due to intermodulation effects is very seldom exhibited.

In order to maintain the previously mentioned level ratios, it is necessary for the transponder to be provided with a control system. This control circuit should have a short rise-time and low fall-time in order to avoid interference due to overload conditions and in order to suppress rapid-level fluctuations. A rise-time of 0.1 s and a fall-time of 2 s have been found to be suitable. The control range was limited to approximately 20 dB so that it would still be possible to operate via the satellite transponder if the control circuit became defective, and in order to ensure that any station using too high an ERP would not be able to completely block the transponder.

3. EFFICIENCY OF THE TRANSPONDER

The previously mentioned considerations are valid for all linear transponders. Several further demands are placed on satellite transponders: The main consideration is that a high efficiency is obtained, and it is well known that linear amplifiers possess principally a poor efficiency for two major reasons:

The demand for linearity requires that the amplifier operates with a certain amount of quiescent current (class AB/B). The efficiency obtained in this manner is considerably lower than when using class C amplifiers. However, the main problem is that the efficiency is directly proportional to the drive level. At the typical mean drive levels of a transponder of -6 to -10 dB, the efficiency will only be 30% to 50% that of the full output level which will be hardly ever achieved. If it were possible to solve this problem, it would be possible to reduce the current requirements or to increase the output power by factor four. This gain was such that it was worthwhile considering matters to achieve this aim.

3.1. ENVELOPE ELIMINATION AND RESTORATION

The method described in (2) where the signal is split into an amplitude (envelope) and phase component which is then restored at the output offers a solution to the above problems. When using this method, it is possible for the output transistor to be operated in class C and the collector voltage will be proportional to the drive due to the modulation; this means that the efficiency will not be dependent on the drive level. Up till now, this method has only been used for SSB applications and it was not known whether this method was at all suitable for wideband transponders. If it was suitable, it was now necessary to establish which design criterion were required.

In order to examine this complex, a computer simulation was made. The results of these calculations were that this method is able to provide good IM ratios with technically reasonable bandwidths. Three main functions are necessary to realize this aim that are not normally available in transponders:

1. A normal diode rectifier is used for obtaining the amplitude information. Attention must be paid that the frequency response of this rectifier is from DC to approximately five times the bandwidth of the transponder.

2. A limitation of the signal is required in order to obtain the phase-modulated component. The limiters used in most FM receivers usually have a time constant that is too slow for this application. The best method is to use two antiphase diodes connected in parallel with a resonant circuit. Two such limiters with intermediate amplification result in a very rapid limiter having a very low conversion from AM into PM.

   Overloaded differential amplifiers have not been found to be successful. It has been found that these amplifiers possess a very high conversion from AM to PM at higher frequencies. This is unfortunate since most integrated limiters operate according to this principle. A combination is used in the H-transponder which is as follows: The first limiter comprises diodes and the second limiter uses an integrated circuit. Since the integrated circuit is provided with a relatively constant voltage, the conversion from AM to PM is no longer of importance.

3. The modulator required to recover the amplitude values at the output stage must be able to handle frequencies from DC to several 100 kHz. In addition to this, it should have a high efficiency. Principally speaking, a pass transistor is suitable, but one will often loose the power gained due to the reduction of the collector voltage.
   The problem was solved by generating a square-wave signal whose on-time is dependent on the momentary amplitude value. The square-wave frequency must be approximately three times higher than the highest frequency to be modulated. Such a pulse-width modulated signal can be obtained by passing the amplitude information to a differential comparitor which compares it to a saw-tooth voltage at the switching frequency (Fig. 2).

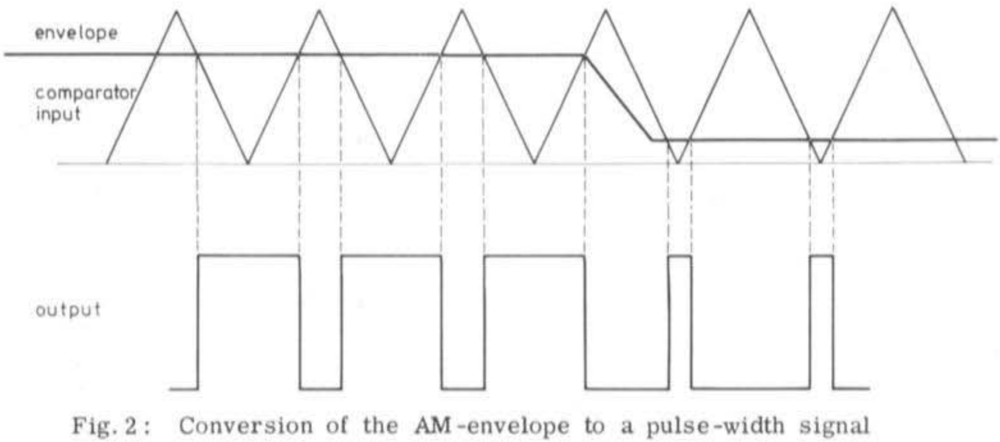


Fig. 2 : Conversion of the AM-envelope to a pulse-width signal

Since higher frequency components of the amplitude signal cause fluctuations of the scanning time-point, interference frequencies will be generated due to the phase modulation. This problem was solved in the case of OSCAR 7 by taking the amplitude value during the lower peak of the square-wave voltage which is then stored. However, recent measurements have shown that this measure is not absolutely necessary. The scanning process generates a time delay that must also be present in the phase channel. This is to ensure that both signal components are restored with the correct phase relationships (see appendix).

The pulse-width, square-wave signal drives a low-loss power switch. Since the current only flows in one direction a diode is suitable for use as switch in the "off" condition. The switched signal is then passed via a low-pass filter for suppression of the switching frequency which only allows the original AM components to pass. The modulator of the H-transponder has an efficiency of approximately 85%.

It is probably advisable for transmitters having a wider bandwidth to use a pulse-width system for lower frequencies and a conventional amplifier in parallel for the higher frequencies. This can be achieved if suitable splitting filters are used (Fig. 3). Calculations have shown that the efficiency in only very slightly lower, but that the upper frequency limit is of one to two orders of magnitude higher.

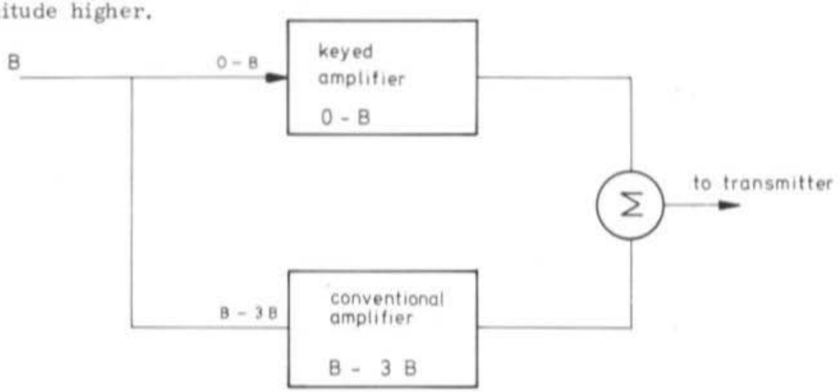


Fig. 3 : Principle of a modulator for very wideband transmitters

4. BLOCK DIAGRAM

A description of how the previously mentioned principles can be obtained in practice is to be described with the aid of the block diagram of the H-transponder given in Figure 4.

The noise power of the receiver is in the order of -150 dBW (100 kHz, 500 K). It will be seen in Figure 1 that a typically loud signal will require 6 dB S/N, or -144 dBW. At an output power of 2 W (7 dB below 10 W), an overall gain of 147 dB is required. Such a high gain is most favourably spread over various frequencies. A very careful construction is required with IF amplifiers having more than 60 dB of gain, in order to obtain neutralization and stability. In principle, it would be possible to construct a transponder using only two mixers. However, since there is the danger that some of the oscillator energy of the transmit mixer is injected to the receive mixer, this could easily lead to self-oscillation (a small signal component of the transmitter is always present at the input of the RX). This danger is avoided by use of three mixers.

The input frequency band of the H-transponder is firstly converted to an intermediate frequency of 50 MHz before being re-converted to 10.7 MHz. The 50 MHz amplifier is controlled. In actual fact, the large increase of gain pre-

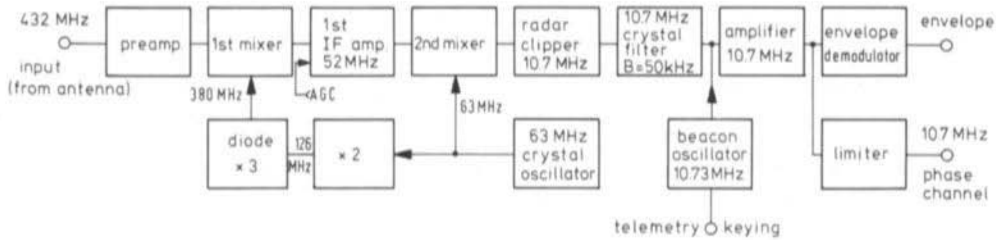


Fig. 4a: Receiver module of the H-transponder (block diagram)

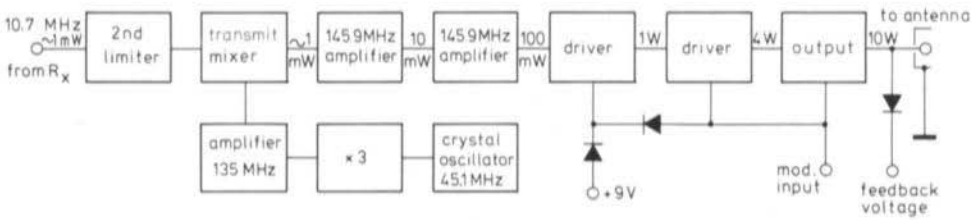


Fig. 4b: Transmit module of the H-transponder

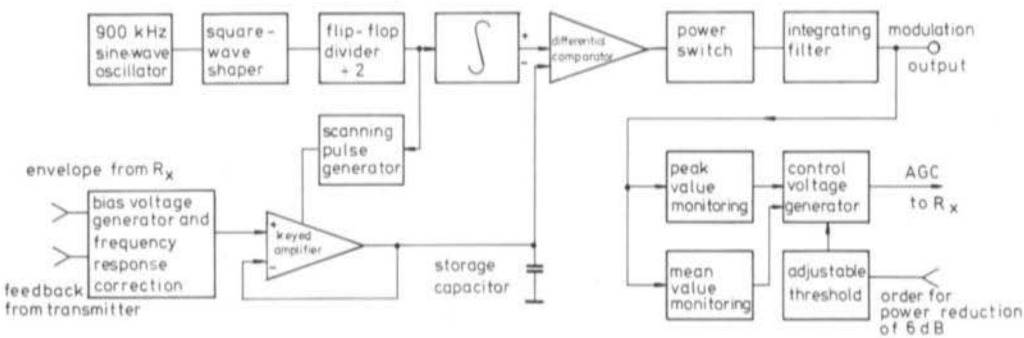


Fig. 4c: Modulator of the H-transponder

vious to the actual selectivity at 10.7 MHz is rather in contrast to the modern design of receivers and with respect to reducing cross-modulation characteristics. However, due to the large distance between the ground station and the satellite, no signals will be received by the satellite that are strong enough to cause cross-modulation.

A bandpass filter with a bandwidth of 200 kHz is used at 10.7 MHz. This is followed by differential amplifier that goes into saturation approximately 6 dB above the full drive level of the transponder. Any radar impulses that could have an interfering effect are limited by this measure. A control voltage is not generated in this manner since the following 40 kHz bandwidth filter will suppress these impulses by further 13 dB (only one fifth of the radar spectrum) so that they are 7 dB below the control threshold. Normal amateur radio signals will pass through the 40 kHz filter at low loss and will not be affected by the limiters since the control circuit ensures that the level remains within the linear range.

Of course, it is possible for this radar limiter to cause strong stations to produce intermodulation outside of the 40 kHz wide transponder band. However, this risk has to be taken to ensure that the transponder was not blocked by ground radar.

The radar clipper is followed by a 40 kHz bandwidth crystal filter and a 10.7 MHz IF amplifier. Sufficient IF voltage is available at the output of the amplifier in order to obtain 10 V (peak-to-peak across 10 kΩ) at the demodulator. A portion of the IF signal is fed to the limiter in order to generate the phase component. The subsequent 10.7 MHz circuits delay the signal in order to compensate for the delay in the modulator.

The signal is provided with a second limiting process in the transmit module before being converted to 146 MHz. After another five stages of gain, approximately 10 W is available at 146 MHz. The amplitude components are re-modulated in the last two amplifier stages. The third stage from the output is also provided with a certain amount of modulation in order to increase the linearity.

A demodulator is provided at the output of the transmitter in order to provide the feedback signal. This feedback is to ensure that modulation distortion and non-linearities of the modulation are suppressed. However, due to the inavoidable phase-shift, this feedback will not be effective in excess of 50 kHz. However, since it is distortions that are within the passband of the transponder that interfere mostly, the feedback does offer an improvement.

A 900 kHz sinewave signal is firstly generated in the modulator. After division by factor 2, a 450 kHz square-wave signal with a ratio of exactly 50% will be available. The square-wave signal is then transposed to a sawtooth signal in a Miller-integrator. At the same time, a sampling pulse is generated that transfers the amplitude value to the intermediate storage. A signal obtained from the IF demodulator minus the feedback voltage is used as amplitude value. A frequency-response correction is also carried out at this point in order to obtain the most favourable characteristics.

The sawtooth voltage is now compared with the amplitude value from the storage which results in the switching voltage. The slope duration of the power switch should be as short as possible since a finite slope duration at the drive limits of the modulator would cause distortion. It is especially unpleasant when these limit ranges of the amplification increase to twice the value. Since the feedback-compensation will no longer be correct at such a point, this can lead to instabilities. A partial compensation of this effect can be achieved when a curve as shown in Figure 5 is used for scanning instead of the square-wave signal. Such a signal can be generated easily by using a resistor in series with the Miller-integration capacitor.

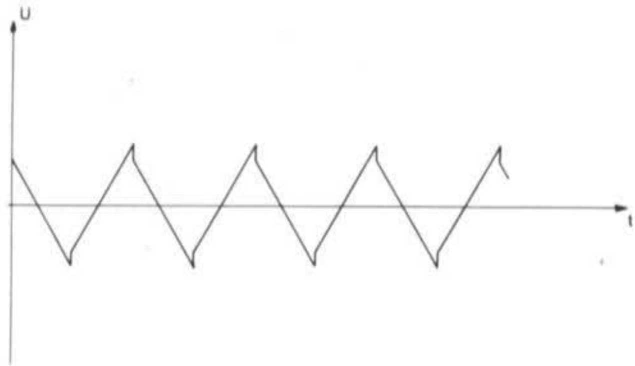


Fig. 5 : The modified scanning curve

Since the output voltage of the modulator is proportional to the drive, this voltage can be used for generating the control voltage. The level is controlled using a peak-value rectification with a rise-time of 0.1 s and a fall-time of 2 s so that the peaks do not overdrive the transponder by more than 3% of time. The resulting distortions do not interfere noticeably.

In addition to this, the mean value of the signal band is monitored with a time constant of approximately 4 s. If the mean value exceeds -6 dB when referred to peak output, the transponder will be controlled. This is to ensure that the current requirements of the transponder remain within reasonable limits when only one strong carrier is driving the transponder.

This type of control also allows a simple reduction of the transponder output power. It is only necessary for the control thresholds to be reduced. Since the efficiency is virtually independent of drive, virtually no disadvantages are present when using this method.

The effectivity of the control was tested by transposing a portion of the 80 m band to the input frequency of the transponder. This allowed virtually realistic operating conditions to be simulated.

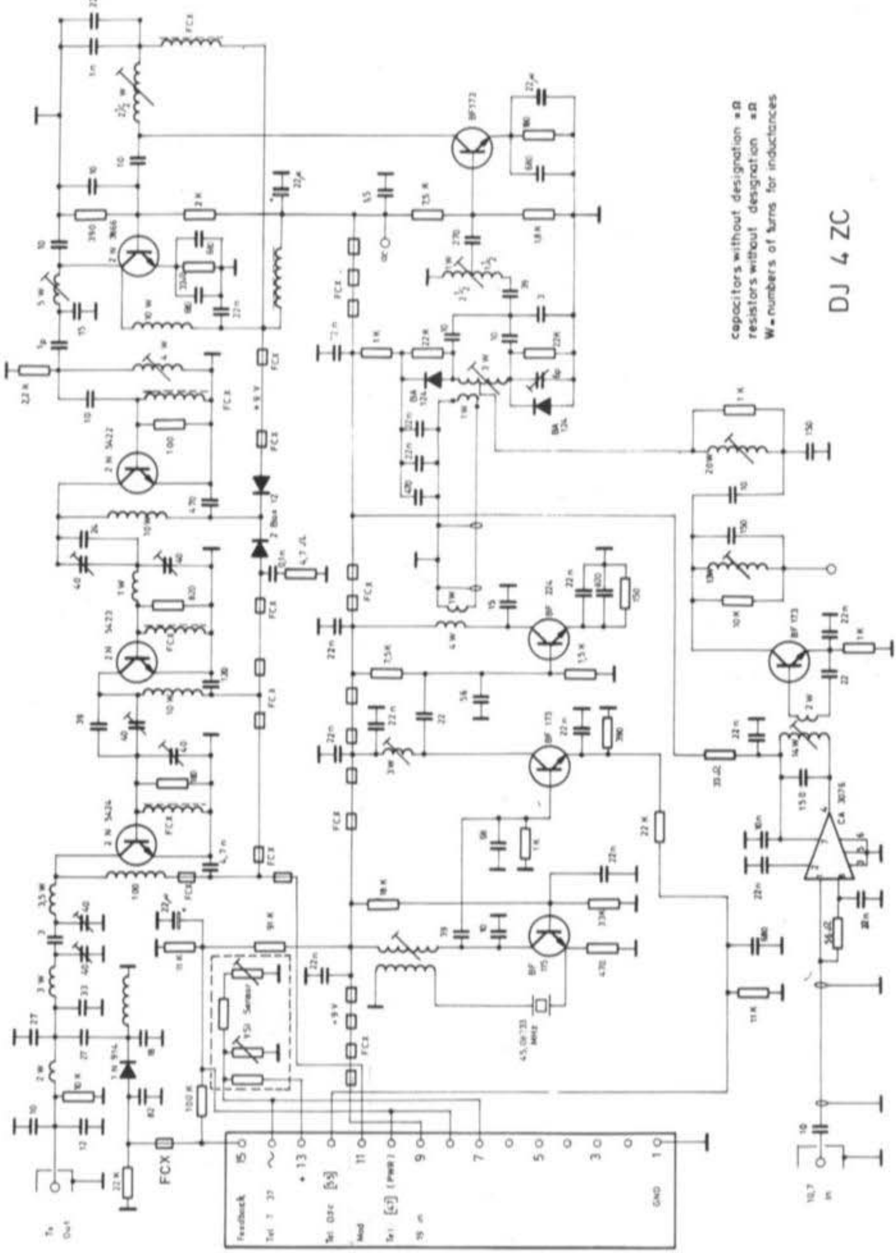


Fig. 6a: Circuit diagram of the H-transponder used in OSCAR 7

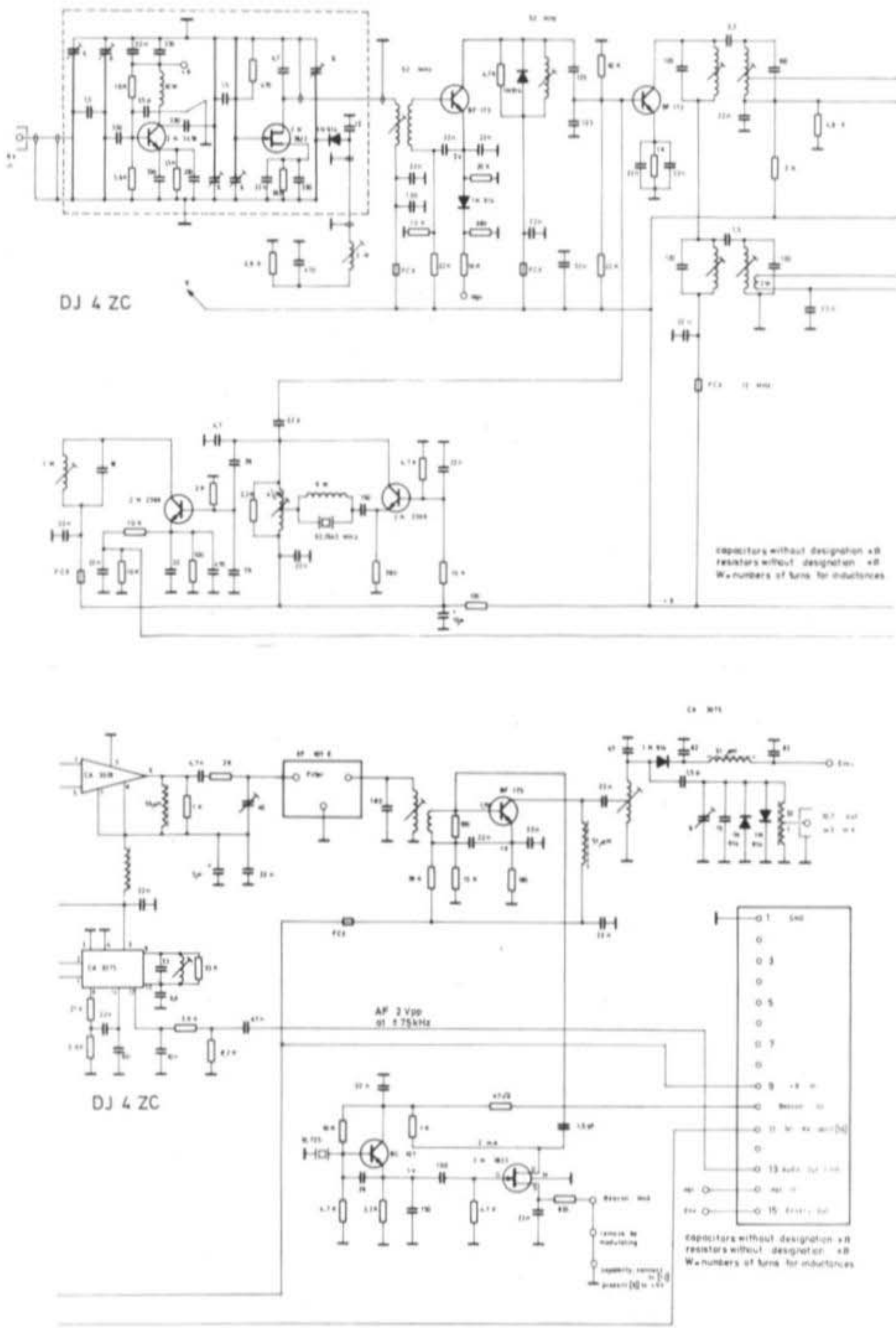


Fig. 6b: Circuit diagram of the receiver of the H-transponder in OSCAR 7

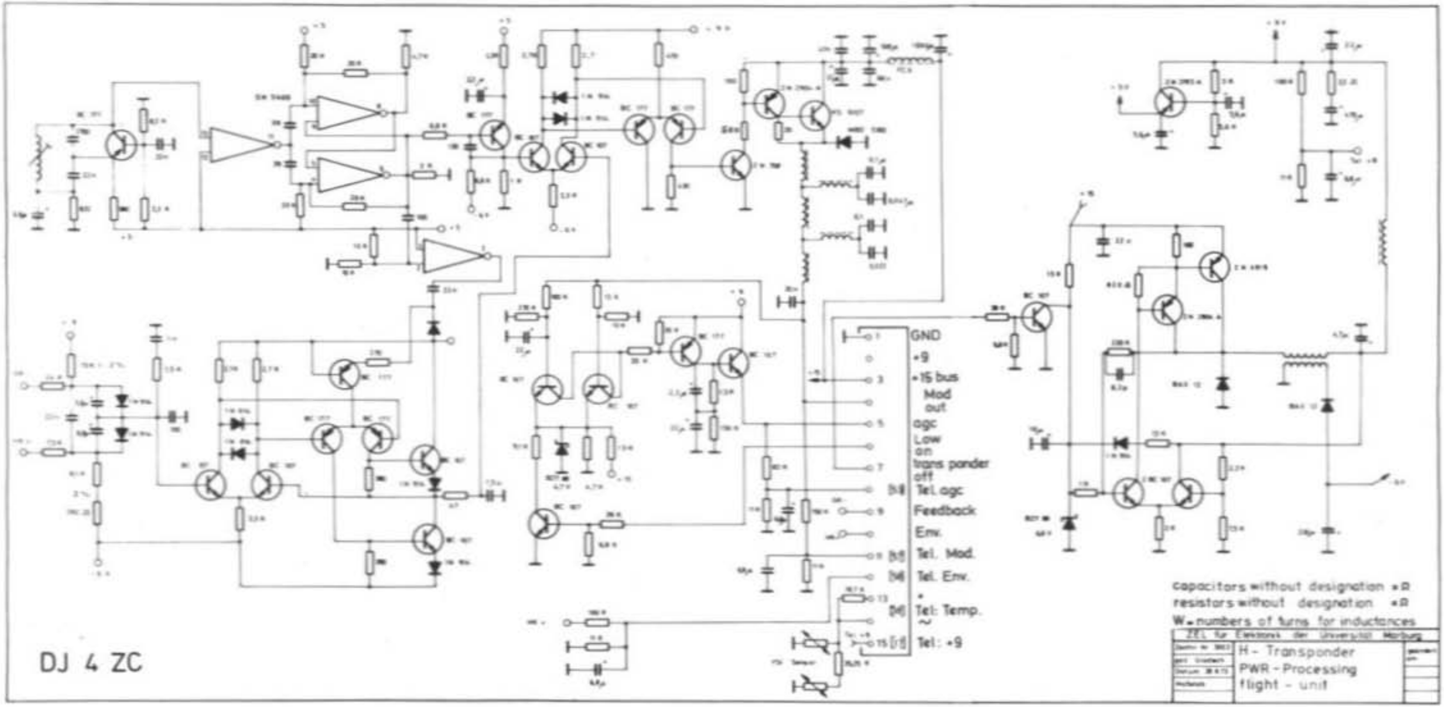


Fig. 6c : Circuit diagram of the power supply for the H-transponder

5. CIRCUIT DETAILS AND MEASURED VALUES

The complete circuit of the H-transponder is given in Figures 6a to 6c. The following measured values were established with the completed unit:

Measured values: $U_B = 13$ V

Drive	P_{out}	Power consumption	Efficiency	Intermodulation during two-tone test			
0 dB	11.2 W	25.3 W	44 %	45 kHz	3rd order	-34 dB	re-
- 3 dB	5.6 W	13.2 W	42 %		5th order	-40 dB	ferr-
- 6 dB	3.0 W	7.5 W	40 %	2 kHz	3rd order	-36 dB	ed to
-10 dB	1.2 W	3.6 W	33 %		5th order	-41 dB	11.2 W

6. APPENDIX

6.1. RESULTS OF THE COMPUTER SIMULATION OF THE ELIMINATION AND RESTORATION PRINCIPLE

A computer program was developed in order to establish the required bandwidth of the phase and amplitude channels. This was based on the model given in Figure A 1.

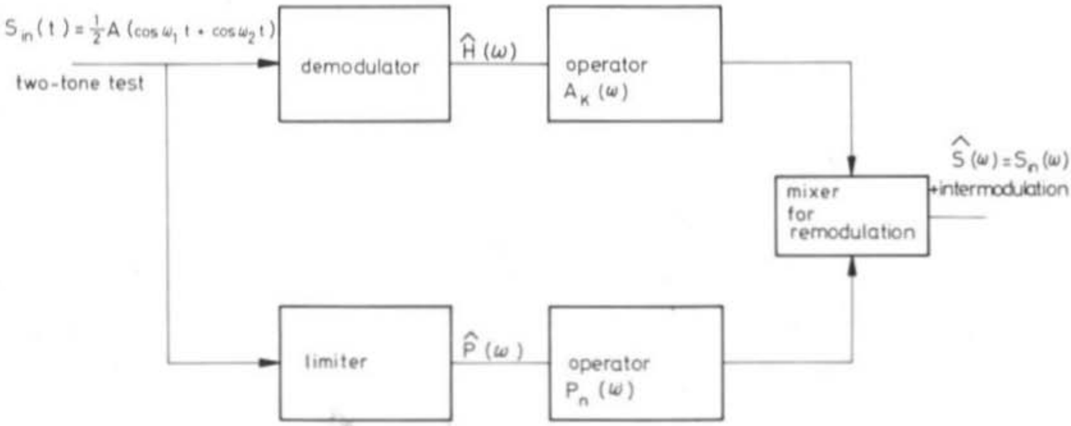


Fig. A1: Model for a computer program for estimating the required bandwidth

By variation of the two operators, the attempt was made to obtain the best possible overall behaviour of the distortion products. Since it was assumed with this model that no unwanted conversion from AM into PM and vice versa would be present, the distortion spectra will be symmetrical. Only one half of this spectrum is displayed. Modulation distortions have not been considered.

Adjacent distortion products have been combined for clarity.

If an abrupt bandwidth (B) limitation is provided in the amplitude channel the characteristics given in Figure A 2 will result.

An abrupt limitation of the bandwidth in the phase channel has a far greater, and more undesirable effect (Fig. A 3).

P - channel unlimited
B envelope channel limited (parameter)

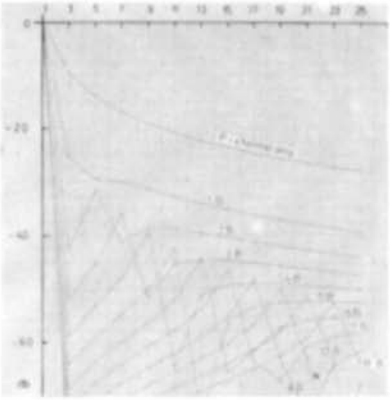


Fig. A2 :
Distortion products with abrupt bandwidth limiting in the amplitude channel

envelope channel: 15 B
P - channel: limited (parameter)

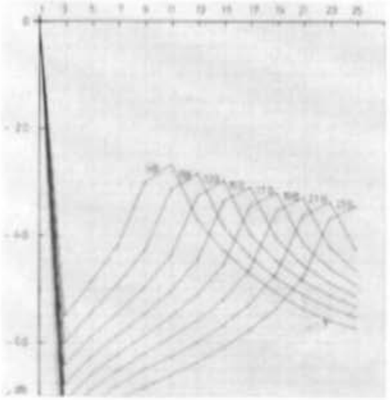


Fig. A3 :
Distortion products with abrupt bandwidth limitation in the phase channel

envelope: 15 B
P - channel unlimited
parameter: 1 res. circuit in P-channel with bandwidth B_S
no delay correction

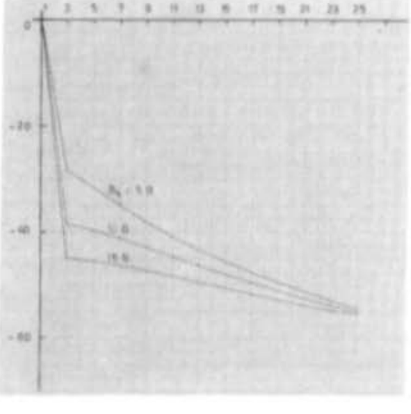
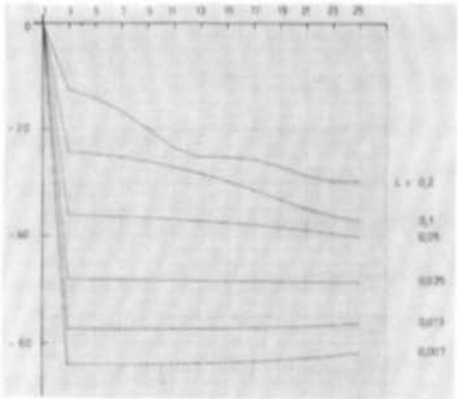
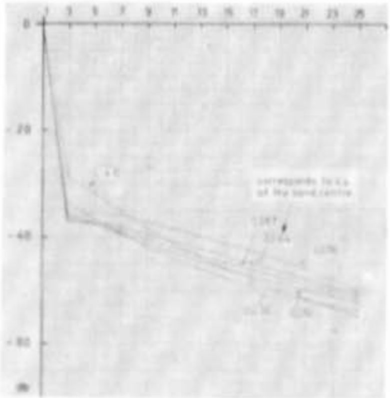


Fig. A4 :
Distortion products with gradual limitation of the bandwidth



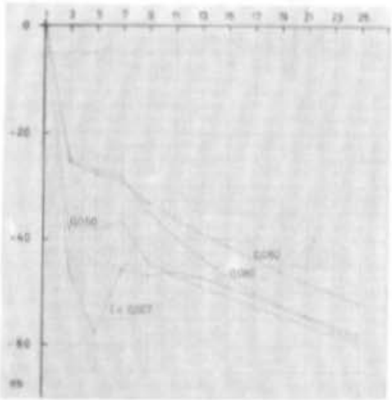
envelope: 15 B
parameter: delay

Fig. A5:
Distortion products as a function of delay differences



envelope: 15 B
1 res. circuit $B_s = 5$ B
parameter: delay compensation

Fig. A6:
Distortion products with the most favourable parameter



envelope: 3 B abrupt
+ 1 lowpass with f = 2.25 B
P-channel: 2 res. circuits of 10 B
parameter: L

Fig. A7:
Distortion products with optimized parameters

However, if this limitation is made gradually with low-pass filters or resonant circuits, the behaviour will be more favourable as can be seen in Figure A 4.

Since differences in the delay time are present between the modulator and phase channel, the effect of the delay time difference is shown in Figure A 5. In this case, the delay time L is equal to 1 x B, where B is the bandwidth, and 1 the absolute delay time difference.

If the delay of the resonant circuit provided to obtain the curve given in Figure A 3 is compensated, this will result in the behaviour given in Figure A 6. It will be seen that lower order distortion products can be reduced, but at the price of higher distortion levels at higher orders.

The various parameters were varied until the most favourable overall behaviour was obtained. This was the case at an amplitude bandwidth of 3 B and using a low-pass filter with a cut-off frequency of 2.25 B. The bandwidth of the phase channel was limited using two resonant circuits of 10 B each. The best behaviour (Fig. A 7) resulted with a delay-time correction of 0.007. With these parameters, the system dependent distortion products have been suppressed to such an extent that it is not the process itself but only the linearity of the modulation that limits matters.

As was mentioned in (2) it is possible for the envelope elimination and restoration principle to be used with simple frequency triplers such as varactor multipliers. In this case, it is necessary for a corresponding frequency division to be made in the IF amplifier in order to ensure the correct phase relationships. Strictly speaking, this method will only operate correctly when the phase is a continuous function of time, which is not the case with the two-tone test. However, this does not seem to have any great effect in practice. Transmitters equipped with frequency dividers and multipliers have shown no disadvantages for SSB and transponder operation.

7. REFERENCES

(1) T. Bittan: Antenna Notebook: OSCAR 7 Antennas
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 248-251

(2) K. Meinzer: A New Method of Frequency Multiplication for VHF, UHF and SSB
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INDEX TO VOLUME 6 (1974) OF VHF COMMUNICATIONS

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Crystal	65.2500 MHz	(HC- 6/U)	) 131 / 131,5 MHz	DM 22.--
Crystal	65.0000 MHz	(HC- 6/U)	....	DM 22.--
Crystal	64.3333 MHz	(HC- 6/U)	for ATV converter (DJ 5 XA)	DM 22.--
Crystal	62.0000 MHz	(HC- 6/U)	for synthesis VFO (DJ 5 HD)	DM 22.--
Crystal	57.6000 MHz	(HC-25/U)	....	DM 33.50
Crystal	57.6000 MHz	(HC- 6/U)	....	DM 22.--
Crystal	38.9000 MHz	(HC- 6/U)	for DJ 4 LB 001 ATV-TX	DM 25.--
Crystal	38.6667 MHz	(HC- 6/U)	for 2-m-converters ..	DM 17.--
Crystal	1.4400 MHz	(HC- 6/U)	for synthesizer	DM 22.50

STANDARD FREQUENCY CRYSTALS

Crystal	1.0000 MHz	(XS 6002)	....	DM 26.--
Crystal	1.0000 MHz	(XS 0605)	for 75° ovens	DM 50.--

Crystal oven	XT-2 (12 V) 75°C	....	DM 82.--
Ceramic filter	455 D for FM IF-strip	DC 6 HL 007	DM 70.--

Crystal socket	for HC- 6/U	horizontal mounting	DM 5.--
Crystal socket	for HC-25/U	horizontal mounting	DM 5.--
Crystal socket	for HC-25/U	vertical mounting	DM 1.50

Crystals 72..... MHz (HC-25/U) .... DM 33.--

Following frequencies available as long as stock lasts:
72.025 / 72.050 / 72.075 / 72.100 / 72.125 / 72.150 / 72.175 /
72.200 / 72.225 / 72.250 / 72.275 / 72.300 / 72.325 / 72.350 /
72.375 / 72.400 / 72.425 / 72.450 / 72.475 / 72.500 MHz

Sideband crystal	XF-901	8.9985 MHz	DM 15.--
Sideband crystal	XF-902	9.0015 MHz	DM 15.--

High Performance VHF-UHF Equipment

SSB/CW 2 m Transceiver SE 300

An incomparable SSB/CW transceiver satisfying the highest demands with compact dimensions continuously variable in the range 144 to 146 MHz.
Outstanding sensitivity and cross-modulation rejection using dual-gate MOSFETs in RF and IF stages.
Two 10.7 MHz KVG 8-crystal double lattice filters in the receiver for extremely high selectivity. Amazing high intermodulation rejection of the transmitter.
Operating voltage 12 V DC. Output power is 10 W. Built-in RIT and loudspeaker. Connectors provided for an external loudspeaker and remote control.

100 Channel 2 m FM Transceiver SE 285

Immediately ready-for-operation on 100 channels with a frequency spacing of 30 kHz between 145 and 148 MHz (North American version) or 80 channels with 25 kHz spacing between 144 and 146 MHz (European version).
Five preprogrammed repeater or simplex channels can be selected on a rotary switch. All other channels can be selected independently for transmit and receive using thumbwheel switches on the front-panel. Digital frequency selection using a frequency synthesizer. Receiver equipped with KVG 10.7 MHz crystal filter and crystal discriminator. Operating voltage 12 VDC. Output power is 10 W RF. Built-in squelch, calling tone, and loudspeaker. Connector provided for an external loudspeaker.

SSB/AM/FM/CW 2 meter Transceiver SE 600 digital

A transceiver that really offers you everything. Extremely low noise figure with excellent selectivity, and high cross and intermodulation rejection.
True transceive or separate operation of transmitter and receiver, which can be switched independently to the CW, LSB, USB, AM and FM modes. This versatility allows problemless operation via repeaters, satellite and balloon-carried translators. Digital frequency readout from the built-in frequency counter using 13 mm Nixie tubes. Direct readout of the transmit and receive frequency; the indication jumps from one to the other on pressing the PTT button etc.
Separate crystal filters for each mode. True AM with plate/screen grid modulation. Built-in speech processor. Product detector for SSB and a crystal discriminator for FM. VOX, antitrip and PTT facilities, as well as RF-output and S-meters. Built-in antenna relay. Built-in power supplies for AC and 12 VDC operation.

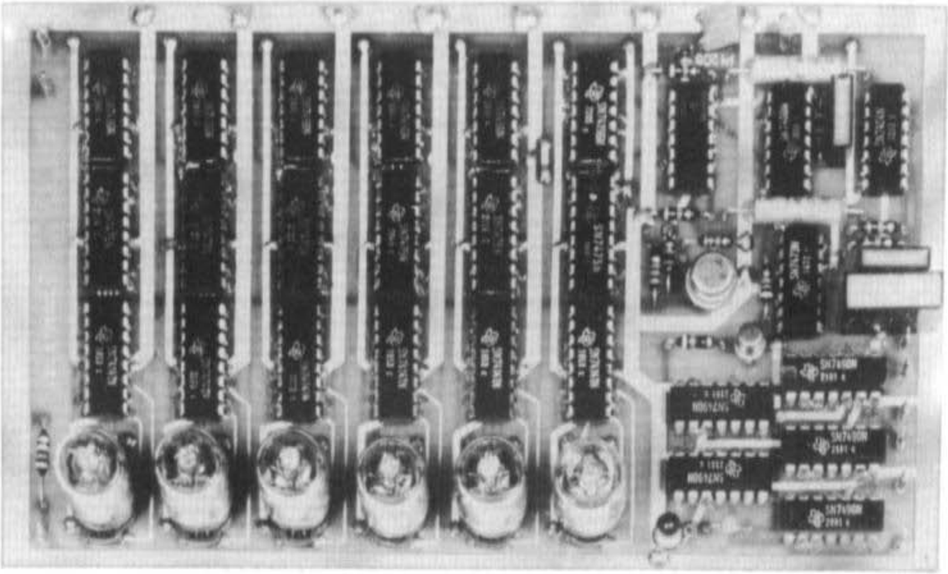
2 m/70 cm Linear Transverter LT 702

An all mode transverter (SSB, AM, FM, CW, RTTY- SSTV) for transposing a 2 m signal to the 70 cm band and vice versa. The full 10 MHz between 430 and 440 MHz is covered in five bands of 2 MHz each. Each of these bands can be selected individually for transmit and receive so that it is especially suitable for operation over repeaters and transponders. The receive converter is synchronized to the transmit oscillator during transceive operation. Several coaxial relays are provided for dual band operation 2 m/70 cm and 70 cm/2 m. Output power is 10 W. Built-in meters for drive and output power. Built-in power supply. Built-in attenuator for input power levels of 1 W to 30 W PEP on 2 m.

Please request our data sheets

KARL BRAUN · Communications Equipment

D-8500 Nürnberg, Deichslerstraße 13, W. Germany



Considerable price reductions on our frequency counter kits:

Due to larger purchasing and rationalization we have been able to reduce the prices of a number of our kits.

250 MHz 6-digit Frequency Counter DM 625.--

Comprising:

DL 8 TM 002	6-digit frequency counter	DM 315.--
DJ 6 TA 001	High-impedance preamplifier	DM 70.--
DJ 6 PI 001	250 prescaler and preamplifier	DM 108.--
DJ 1 JZ 001	Time-base (with crystal oven DM 174.--)	DM 70.--
DL 3 YK 002	Power supply	DM 70.--

Complete 250 MHz frequency counter .... DM 625.--
As above but with crystal oven and precision crystal .... DM 725.--

500 MHz 6-digit Frequency Counter .... DM 865.--

Comprising above kits but with 500 MHz prescaler DJ 5 HD 003 instead of DJ 6 PI 001

Complete 500 MHz frequency counter without crystal oven........ DM 865.--
Complete 500 MHz frequency counter with crystal oven........ DM 965.--

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West-Germany - Telephone (0 91 91) 91 57 or (0 91 33) 33 40

Bank accounts: Raiffeisenbank Erlangen 22411, Postscheckkonto Nürnberg 30455-858

TWO-METER ANTENNAS

Jaybeam Limited

HALO

A Broad Band Halo type antenna with no capacity loading and a correct Gamma Match to coaxial termination.

Width 12" (30½ cm)

Head only to fit 5/16" — 1" diam. Mast **Cat. No. HO/2M**

Complete with ½" diam. Mast **Cat. No. HM/2M**

Weight 8 ozs.

Wind loading 10 lbs. at 100 m.p.h.

5 ELEMENT YAGI Cat. No. 5Y/2M

Gain 7.8dB

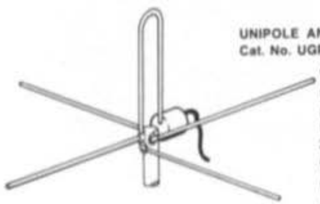
Length 63½" (161 cm)

Width 40½" (103 cm)

Horizontal Beamwidth between half power points 52°

Weight 3 lbs.

Wind loading 30 lbs. at 100 m.p.h.



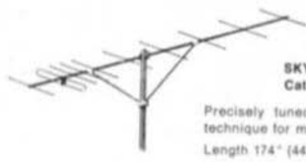
UNIPOLE AND GROUND PLANE
Cat. No. UGP/2M

Gain : Unity

Unipole and ground plane aerial with clamp to fit to masts up to 2" O.D.

Weight 3 lbs.

Wind loading 12 lbs. at 100 m.p.h.



SKYBEAM 10 ELEMENT YAGI
Cat. No. 10Y/2M

Precisely tuned using the "Long Yagi" technique for maximum gain 13.2dB.

Length 174" (443 cm)

Width 40½" (103 cm)

Horizontal Beamwidth between half power points 33°

Weight 12 lbs.

Wind loading 72 lbs. at 100 m.p.h.

8 ELEMENT YAGI
Cat. No. 8Y/2M

Gain 10dB

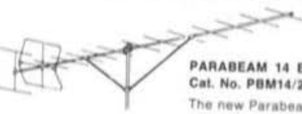
Lenght 102" (260 cm)

Width 40½" (103 cm)

Horizontal Beamwidth between half power points 45°

Weight 4 lbs.

Wind loading 48 lbs. at 100 m.p.h.



PARABEAM 14 ELEMENT YAGI
Cat. No. PBM14/2M

The new Parabeam with increased gain — 15.2dB — and broader bandwidth.

Length 234" (595 cm) Width 41" (104 cm)

Horizontal Beamwidth between half power points 24°

Weight 14 lbs.

Wind loading 91 lbs. at 100 m.p.h.

FIVE OVER FIVE
Cat. No. D5/2M

Gain 10.8dB

Slot Fed Double 5 Yagi

Length 63½" (161 cm)

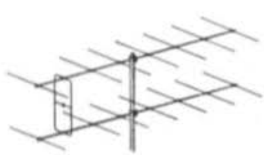
Width 40½" (103 cm)

Height 46" (116 cm)

Horizontal Beamwidth between half power points 52°

Weight 7 lbs.

Wind loading 62 lbs. at 100 m. p.h.



EIGHT OVER EIGHT
Cat. No. D8/2M

Gain 12.6dB

Slot Fed Double 8 Yagi

Length 102" (260 cm)

Width 40½" (103 cm)

Height 46" (116 cm)

Horizontal Beamwidth between half power points 45°

Weight 9 lbs.

Wind loading 90 lbs. at 100 m.p.h.

Mounting Kit for Slot Fed Aerials Vertical Polarisation **Cat. No. SVMK/2M**

HARNESSES FOR STACKED ARRAYS

PMH2/2M	Moulded Waterproof Harness to match and phase any two 2 meter antennas.	**PMH4/2M**	Moulded Waterproof Harness to match and phase any four 2 meter antennas.
PMH/2C	Moulded Waterproof Harness to fit crossed yagis for circular polarisation.	**PMH2/4M**	Moulded Waterproof Harness to match and phase two 4 meter antennas.

RECOMMENDED STACKING SPACING BETWEEN CENTRES

HO/2M	Halo	41" (104 cm)	10Y/2M	10 Element Yagi	132" (335 cm)
XD/2M	Crossed Dipoles	41" (104 cm)	PBM14/2M	14 Element Yagi	144" (366 cm)
5Y/2M	5 Element Yagi	82" (208 cm)	D5/2M	Double 5 Slot	147" (373 cm)
8Y/2M	8 Element Yagi	100" (254 cm)	D8/2M	Double 8 Slot	160" (405 cm)