VHF COMMUNICATIONS

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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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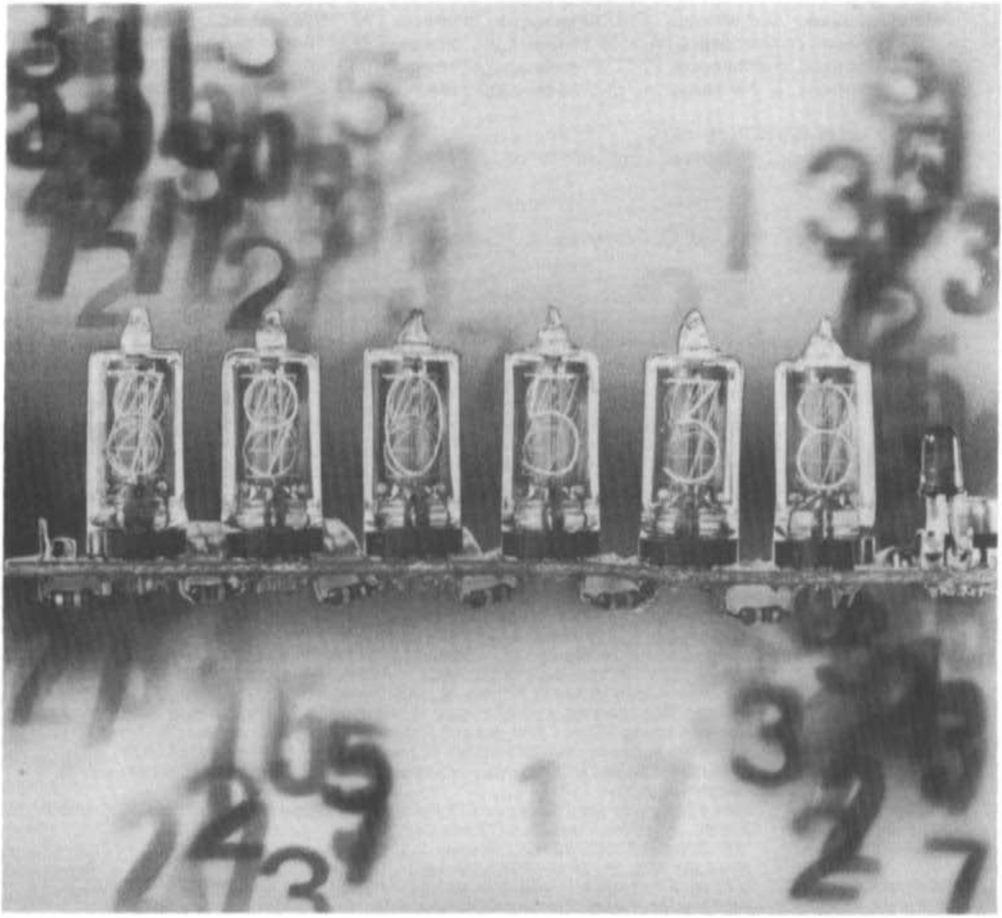

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The next editions of VHF COMMUNICATIONS will include:

ATV: A pattern generator to match the pulse center

Measuring technology: A digital voltmeter
A wideband reflectometer for 50 Ohms

2 m: An 80 channel FM transceiver with synthesizer (25 kHz spacing, Europe)
An 400 channel FM transceiver with synthesizer (10 kHz spacing, 144-148 MHz)
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A highly selective converter for repeaters

70 cm: Matching receive converter for the DJ 6 ZZ 002
Linear transverter 144/432 MHz
Transistorized linear amplifier for 10 W and 20 W at 12 V

23 cm: Linear transverter 144/1296 MHz with 2 C 39 mixer in microstripline technology

SHF: We are still waiting for you to tell us about your designs !

A MODULAR ATV TRANSMITTER

Part II: Modules DJ 4 LB 003, 004 and 005

by G. Sattler, DJ 4 LB

3.4. NBFM WITH THE SOUND IF-MODULE DJ 4 LB 002

The sound IF module DJ 4 LB 002 operates according to the CCIR-standard with a frequency deviation of $\pm$ 40 kHz, so that a maximum bandwidth of 120 kHz results for the sound signal. If this module is to be used for normal amateur NBFM transmission, it is only necessary for the varactor diode BA 124 (D 203, approx. 50 pF at 2 V) to be exchanged for a type BA 110 or BA 149 (approx. 10 pF or 8 pF at 2 V). The RF-output power and the adjustment range remain practically unaltered, however, the oscillator will operate at a frequency approximately 6 MHz higher. If the same local oscillator frequency of 473.15 MHz required for ATV operation is also used for conversion of the NBFM, it will be possible by adjustment of the core of inductance L 201 to obtain any required output frequency in the voice transmission portion of the 70 cm band (430 MHz to 433.5 MHz). The frequency stability of the modified sound-IF oscillator is better than that of the wideband FM version, since the capacitance of the BA 110 diode only represents a small part of the total resonant circuit capacitance. Of course, a good long-time constant of the transmit frequency can only be guaranteed when operating the module DJ 4 LB 002 with automatic frequency control (AFC via Pt 206).

4. LOCAL OSCILLATOR MODULE DJ 4 LB 003

Module DJ 4 LB 003 generates a crystal-controlled, local oscillator frequency of 473.15 MHz for the transmit mixer DJ 4 LB 004 (see Fig. 1 in Part I). A connection is also provided for a receive converter (transceive operation).

As can be seen in the block diagram (Fig. 14), the crystal-controlled frequency of 78.858 MHz is multiplied by six. The bandpass filters at the output of the tripler and doubler stages efficiently suppress spurious signals which are always generated during the frequency multiplying process. The subsequent amplifier stage provides an output power of approximately 10 mW to 15 mW and ensures an isolation between the output socket and the bandpass filter.

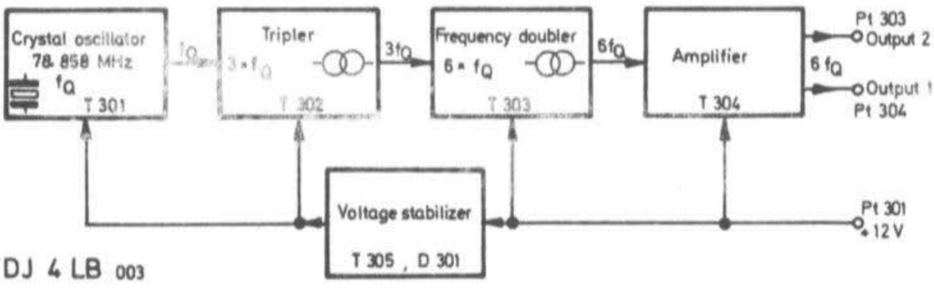

Fig. 14: Block diagram of the local oscillator module DJ 4 LB 003

4.1. CIRCUIT DETAILS

Figure 15 gives the circuit diagram of the local oscillator module. Transistor T 301 operates as crystal oscillator and the resonant circuit comprising L 301/ C 303 is tuned to the overtone frequency of the crystal (in our case 78.8 MHz). The subsequent transistor T 302 generates strong harmonics when operating in Class C and the bandpass filter comprising inductances L 302 and L 303 filter out the required frequency of three times that of the crystal oscillator frequency. This signal is now fed to the doubler stage equipped with transistor T 303 which operates in class AB and therefore generates mainly even harmonics. The bandpass filter comprising inductances L 304 and L 305 filters out the doubled frequency which is then six times the original crystal-controlled frequency. This signal is then fed to the amplifier stage comprising transistor T 304. The output circuit of this stage is in the form of a Pi-filter which transforms the output signal to an impedance of 60 Ω at Pt 304. Connection Pt 303 is an additional RF output having approximately 20% of the output power for driving a receive converter. The crystal-controlled oscillator (T 301) and the subsequent tripler stage (T 302) are fed via transistor T 305 (and D 301) with a stabilized voltage of approximately 8.5 V so that no frequency variations are caused by fluctuations of the operating points.

4.2. OTHER APPLICATIONS FOR MODULE DJ 4 LB 003

4.2.1. DIFFERENT FIXED FREQUENCIES

Module DJ 4 LB 003 can be used for generating crystal-controlled frequencies in the range of approximately 400 MHz to 500 MHz. Due to the relatively large adjustment range of the trimmer capacitors, it is only necessary for the crystal and the appropriate resonant circuit capacitor C 303 to be exchanged. The components list offers some examples. An interesting combination would be as an exciter for 70 cm.

4.2.2. HIGHER OUTPUT VOLTAGE

Approximately 5 mW are sufficient as local oscillator signal for the mixer module DJ 4 LB 004. It will be seen that module DJ 4 LB 003 provides more than enough output power for this application. However, if a higher output power is required for other applications, it is possible to replace the transistor used in the output stage (T 304) with one for a higher output power. The BF 223 allows an output power of up to 40 mW to be obtained in the above frequency range.

4.3. MECHANICAL CONSTRUCTION

The described local oscillator module DJ 4 LB 003 is accommodated on a single-coated PC-board having the dimensions 135 mm x 50 mm. Figure 16 shows this PC-board and the associated component location plan. A photograph of the author's prototype is given in Figure 17. The only soldered connection to be made on the component side of the board is the soldering of the ceramic capacitor C 319 to the coil tap on inductance L 305. It is advisable to also mount this module in a TEKO-box 4 B in order to screen it against UHF injection from the transmitter. Details regarding the mounting of the PC-board into the case were given in Part I of this article.

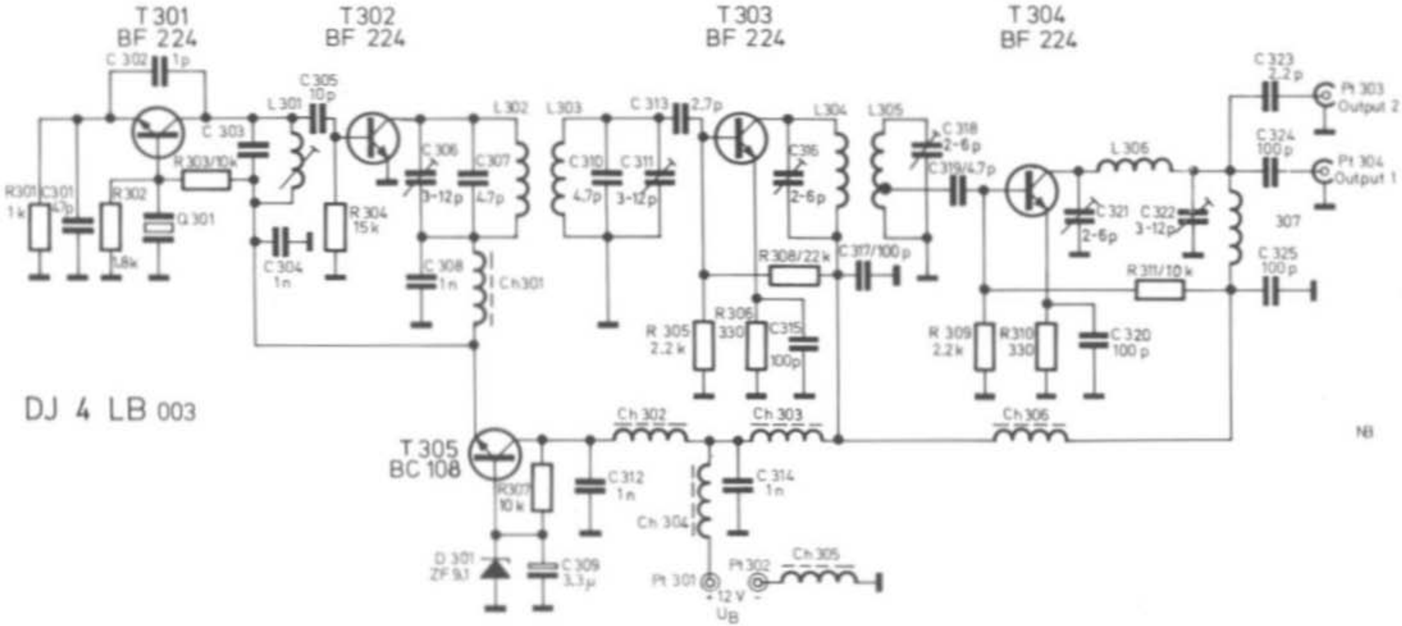

Fig. 15: Circuit diagram of the local oscillator module DJ 4 LB 003

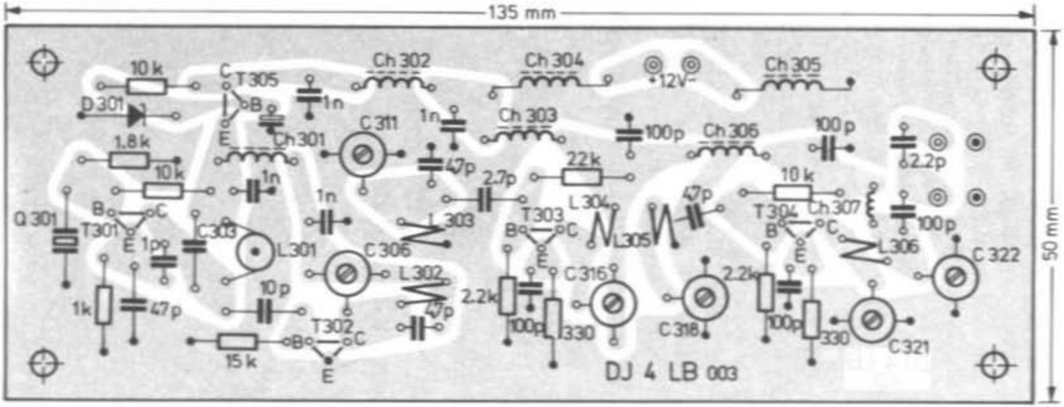

Fig. 16: Printed circuit board and component location plan of the local oscillator module

Fig. 17: Author's prototype of module DJ 4 LB 003

4.3.1. SPECIAL COMPONENTS

T 301 - T 304: BF 224, BF 173
T 305: BC 108 or similar

D 301: BZY 85/C9V1 or similar 9.1 V zener diode

L 301: 4.75 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 5 mm dia. coil former with VHF core (brown). Coil length approx. 7 mm, facing the collector side of the board.
L 302-L 306: 1.75 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire self-supporting. L 302, L 303: 5 mm inner diameter, approx. 3 mm spacing between coil and board. L 304, L 305: 4 mm inner diameter, 1 to 2 mm spacing between coil and board. L 306: 5 mm inner diameter, spaced 2 mm from the board. The direction of the coil and coil length are given by the holes in the PC-board. Coil tap for L 305: 0.75 turns from the ground end.

Ch 301, Ch 302, Ch 303, Ch 306: 3.5 turns of 0.4 mm dia. (26 AWG) enamelled copper wire placed through a ferrite bead of 3.5 mm dia., 5 mm long (Philips)
Ch 304, Ch 305: Wideband ferrite choke 6 mm dia., 10 mm long. Z = 800 Ω (Philips).
Ch 307: 3 turns of 0.4 mm dia. (26 AWG) enamelled copper wire wound on a 3 mm former, length approx. 3 mm, self-supporting

Q 301: 78.858 MHz, HC-25/U with holder (vertical) or HC-6/U without holder.

C 306, C 311, C 322: 3 - 12 pF ceramic disc trimmer, 10 mm dia.
C 316, C 318, C 321: 2 - 6 pF ceramic disc capacitor, 10 mm dia.
C 309: 3.3 μF/16 V tantalium drop-type electrolytic

$$\left.\begin{array}{l}\text{C 301: 47 pF}\\ \text{C 303: 33 pF}\\ \text{C 305: 10 pF}\\ \text{C 313: 2.7 pF}\end{array}\right\}\text{ ceramic tubular capacitor for 10 mm spacing}$$

All other capacitors: Ceramic disc capacitors, spacing 5 mm.

All spacing of 12.5 mm is available for the resistors.

Modifications for other output frequencies:

404 MHz: Q 301: 67.333 MHz; C 303: 47 pF
432.··· MHz: Q 301: 72.··· MHz; C 303: 39 pF

Modifications for higher output power levels:
T 304: BF 223 (AEG-Telefunken)
R 310: 68 Ω

4.4. ALIGNMENT AND TESTING OF MODULE DJ 4 LB 003

A reflectometer can be used for indicating the relative output power during the alignment process. The stripline reflectometer DK 2 VF 002 as described in (3) is suitable for this. It should be connected between the RF-output 1 (Pt 304) of module DJ 4 LB 003 and a 60 Ω terminating resistor. However, it is advisable to use a RF voltmeter (multimeter with a diode input) or a tube voltmeter (VTVM with RF-probe) for the preliminary alignment steps.

The RF-voltmeter is firstly loosely coupled to the resonant circuit of the crystal oscillator and the core of inductance L 301 should be adjusted until RF is indicated. The oscillator will now oscillate at the correct frequency since the feedback conditions do not favour any spurious oscillation. This is followed by aligning the resonant circuits of the subsequent stages to resonance by adjusting the variable capacitors. This can also be checked by loosely coupling the RF-voltmeter to the resonant circuit in question.

The described preliminary alignment is repeated until the reflectometer at the output indicates a reading. All resonant circuits are then aligned for maximum reading on the reflectometer and the alignment is repeated until no increase of the output power is possible. For reasons of stability, the core of inductance L 301 should then be slightly extracted until the output power is reduced slightly.

The module is checked by removing the crystal from the holder and ensuring that the circuit no longer provides any RF voltage. In addition to this, the 60 Ω terminating resistor should be removed in order to obtain any required mis-match conditions with the aid of various unterminated coaxial cables. If no spurious oscillations occur, the module will be ready to operate even when the output termination is not exactly obtained. Any tendency to oscillation with the version with a higher output power can be neutralized by increasing the coupling of inductances L 304 and L 305 to another (decreasing the distance between them).

5. TRANSMIT MIXER AND AMPLIFIER MODULE DJ 4 LB 004

As can be seen in the block diagram given in Figure 18, module DJ 4 LB 004 is provided with the local oscillator frequency of 473.15 MHz and the combined video and sound-intermediate frequency of 38.9 MHz and 33.4 MHz. The required output frequency is obtained by conversion of the frequency differences 434.25 and 439.75 MHz which is amplified in a three-stage linear amplifier to approximately 100 mW.

A push-pull mixer stage equipped with field effect transistors is used which virtually completely suppresses the push-push local oscillator signal. This is especially important due to the relatively small frequency spacing between the local oscillator and required output frequency.

The FET push-pull mixer is driven via a differential amplifier which has been dimensioned as a phase-reversal stage so that two equal-amplitude IF voltages are formed that are phase-shifted by 180° to another. Since this phase-reversal stage and the input circuit for the local oscillator frequency do not contain any resonant circuits, the module DJ 4 LB 004 can be used for mixing other frequency combinations without modification, such as 28 to 30 MHz and 404 MHz to 432 - 434 MHz.

5.1. CIRCUIT DETAILS

Figure 19 gives the circuit diagram of the mixer and linear amplifier module.

5.1.1. PHASE-REVERSAL STAGE

Resistor R 409 forms, together with the input impedance of the differential amplifier (T 403, T 404) the 60 Ω termination for the IF-signal at connection

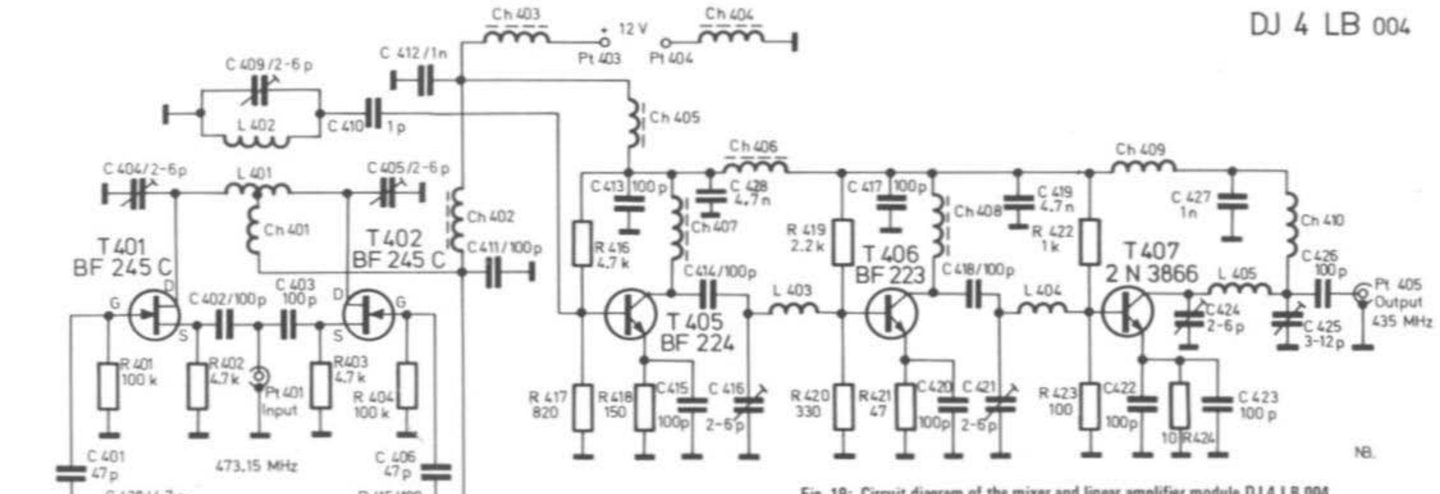
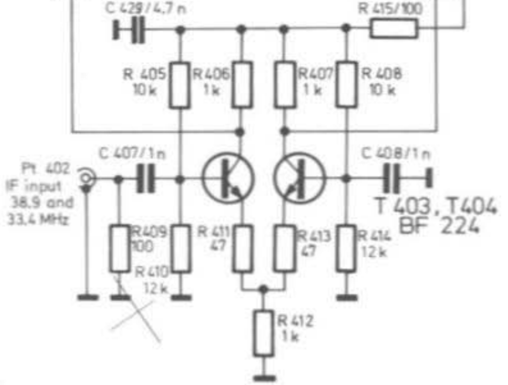

Fig. 19: Circuit diagram of the mixer and linear amplifier module DJ 4 LB 004

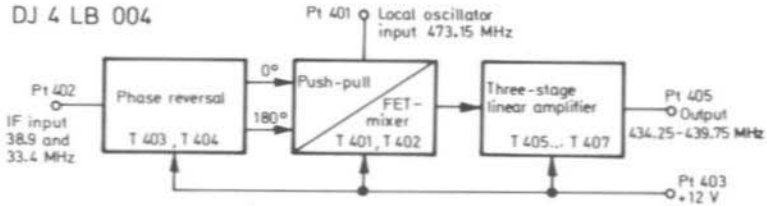

Fig. 18: Block diagram of the mixer and linear amplifier module DJ 4 LB 004

Pt 402. Since the IF input level of approximately 0.6 V (peak-to-peak) is already sufficient for driving the mixer, the voltage gain of the differential amplifier is adjusted to a value of approximately 1.5 by the emitter resistors R 411 and 413. This feedback increases the linearity of the amplifier and reduces the influence of component tolerances on the balance and driveability of the circuit. Two virtually equal-amplitude IF voltages of max. 1 V (peak-to-peak) each is available at the collectors of the two transistors which are phase shifted by 180° to another.

5.1.2. PUSH-PULL FET MIXER

The two phase-shifted IF-signals are fed via capacitors C 401 and C 406 to the high impedance gate connections of the field effect transistors T 401 and T 402. The parallel-connection of the source connections represents the low-impedance termination for the local oscillator voltage. The output circuit of the mixer comprises the centre-tapped inductance L 401 and the series connection of the trimmer capacitors C 404 and C 405. The inductive coupling to the resonant circuit L 402/C 409 forms a bandpass filter which, due to its relatively high Q (low damping) mainly determines the passband characteristics of this module in the 70 cm band.

5.1.3. LINEAR AMPLIFIER

The linear amplifier consists of three amplifier stages equipped with transistors T 405, T 406 and T 407. The virtually constant DC operating points of all transistors (class A) result in a good linearity. It is also ensured that fluctuating signal amplitudes will not noticeably detune the resonant circuits of the amplifier due to the transistor capacitances that are dependent on the operating points. A special UHF filtering also has a good effect on stable operation of the amplifier. The additional higher-value by-pass capacitors C 419 and C 428 ensure that no parasitic oscillations can occur in the shortwave region. The Pi-filter at the output of the amplifier comprising C 424, L 405 and C 425 allows an exact power-matching to a 60 Ω terminating resistor.

5.2. CHARACTERISTICS

5.2.1. LINEARITY

A low-distortion conversion of the complex ATV signal from the intermediate frequency level to UHF and its subsequent amplification places high demands on the linearity of all stages in the signal path. Figure 20 indicates the virtually linear relationship between the IF input voltage and the UHF output power which was measured on a prototype of the module DJ 4 LB 004. These characteristics were measured with the following measuring instruments:
Signal generator hp 608 D, powermeter hp 431 B, VHF attenuator: hp 355 D.

5.2.2. BANDWIDTH

The ATV signal requires a bandwidth of approximately 6.5 MHz which results from the frequency spacing between the video and sound carriers (5.5 MHz) plus the approximate 1 MHz of the residual lower sideband (video modulation spectrum). The reason why a suppressed lower sideband is transmitted was given in (1). Figure 21 gives the measured passband curve. As can be seen in Figure 20, the output power is most certainly in the linear portion of the characteristic curve so that measuring errors due to clipping effects cannot occur.

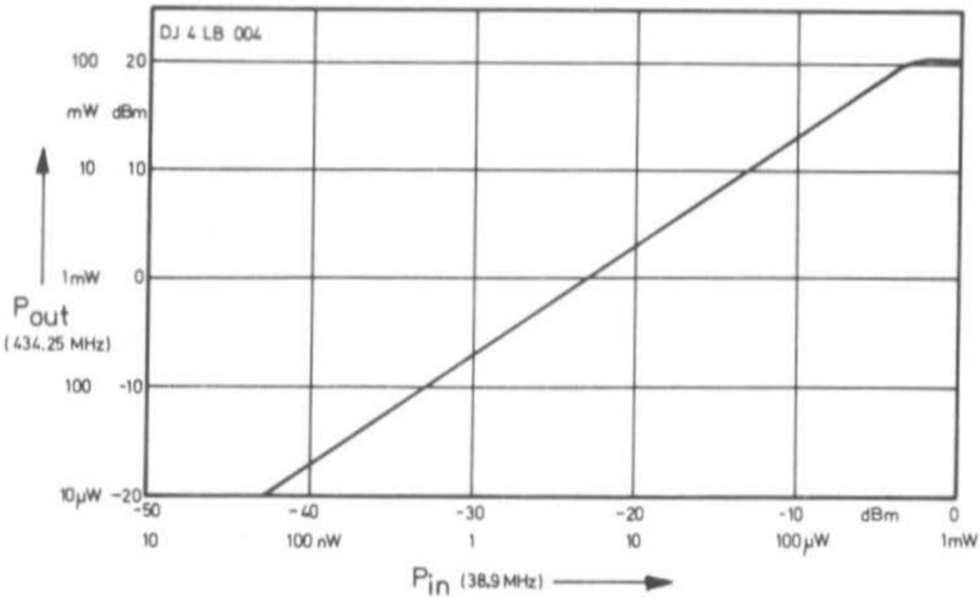

Fig. 20: Amplitude response of module DJ 4 LB 004

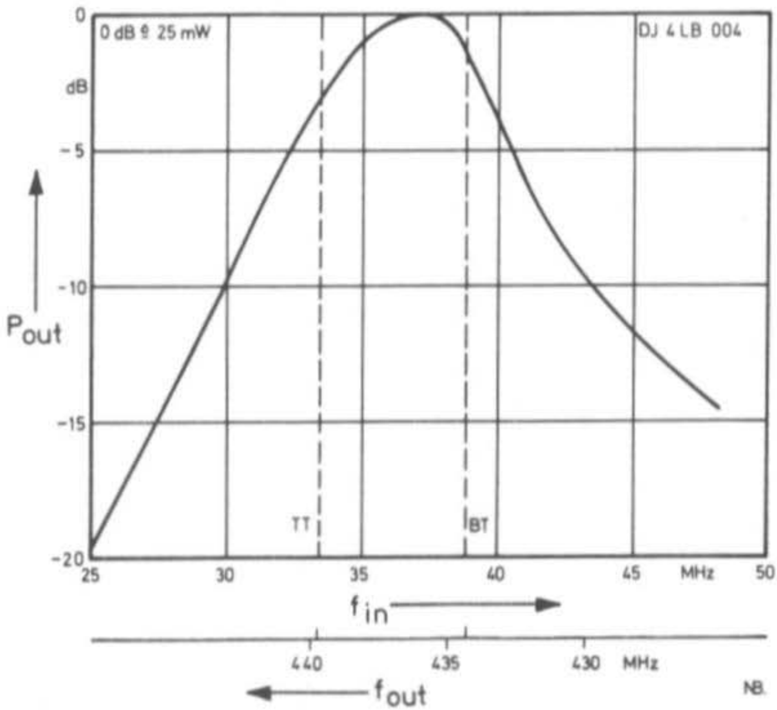

Fig. 21: Frequency response of module DJ 4 LB 004

Whereas the passband curve measured in Figure 21 resulted after alignment without any special measuring instruments, Figure 22 shows a photograph of an ideal passband curve aligned with the aid of a sweep frequency generator at IF level: TV-sweep generator Heathkit IG-52. The UHF demodulator DJ 4 LB 005, which is to be described later, supplies the measuring voltage for the Y-input of the oscilloscope.

The two bandpass characteristics given cannot be directly compared to another due to the different representation (logarithmic or linear scale). However, practice has shown that the effects of the various passband curves on the picture quality is hardly noticeable even when observing special test cards.

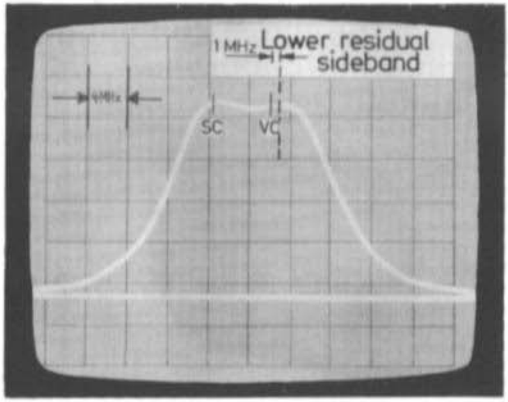

Fig. 22:
Swept frequency response of module DJ 4 LB 004

5.2.3. SUPPRESSION OF THE LOCAL OSCILLATOR AND IMAGE FREQUENCIES

The local oscillator frequency of 473.15 MHz is present with approximately 5 μW at the output of module DJ 4 LB 004. This corresponds to a suppression of 43 dB referred to the power level of the required signal (100 mW). The image frequency of 512 MHz (473.15 MHz + 38.9 MHz) is suppressed by more than 60 dB.

5.3. CONSTRUCTION OF DJ 4 LB 004

The described module DJ 4 LB 004 is accommodated on a single-coated PC-board having the dimensions 135 mm x 50 mm (Fig. 23), which has been designated DJ 4 LB 004. Figure 24 shows a photograph of the author's prototype. The higher TEKO-box 4 B should also be used for this module so that the resonant circuits are not detuned on mounting the cover. Due to the use of only single-coated PC-boards, stable operation of this UHF module is only possible when the board is provided with metal spacing bushings of approximately 5 mm in length between all six mounting positions and the base of the TEKO box, or similar metal surface. The PC-board should be tinned where the spacer bushings touch the PC-board in order to provide a good ground connection since the PC-board is provided with a protective coating.

5.3.1. SPECIAL COMPONENTS

T 401, T 402: BF 245 C (TI), W 245 C (Siliconix)
T 403 - T 405: BF 224, BF 173
T 406: BF 223 (AEG-Telefunken)
T 407: 2 N 3866

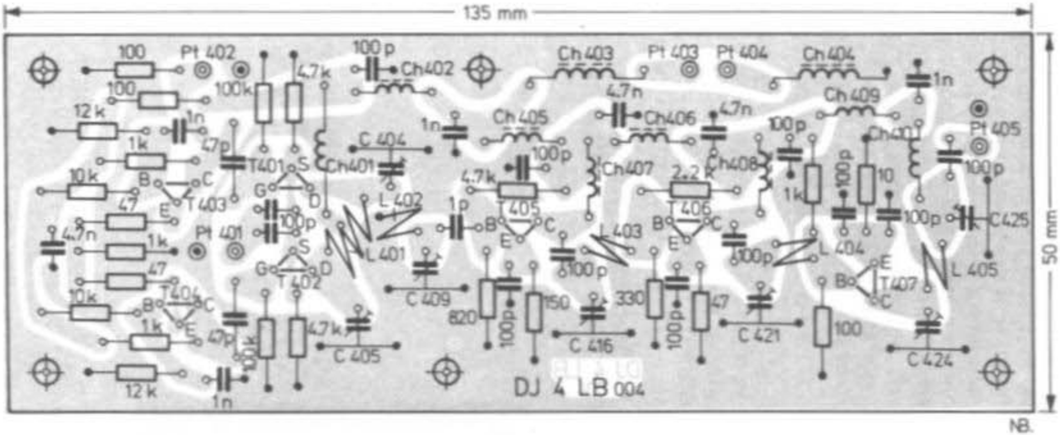

Fig. 23: Printed circuit board and component locations of the mixer and linear amplifier module

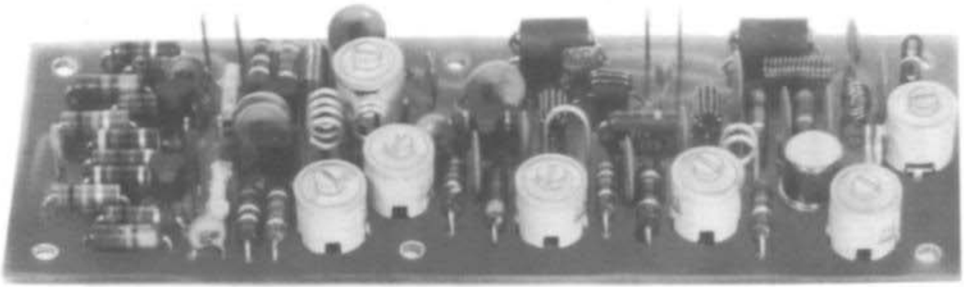

Fig. 24: Author's prototype of module DJ 4 LB 004

All inductances are made of 0.8 mm dia. (20 AWG) silver-plated copper wire as given, self-supporting.

L 401: 3.75 turns, 4 mm inner diameter, spaced 2-3 mm from the board, centre tap.
L 402: 1.75 turns, inner diameter and spacing as L 401
L 403: 0.75 turns, U-shaped, spacing between the ends of the wire: 5 mm .Top of the "U": approx. 10 mm above the board.
L 404: 1.75 turns, inner diameter 4 mm, spacing to board: approx. 1 mm
L 405: 1.75 turns, inner diameter 5 mm, spacing to board: 2-3 mm

Ch 401, Ch 409: Approx. 17 cm of 0.4 mm dia. (26 AWG) enamelled copper wire, self-supporting, 3 mm inner diameter, coil length 10 mm
Ch 402, Ch 405, Ch 406: 3.5 turns of 0.4 mm dia. (26 AWG) enamelled copper wire pulled through a ferrite bead (3 mm dia., 5 mm length)
Ch 407, Ch 408: 6.5 turns, otherwise as Ch 402
Ch 403, Ch 404: Wideband ferrite choke Z = 800 Ω, 2.5 turns, 6 mm diameter 10 mm long (Philips)
Ch 410: 3 turns of 0.4 mm dia. (26 AWG) enamelled copper wire, self-supporting, 3 mm inner diameter, coil length approx. 3 mm.

C 404, C 405, C 409, C 416, C 421, C 424: 2-6 pF ceramic disc trimmer, 10 mm dia.
C 425: 3 - 12 pF ceramic disc trimmer 10 mm dia.
C 401, C 406: 47 pF ceramic tubular capacitor for 10 mm spacing.

All other capacitors: Ceramic disc types for 5 mm spacing.
A spacing of 12.5 mm is available for all resistors.

5.4. ALIGNMENT OF MODULE DJ 4 LB 004

The local oscillator frequency is now connected to connection Pt 401 and the video IF signal to input Pt 402. A reflectometer for indicating the output power can be connected between connection Pt 405 and the 60 Ω terminating resistor as has been already described for the alignment of module DJ 4 LB 003. The 435 MHz resonant circuits of the amplifier are aligned to resonance with the aid of a RF voltmeter until the reflectometer indicates RF-power at the output. The IF input power is now increased in steps (P 102 of module DJ 4 LB 001), until the UHF output voltage does not increase noticeably in spite of the adjustment of the resonant circuits. The alignment of the Pi-filter at the output is made for maximum output power by alternate adjustment of the two appropriate trimmer capacitors.

On touching the various turns of inductance L 401 (e.g. with a screw-driver) it is possible to easily find the electrical centre point where the lowest reduction of the output power is obtained. This electrical point can be shifted to the connection point of choke Ch 401 by appropriate adjustment of trimmers C 404 and C 405.

An equal sound IF voltage (adjustable with P 202 of module DJ 4 LB 002) is now fed to the module instead of the video IF voltage. Trimmers C 404 and C 405 are tuned in the same direction so that the sound output power increases and approximates the value of the previously measured video level. The tuning of the other resonant circuits remains unchanged. If the module produces the same output power when fed alternately with the video and sound intermediate frequency, the passband characteristic will correspond approximately to that given in Figure 21.

A fine alignment with the aid of a sweep frequency generator allows the passband characteristic to be aligned to that shown in Figure 22. In this case, the IF voltage from the sweep frequency generator is fed to the connection Pt 402; the UHF demodulator DJ 4 LB 005 is capacitively coupled to the UHF output Pt 405 and is terminated with 60 Ω so that it will supply the measuring voltage for the Y-input of the oscilloscope. As has been previously mentioned, the bandpass filter comprising inductances L 401 and L 402 has a large effect on the passband characteristics of the whole module. The single-link characteristic caused by the extremely close coupling of the inductances (spacing approx. 1 mm) is advisable for the alignment with meters to indicate maximum output power. If an oscilloscope is available, the passband characteristic can be aligned for a critical or transitional bandpass filter coupling which is indicated by the formation of the humps. This is obtained by increasing the spacing between the inductances to approximately 4 mm and shifting inductance L 402 so that the axis of the two inductances form an angle of approximately 45° to another. Finally, the appropriate trimmer capacitors of the bandpass filter are realigned. The adjustment of the other resonant circuits can also be corrected and a slight shaping of the passband curve on altering the sweep frequency voltage is advisable. The video and sound carriers are marked in Figure 22 with a spacing of 5.5 MHz, which can be seen when using the frequency markers of the sweep frequency generator. The position of the video carrier within the passband curve is selected so that the lower, residual sideband of the video modulation spectrum is not attenuated for approximately 1 MHz at UHF level, as is laid down in the CCIR standard.

If all other resonant circuits of the amplifier are aligned at 435 MHz for maximum output power and symmetrical passband characteristic is adjusted subsequently with the bandpass filter, the passband ripple of the output voltage in the range of 430 - 440 MHz will be less than 1 dB. With the module aligned in this manner, it would be possible, for instance, for a 28 - 30 MHz SSB signal to be fed to the IF input and mixed with an appropriate local oscillator signal of 402 or 404 MHz and mixed to obtain any output frequency on the 10 MHz wide 70 cm band.

6. UHF DEMODULATOR DJ 4 LB 005

The higher the output power, the more difficult it will be to monitor the ATV signal radiated on 70 cm. It is not possible to establish whether any picture distortion is due, for instance, to incorrect modulation of the transmitter, or to overloading of the RF amplifier of the TV receiver. The monitor circuit DJ 4 LB 005 operates without overloading up to an UHF input voltage of several volts and is therefore suitable for checking the quality of the radiated video signal. As the circuit diagram given in Figure 25 indicates, the module comprises a 435 MHz resonant circuit in the input which is followed by a demodulator diode. The resulting composite video signal is fed through a two-stage emitter follower which allows a low impedance termination.

6.1. CIRCUIT DETAILS

Usually, several centimeters of insulated wire are sufficient at connection Pt 501 when it is placed in the vicinity of the final transistor T 407 or the output tube

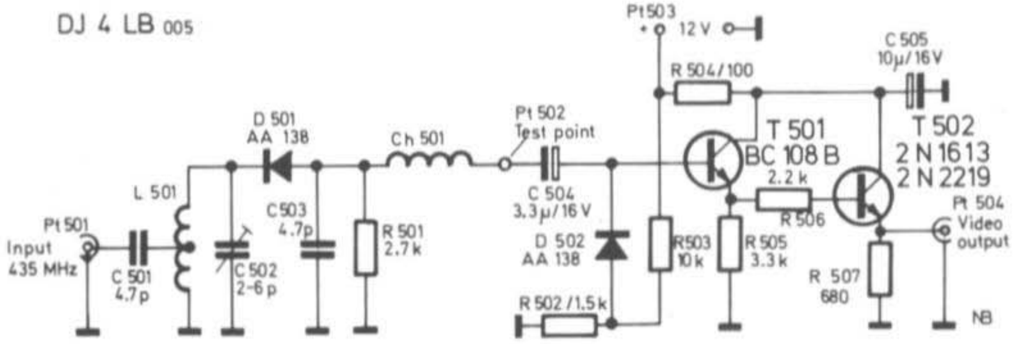

Fig. 25: Circuit diagram of the UHF demodulator DJ 4 LB 005

of a linear amplifier. The resonant circuit L 501/C 502 has a bandwidth of approximately 20 MHz which means that the tuning is not critical for the exact reproduction of the video frequency response. Pt 502 is a test point provided for connection of an oscilloscope. As has been already described, an oscilloscope can be connected during the swept frequency measurement and the measured voltage fed to the oscilloscope. It is also possible for the modulation signal regained by UHF demodulation to be compared with the original input signal at Pt 101 of the video IF module. The operation of the subsequent clamper circuit ensures that the composite video signal is added to the DC-voltage generated by the voltage divider R 502/R 503. This leads to a low quiescent current of the two-stage emitter follower and to a low power dissipation of the final transistor T 502, which is mainly dependent on the amplitude of the signal. If a short-circuit is made at the output (Pt 504), the protective resistor R 504 will reduce the collector voltage of the transistors so that they cannot be damaged by overloading. The base bias resistor R 506 limits the maximum current flowing via the output transistor. The output of the circuit (Pt 504) is at low impedance, a signal with a low-load voltage of 2 V will be reduced with a 60 Ω termination to approximately 1.5 V (both values peak-to-peak). The maximum, undistorted output voltage amounts to approximately 6 V (peak-to-peak).

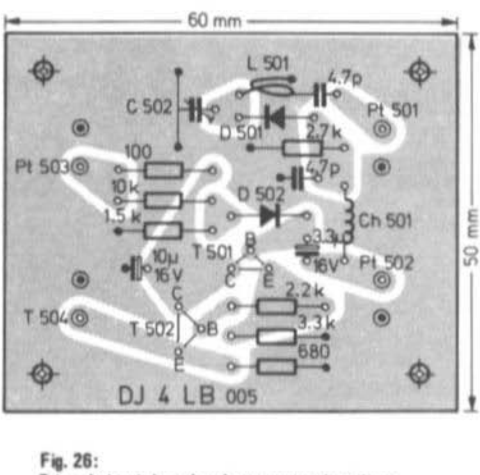

Fig. 26:
Printed circuit board and component locations of the UHF demodulator module

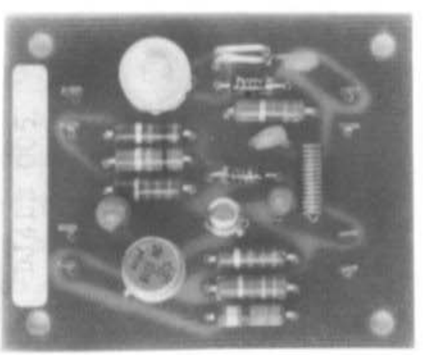

Fig. 27:
Author's prototype of module DJ 4 LB 005

6.2. NOTES REGARDING THE CONSTRUCTION AND ALIGNMENT

The UHF demodulator circuit is accommodated on a single-coated PC-board designated DJ 4 LB 005 (Fig. 26). The dimensions of this PC-board are 60 mm x 50 mm which allows the module to be accommodated in a TEKO-box 2 A or 2 B. The larger case provides sufficient room, e.g. for a RF connector if the module is to be installed in a TV receiver and the UHF input signal injected with the aid of a coaxial cable. Figure 27 shows the author's prototype of the UHF demodulator.

The alignment of the monitor circuit only comprises the alignment of the 435 MHz resonant circuit which is adjusted with the aid of trimmer C 502 for maximum (max. contrast).

6.3. SPECIAL COMPONENTS

T 501: BC 108 or similar (B min. 100)
T 502: 2 N 1613, 2 N 2219 or similar

D 501, D 502: AA 138 (AEG-Telefunken)

C 502: 2 - 6 pF ceramic disc trimmer 10 mm dia.

L 501: 1.75 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire, self-supporting, 6 mm inner diameter, coil tap 1 turn from the cold end, spaced 2 to 3 mm from the PC-board.

Ch 501: Approx. 17 cm of 0.4 mm (26 AWG), 3 mm inner diameter, self-supporting, coil length 10 mm.

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VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 89-92

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A 144 MHz LINEAR AMPLIFIER WITH 25 W OUTPUT AT 12 V TO 14 V

by G. Otto, DC 6 HL

The following linear amplifier has been developed in order to increase the output power of the DC 6 HL SSB transceiver described in (1) for mobile and fixed operation. It is equipped with two transistors and offers a maximum output power of 25 W (single-tone, linear operation) at an operating voltage of 12 V to 14 V. The required drive power of 2.5 W can, of course, be obtained from any other exciter, in any mode (SSB, AM, FM, CW).

One of the special features of this linear amplifier is the RF-VOX circuit and a lowpass filter. The RF-VOX automatically switches the linear amplifier from the "switched-through" condition to "amplifier" operation. No control line is required between the transceiver and the linear amplifier. The lowpass filter at the output is provided to suppress the harmonic content which is more prevalent with transistor amplifiers than with tubed types. This filter represents a very necessary part of the amplifier module.

The linear amplifier, including RF-VOX and lowpass filter are accommodated on a printed circuit board that is mounted in a TEKO box. This means that construction is very easy. Figure 1 shows the author's prototype, but without heat sink.

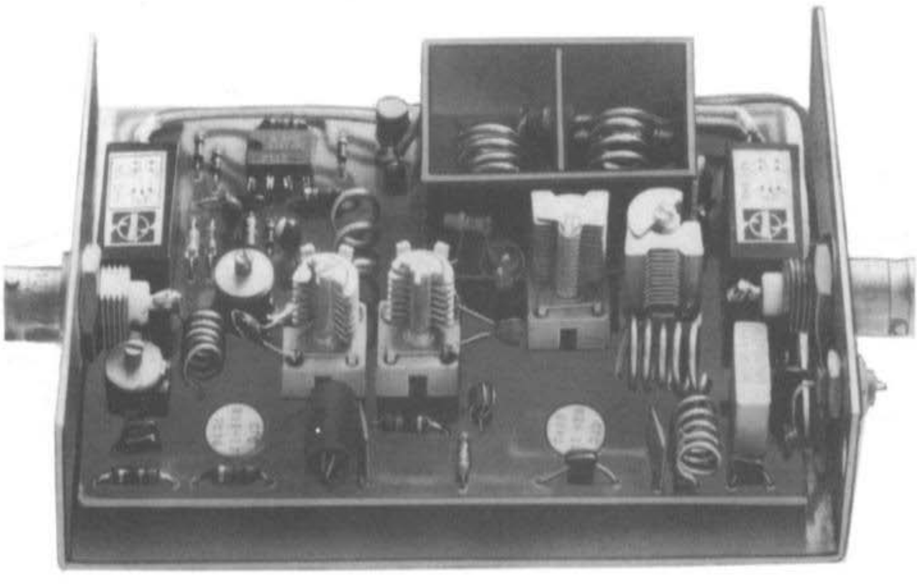

Fig. 1: A 25 W Linear amplifier for 144 MHz

1. CIRCUIT DETAILS

The circuit diagram of the linear amplifier, RF-VOX and lowpass filter is given in Figure 2. The three parts of the complete linear amplifier are now to be described:

1.1. LINEAR AMPLIFIER CIRCUIT

Trimmer capacitors C 902, C 903 and inductance L 901 transform the complex input impedance of transistor T 901 to 50 Ω to 60 Ω. In order to obtain a good intermodulation rejection even at low drive levels, a positive base bias voltage is provided so that a low, quiescent collector current flows. Resistors R 901 and R 902 and diode D 901 (temperature-dependent resistor) form the voltage divider for the bias voltage. The diode has a direct thermal contact to the transistor case and ensures that the collector current cannot increase beyond a certain point.

The transformation link comprising C 905, C 906 and L 902 matches the output of the first stage to the input of the second. The high-pass filter characteristic of the transformation link ensures that any low-frequency oscillation due to resonance conditions on the chokes is suppressed so that no interaction occurs.

The temperature-compensated, bias voltage supply for the final transistor is identical to that of transistor T 901. The operating voltage for the final amplifier is fed via inductance L 904. This inductance is in parallel resonance at 145 MHz together with the dynamic output capacitance of the transistor and the reactive component of the transformed load impedance caused by L 903, C 908 and C 909 in the collector circuit. This circuit configuration avoids the use of a choke, which, at higher power levels would either possess too much loss, or would be expensive.

1.2. HARMONIC FILTER

As is known, transistor amplifiers generate a higher harmonic content than tubed amplifiers. For this reason, a double, lowpass filter comprising C 910, C 911, C 912, L 905 and L 906 has been provided at the output. If the value of C 910 is 20 pF, the input and output impedance will be 60 Ω. In the described circuit, the value of C 910 has been increased to 39 pF so that a value of maximum 30 pF is sufficient for the parallel trimmer C 909.

1.3. RF-VOX

If no RF voltage is passed from the input Pt 901 via capacitor C 901 to the rectifier circuit of the RF VOX, no voltage will be present between the two inputs of the operational amplifier I 901. The output of the amplifier is kept at 0 V by connecting a resistor of 680 kΩ between the offset compensation input (connection 5) and the operating voltage $+U_b$. Any RF voltage at the input will cause a voltage difference between the two inputs of the differential amplifier. This will mean that the output voltage of the amplifier will increase to $+U_b$ due to the high gain. The subsequent emitter follower is then switched, which will in turn switch the two miniature relays.

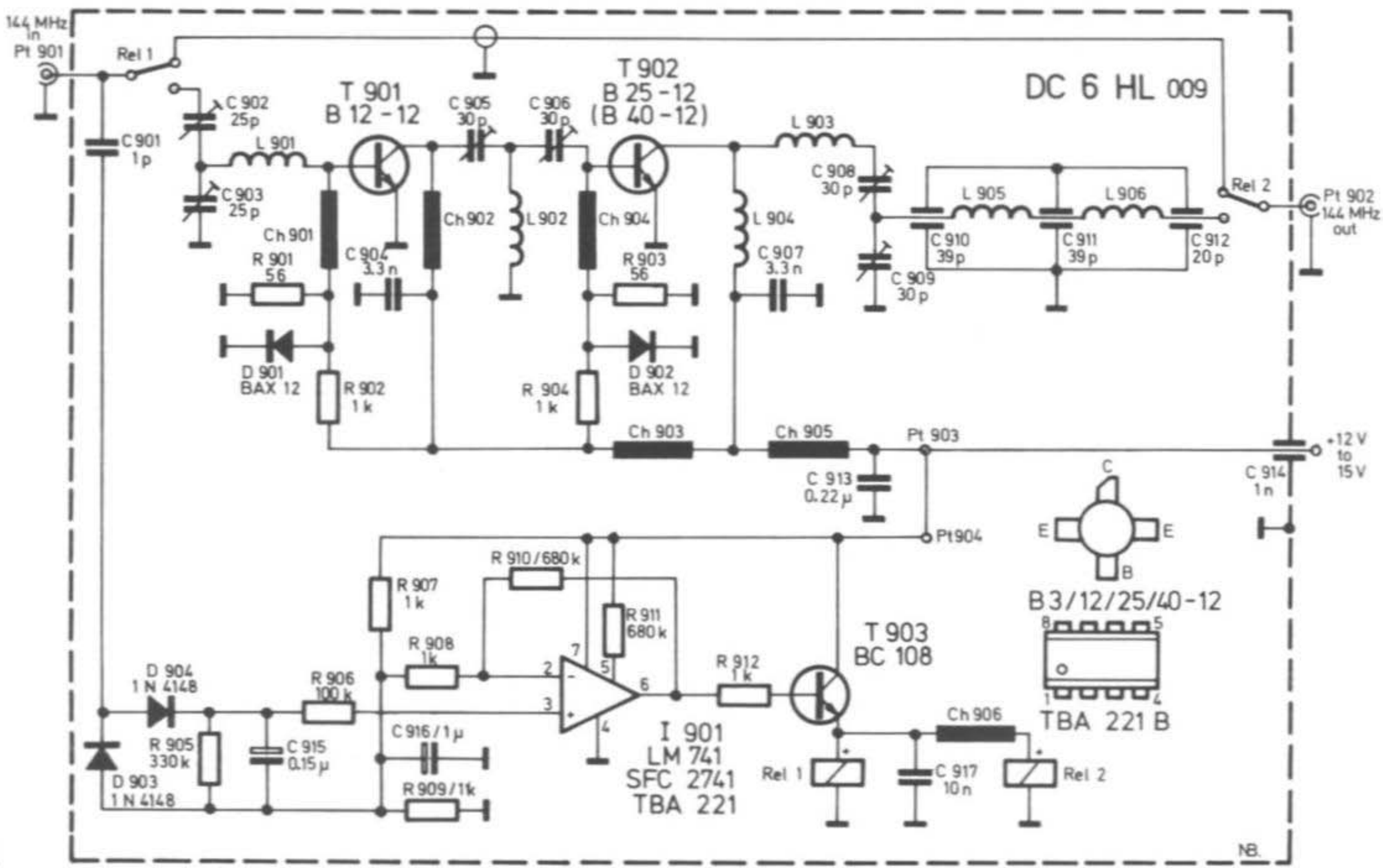

Fig. 2: Circuit diagram of the 25 W linear amplifier, with HF–VOX and lowpass filter

The described arrangement possesses such a high sensitivity that even the suppressed carrier of the SSB signal will switch the linear amplifier into the linear amplifier mode. Since the amplifier will therefore not be switched off in the pauses between words (since the suppressed carrier will still be present in the PTT mode), it is possible for very short fall times to be used for the RF-VOX so that no information is lost on switching from transmit to receive.

2. CHARACTERISTICS

The author was able to carry out exact power, intermodulation and suppression measurements under laboratory conditions.

2.1. POWER

Figure 3 gives the output power P_{out} of the linear amplifier as a function of the drive power P_{in} for two different operating voltages U_b. With the transistors type B 12-12 and B 25-12 used by the author, approximately 20 W was obtained in the linear range of the characteristic at U_b = 12.5 V and approximately 25 W at U_b = 14 V (see section 2.2.). The drive power was in the order of 2 W to 2.5 W. The DC 6 HL SSB transceiver provides an output power of 1.5 W to 2 W at 14 V.

At an operating voltage of U_b = 12.0 V and an output power P_{out} = 25 W, the total DC drain I_{tot} amounts to 4.2 A, which indicates a total efficiency of 49%.

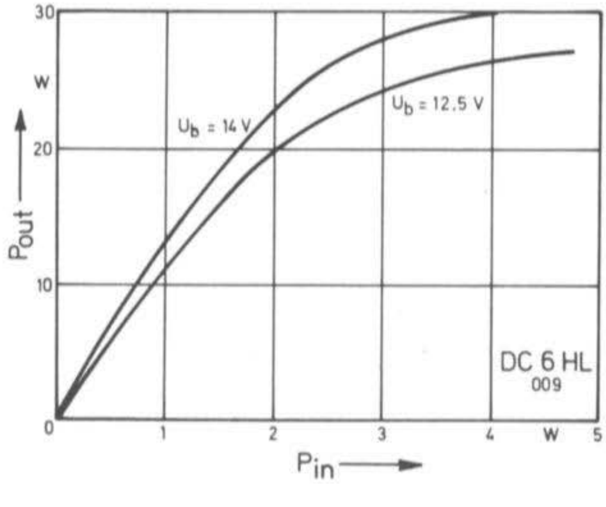

Fig. 3: Output power P_{out} as function of the drive power P_{in} at two different operating voltages

Other transistor types or individual transistors of the same type were not available at the time the tests were made. The same manufacturer offers a higher power version of the B 25-12 which is designated B 40-12, and provides an output power in the order of 40 W to 45 W. This transistor can also be driven by the same B 12-12. It requires a drive power of approximately 2.5 W. The smallest transistor of this family is the B 3-12, which can be used as driver for type B 12-12 as final amplifier. This combination offers approximately 12 W for a drive power of approximately 0.3 W. However, it has not been established whether the adjustment ranges of the transformation links are sufficient. The same is valid for similar transistor types manufactured by different companies. It would be, for instance, possible to combine a BLY 87 (Philips) with a 2 N 5591 (Motorola) which should provide an output power of approximately 28 W at a drive power of approximately 1.5 W.

2.2. INTERMODULATION SUPPRESSION

The intermodulation suppression of the described linear amplifier was measured in the test set-up given in Figure 4 under the following conditions:

Operating voltage $U_b = 12.5$ V
Frequencies: $f_1 = 145.1$ MHz; $f_2 = 145.3$ MHz
Drive powers: $P_{in}1 = P_{in}2$;

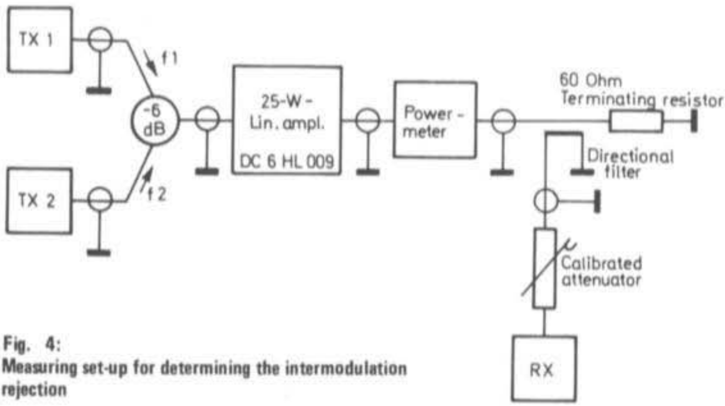

Fig. 4:
Measuring set-up for determining the intermodulation rejection

Measured was the attenuation of the intermodulation product $f_2 + (f_2 - f_1) =$ 145.5 MHz or $2f_2 - f_1 = 145.5$ MHz referred to the two wanted output powers $P_{out\,f1}$ or $P_{out\,f2}$.

$P_{out\ tot}$	Intermodulation suppression
5 W	30 dB
20 W	28 dB
25 W	26 dB

2.3. HARMONIC SUPPRESSION

The suppression characteristics of the built-in, double-link lowpass filter was measured in the measuring set-up shown in Figure 5. Figure 6 shows the result of the measurement. The first harmonic (288 MHz is suppressed by more than 30 dB and the second by more than 45 dB, without filter.

After passing the signal through the filter, the first harmonic is suppressed by more than 60 dB at an output power of 20 W and the second and third har-

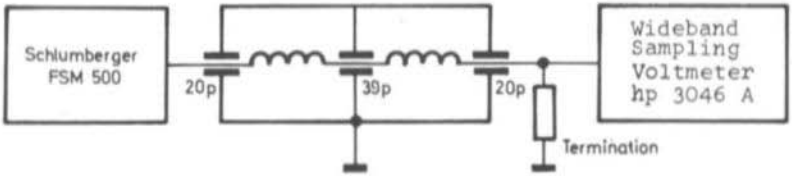

Fig. 5: Measuring set-up for determining the harmonic suppression

monics by more than 80 dB. The insertion loss of the filter at 145 MHz amounts to 0.3 dB. Of course, the filter is just as suitable for use with other transmitters. The two chambers can be made, for instance, from PC-board material. In this case, the input capacitor C 910 must have a value of 20 pF; the input and output impedance is approximately 60 Ω.

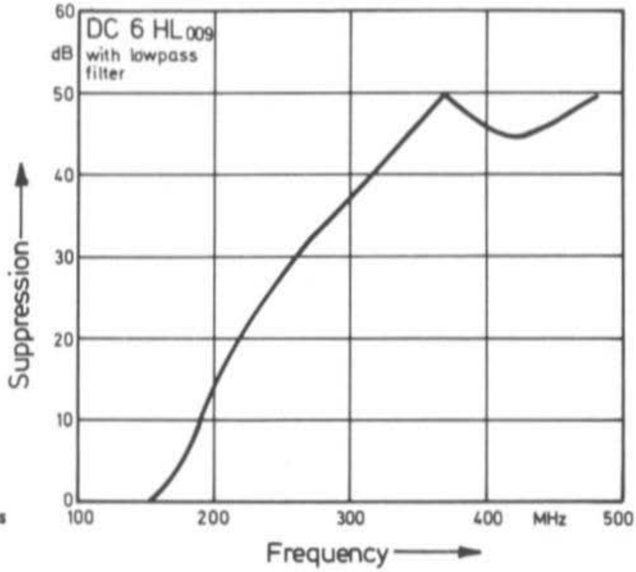

Fig. 6:
Stopband characteristics of the lowpass filter

3. CONSTRUCTION

With the exception of the feedthrough capacitor C 914, all components are accommodated on a single-coated printed circuit board. The printed circuit board, which has been designated DC 6 HL 009, is 95 mm x 65 mm. The printed circuit board and the component locations are given in Figure 7. The whole linear amplifier module can be installed in a TEKO box 3 A and screwed complete with box onto a heat sink. The photograph of the author's prototype given in Figure 1 gives further details as to the construction.

The two power transistors are inserted from the conductor side of the PC-board into the two 10 mm holes. The connection strips should be shortened to approximately 3 to 4 mm and directly soldered to the conductor lanes. The tapered connection strip is the collector connection (see Fig. 2). The transistor bolts are placed through the corresponding holes in the TEKO box and the heat sink, and is used to mount these together. The contact surfaces should be provided with heat-conductive paste. Diodes D 901 and D 902 are glued to the associated transistor with heat-conductive paste and soldered to the corresponding conductor lanes of the PC-board. The small 2-chamber case for the lowpass filter is 18 mm in height and should be soldered together from brass or tin plate, or from PC-board material. The dimensions are given in the component location plan (Fig. 7). The filter capacitors C 910, C 911 and C 912 are soldered half-way up the walls and the two coils are soldered, self-supporting between the capacitors. After the preliminary alignment, the filter is mounted onto the ground surface of the board with a wire at the input and output.

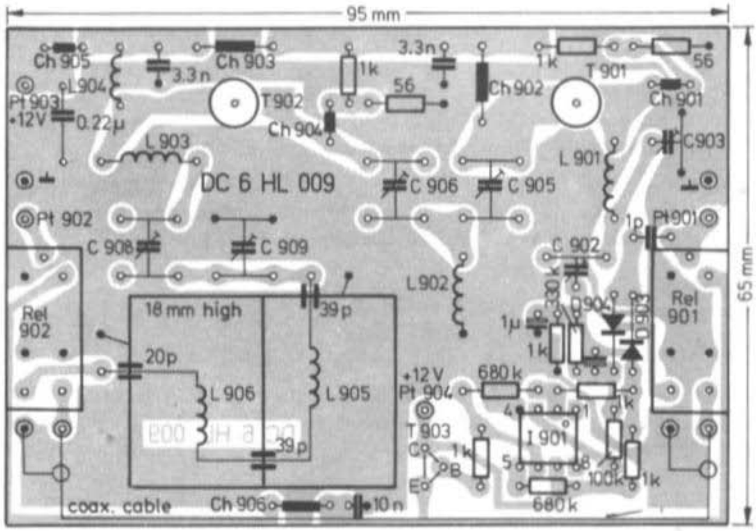

Fig. 7: Component locations on PC-board DC 6 HL 009

The two coaxial connectors are mounted at a suitable position on the narrow side of the TEKO box and connected to the ground of the PC-board, and feed-through capacitor C 914. The box is completed by screwing on the cover.

3.1. SPECIAL COMPONENTS

T 901: B 12-12 Communications Transistor Corporation
T 902: B 25-12 Communications Transistor Corporation
T 903: BC 108 or similar

I 901: TBA 221 B (Siemens), LM 741 (National Semiconductor), SFC 2741

D 901, D 902: BAX 12 (Philips) or 1 N 914, 1 N 4148
change R 902 and R 904 if necessary
D 903, D 904: 1 N 4148, 1 N 914 or similar

L 901: 5 turns of 0.6 mm dia. (23 AWG) silver-plated copper wire, inner diameter 4 mm, self-supporting, spaced one wire diameter.
L 902: 3 turns of 1 mm dia. (18 AWG) silver-plated copper wire, inner diameter 4 mm, coil length 13 mm, self-supporting.

L 903: 5 turns of 1 mm dia. (18 AWG) silver-plated copper wire, inner diameter 7 mm, self-supporting, winding spaced one wire diameter.
L 904: 5 turns of 0.6 mm dia. (23 AWG) silver-plated copper wire, inner diameter 4.5 mm, self-supporting, coil length 8 mm.
L 905, L 906: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire, 7 mm inner diameter, self-supporting, winding spaced approx. 1.5 mm.

Ch 901: 2 turns of approx. 0.5 mm dia. (24 AWG) enamelled copper wire passed through a ferrite bead
Ch 902: Wideband ferrite choke with 6 holes, Z = 800 Ω (Philips)
Ch 903 - Ch 906: As Ch 901.

Rel 1, Rel 2: Encapsuled miniature relays for PC-board mounting, National RH - 12 V (available from the publishers).

C 902, C 903: 2 - 24 pF plastic-foil trimmer, 7 mm dia.
C 905, C 906, C 908, C 909: 3 - 30 pF air-spaced trimmer.
C 910, C 912: 20 pF tubular feedthrough capacitor for solder mounting.
C 911: 39 pF tubular feedthrough capacitor

Heat sink: 100 mm x 120 mm, 25 mm high with 10 fins, mounted asymmetrically to the PC-board so that the power transistors can be mounted between the two groups of five fins.

4. ALIGNMENT

4.1. LOWPASS FILTER

Firstly construct the lowpass filter so that C 910 has a value of 20 pF.

Align a 2 m transmitter to 146 MHz and connect it to the input of the filter. The output of the filter should be connected to a wattmeter (or reflectometer) with terminating resistor.

Inductances L 905 and L 906 should be pulled out until no increase of power is indicated.

Increase the value of C 910 to 39 pF by adding a parallel capacitor and solder the filter onto the PC-board.

4.2. RF-VOX

No alignment is necessary for this part of the circuit. If required, the coupling capacitor C 901 can either be reduced or increased for different power levels to ensure correct operation of the automatic circuit.

4.3. AMPLIFIER STAGES

The linear amplifier is connected in the set-up given in Figure 8 with exciter, wattmeter (or reflectometer) and terminating resistor and fed via a current limiting circuit with 12 V.

Point "P" should be temporarily grounded and the short-circuit current of the 500 Ω trimmer resistor adjusted to approximately 1 A.

Remove the short-circuit and drive tne amplifier with approximately 100 mW. The RF-VOX circuit should switch.

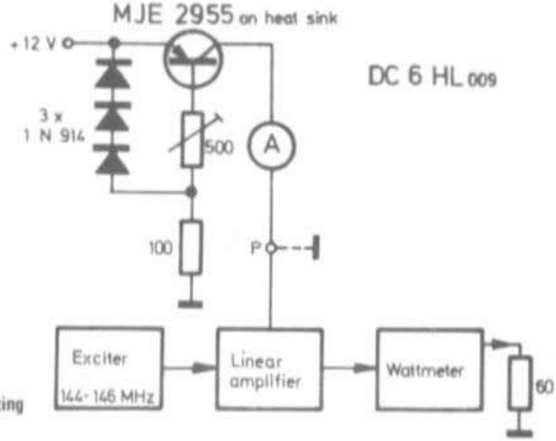

Fig. 8:
Measuring set-up for alignment with current limiting

Align the input trimmer capacitors C 902 and C 903 for maximum current drain to the amplifier. The same is valid for C 905 and C 906 of the final transistor.

The output trimmers C 908 and C 909 should be adjusted for maximum output power.

Remove the current limiting circuit and connect the power supply (e.g. such as that given in Fig. 9) directly to the amplifier.

Increase the RF drive power to a maximum of 3 W and align all trimmers from output to input for maximum output power. If the adjustment ranges of the capacitors are not sufficient, the turns of inductance L 901 should be pressed closer together and those of L 902 pulled further apart.

If the power supply given in Figure 9 is to be used, it is important that the output is loaded with at least 50 mA and bridged with a 0.1 μF capacitor.

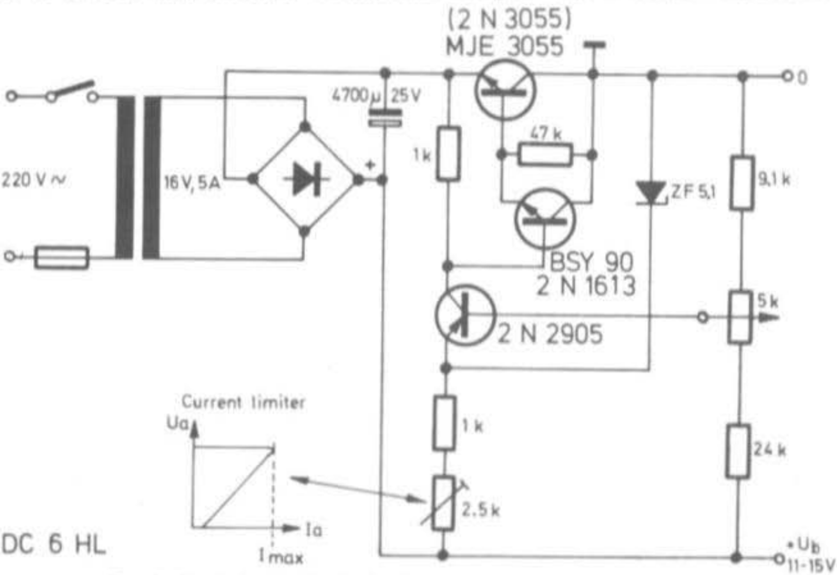

Fig. 9: Circuit diagram of a simple voltage stabilizer for max. 5 A with current limiting

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VHF COMMUNICATIONS 4 (1972), Edition 4, Page 207

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A DUAL-INPUT PREAMPLIFIER WITH 2:1 PRESCALER FOR FREQUENCY COUNTERS FROM 1 Hz TO A MINIMUM OF 100 MHz

by W. R. Kritter, DL 8 TM

This preamplifier/prescaler is designed for operation with the frequency counter described in (1). To be described is a dual-input preamplifier with pulse shaper and 2 : 1 prescaler. This module has been designated DL 8 TM 003 and can be used in conjunction with other frequency counters (2), (3). Due to the 2 : 1 division of the frequency to be measured, it is necessary for the clock frequency of the counter in question to also be divided by 2 : 1. A free flip-flop is available in module DL 8 TM 002 for this purpose. The frequency counter DL 8 TM 002 is already designed for use with this prescaler and no modification is required. Figure 1 in (1) shows how the two modules are interconnected.

1. CHARACTERISTICS

Both amplifiers operate in the voltage range of approximately 100 mV to 1 V (RMS) without requiring any adjustment. Since the majority of measurements are made in this voltage range, the author is of the opinion that this is suitable for most applications. If lower voltage levels are to be measured, it will be necessary for a preamplifier to be provided, for higher levels, an attenuator can be placed in front of the input. Such input networks with a limited frequency range can be made without difficulty.

The AF preamplifier amplifies and shapes frequencies between 1 Hz and approximately 1 MHz. The RF preamplifier is able to process frequencies from approximately 100 kHz. The upper frequency limit is determined by the two Schottky TTL integrated circuits SN 74 S00 N and SN 74 S 112 N. The manufacturer gives a value of typically 125 MHz for both types. However, if the first counting decade (I 216: SN 74 196 N) in the frequency counter only obtains its guaranteed minimum of 50 MHz, this will limit the upper measuring frequency limit (100 MHz). If this counting decade is better, it will be possible for the high limit frequency of the Schottky ICs to be fully utilized; it may even be worthwhile using the selected type SN 74 S 112 N S 1.

2. CIRCUIT DETAILS

Figure 1 shows the circuit diagram of the dual input preamplifier with 2 : 1 prescaler. The circuit comprising transistors T 301 to T 305 processes the higher frequencies. The last three stages represent a further development of the preamplifier for frequency counters up to 60 MHz as was described in (4). In order to ensure that higher frequencies can be processed, the previously used first stage (T 302) is equipped with a base voltage divider and a feedback resistor in the emitter circuit. An additional amplifier stage equipped with the UHF transistor T 301 is used to compensate for the loss in sensitivity. This stage operates as an amplifier for low input voltages and a limiter for higher input voltage levels. In addition to this, several resistance values have been changed with respect to the original design.

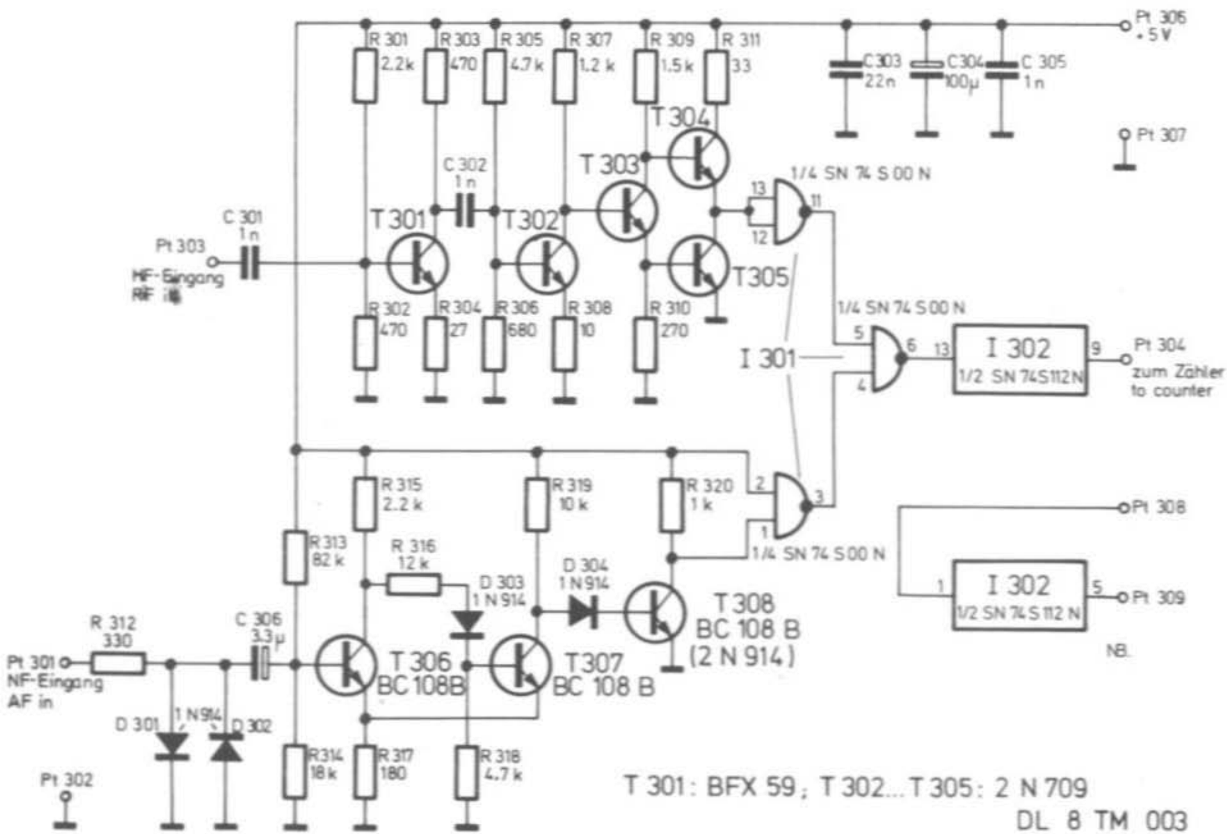

Fig. 1: Circuit diagram of the dual-input preamplifier with 2:1 prescaler

The lower part of the circuit diagram shows the AF preamplifier equipped with transistors T 306 to T 308. Two silicon diodes are connected in antiphase at the input which limit the input voltage to a peak value of approximately 0.7 V. The input signal is amplified in transistor T 306. The subsequent transistor T 307 is only driven with the positive half wave via diode D 303 and operates as a switch. The last transistor (T 308) supplies the required pulses for the TTL logic circuits. Diode D 304 ensures that it is completely blocked.

Both amplifiers possess a logic-0 (less than 0.7 V) when no input signal is present. The 0-level is fed via an OR-gate which is connected as an inverter. The logic 1-signal is then fed to a further OR-gate (connections 4, 5 and 6 of I 301). The OR-gate passes the pulses from one input to the output when the other input is at logic-1 level. According to this principle, it connects the two inverted outputs of the amplifier with the subsequent flip-flop, which divides the input frequency by 2 : 1. The second flip-flop in the integrated circuit SN 74 S 112 N is not used and can be used for division of the time base frequency of a counter or for further division of the input frequency, if this is required for the counter in use. These connections are fed to suitable points on the printed circuit board.

3. SPECIAL COMPONENTS

T 301: BFX 59 (Siemens)
T 302 - T 305: 2 N 709 (Texas Instruments)
T 306 - T 308: BC 108 B or similar (B min. 100)

D 301 - D 304: 1 N 914, 1 N 4148 or similar silicon diode

I 301: SN 74 S 00 N
I 302: SN 74 S 112 N (S 1)

All resistors with 5% tolerance, spacing 10 mm .
All unipolar capacitors are ceramic disc types for 5 mm spacing.

C 304: 100 μF/6 V tantalium electrolytic
C 306: 3.3 μF/6 V tantalium drop-type electrolytic

4. CONSTRUCTION

The dual-input preamplifier with 2 :1 prescaler is accommodated on a PC-board whose dimensions are 85 mm x 55 mm. Figure 2 shows the printed circuit board and component locations of this board, which has been designated DL 8 TM 003. Due to the high input frequencies that can be processed in this module and the steep rise times in the RF-amplifier, it is necessary for the connection wires of all components to be kept as short as possible. The interconnection between the output of this module and the counter should also be as short as possible since even half the input frequency possesses fast rise times. In spite of this, it is advisable for sockets to be provided for the two integrated circuits so that it is possible to select the IC with the highest input frequency. Practice has shown that this is more advantageous than the extra length of the connections. Figure 3 shows a photograph of the author's prototype.

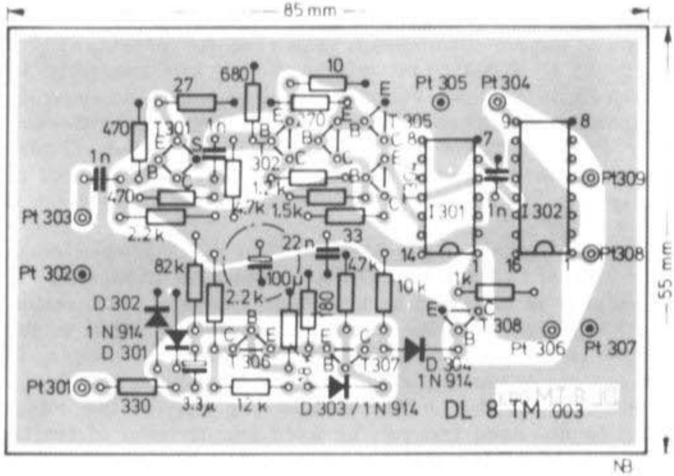

Fig. 2: PC-board DL 8 TM 003

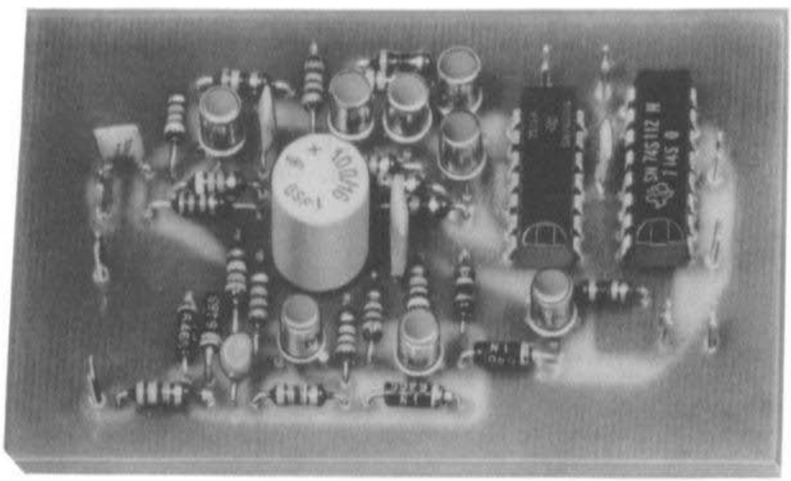

Fig. 3: Author's prototype of module DL 8 TM 003

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(1) W.R.Kritter: Six-Digit Frequency Counter for Frequencies between 1 Hz and Typically 100 MHz
In this edition of VHF COMMUNICATIONS

(2) F.Weingärtner: A Four-Digit Frequency Counter Module for Frequencies up to 30 MHz
VHF COMMUNICATIONS 3 (1971), Edition 3, Pages 159-171

(3) F.Weingärtner: Further Development of the Four-Digit Frequency Counter
VHF COMMUNICATIONS 4 (1972), Edition 4, Pages 229-234

(4) W.R.Kritter: Wideband Preamplifier for Frequency Counters up to 60 MHz
VHF COMMUNICATIONS 3 (1971), Edition 3, Pages 156-158

A SIX-DIGIT FREQUENCY COUNTER FOR FREQUENCES BETWEEN 1 Hz AND TYPICALLY 100 MHz

by W. R. Kritter, DL 8 TM

The following frequency counter is equipped with TTL-circuits of the inexpensive SN 74 ... N series. Other families of integrated circuits, such as, for instance, ECL-circuits have not been used, mainly due to the extensive level converters that must be used and in order to guarantee simple construction. The present technological limits of TTL-circuits are utilized to the full due to the use of Schottky-TTL-circuits (SN 74 S ... N) in the first divider and counting gate.

1. THE WHOLE FREQUENCY COUNTER AND CHARACTERISTICS

The frequency counter comprises four modules: Power supply, time base, frequency counter and preamplifier with pulse shaper and 2 : 1 prescaler. Figure 1 gives the block diagram of a combination of these modules.

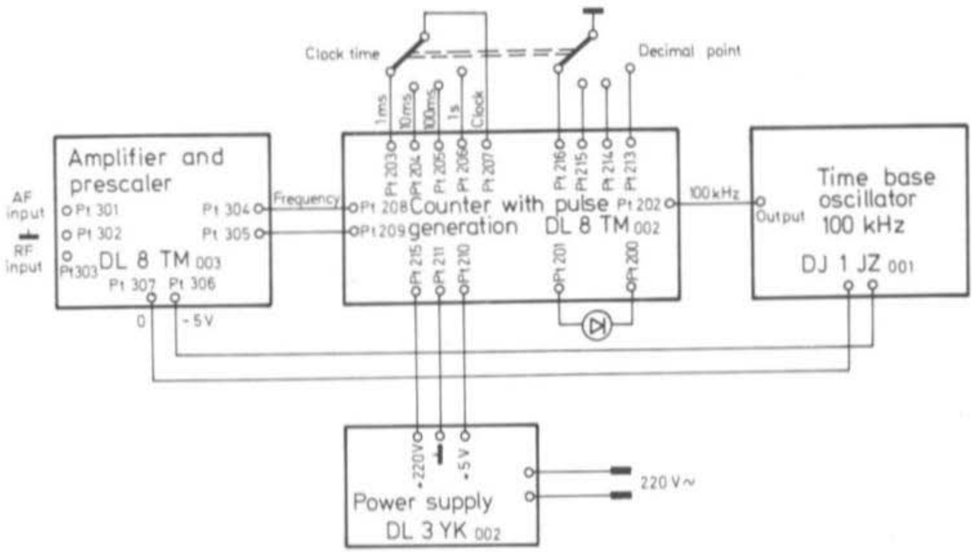


Fig. 1: Block diagram of the 6-digit frequency counter

The universal power supply with integrated control circuit described in (1) is used as power supply. This circuit provides 5 V ± 5% at approximately 0.7 A and 170 - 250 V at approximately 20 mA.

The extremely stable crystal oscillator described in (2) is used as time base. Due to the use of a crystal oven and special crystal, and/or synchronization with a standard frequency transmitter, it is possible for the upper counting limit to be indicated with a frequency resolution of 1 Hz. If the time base frequency were to differ by only $\pm 1 \times 10^{-7}$, this would mean that an indicated frequency of 100 MHz would be incorrect by ± 10 Hz. For this reason, the stability and accuracy of the time base must increase together with the resolution and the highest frequency to be measured.

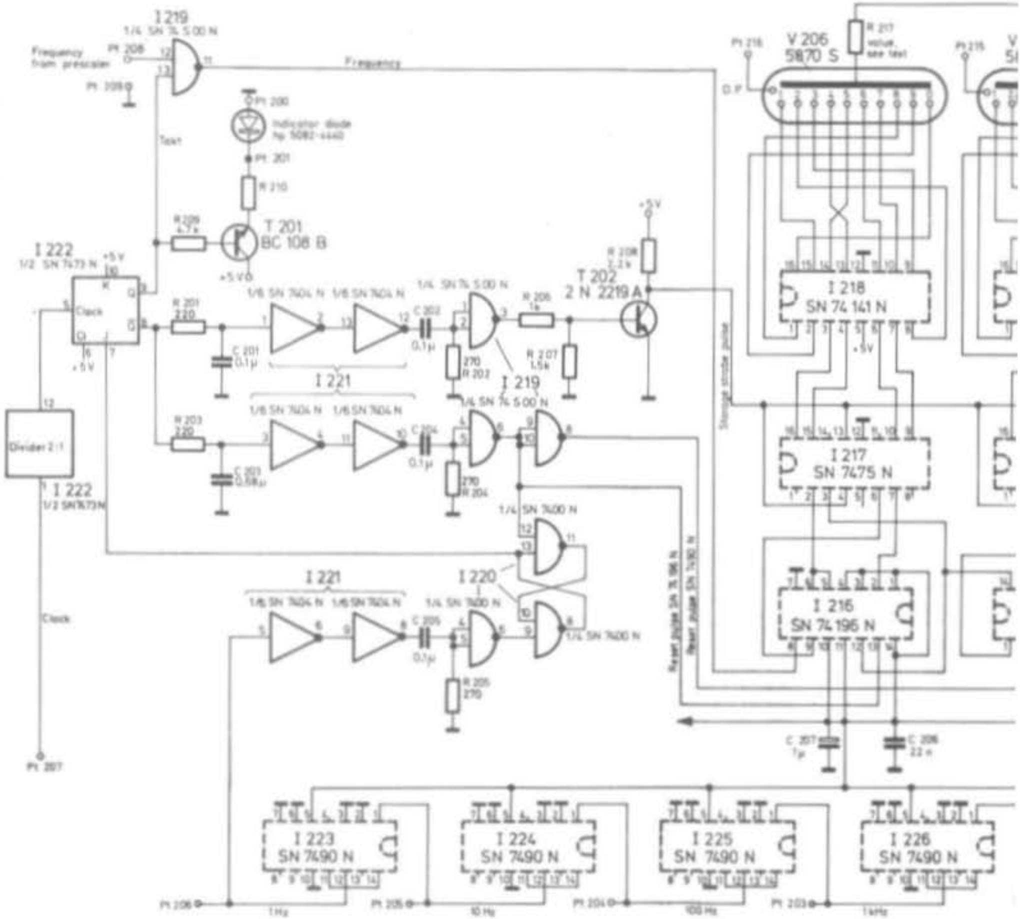
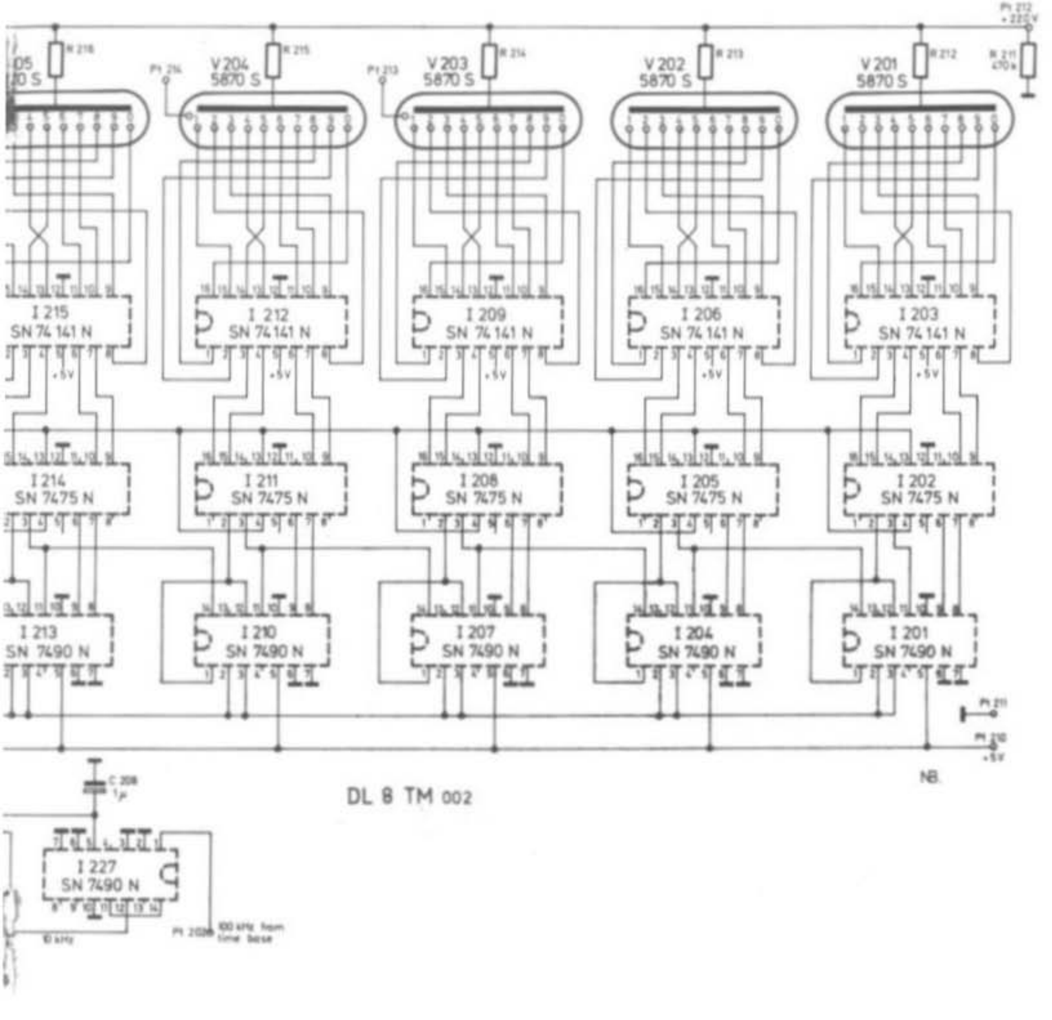


Fig. 3: Circuit diagram of t



ae frequency counter DL 8 TM 002

The module: preamplifier, pulse shaper and 2 : 1 prescaler is accommodated on PC-board DL 8 TM 003 which is described in this edition of VHF COMMUNICATIONS. This module is equipped with two preamplifiers, one for low frequencies and one for high frequencies. The output signals are fed via an OR-circuit to the prescaler. The manufacturer of the flip-flop used as 2 : 1 divider (type SN 74 S 112 N) gives a minimum upper frequency limit of 80 MHz. The selected type SN 74 S 112 N S1 (twice the price) is guaranteed for minimum 120 MHz. In the author's prototype of the preamplifier, the input divider was triggered by an input voltage of 100 mV at 140 MHz.

The actual frequency counter board is equipped with 27 integrated circuits of the TTL series SN 74 ... N, 2 transistors, 1 light-emitting diode, 6 indicator tubes as well as several resistors and capacitors. The relatively new miniature Nixie-type tubes are used for indication of the frequency. The digit-height is 13.5 mm and the tubes are only 30.5 mm high with a diameter of max. 13 mm. These tubes possess built-in decimal points and their connection wires allow them to be soldered into place so that only 5 mm spacing is present between the base of the tubes and the PC-board. Figure 2 shows a photograph of the author's prototype, which was not equipped with through-contacts.

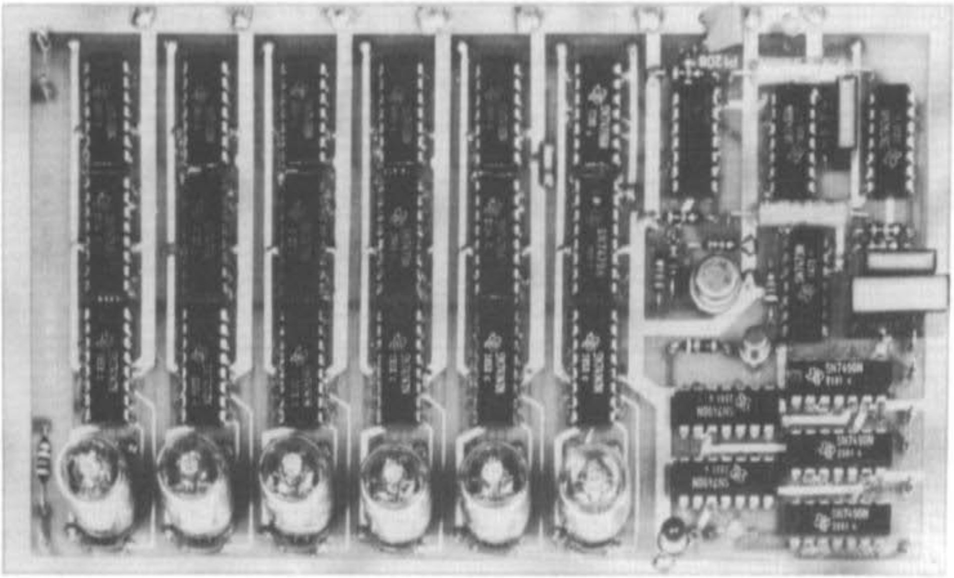


Fig. 2: Author's prototype of the frequency counter (PC-board did not have through-contacts)

2. FREQUENCY COUNTER MODULE DL 8 TM 002

The frequency counter module comprises the six counter and indicator decades, the pulse control, and divider chain for the time base. Figure 3, which is given on the centre pages, gives the circuit diagram of this module.

2.1. CIRCUIT DETAILS

The six counting decades each comprise a Nixie-tube, decoder-driver, storage and counter. A counter type SN 74196 N is used in the first decade (I 216). The upper frequency limit of this type is guaranteed to be 50 MHz. Due to the use of the 2 : 1 divider, the counter can be assumed to read typically 100 MHz and if the flip-flop in the prescaler operates at a frequency higher than the guaranteed minimum of 80 MHz and obtains its typical frequency limit. If both circuits possess an upper limit equal to their typical or even maximum values, the counter will be able to indicate frequencies up to 120 MHz or even 140 MHz.

The decade counter type SN 74 196 N can be used as replacement for type SN 7490 N without difficulty. In the application used here, attention must be paid that the reset pulse must have reverse polarity.

Due to the 2 : 1 division of the frequency to be measured, it is necessary for the clock time to be lengthened by a factor of 2 so that the indicated frequency corresponds to that of the input signal. This is obtained using one half of I 222 (SN 7473 N).

The counting principle is the same as was described in (3). The counting gate (I 219) is opened for a predetermined (switchable) time. The frequency to be counted has been shaped to corresponding pulses and is passed to the counting decades (I 201, I 204, I 207, I 210, I 213, I 216). After the counting gate has closed, the binary coded condition of the counter is passed to the storage (I 202, I 205, I 208, I 211, I 214, I 217). The result is indicated by the Nixie-tubes (V 201 - V 206) via the decoder and driver stages (I 203, I 206, I 209, I 212, I 215, I 218). The counting decades then return to zero and the frequency remains indicated with the aid of the storage until the next counting cycle has been completed.

The counter operates with a fixed counting time of 1 s which means that one count is made every second and the result is indicated independent of the time the gate is open. This counting frequency has been found to be satisfactory.

The operation of the pulse control of the counter is to be described with the aid of the pulse diagram given in Figure 4. It is assumed that J = 0 (I 222, connection 7; line 7 in Fig. 4) and the reset pulse I = 1 (I 216, connection 13; line 4 in Fig. 4) is valid so that a 1-condition is present at I 220, connection 11. This 1-condition results in a 0-condition together with the 1-condition at output 8 of I 220. This means that J is actually 0. The next rising slope of the 1 Hz signal (Pt 206, line 1 in Fig. 4) differentiated by C 205/R 205 places output 6 of I 220 temporarily to 0 (line 2 in Fig. 4). This means that output 8 of I 220 will temporarily correspond to 1. This means that J = 1 is valid. At the same time, connection 11 of I 220 will become 0. Since connection 9 of I 220 is at zero-potential, this condition will be maintained. Further 0 to 1 slopes of the 1 Hz clock signal will have no effect.

The next 1-0 slope that is fed from the clock divider (line 6 in Fig. 4) to the input (connection 5) of I 222 switches the flip-flop to the other position; 1 will appear at output Q, and 0 will appear at $\bar{Q}$ (lines 8 and 9 in Fig. 4). The frequency to be counted is fed from the prescaler of module DL 8 TM 003 and controls the counting gate I 219 at connection 12 until the next 1-0 slope of the

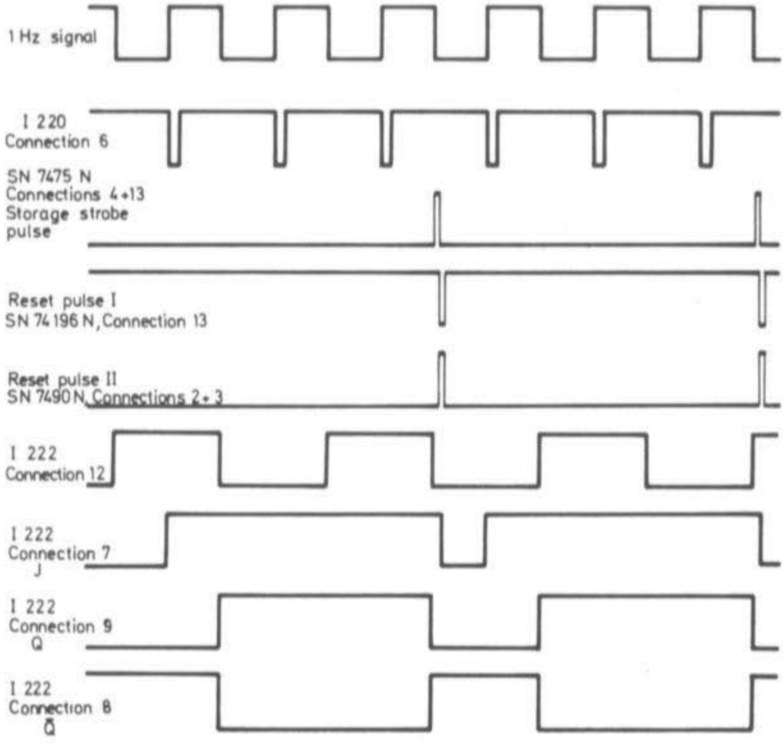


Fig. 4: Pulse diagram of the control pulses

clock frequency (line 6 in Fig. 4). This means that the flip-flop will be actuated once again; Q will become 0, $\bar{Q}$ will be 1, which means that the gate will be blocked and the counting process is completed. The 0-1 slope from the $\bar{Q}$-output of flip-flop I 222 (connection 8) is delayed via two integrating links. This corresponds to approximately 12 µs for the storage-strobe pulse by R 201/ C 201 and by approximately 70 µs for the reset-pulse by R 203/C 203. The subsequent inverter reshapes the pulses distorted due to the delay network. After this, the RC-link C 202/R 202 differentiates the pulse for the storage (line 3 in Fig. 4). The last stage of this chain is transistor T 202 which is used as inverter. This transistor must be able to accept the return current of the 6 storage circuits (approx. 40 mA).

The reset-pulses are differentiated by C 204/R 204, inverted again and amplified. Connection 6 of I 219 provides the reset-pulse I for the first counter I 216 (line 4 in Fig. 4). This pulse simultaneously sets the output (connection 11) of I 220 to 1. J (line 7 in Fig. 4) is set to 0 via connection 8 of I 220. The reset-pulse II (line 5 in Fig. 4) for the counters type SN 7490 N is obtained by inverting the reset-pulse I. The process is recommenced with the next 0-1 slope of the clock signal.

Another special feature is the indication of the period when the gate is open using a light-emitting diode. A higher level will be present at connection 13 of I 219 as long as the counting gate is open. This will mean that transistor T 201 will conduct and the diode will light. This indication allows the operation of the clock circuit to be checked.

2.2. COMPONENTS

V 201 - V 206: 5870 S (ITT), ZM 1330 (Siemens)
Indicator diode: hp 5082-4440 or similar

I 201: SN 7490 N	I 210: SN 7490 N	I 219: SN 74 S 00 N
I 202: SN 7475 N	I 211: SN 7475 N	I 220: SN 7400 N
I 203: SN 74141 N	I 212: SN 74141 N	I 221: SN 7404 N
I 204: SN 7490 N	I 213: SN 7490 N	I 222: SN 7473 N
I 205: SN 7475 N	I 214: SN 7475 N	I 223: SN 7490 N
I 206: SN 74141 N	I 215: SN 74141 N	I 224: SN 7490 N
I 207: SN 7490 N	I 216: SN 74196 N	I 225: SN 7490 N
I 208: SN 7475 N	I 217: SN 7475 N	I 226: SN 7490 N
I 209: SN 74141 N	I 218: SN 74141 N	I 227: SN 7490 N

T 201: BC 108 B or similar
T 202: 2 N 2219 A, 2 N 1613 or similar

C 201: 0.1 μF $\pm$10%/100 V, plastic-foil capacitor, spacing 10 mm
C 202: 0.1 μF $\pm$10%/100 V, plastic-foil capacitor, axial connection max. 25 long
C 203: 0.68 μF $\pm$10%/100 V, plastic-foil capacitor, spacing 15 mm
C 204: 0.1 μF $\pm$10%/100 V, plastic-foil capacitor, axial connections max. 25 long
C 205: 0.1 μF $\pm$10%/100 V, plastic-foil capacitor, spacing 10 mm
C 206: 22 nF ceramic capacitor, spacing 5 mm
C 207, C 208: 1 μF/10 V tantalium capacitor, drop type

R 201: 220 Ω	R 205: 270 Ω	R 209: 4.7 kΩ
R 202: 270 Ω	R 206: 1 kΩ	R 210: approx. 100 Ω
R 203: 220 Ω	R 207: 1.5 kΩ	R 211: 470 kΩ
R 204: 270 Ω	R 208: 2.2 kΩ	

R 212 - R 217: $R = \frac{U_b - 150}{3}$; R in kΩ; U_b in **V** (at Pt 212);

U_b	R
170 V	7.5 kΩ
200 V	18 kΩ
250 V	36 kΩ

All resistors for 10 mm spacing.
R 201 - R 207 = ± 5% tolerance.

2.3. CONSTRUCTION

The frequency counter module is accommodated on the double-coated printed circuit board designated DL 8 TM 002, whose dimensions are 170 mm x 100 mm. Since the printed circuit board possesses through-contacts, no bridges are necessary. This means that the construction according to the component location plan given in Fig. 6 is virtually fool-proof. The dropper resistors for the anodes of the indicator tubes are soldered to the lower side of the board.

The interconnections between the various modules are given in the block diagram given in Figure 1. It is important that the input connection between the prescaler and the counter input (Pt 208) should be short and straight so that the highest frequency pulses are not distorted, which could otherwise reduce the upper frequency limit. This is the reason why the 2 : 1 prescaler is mounted directly behind the preamplifier on PC-board DL 8 TM 003.

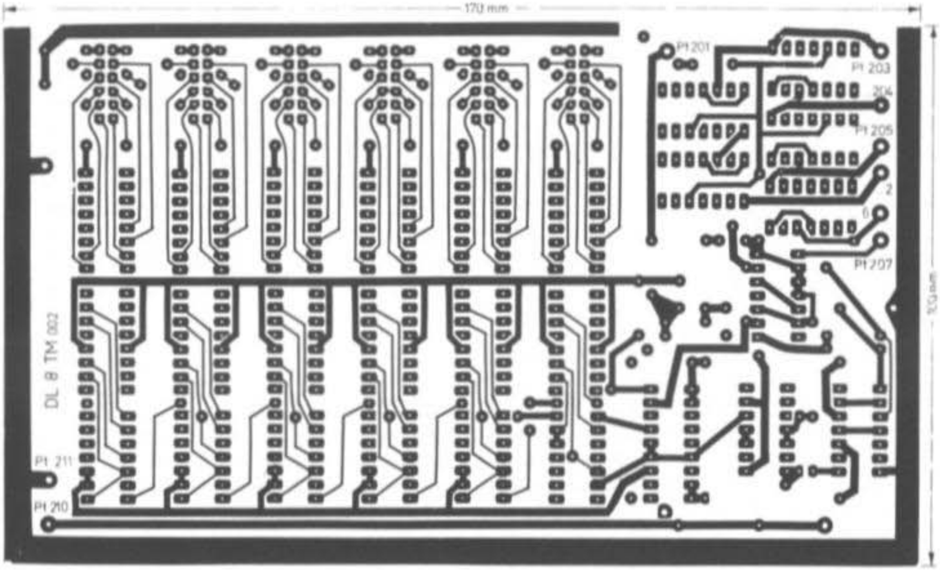


Fig. 5: Printed circuit board DL 8 TM 002

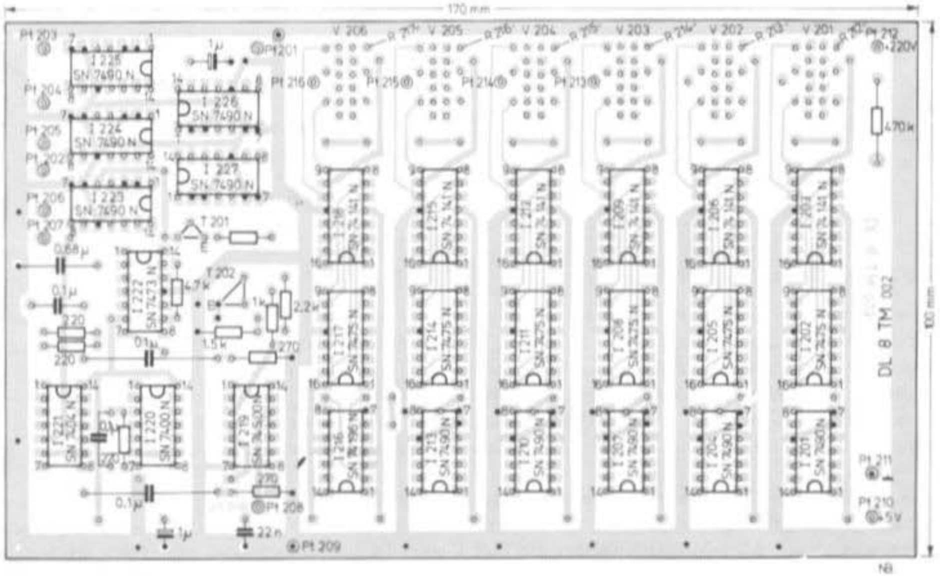
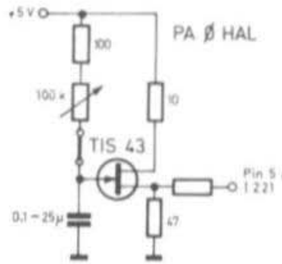


Fig. 6: Component location plan DL 8 TM 002

The connection points Pt 213 - Pt 216 are alternately grounded with a second wafer of the time base switch. The decimal points built into the indicator tubes will then light so that the indication is always given in kHz. This means that the frequency 98.7654 MHz will be indicated as 98765.4 kHz. On switching the clock frequency, an overflow will occur for the first positions so that finally the previously mentioned frequency could be indicated as 765.432 (kHz).

PA Ø HAL has brought a modification of this frequency counter to our notice which was also present in the author's first counter that was not published:

It is possible to vary the period of indication by connecting a simple unijunction transistor oscillator to pin 5 of I 221 (disconnect the connection to Pt 206). If an on/off switch is provided, this will allow the indication to be even "frozen". The editors do not think that there are many applications for such a modification but give the circuit for those that may be interested:



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(4) F. Weingärtner: Further Development of the Four-Digit Frequency Counter
VHF COMMUNICATIONS 4 (1972), Edition 4, Pages 229-234

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CIRCULAR POLARIZATION ON 2 METRES

by T. Bittan, G 3 JVQ/DJ Ø BQ

A theoretical consideration of circular polarization as well as a practical example of a helical antenna for 1296 MHz were described in (1). This article is to discuss the experience gained with circular polarization over a considerable period as well as some surprising results. It also includes construction details for a double-five crossed Yagi.

1. ADVANTAGES AND LIMITATIONS

Before commencing experiments with circular polarization, the author studied the polarity of signals received at his location, on a hill surrounded from North over East to South by mountains (approx. 200 metres higher) and from North to South West by forest (approx. 10-20 metres higher than the antenna). The observations made during this period indicated that extremely few of the received signals were received with their original polarization, only those that were truely line-of-sight and those signals not under going any considerable diffraction (less than 1^{o}). The majority of signals possessed diagonal polarity, e.g. an intermediate value between the vertical and horizontal plane. However, other signals exhibited a slow circular polarization which is undoubtably caused by non-constant diffraction characteristics, since it has been found that reflected signals virtually maintain the same polarization shift as long as the reflecting surface is fixed.

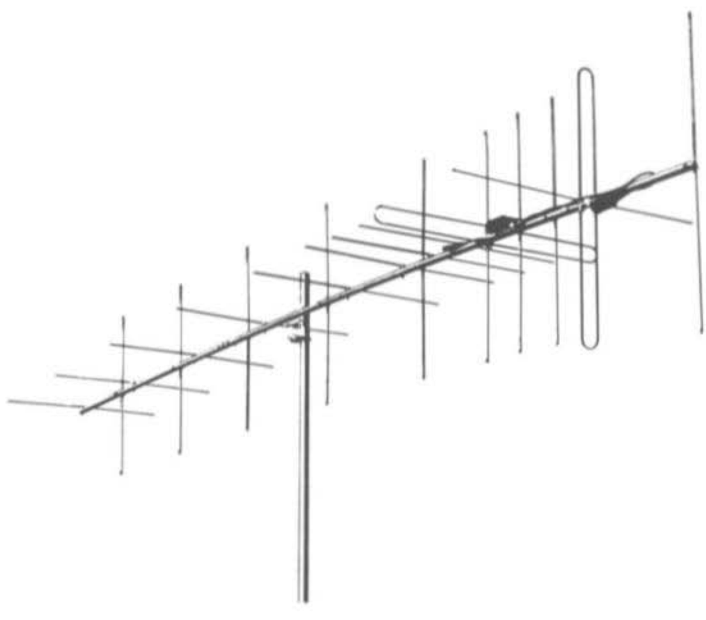


Fig. 1: Photograph of a double-ten crossed Yagi

It is known that a circularly polarized antenna exhibits a loss of 3 dB over a comparable linearly polarized antenna when a linearly polarized signal of the correct polarization is received. However, this is a theoretical value since it has already been stated that truely vertical or horizontal polarization is only received at the authors location with line-of-sight and virtually non-diffracted signals. Since such signals are usually of sufficient strength, the loss of 3 dB can be accepted.

On the other hand, this loss of 3 dB is valid with circular polarization for all vectors. Let us assume the worst case of say a vertically polarized signal which has been phase shifted on the transmission path by 90° so that it is now horizontally polarized. Normally, a radio amateur will attempt to receive this signal on a vertically polarized antenna. In theory, the signal should completely disappear; however, this is not the case in practice where incorrect polarity usually causes a loss of 15 to 20 dB. This means that the originally vertically polarized signal will be received 15 dB - 20 dB down due to the polarization shift in comparison to the maximum of 3 dB down when using a similar circularly polarized antenna. Thus under worst-case conditions, circular polarization can bring a gain of 12 dB to 17 dB over linear polarization.

With the other, non-constant polarization shifts causing slow, circular polarization, a regular fading of the received signal will occur; the maximum signal will be present when the received signal is in phase with the receiving antenna and will be down by at least 15 dB when the signal is 90° out-of phase. This fading disappears completely when using circular polarization since the antenna provides the same RF energy at any polarization angle.

After an antenna was constructed for circular polarization, the following advantages were found in addition to the theoretical considerations given above: It was found that circular polarization produced a far more homgeneous coverage than linear polarization and possessed the property of being able to penetrate into distant valleys and completely screened areas where no communication is possible with linear polarization. This is no doubt due to the fact that the multiple reflections that are present favour circular polarization. Another very considerable advantage is for communication with mobile stations. A mobile station using vertical polarization is virtually only received by reflections of one type or another, which means that the signal is undergoing a continuous polarization shift due to its movement. On a large number of occasions, the author has been able to accompany mobile stations over far greater distances than would even be possible over the local repeaters. The majority of the flutter-type fading disappears completely with circular polarization since this is mainly caused by polarization shifts. The remaining fading is only due to obstructions in the signal path. This means that circular polarization should be used to great effect by repeater stations; this would also offer another advantage since additional isolation could be obtained by using, for instance, clockwise circular polarization for reception and anticlockwise for transmit. This would offer an additional isolation between the input and output antennas of up to 30 dB.

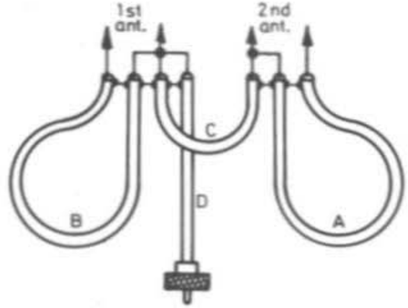
2. PRACTICAL ANTENNAS

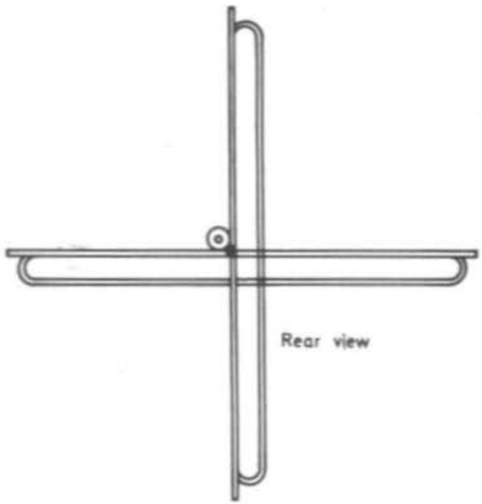
2.1. A CROSSED YAGI FOR TWO METRES

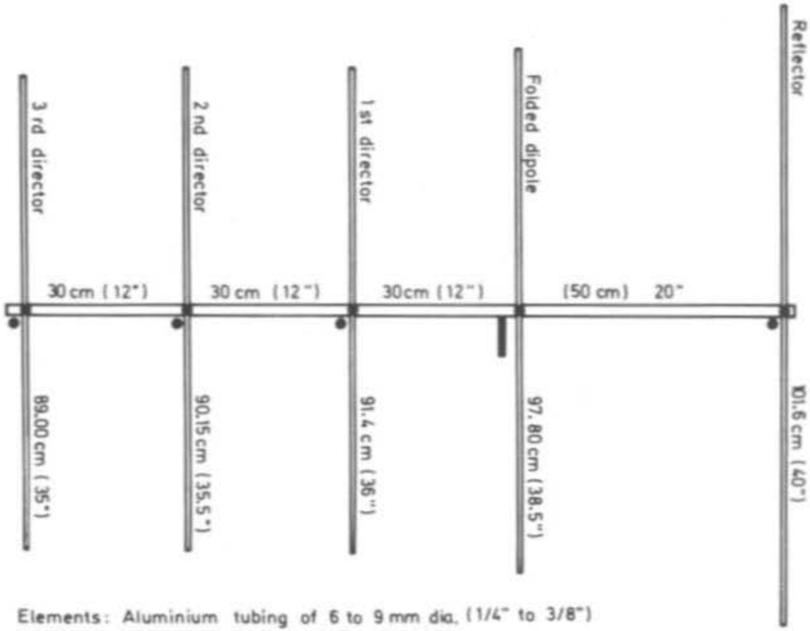
There are two main methods of feeding a crossed dipole or Yagi to form circular polarization. The most common method is for two independent Yagis to

Coaxial cable lengths (solid dielectric):

A and B :	68.3 cm (26.9") 60 to 75 Ohm
C :	34.2 cm (13.45") 75 Ohm
D :	34.2 cm 52 Ohm (for 75 Ohm feeder)
	34.2 cm 47 Ohm (for 60 Ohm feeder)







Elements: Aluminium tubing of 6 to 9 mm dia. (1/4" to 3/8")

Fig. 4: Construction details of the G 3 JVQ-Twister

be mounted on a common boom and connected together with a $1/4\ \lambda$ phasing line. Such a configuration, as shown in Figure 4, can be obtained using two identical Yagi antennas and mounting the elements on one boom in a cross-configuration so that the elements form an angle of 90^{o} to another. The elements should be mounted directly adjacent to another along the boom. Fig. 4 gives the dimensions for such a crossed Yagi for two metres, which the author has named the "G 3 JVQ-Twister".

2.1.1. PHASING LINE

Each of the Yagi antennas should be matched to coaxial cable with the aid of a balun transformer. The normal method of obtaining the 90^{o} polarity shift is by use of a $1/4\ \lambda$ phasing line of coaxial cable. The length 1 of this line can be calculated using the formula:

$$1 = \lambda/4 \times V$$

Where V is the velocity factor of the cable in question. The velocity factor is approximately 0.66 for coaxial cables with solid dielectric and roughly 0.85 for semi-airspaced types (including foam dielectrics). For two metres, this results in lengths 1 of:

$$1 = 51.75 \text{ cm} \times 0.66 = 34.15 \text{ cm } (13.45'') \quad (\text{ solid dielectrics })$$

or

$$1 = 51.75 \text{ cm} \times 0.85 = 43.98 \text{ cm } (17.32'') \quad (\text{ semi airspaced })$$

This phasing line is used to interconnect the two dipoles. The direction of the circular polarization (clockwise or anticlockwise) depends on which dipole is directly energized by the feeder, and which is energized by the phasing line. This means that it is possible to switch from clockwise to anticlockwise circular polarization and vice versa with the aid of a coaxial relay at the antenna. Full details regarding the construction of the $\lambda/4$ phasing line and the baluns are given in Figure 4.

Since both antennas are connected in parallel by the phasing line, the impedance presented to the feeder is half the feedpoint impedance of each antenna. If the feedpoint impedance of each feeder is assumed to be 75 Ω, the value presented to the feeder will be approximately 37 Ω. This low impedance can be transformed back to a value suitable for the feeder in a quarterwave transformer. The required impedance of the $\lambda/4$ line can be obtained by the formula:

$$Z_l = \sqrt{Z_a \times Z_f}$$

where Z_l is the impedance of the $\lambda/4$ line; Z_a the feedpoint impedance to the antenna; and Z_f the characteristic impedance of the feeder. This means that a reasonable match to a 60 Ω to 75 Ω cable will be provided when a $\lambda/4$ transformer made from 48 Ω to 52 Ω coaxial cable is used. The dimensions of the $\lambda/4$ transformer are given in Figure 4.

2.2. ALTERNATIVE METHOD

A more versatile method of obtaining circular polarization is the principle used with the famous MOONBOUNCER manufactured by the British company of J-Beam. A photograph of this antenna is given in Figure 1. With this antenna, the phase shift of 90^{o} is not obtained with a phase line as was described in Section 2.1.1. but by spacing the dipole and other elements mechanically $\lambda/4$

from another. This means that circular polarization is obtained when both feeders are of equal length. If only clockwise circular polarization is required, the two feeders of equal length can be connected together and matched to the common feeder with a $\lambda/4$ transformer as explained in 2.1.1.

However, such a crossed Yagi offers a considerable versatility when both feeders (of equal length) are fed down to the station. In this case, the circular polarization direction can be switched, as shown in Figure 2, by adding a phase line of $\lambda/2$ in order to obtain anticlockwise circular polarization (2).

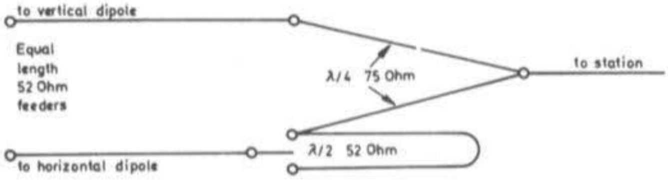


Fig. 2:
Circuit of a switching unit for selecting either clockwise or anticlockwise circular polarization

The switching unit at the station can be extended still further (see Fig. 3) so that the following polarization modes are possible (2):

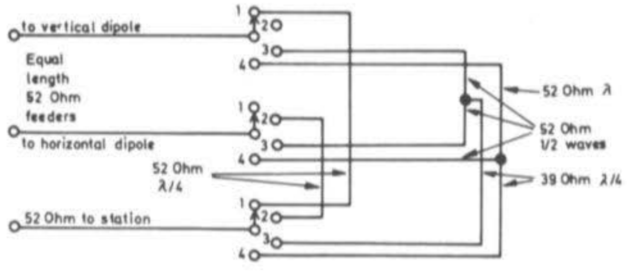


Fig. 3: A more extensive switching unit for four polarization modes

Pos. 1: Vertical, linear polarization
Pos. 2: Horizontal, linear polarization
Pos. 3: Clockwise, circular polarization
Pos. 4: Anticlockwise, circular polarization

2.3. CONSTRUCTION DETAILS

The following section is to describe a double, five element crossed Yagi for 2 metres together with balun phasing line and matching transformer details. The given values are valid for element diameters of 6 to 9 mm (1/4" to 3/8"). All dimensions of the antenna and matching/phasing lines are given in Figure 4. Since this antenna is designed for use as a circularly polarized antenna, it is advisable to mount the antenna so that each of the elements are diagonal and not parallel to the mast. This ensures the minimum influence of the mast on the polar diagram which is an important point when the antenna is not to be mounted at the top of the mast, or when several such antennas are to be stacked.

The elements are mounted onto the boom with suitable mounting screws, and the boom to mast mounting can be made with a TV-antenna clamp.

2.4. PERFORMANCE

Since the boom length of the antenna (140 cm) is just under $3/4\ \lambda$, the gain of the antenna was estimated to be between 8 and 8.5 dB. This value has been proved in practice.

The horizontal and vertical beamwidths (-3 dB) are approximately 60°.

3. EXTENSIONS

Of course, the antenna can be extended to form a longer crossed Yagi. The author, for instance, is using a double 10 element crossed Yagi constructed according to the same principle as given in 2.3.

The double five can also be stacked and bayed to form larger arrays for EME and satellite communications. Since a large number of OSCAR satellites are planned, it would most certainly be worthwhile constructing such an antenna.

3.1. 70 cm VERSION

There are some slight difficulties with a 70 cm version of the described antenna: If a strong boom is used, this will usually represent quite a fraction of a wavelength at 70 cm. This means that it can no longer be neglected since the two centres of the vertical and horizontal elements would be spaced too far from another. It is necessary for either a smaller diameter boom to be used or for the elements to be centered to the boom. Another possibility would be for a glass-fibre or PVC tubing to be used for the boom. The author is willing to provide information regarding the design of a such antenna for 70 cm if there is sufficient interest. Please inform your local representative of VHF COMMUNICATIONS of your interest.

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In this edition of VHF COMMUNICATIONS

THEORY, ADVANTAGES AND TYPES OF ANTENNAS FOR CIRCULAR POLARIZATION AT UHF

by Dr. Ing. A. Hock, DC 0 MT

It is unfortunate that circular polarization does not enjoy the popularity it deserves. The main reason why occasional experiments have not been very successful and why it has not gained in popularity is due to the lack of stations that are also equipped for this polarization mode. It is true, that the full advantages for circular polarization can only be utilized when both stations are suitably equipped for this polarization. The advantages of circular polarization have been well known in professional communication fields for some time. The experience gained on 2 metres with circular polarization, as well as a practical example on how circular polarization can be realized at VHF were described in (1). A practical realization was also described in (2).

This article is to discuss the theory and advantages of circular polarization together with a practical example on how circular polarization can be achieved on 70 cm.

Circular polarization offers considerable advantages in the UHF and SHF range and it is hoped that this article will forward this type of polarization.

1. CIRCULAR POLARIZED WAVES

The most common forms of antenna are designed for radiation or reception of linearly polarized waves. The given polarization refers to the position in space of the electrical field lines. These are constant with linear polarization, such as the horizontal or vertical modes. Linear polarization is indicated with the aid of the dipole radiator given in Figure 1. The vertical dipole elements generate electrical field lines that are vertical in space and maintain this position as long as no polarization shift is caused by an obstruction in the transmission path. The direction of the field lines alters by 180° for each half wave length, thus altering polarity.

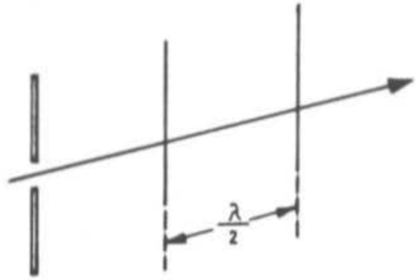

Fig. 1:
Linear polarization (vertical)

With circular polarization, a vector of the electrical field lines occurs in space so that the propagation direction continuously rotates around an axis, which is the direction of propagation. Figure 2 shows a perspective diagram. The vector is continuously changing its direction. The vector will have made a complete rotation and obtained the original position after one wavelength in space. This means that the electrical field lines will have changed direction by 180° in a spacing of one half wave length.

Such a vector can be assumed to be generated by the superimposition of a sine and cosine oscillation as shown in Figure 3. This indicates the first type of antenna that can be used: Two crossed dipoles which are fed with voltages phase-shifted by 90°. Each dipole radiates a signal into space where they are superimposed to form a circular-polarized wave by generation of a rotating electrical vector in the direction of propagation.

Fig. 2:
A circular polarized wave

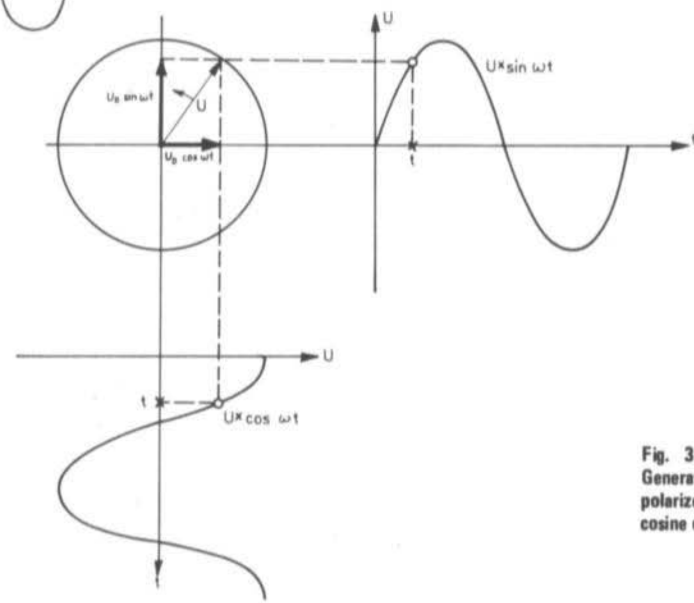

Fig. 3:
Generation of a circularly polarized wave from a sine and cosine oscillation

2. ADVANTAGES OF CIRCULAR POLARIZATION

In order to compare linear and circular polarization, it is assumed that the circular polarization is completely circular: This means that the length of the vector remains constant during the complete cycle, in other words, that the basic sine and cosine wave are of equal amplitude.

2.1. TRANSMISSION PATH WITHOUT OBSTRUCTIONS

When used for reception, a circularly polarized receive antenna will produce a RF power from any vectoral direction. The maximum power will be produced when a truly circularly polarized wave is received. If the polarization of the received signal is elliptical, this will mean that the RF voltage regained in the antenna will be reduced and will obtain a minimum with purely linear polarized waves. In this case, the antenna will produce 3 dB less than when receiving a truly circularly polarized wave; this is independent of the polarization plane. The main practical advantage of a circularly polarized antenna for reception is that any polarization can be received equally well without switching and that only 3 dB (0.5 S-points) is lost under the most unfavourable condition, e.g. when receiving a linearly polarized wave.

2.2. PROPAGATION PATH WITH OBSTRUCTIONS

Completely unobstructed transmission paths are practically never present on the earth surface and atmosphere. Additional attenuation and polarization shifts occur.

An attenuation is often made by obstructions in the propagation path (Fresnel zone) which mainly extends in the direction of the electrical field lines. A typical example of this is the considerable attenuation of vertically polarized waves by trees. This is where the use of circular polarization would cause a considerable increase of the field strength, since practically only the vectors are suppressed in which direction the obstruction stands; all other components will pass through the obstruction without considerable attenuation. Measurements on a model test field have shown that a vertically polarized wave with a wavelength of 10 cm was attenuated by approximately 40 dB on passing through a model forest of damp matches (length approx. $\lambda/2$), whereas a circularly polarized wave was only attenuated by approximately 3 dB.

With all UHF and SHF communications that are not made under true line of sight conditions, it is to be assumed that certain reflections take place. In some circumstances, the reflected waves can form the main part of the received field strength. Polarization shifts occur to a higher or lower degree with every reflection. It has been proved that an originally horizontally polarized vector which has been shifted by 45° will provide a signal into the horizontally polarized receive antenna that is approximately 3 dB less. Greater polarization shifts will reduce the received energy still further until the signal will theoretically completely disappear at a shift of 90° (e.g. due to reflections on mountains). However, slight variations from the 90° ensure that the signal is not completely cancelled out. In practice, a 90° polarization shift will cause an attenuation in the order of 15 to 20 dB. Circularly polarized waves, will, of course, also be shifted by the reflection, however, the receive antenna will produce the same RF energy.

3. ANTENNAS FOR CIRCULAR POLARIZATION

In practice, only a few types of antennas can be used for circular polarization: The first of these is a crossed dipole (or Yagi) that is especially suitable for the 70 cm band, and the helical antenna for higher frequencies.

3.1. CROSSED DIPOLE

These dipoles are dimensioned in the same manner as for a single dipole. Both simple and folded dipoles can be used and no modifications to the dimensions are required. The phase shift of 90° between the dipoles is made using a $\lambda/4 \times V$ phase line between the dipoles (V = velocity factor = $1/\sqrt{\varepsilon}$). Both coaxial or balanced feeders can be used. This means that one antenna is directly connected to the feeder and the other via a quarter wave phase line.

The usual methods of matching the antenna impedance to the feed line can be used as for individual dipoles, e.g. Balun-transformers. Since two antennas are to be connected in parallel with the aid of the phase line, the common impedance of this arrangement has half the impedance of a single antenna. This impedance can be increased by placing the first director near to the dipole radiator. Of course, these crossed dipoles can be extended to form crossed Yagis or antenna arrays constructed from linearly polarized antennas.

3.2. HELICAL ANTENNA

Helical antennas have found great popularity in professional and satellite communications. This antenna has a very wide bandwidth which means that the dimensions are not critical. The dimensions of such an antenna for the 24 cm band are very small and allow several helical antennas to be combined on a common reflector to form an efficient array.

The following rules must be observed during the design:
The greater the number of turns, the greater will be the gain of the antenna; however, no great increase of the gain is achieved when more than 12 turns are used.

Wire diameter d: $0.006 \leq d/\lambda \leq 0.05$

Circumference $C \approx \pi \times D$ of one turn: $3/4 < C/\lambda < 4/3$

Pitch β of the helix: $12^\circ < \beta < 15^\circ$

Figure 4 shows the above mentioned dimensions of the helical antenna.

A reflector must be provided behind the helix in order to obtain an omnidirectional characteristic. The dimensions of this reflector must be at least $\lambda/2$. In this case, the feedpoint impedance Z of the antenna will then amount to:

$$Z \approx \frac{140 \times C}{\lambda} \;;$$

a $\lambda/4$ transformer can be used for matching the antenna to the feeder.

Due to the described physical characteristics of circular polarization, it would be advisable if more amateur stations would experiment with this propagation mode at UHF. However, it is extremely necessary that a common direction of rotation is used. The author suggests that clockwise circular polarization should be used, when seen from the direction of the transmit to receive antenna.

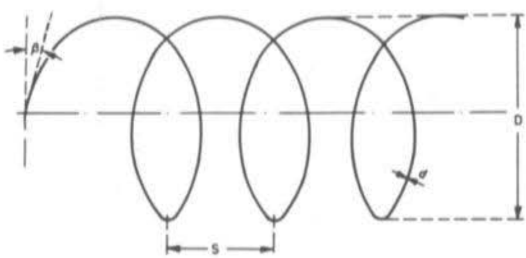

Fig. 4: Dimensions of a helical antenna

4. PRACTICAL EXAMPLE

The design of a helical antenna for 23 cm is to be given as an example:

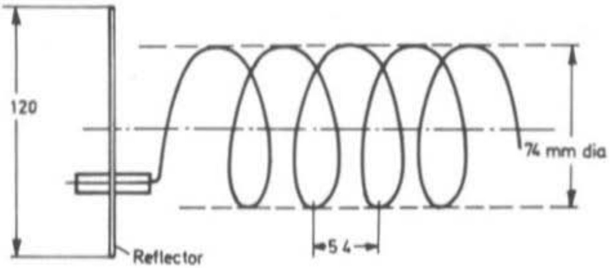

Fig. 5: Constructional drawing of a helical antenna for 23 cm

1. Wire diameter d: selected is $d/\lambda = 0.01$
   $d = 0.01 \times \lambda = 0.01 \times 23.2 = 0.23$ cm $\approx$ 2 mm diameter (12 AWG)
   where $\lambda = 23.2$ cm at $f = 1296$ MHz

2. Circumference C: Selected is $C/\lambda = 1$
   $D = \lambda$; $C = 23.2$ cm ($D \approx \frac{23.2}{\pi} \approx 7.4$)

3. Pitch ß: Selected is $\beta = 13^{o}$
   $\tan \beta = 0.2309$
   $S = C \tan \beta = 23.2 \times 0.231 = 5.36$

4. Feedpoint impedance Z:
   $$Z = \frac{140 \times C}{\lambda} = \frac{140 \times 23.2}{23.2} = 140\ \Omega$$

5. Reflector diameter D_{refl}:
   $D_{refl.} > 11.6$ cm

6. Number of turns n:
   $n > 3$; the greater the number of turns, the greater will be the gain of the antenna.

7. Matching of the antenna to 60 Ω coaxial cable with the aid of a $\lambda/4$ transformer:
   $$Z_{tr} = \sqrt{60 \times Z} = \sqrt{60 \times 140} = 91.6\ \Omega$$

The wire of the helical antenna can be used as inner conductor of the $\lambda/4$ coaxial transformer. The external conductor is then:

$$Z_{tr} = 60 \times \ln \frac{D}{d} \ ; \ \frac{D}{d} = e^{91.6/60} = 4.25$$

$$D = 4.25 \times d = 8.5 \text{ mm}$$

Length of the coaxial transformer (air as insulator): $l = \lambda/4 = 5.75$ cm;

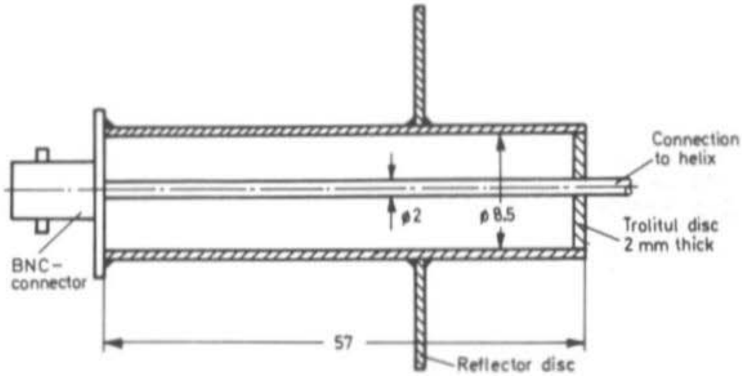

Fig. 6: Quarter wave transformer for matching to 60 Ω

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TEMPERATURE-COMPENSATED OSCILLATOR WITH VARACTOR TUNING

by T. Schad, DJ 8 ES

Oscillators in transmitters and receivers that are equipped with variable capacitors or variable inductance tuning, possess the disadvantage that extensive tuning devices such as gears must be accommodated in the vicinity of the oscillator. It is difficult to obtain a rapid frequency change between predetermined fixed frequencies which means that two oscillators must often be provided in transmitters and receivers. This is especially valid for the VHF range where simplex operation on the same frequency is not all that prevalent for operation via repeators where the transmit and receive frequency do not coincide. These disadvantages are avoided when oscillators are used whose frequency is determined by the capacitance of a diode biased into its blocking range. The disadvantage of this is that a potentiometer with a good resolution is required for tuning, and, furthermore, a very constant DC voltage. The latter can be obtained with the aid of an integrated voltage stabilizer. When considering circuits for temperature compensation, it is also necessary for the temperature response of this stabilized voltage to be taken into consideration

1. PRELIMINARY CONSIDERATIONS

The temperature response TC_c of the capacitance of a biased silicon diode is positive which means that it is necessary for the tuning voltage U_{tune} (= inverse voltage of the diode) to be increased on increasing temperature in order to compensate the temperature effects. According to the type of diode, and also somewhat dependent on the temperature and the connected inverse voltage, the required voltage variation amounts to 2 to 5 mV/°C (Fig. 1).

Figure 2 gives the forward voltage of a silicon diode as a function of the forward current at two different temperatures. It will be seen that the voltage is reduced on increasing temperature, and more so the lower the forward current. This fact is used in the described compensation circuits.

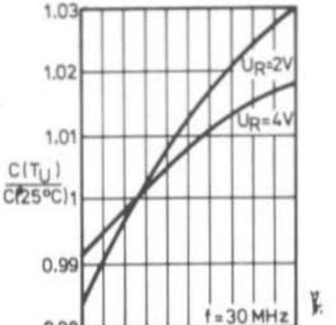


Fig. 1: Temperature dependence of the junction capacitance of a varactor diode BA 110

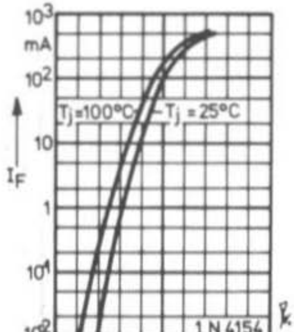


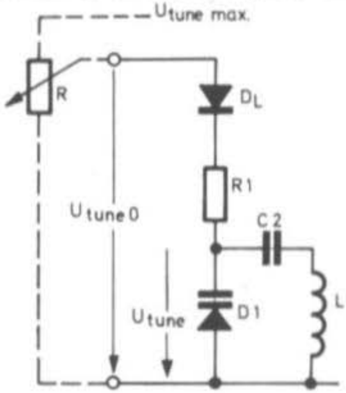
Fig. 2: Forward current characteristics of a silicon planar diode at a junction temperature of 25°C and 100°C

2. POSSIBILITIES OF TEMPERATURE COMPENSATION

2.1. SIMPLE COMPENSATION CIRCUITS

A simple temperature compensation circuit is given in Figure 3 that has resulted from the considerations of the previous section 1. Resistor R 1 is used to separate the RF current, capacitor C 2 for blocking the direct current. In this circuit, the current flowing via diode D 2 is equal to the inverse current flowing via the tuning diode D 1 ($\approx 10^{-4}$ mA), which does not vary greatly in the considered range of the inverse voltage. The tuning voltage U_{tune} must be greater than the peak value of the RF voltage across D 1 so that no distortion occurs. In addition to this, U_{tune} must always be less than the breakdown voltage of the diode minus the peak value of the RF voltage. By extrapolation from Figure 2, it will be seen that the frequency response of the tuning voltage of the circuit given in Figure 3 is approximately 2.7 mV/^{o}C at I = 10^{-4} mA. This voltage variation will only be correct in a very few cases. However, this circuit can be used, for instance, for portable FM equipment due to its low current requirements and sufficient stability. The tuning voltage could then be taken directly from the operating voltage as long as no great variations of load occur. In other cases, e.g. for AM transmissions, it will be necessary for a separate battery to be provided for the tuning voltage.

One possibility of decreasing the voltage variation is to increase the current flowing via diode D 2 as is indicated in Figure 4. The current flowing via diode D 2 is adjusted here using resistor R 2. One disadvantage of this is that the current varies together with the tuning voltage. This disadvantage is not exhibited by the circuit given in Figure 5.



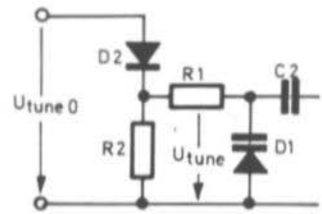


Fig. 4: Compensation with a greater forward current via D 2

Fig. 3: Simple temperature compensation of a varactor diode; Potentiometer R for tuning; voltage drop across R 1 is negligible.

2.2. VARIABLE COMPENSATION WITH A CONSTANT-CURRENT SOURCE

In Figure 5, transistors T 1 and T 2 represent a constant-current source in conjunction with resistor R. This circuit will be temperature compensated as long as both transistors are equal and possess the same temperature. The collector current of T 1 and the base current of both transistors flow via R. Since the characteristics of both transistors and their temperature must be equal, the base currents must also be equal. This means that the following is valid:

$$I = I_C + 2\,I_B \qquad \text{and thus:} \quad I_C/I = \frac{B}{B+2} \approx 1$$

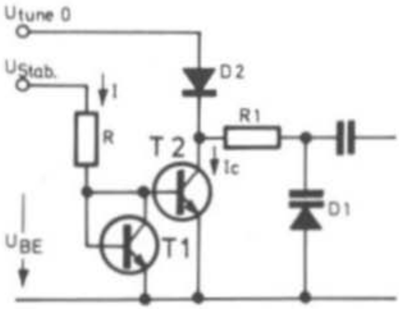


Fig. 5: Adjustable compensation with a constant-current source.

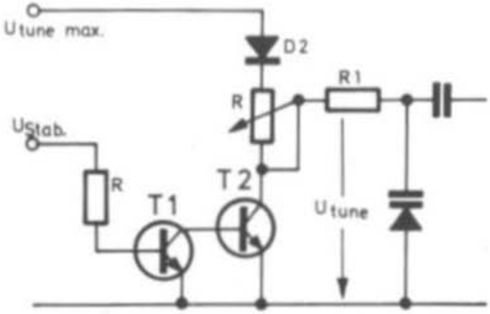


Fig. 6: Tuning circuit for compensation with a constant-current source.

Whereby B is the DC gain of the transistors in a common-emitter circuit with $B \gg 1$.

This means that $I_c \approx I = \dfrac{U_{stab.} - U_{BE}}{R}$

The current flowing via diode D 2 can be varied with resistor R independent of the tuning voltage and independent of the temperature since the variation of U_{BE} is negligible with respect to $U_{stab.} - U_{BE}$. The stabilized voltage $U_{stab.}$ can also be used as operating voltage for the oscillator. The tuning arrangement is shown by dashed lines in Figure 3.

Due to the impressed current, a tuning circuit as shown in Figure 6 will also be possible. In this case, a voltage drop will be generated across the potentiometer R. This will be:

$$U_{tune\ max.} - I_C \times R = U_{tune}$$

By forming the resistance R from fixed resistors and a potentiometer, it is possible for a certain dependence of the frequency on the adjustment angle of the potentiometer to be obtained. A disadvantage is that R must be selected according to the impressed current.

If several diodes are used instead of D 2, it is possible for even greater (more positive) temperature coefficients to be compensated.

2.3. COMPENSATION OF A NEGATIVE TEMPERATURE COEFFICIENT

Figure 7 indicates the possibility of compensating a negative temperature coefficient. The operation is similar to that previously described. A constant current will flow via the potentiometer since the inverse current via diode D 2 will be adjusted with the total resistance R of the potentiometer.

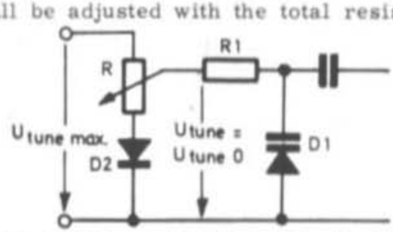


Fig. 7: Compensation of a negative temperature coefficient of the circuit capacitance.

However, attention should always be paid that the resonant circuit of the oscillator is compensated as well as possible since a TC_c-free capacitance is required for D 1. This is also valid for the previously mentioned circuits. This means that the possibilities given in Figure 5 or 6 should always be considered.

3. CONSTRUCTION OF AN OSCILLATOR FOR 23 to 25 MHz

An oscillator for the frequency range of 23 to 25 MHz was constructed according to the considerations described in Sections 1 and 2. The circuit of this oscillator is given in Figure 8. Figure 9 illustrates the printed circuit board DJ 8 ES 001 with component locations which is suitable for installation in a TEKO-box, size 2 A.

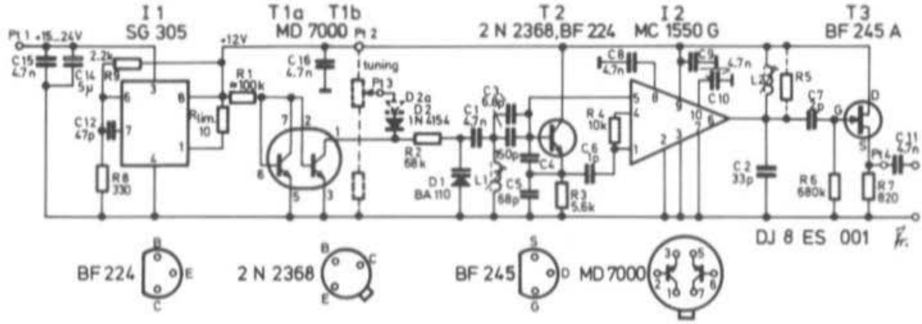


Fig. 8: Circuit diagram of a diode-tuned VFO with integrated voltage stabilizer and temperature compensation

Transistors T 1a and T 1b form the compensated, constant-current source. A dual-transistor type MD 7000 (Motorola) is used. Of course, any other dual-transistor can be used. Transistor T 2 is the oscillator transistor. The integrated circuit MC 1550 G amplifies the oscillator signal and also isolates the oscillator from the load. This integrated circuit is very inexpensive. However, it can be replaced by any RF transistor and sufficient space is provided for this on the PC-board.

Transistor T 3 represents an impedance converter. Although the output impedance of a bipolar transistor in a collector circuit would have been less due to the higher slope, a field effect transistor has been preferred due to the better linearity and the simpler DC-circuitry. The transistor BF 245 A can be used for T 3 when a output voltage of approximately 100 mW is sufficient since the low drain current does not load the voltage stabilizer. This has a favourable effect on its temperature response.

Resistor R 1 is used for adjustment of the temperature coefficient as was described in Section 2.2. The temperature compensation is dimensioned so that the TC of the voltage stabilizer is compensated for.

Resistor R 5 is used to dampen the resonant circuit of the oscillator L 2/C 2 if this should be necessary in order to obtain a constant output voltage amplitude over the whole tuning range. An integrated circuit SG 305 (LM 305) is used for generation of the very stable voltage. With a little care, it is possible for the less expensive type ICB 8723 C (manufactured by Intersil in metal case) to be used, whose characteristics are only slightly inferior. On installing the oscillator, the current-limiting resistor R_{lim} ($\approx 10\,\Omega$) should be inserted. After testing and finding that the operation is satisfactory, R_{lim} can be removed and replaced by a bridge across connections 1 and 8 of the SG 305. This will improve the temperature response of the tuning voltage considerably. If no accurate measurements can be made, R 1 can be assumed to be 1 kΩ. The author uses a 10-turn helical potentiometer for tuning. If a 20-turn potentiometer is used, the tuning will be sufficiently fine that no gearing is required.

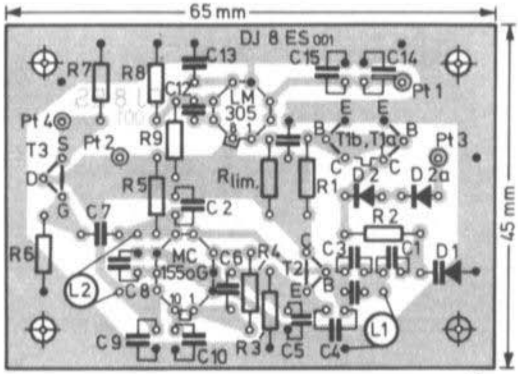


Fig. 9: Printed circuit board DJ 8 ES 001 and component plan for the circuit given in Fig. 8

3.1. SPECIAL COMPONENTS

I 1: SG 305 (Silicon General), LM 305 (National Semiconductor)
I 2: MC 1550 G (Motorola)

T 1a, 1b: MD 7000 (Motorola)
T 2: BF 224 (TI), BF 173 or similar
T 3: BF 245 A (TI)

D 1: BA 110 (ITT), BA 149 (AEG-Tfk)
D 2: 1 N 4154, 1 N 4148 (1 N 914)

All capacitors are for 2.5 mm spacings (Philips ETBU/0.6 type 1 B)

C 3: 2 x 6.8 pF, N 150 from series SDPN type 1 B
(or similar miniature type)
C 4, C 5: N 750

L 1: 17 turns of 0.15 mm dia. (34 AWG) enamelled copper wire
wound on a 4 mm coilformer with SW core

L 2: 14 turns, otherwise as L 1

Potentiometer: 20 kΩ helical potentiometer, e.g. Amphenol 2151 B
or Megatrom 2510.

4. MEASURED VALUES AND DISADVANTAGES

Table 1 shows the results obtained with such an oscillator that has also proved itself in the SSB mode.

The unfavourable behaviour shortly after switching on is a disadvantage over oscillators with capacitor or inductance tuning due to the various currents flowing through the frequency and voltage determining components. A further great disadvantage is the non-linear dependency of the frequency on the shaft angle of the potentiometer. This disadvantage could be avoided using the tuning circuit given in Figure 6 or by connecting a capacitance in parallel to the diode. The first possibility is very extensive and the second will limit the tuning range.

Table 1: Measured values of the oscillator

U_{tune} (V)	4	6	12
Δf(Hz/°C)	60	50	130

Output voltage: 100 mV into 60 Ω
Impedance: ≈ 80 Ω
Suppression of the 1st harmonic: > 20 dB
Suppression of the 2nd harmonic: > 30 dB

Transient behaviour: In the first 5 minutes $\Delta f < 2000$ Hz, subsequently less than 100 Hz per hour measured at $U_{tune} = 12$ V in an aluminum case at room temperature.

5. MODIFYING THE OSCILLATOR TO OTHER FREQUENCY RANGES

The oscillator can easily be modified for other frequency ranges. This can be achieved by replacing the tuning diode with a TC=0 capacitor having a capacitance equal to that provided by the diode at the mean tuning voltage of approximately 6 V. Series and parallel capacitors are now used in order to obtain the lowest possible temperature response. The diode is now installed and R 1 once again adjusted for the lowest temperature response. Room enough for a second compensating diode D 2a is provided for this purpose.

If higher output voltages than 100 mV are required, it is possible for the buffer stage I 2 to be more tightly coupled to the oscillator and for a BF 245 B or C to be used as impedance converter. These types can be driven at higher values since U_{GS} may be higher. However, this method has the disadvantage that the voltage stabilizer is more highly loaded on increasing the drain current. This increases the dissipation power and thus the heating and will also cause the temperature response to be deteriorated. A bootstrap-circuit as shown in Figure 10 would be better.

This allows the drive range to be increased whilst maintaining a low drain current. Resistor R_S is adjusted for the required drain current and should be of a value as shown in Figure 8. The voltage at the source connection is adjusted with R 1 to approximately half the operating voltage U_b. This stage can then be driven virtually linearly by $\pm 0.5\ U_b$. The higher input impedance obtained in this manner is not important since the circuit L2/C2 is dampened with a far lower resistance.

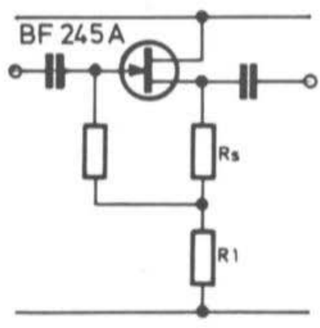


Fig. 10: Bootstrap-circuit for increasing the drive range

With the modern diodes now coming onto the market that are manufactured according to ionimplantation, it is assumed that far better results will be obtainable. Also the heat sinks that are now available for integrated circuits are of great assistance in the generation of highly constant voltages. The author is experimenting in both of these spheres and will report on his results in a later article.

6. AVAILABLE PARTS

See material price list.

7. REFERENCES

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(2) H. J. Franke and H. Kahlert: A Universal Power Supply Using an Integrated DC-Voltage Stabilizer
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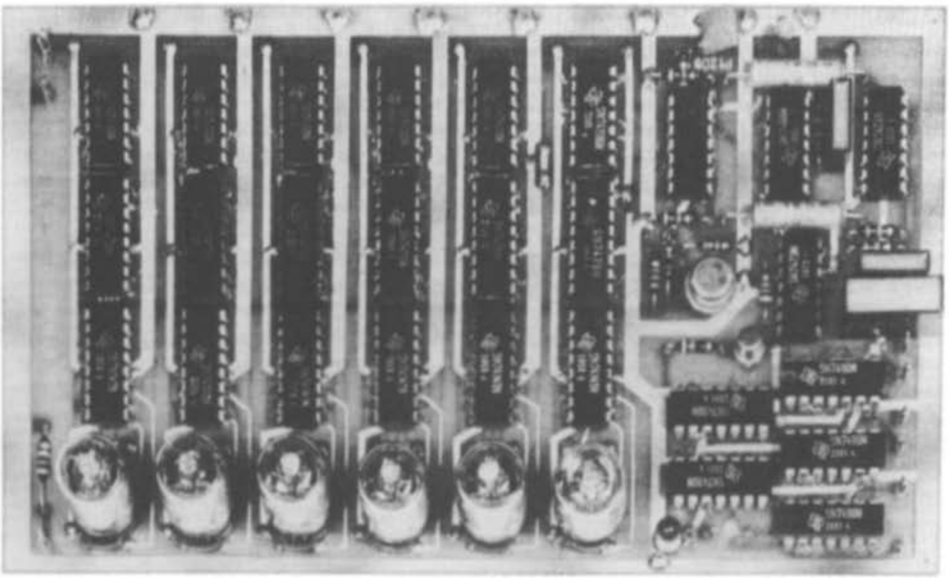

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A MINIATURE AM/CW/FM TRANSMITTER FOR 144 MHz

by B. Dietrich, DJ 8 PG

Miniature, low-power transmitters are required for DF meetings and foxhunts as well as for antenna measurements. Output powers in the order of 20 mW are able to provide a sufficiently strong signal for such purposes. All operating modes used on the 2 m band should be possible; the transmitter ("fox") should be able to provide either a CW, amplitude or frequency-modulated signal. Transmitters used for DF-meetings or foxhunts should be inexpensive, lightweight and so small that they can be hidden easily. Accumulators or dry batteries can be used as power supply. Since the operating period is usually short, it is usually more economical to use the cheapest batteries.

A transmitter is now to be described that operates from a miniature 9 V battery (as used for small transistor radios) and is constructed with miniature components. After being installed into its case, the transmitter will be 80 mm by 55 mm by 30 mm and will weigh 220 g including battery. Further miniaturization by use of sub-miniature components is not economical. The transmitter will then only need a $\lambda/4$ antenna which can be made from suitable wire. Of course, it would also be possible to use the antenna to suspend the transmitter from a tree, etc. Figure 1 shows the author's prototype.

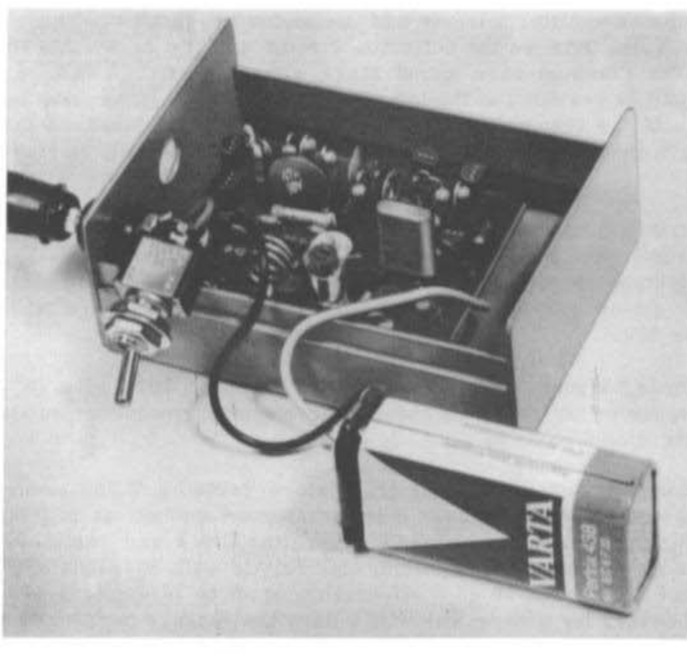

Fig. 1: Miniature transmitter for 2 m

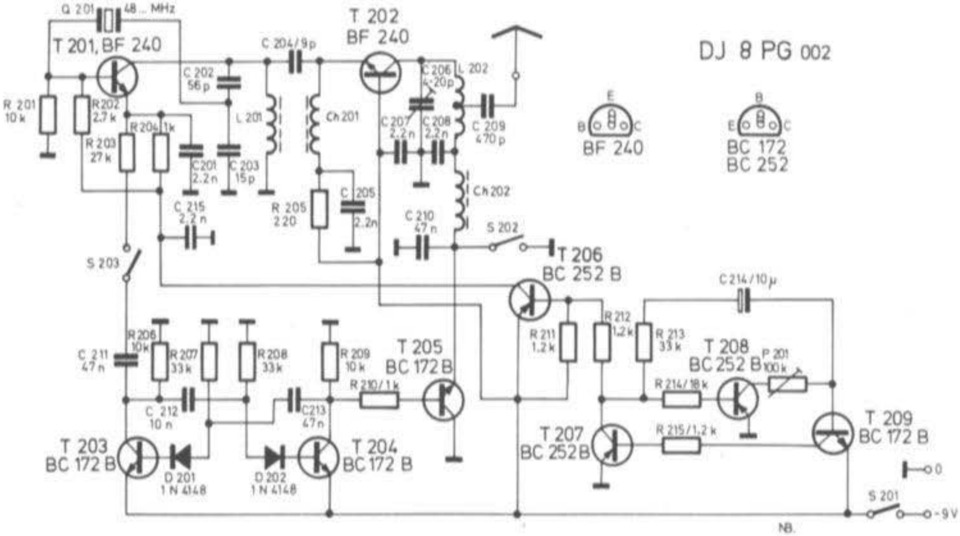

Fig. 2: Circuit diagram of the miniature AM, FM and CW transmitter

1. CIRCUIT DETAILS

The circuit diagram is given in Figure 2. It will be seen that the RF-portion consists of a 2-stage transmitter. The oscillator comprising transistor T 201 is in a common-emitter circuit and oscillates at 48 MHz. The crystal is in the feedback link between the collector circuit and the base. The frequency is tripled in the common-base output stage with transistor T 202. A simple resonant circuit is provided at the output onto which the $\lambda/4$ antenna is capacitively coupled. If the transmitter is to be used for other purposes than as a DF (foxhunt) transmitter it will be advisable for an external, harmonic filter to be used.

The amplitude modulation is made with the aid of transistor T 205 in the collector DC circuit. The modulator transistor is driven by the multivibrator circuit comprising transistors T 203 and T 204. When switch S 202 is closed, the modulator transistor will be short-circuited and the output stage will operate at peak-power level.

The modulating signal can also be fed via switch S 203 and a RC-link to the crystal oscillator in order to obtain narrow-band frequency modulation in a very simple manner.

The switching circuit comprising transistors T 206 to T 209 ensures that the oscillator, and thus the transmitter is switched on and off at regular intervals. The switching period is determined by capacitor C 214 and resistors R 213 and P 201. The given drop-type tantalium electrolytic will be sufficient for periods in the order of 10 s to 20 s. If off-periods of up to 1 minute are required, it will be necessary for a capacitor with a very low leakage current to be selected. The transistors should possess a high current gain and have a low residual current.

2. CONSTRUCTION

The miniature transmitter can be built up on a printed circuit board of 55 mm by 47.5 mm. Figure 3 shows the component locations and conductor lanes of the PC-board which has been designated DJ 8 PG 002. The crystal (HC-25/U) can either be directly soldered onto the PC-board or made plugable using two contact springs from an octal-tube socket, or similar. Even switches S 202 and S 203 are home-made. They are both made from approximately 1 cm pieces of springy steel wire that are soldered into the holes of the PC-board and bent so that they fit into the open eyelet of the other connection. This means that the transmitter can be set up for the required modulation mode before commencement of the DF (foxhunt)-meeting. The resistors should stand vertically on the PC-board.

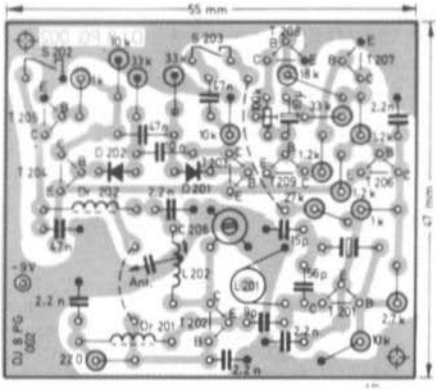

Fig. 3: Printed circuit board DJ 8 PG 002

The completed PC-board is mounted into a small case with the aid of two screws. The author constructed the case from 1 mm brass plate having an internal length of approximately 75 mm. By the way, the battery connection can be made by dismantling an old battery if such connectors are not readily available.

3. COMPONENTS

T 201, T 202: BF 123, BF 125, BF 240 (ITT) or BF 173,
BF 240 (AEG-Tfk, Siemens), BF 224 (TI)
T 203, T 204, T 205, T 209: BC 172 B (ITT) or BC 168 B, BC 183 B,
BC 238 B, BC 108 B or similar.
T 206 - T 208: BC 252 B (ITT) or BC 213 B, BC 178 B or similar PNP.

D 201, D 202: 1 N 914, 1 N 4148 or similar.
L 201: 11 turns of 0.5 mm dia. (24 AWG) silver-plated copper wire close-wound on a 4.3 mm coilformer with SW-core (red).
L 202: 4 turns of 0.8 mm dia. (20 AWG) wound on a 6 mm former, self-supporting.

Ch 201, Ch 202: 20 turns of 0.5 mm dia. (24 AWG) enamelled copper wire wound on a ferrite pin, 10 mm long.
C 206: 4-20 pF ceramic or myhar trimmer of 7 mm dia.
P 201: 100 kΩ trimmer potentiometer, vertical mounting, pin spacing 5/2.5 mm.
All ceramic capacitors: 5 mm spacing.

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 2/1973 of VHF COMMUNICATIONS

DJ 4 LB 001	ATV TRANSMITTER, Module 1		Ed. 1/1973
PC-board	DJ 4 LB 001	(with printed plan)	DM 11.--
Semiconductors	DJ 4 LB 001	(6 transistors, 2 diodes)	DM 16.90
Minikit 1	DJ 4 LB 001	(2 coilsets, 2 ferrite beads, 2 ferrite chokes, 1 TEKO box 4 B)	DM 12.--
Minikit 2	DJ 4 LB 001	(with all other components: 16 cap., 17 resistors, 2 trimmer potentiometers, 5 feedthroughs, 10 solderpins)	DM 24.50
Crystal	38.9000 MHz	(HC-6/U)	DM 25.--
Kit	DJ 4 LB 001	(complete with all components)	DM 84.50
DJ 4 LB 002	ATV TRANSMITTER, Module 2		Ed. 1/1973
PC-board	DJ 4 LB 002	(with printed plan)	DM 11.--
Semiconductors	DJ 4 LB 002	(8 transistors, 4 diodes)	DM 26.30
Minikit 1	DJ 4 LB 002	(1 coilset, 1 ferrite bead, 2 ferrite chokes, 1 TEKO box 4 B)	DM 9.80
Minikit 2	DJ 4 LB 002	(with all other components: 26 capacitors, 26 resistors, 2 trimmer potentiometers, 6 feedthroughs, 10 solderpins)	DM 52.--
Kit	DJ 4 LB 002	(complete with all components)	DM 98.--
DJ 4 LB 003	ATV TRANSMITTER, Module 3		Ed. 2/1973
PC-board	DJ 4 LB 003	(with printed plan)	DM 11.--
Minikit 1	DJ 4 LB 003	(5 transistors, 1 diode, 1 coilformer with core, 4 ferrite beads, 6 ceramic trimmers, 1 TEKO box 4 B)	DM 28.20
Minikit 2	DJ 4 LB 003	(21 capacitors, 11 resistors, wire for coils, 2 feedthroughs, 7 connection pins)	DM 25.20
Crystal	78.858 MHz	(HC-6/U)	DM 26.--
Kit	DJ 4 LB 003	(complete with all components)	DM 89.50
DJ 4 LB 004	ATV TRANSMITTER, Module 4		Ed. 2/1973
PC-board	DJ 4 LB 004	(with printed plan)	DM 11.--
Minikit 1	DJ 4 LB 004	(7 transistors, 5 ferrite beads, 2 ferrite chokes, 7 ceramic trimmers, 1 TEKO box 4B)	DM 49.55
Minikit 2	DJ 4 LB 004	(24 capacitors, 24 resistors, wire for coils, 3 feedthroughs, 10 connection pins)	DM 32.40
Kit	DJ 4 LB 004	(complete with all components)	DM 91.--
DJ 4 LB 005	ATV TRANSMITTER, Module 5		Ed. 2/1973
PC-board	DJ 4 LB 005	(with printed plan)	DM 7.--
Minikit 1	DJ 4 LB 005	(2 transistors, 2 diodes, 1 ceramic trimmer, 1 TEKO box 2 A)	DM 14.30
Minikit 2	DJ 4 LB 005	(13 caps., 7 resistors, 2 feedthroughs)	DM 14.50
Kit	DJ 4 LB 005	(complete with all components)	DM 35.--

TOTAL KIT	Price only	DM 396.--
PC-BOARDS	Price only	DM 50.--

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DC 6 HL 009	25 W LINEAR AMPLIFIER		Ed. 2/1973
PC-board	DC 6 HL 009	(with printed plan)	DM 11.--
Power transistors	DC 6 HL 009	(transistors B 12-12 and B 25-12)	DM 108.--
Minikit 1	DC 6 HL 009	(1 transistor, 4 diodes, 1 IC, 5 ferrite beads, 1 ferrite choke, 6 trimmer caps., 3 feed-through caps., 1 heatsink, 2 miniature relays, 1 TEKO 3 A case)	DM 68.75
Minikit 2	DC 6 HL 009	(9 capacitors, 12 resistors, wire for coils, 2 BNC connectors)	DM 28.50
Kit	DC 6 HL 009	(complete with all components)	DM 215.--

DL 8 TM 002	6-DIGIT FREQUENCY COUNTER (Min.100 MHz)		Ed. 2/1973
PC-board	DL 8 TM 002	(double-coated, through contacts)	DM 48.--
Minikit	DL 8 TM 002	(27 ICs, 2 transistors, 2 tantalium capacitors, 1 ceramic capacitor)	DM 170.--
Indicator tubes	DL 8 TM 002	(6 Nixie types ZM 1330, 1 LED)	DM 111.--
Kit	DL 8 TM 002	(complete with all components)	DM 315.--

DL 8 TM 003	100 MHz PREAMPLIFIER/PRESCALER		Ed. 2/1973
PC-board	DL 8 TM 003	(with printed plan)....................	DM 9.50
Minikit	DL 8 TM 003	(8 transistors, 2 ICs, 4 diodes, 4 ceramic capacitors, 2 electrolytics)	DM 60.--
Kit	DL 8 TM 003	complete with all components	DM 69.--

---

Centimetre / inches

		0	1	2	3	4	5	6	7	8	9	10 cm
cm	0	0	.394	.787	1.181	1.575	1.969	2.362	2.756	3.150	3.543	3.937
"	.1	.039	.433	.827	1.221	1.614	2.008	2.402	2.795	3.189	3.583	
"	.2	.079	.472	.866	1.260	1.654	2.047	2.441	2.835	3.228	3.622	
"	.3	.118	.512	.906	1.299	1.693	2.087	2.480	2.874	3.268	3.661	
"	.4	.158	.551	.945	1.339	1.732	2.126	2.520	2.913	3.307	3.701	
"	.5	.197	.591	.984	1.378	1.772	2.165	2.559	2.953	3.347	3.740	
"	.6	.236	.630	1.024	1.417	1.811	2.205	2.598	2.992	3.386	3.780	
"	.7	.276	.669	1.063	1.457	1.850	2.244	2.638	3.032	3.425	3.819	
"	.8	.315	.709	1.102	1.496	1.890	2.284	2.677	3.071	3.465	3.858	
"	.9	.354	.748	1.142	1.535	1.929	2.323	2.717	3.110	3.504	3.898	

Millimetre / inches

		0	1	2	3	4	5	6	7	8	9	10 mm
mm	0	0	.0394	.0787	.1181	.1575	.1969	.2362	.2756	.3150	.3543	.3937
"	.1	.0039	.0433	.0827	.1221	.1614	.2008	.2402	.2795	.3189	.3583	
"	.2	.0079	.0472	.0866	.1260	.1654	.2047	.2441	.2835	.3228	.3622	
"	.3	.0118	.0512	.0906	.1299	.1693	.2087	.2480	.2874	.3268	.3661	
"	.4	.0158	.0551	.0945	.1339	.1732	.2126	.2520	.2913	.3307	.3701	
"	.5	.0197	.0591	.0984	.1378	.1772	.2165	.2559	.2953	.3347	.3740	
"	.6	.0236	.0630	.1024	.1417	.1811	.2205	.2598	.2992	.3386	.3780	
"	.7	.0276	.0669	.1063	.1457	.1850	.2244	.2638	.3032	.3425	.3819	
"	.8	.0315	.0709	.1102	.1496	.1890	.2284	.2677	.3071	.3465	.3858	
"	.9	.0354	.0748	.1142	.1535	.1929	.2323	.2717	.3110	.3504	.3898	

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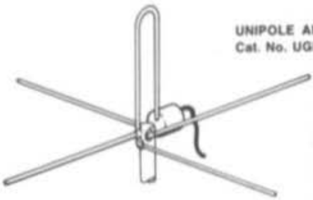

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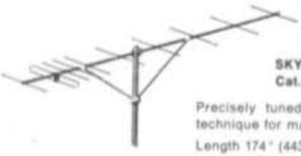

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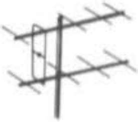


FIVE OVER FIVE
Cat. No. D5/2M

Gain 10.8dB

Slot Fed Double 5 Yagi

Length 63¼" (161 cm)

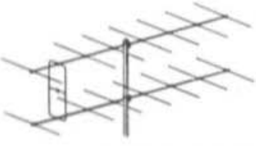
Width 40½" (103 cm)

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)

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Height 46" (116 cm)

Horizontal Beamwidth between half power points 45°

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

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Type U 3   Type U 4   Type U 5

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TA 4	$1/4\ \lambda$	144-175 MHz	0 dB	130 g	Stainless steel (PH 17-7)
U 3	$5/8\ \lambda$	400-470 MHz	3 dB	100 g	Silver-plated, epoxy coated
U 4	Colinear	420-470 MHz	4 dB	150 g	Stacked $\lambda/4$ and $5/8\ \lambda$
U 5	Colinear	420-470 MHz	5 dB	175 g	Stacked $5/8\ \lambda$ and $5/8\ \lambda$

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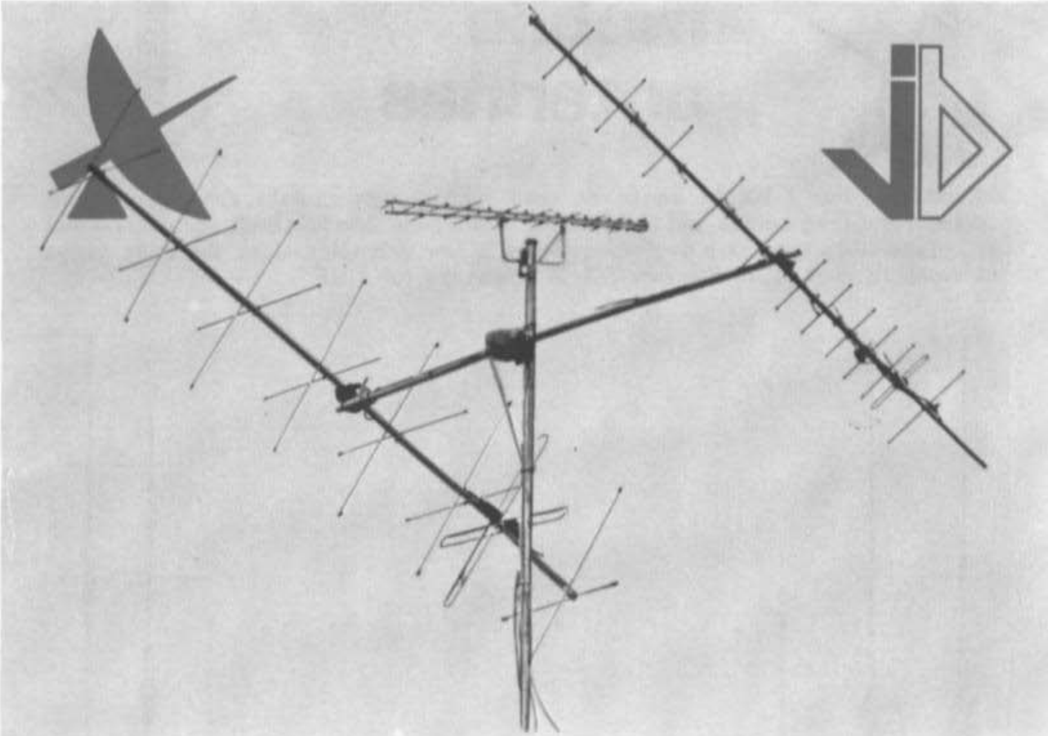

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8XY/2 m	2 x 8	12.2 dB (10.0 dB)	45°	2.85 m
10XY/2 m	2 x 10	14.2 dB (12.0 dB)	33°	3.65 m
12XY/70 cm	2 x 12	15.2 dB (13.0 dB)	35°	2.60 m