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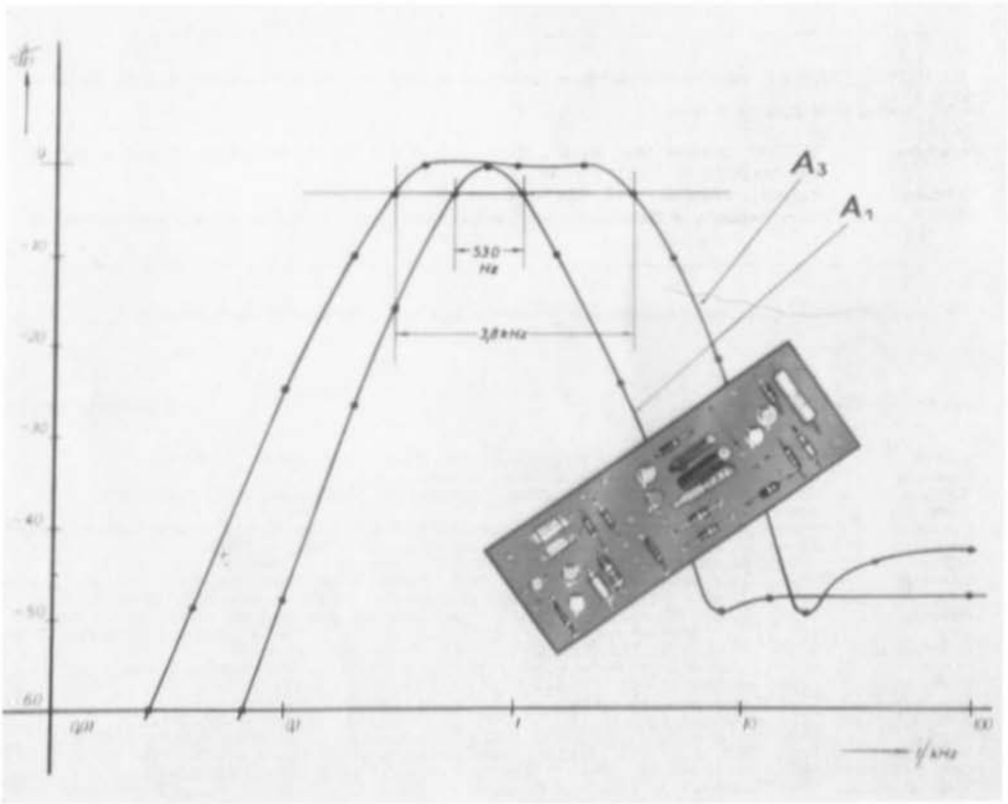

ACTIVE AUDIO FILTERS

CONTENTS:

F. Weingärtner, DJ 6 ZZ	A 28 MHz/144 MHz Transistorized Transverter	Page 189—195
D. E. Schmitzer, DJ 4 BG	Linear Integrated Circuits for Amateur Applications	Page 196—204
K. Maiwald, DJ 4 KH	A Bandpass Filter for 145 MHz	Page 205—208
G. Hoffschildt, DL 9 FX	A Three-Stage VFO for 48.0 to 48.7 MHz	Page 209—214
H. Wilhelm, DL 8 AT	Automatic Search Oscillator for Two Metre Converters	Page 215—217
D. E. Schmitzer, DJ 4 BG	Active Audio Filters Part I	Page 218—225
D. E. Schmitzer, DJ 4 BG	Active Audio Filters Part II	Page 226—235
K. Meinzer, DJ 4 ZC	A Balloon-Carried Translator	Page 236—242
H. J. Franke, DK 1 PN	A Ten Watt Transmitter for 70 cm	Page 243—248

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EDITORS: Robert E. Lentz, DL 3 WR
Terry D. Bittan, G 3 JVQ
DJ Ø BQ

A 28 MHz/144 MHz TRANSISTORIZED TRANSVERTER

by F. Weingärtner, DJ 6 ZZ

1. INTRODUCTION

The described transverter allows any ten metre or multiband transceiver to be extended for operation on the two metre band. The transverter is fully transistorized and is equipped with field effect transistors in the more critical stages. A printed circuit board with the dimensions 117 mm by 123 mm accommodates the whole unit. An input voltage of 0.5 V at 28 MHz is sufficient for an output of 200 mW at 144 MHz. The transverter can be used "barefoot" for portable or mobile operation or used in conjunction with a linear amplifier.

With the exception of the somewhat more elaborate oscillator section, the receive converter corresponds to the well-known DL 6 SW FET converter.

2. CIRCUIT DESCRIPTION

The complete circuit diagram is given in Fig. 1. Since the receive converter was described extensively in (1), this will not be explained in detail. It will be seen that junction field effect transistors (JFET) are used in the two neutralized RF amplifier stages and in the mixer stage (T 7, T 8, T 9). The required auxiliary frequency of 116 MHz is derived from a 38.667 MHz quartz crystal. The crystal oscillator is equipped with a bipolar transistor (T 4) and operates in a different manner to that used for the DL 6 SW converter. It uses the well-known, effective, common base circuit with feedback to the emitter via the crystal. The crystal oscillator is followed by a frequency tripler with the transistor T 5 in a common emitter circuit and a buffer stage equipped with the low-reactive transistor T 6. This stage suppresses undesired harmonics and supplies enough power for the transmit and receive mixer. It also ensures that any load impedance variations do not have an effect on the frequency.

The auxiliary oscillator signal is fed via the loosely coupled filter comprising inductances L 10 and L 11 to the mixer stage of the receive converter and via a low impedance tap on inductance L 10 to the mixer transistors T 1a and T 1b of the transmit converter.

The transmit mixer and the two subsequent linear amplifier stages (T 2a/T 2b and T 3a/T 3b) are built up in a push-pull configuration. Whereas the auxiliary frequency signal is fed in push-push to the two base connections of the mixer stage, the modulated signal (28-30 MHz) is fed via the resonant circuit with the inductance L 1 and the two coupling links in push-pull to the same connections. The operating points of each of the two mixer transistors are then individually adjusted using potentiometer P 1 and P 2. It is thus possible to achieve the maximum conversion gain and most favourable balancing. The auxiliary frequency signal is therefore suppressed by approx. 45 dB with respect to the maximum output power of the required signal (measured at the output of the transmit converter, points Pt 2a/2b).

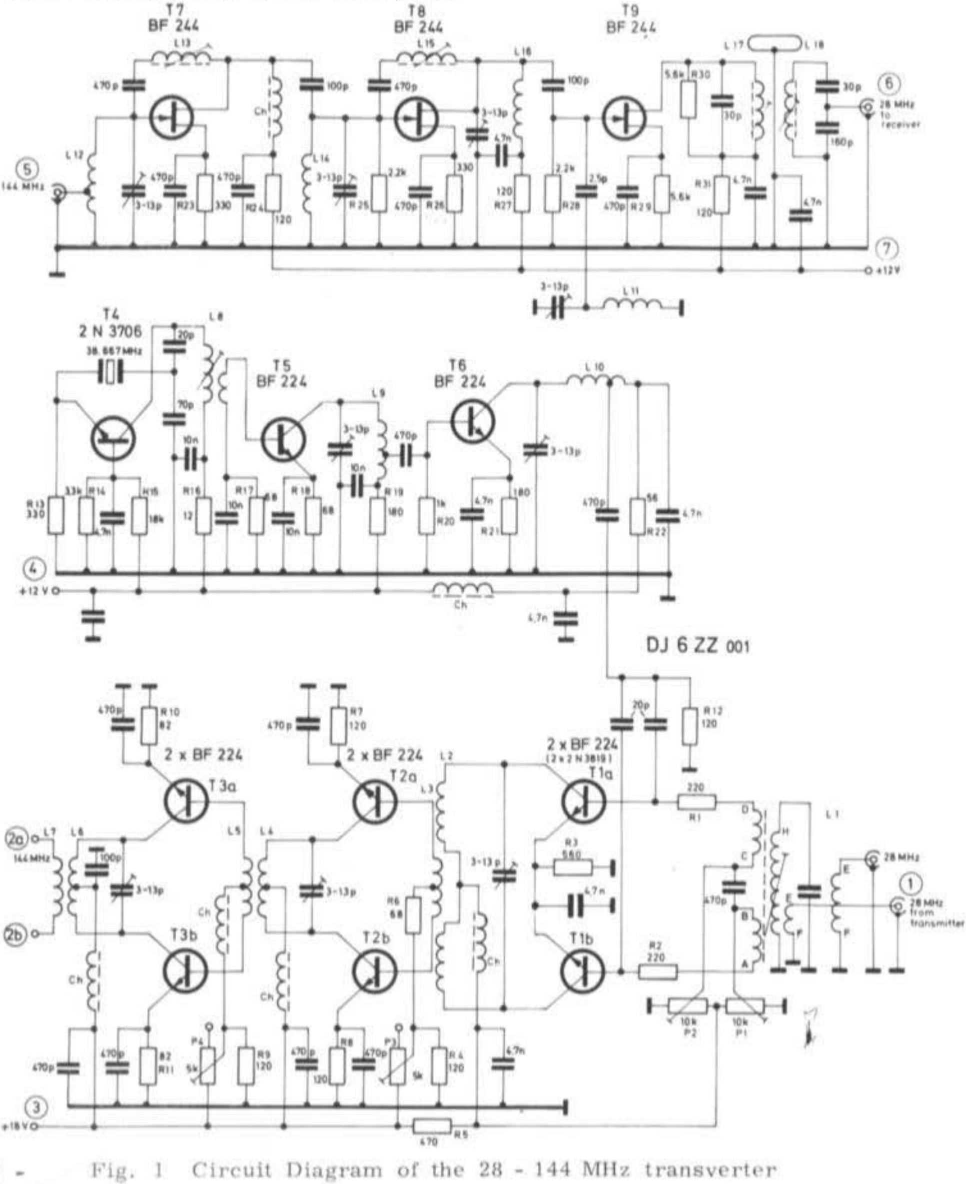


Fig. 1 Circuit Diagram of the 28 - 144 MHz transverter

The operating points of the two push-pull linear amplifier stages are adjusted with potentiometers P 3 and P 4 to the most favourable compromise between amplification and linearity.

It is also possible for the transmit mixer to be equipped with junction field effect transistors. To do this, it is merely necessary to exchange the transistors and to connect them in the following manner:

Gate connections to RF ground (R 3), the source connections are connected to the original base connections, and the drain connections lead to the resonant circuits with inductance L 2. It is possible to maintain all component values; however, the output power of the transmit converter will be reduced to approx. 50 mW after the alignment for maximum suppression of the auxiliary frequency is completed. This is due to the lower transconductance of the junction field effect transistors.

3. MECHANICAL ASSEMBLY

The complete transverter is accommodated on a printed circuit board having the dimensions 117 mm x 123 mm. Figure 2 shows a photograph of the described unit. The transmit converter is located on the left (output upper left), the crystal oscillator with frequency tripler and buffer stage (below) in the middle and the receive converter to the right (antenna input upper right). The screening plates (27 mm high, 0.5 mm brass plate) screen the various stages from each other.

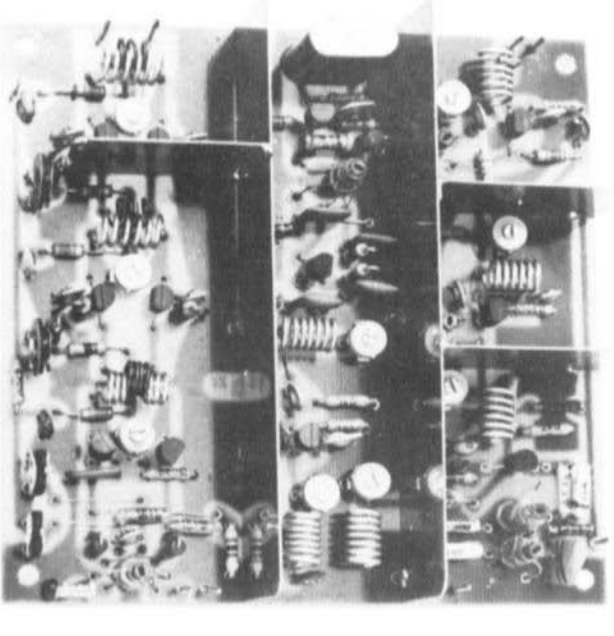


Fig. 2 Photograph of the completed transverter

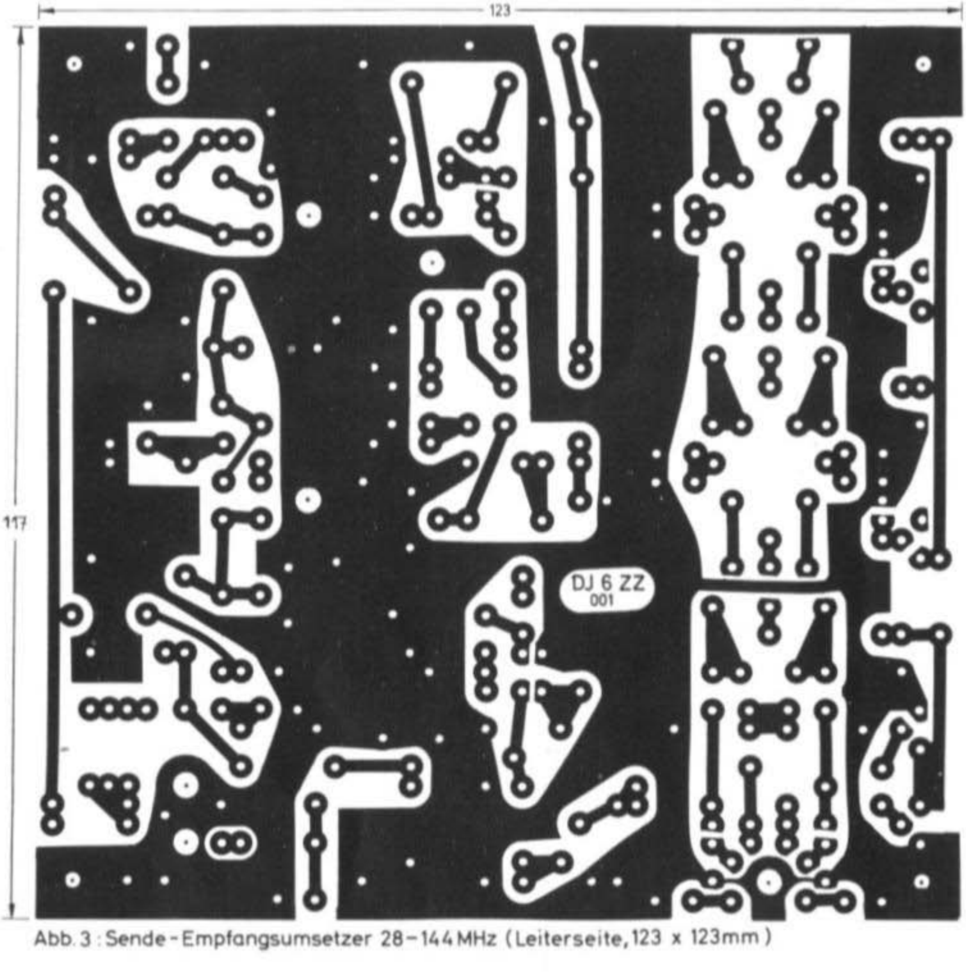


Abb. 3 : Sende - Empfangsumsetzer 28 – 144 MHz (Leiterseite, 123 x 123mm)

Fig. 3 PC-board DJ 6 ZZ 001 of the 28 - 144 MHz transverter

Figure 3 shows the conductor side of printed circuit board DJ 6 ZZ 001. The corresponding component location plan is given in Fig. 4.

The 100 pF capacitor for bypassing inductance L 6 is accommodated on the conductor side of the PC-board. The resistors R 1, R 2, R 4, R 7 to R 11 and R 17 to R 19 are mounted vertically. The quartz crystal is directly soldered, its casing connected to ground.

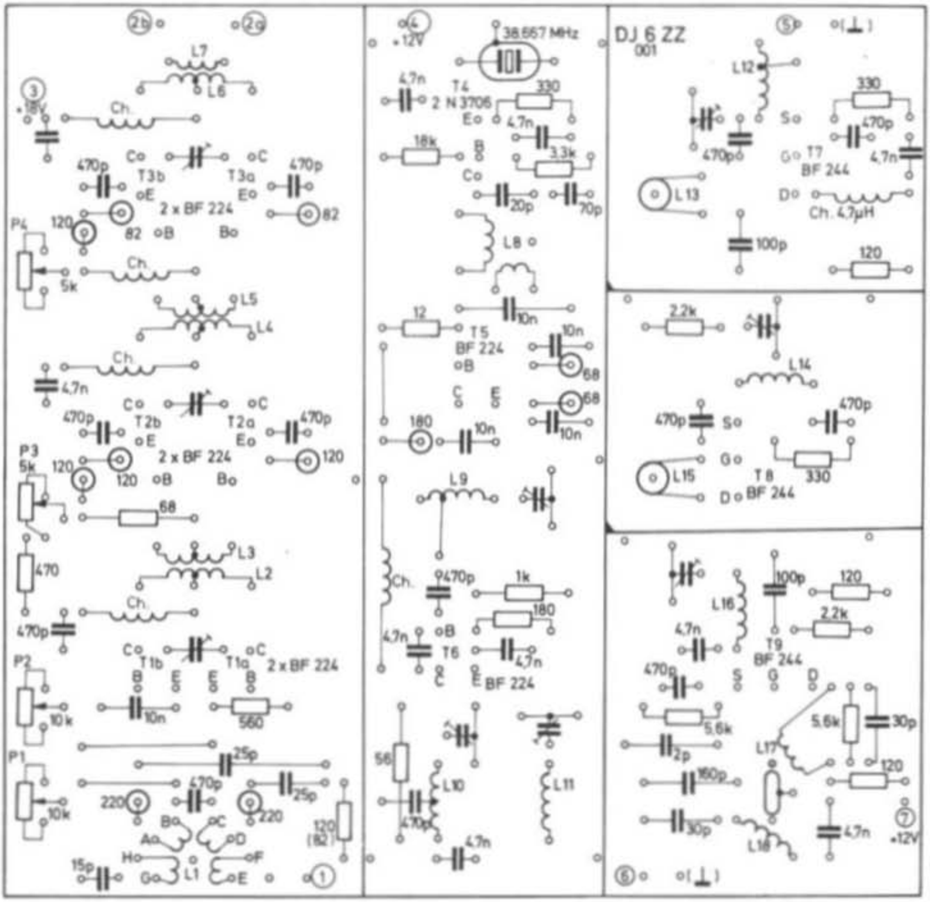


Fig. 4 Component location plan of the 28 - 144 MHz transverter

The coupling capacitor of 470 pF between inductance L 10 and the transmit mixer has only one connection to the PC-board. The other connection is soldered directly to the first turn of L 10 (from the cold end). The only further feature is the three trimmer capacitors in the transmit converter (ceramic disc trimmers): one of the two rotary connections are removed and the other is rotated so that it is opposite the stator connection.

3.1. COIL DATA

L 1 — 19 turns of 0.4 mm dia. (26 AWG) silk-covered enamelled copper wire wound on a 4 mm dia. coil former with SW core. A paper sleeve covers the inductance onto which three coupling links are wound from 0.2 mm dia. (32 AWG) enamelled copper wire; the centre link has four turns, the two outer links 5 turns. See Fig. 5 for further details.

L 2 — 6 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 5 mm former. Self-supporting. With centre tap.

L 3 — 4 turns of insulated copper wire wound on a 5 mm former. Self-supporting and coupled between the turns of L 2. With centre tap.

L 4 — 6 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 5 mm former. Self-supporting. With centre tap.

L 5 — 3 turns of insulated copper wire wound on a 5 mm former. Self-supporting and coupled between the turns of L 4. With centre tap.

L 6 — As L 5

L 7 — As L 5 but without centre tap.

L 8 — 11 turns, wire and coil former as for L 1. Coupling link: 2 turns of the same wire wound onto the cold end of L 8.

L 9 — 7 turns, wire and coil former as for L 2. Coil tap: 1 turn from the cold end.

L 10 — As L 9

L 11 — As L 9 but without coil tap.

L 12 — 6 turns otherwise as L 9. Coil tap: 1.5 turns from cold end.

L 13 — 10 turns, wire and coil former as for L 1.

L 14 — As L 12 but without coil tap.

L 15 — As L 13

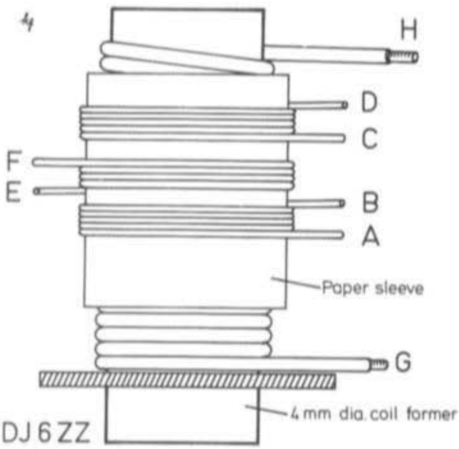
L 16 — As L 12 but without coil tap.

L 17 — 20 turns, wire and coil former as for L 1.

L 18 — As L 17. With coupling link between the hot ends of L 17 and L 18.

RF choke — 4.7 μH or 1 metre of 0.2 mm dia. (32 AWG) enamelled copper wire wound on a 4 mm coilformer.

All inductances are wound in the same direction. The self-supporting coils should be varied to suit the spacings on the PC-board.



Each winding is fixed with a universal adhesive

Fig. 5 Build-up of inductance L1 with the three coupling links A-B, C-D, E-F

4. ALIGNMENT INSTRUCTIONS

The oscillator is firstly brought into operation by connecting 12 V to point Pt 4. After aligning and checking the frequency, a voltage of approximately 0.3 V should be measured with an RF voltmeter across resistor R 12. It is possible to correct the amplitude by altering the value of R 19. If the amplitude of the auxiliary frequency is too great, this could cause unwanted conversion products.

This is followed by adjusting potentiometers P 1 to P 4 so that the associated transistors draw as little current as possible (wiper towards ground) and by connecting 18 V via a wideband choke to point Pt. 3. An RF power meter or 3.8 V/0.07 A lamp is now connected to the output (Pt 2a/2b) and trimmer potentiometers P 3 and P 4 aligned so that the associated stages draw a current of 10 mA each. Potentiometers P 1 and P 2 are now adjusted until a voltage drop of 2 V can be measured across the emitter resistor R 3. The wiper positions of these two potentiometers should firstly be identical. In this condition, an unmodulated carrier signal of 29 MHz with an effective voltage of approximately 0.5 V is fed to the input Pt. 1. The resonant circuits are now aligned, whereby a VTVM with probe would prove advantageous. The lamp should now glow slightly. The coupling between the stages and the output may now be realigned for the most favourable results. This is followed by carefully varying the positions of potentiometers P 1 to P 4 in order to find the most favourable operating point. This procedure is repeated several times until no further improvement is possible.

Finally, the mixer is balanced by alternately aligning potentiometers P 1 and P 2 in the absence of a drive signal. This is made by aligning for minimum 116 MHz signal at the output resonant circuit.

The signal is now monitored on a good two metre receiver to ensure that no noticeable non-linearities are present. If necessary, the operating points can be adjusted more towards class A operation or the drive level reduced.

This rather extensive description may tend to indicate a complicated alignment. However, this is not the case and all that is required is a little patience.

5. FIELD EFFECT TRANSISTORS IN THE TRANSMIT MIXER

If desired, it is possible to equip the mixer stage with field effect transistors subsequent to the alignment. Good results were obtained by the author using the type 2 N 3819. It is necessary for the alignment to be corrected but the settings only vary very slightly. The output power is lower but the intermodulation ratio greater, which is a great advantage if the signal is to be subsequently amplified to a high power level.

6. AVAILABLE PARTS

The printed circuit board DJ 6 ZZ 001, coil formers and trimmers as well as a complete kit of parts are available from the publishers or their national representatives. Please see the advertising page.

7. REFERENCES

(1) W. v. Schimmelmann: A 2 Metre Converter with Field Effect Transistors VHF COMMUNICATIONS 1 (1969), Edition 1, Pages 2-10.

LINEAR INTEGRATED CIRCUITS FOR AMATEUR APPLICATIONS

by D. E. Schmitzer, DJ 4 BG

Part II

Continuation from VHF COMMUNICATIONS, Edition 3.

3. APPLICATIONS USING THE INTEGRATED CIRCUIT CA 3005

3.1. CIRCUIT DIAGRAM

As can be seen in Fig. 8, this amplifier is similar to type CA 3028. It will be seen that the integrated circuit possesses an additional resistor in the emitter lead of transistor system Q 3, as well as a combination of two diodes and a resistor (R 3, D 1, D 2) together with the associated connections. According to which connections are bridged, the two emitter resistors (R 4, R 5) allow three different operating points to be adjusted. The two diodes, whose forward characteristics have the same temperature response as the emitter-base path of transistor system Q 3, allow the temperature stability to be improved.

Connection 8 is connected to the case and substrate and should always be the most negative point of the circuit. Connection 9 is connected to the highest positive potential of the operating voltage.

This integrated circuit is especially suitable for:
Narrow and wideband amplifiers, mixers, oscillators, limiters, product detectors, sawtooth generators.

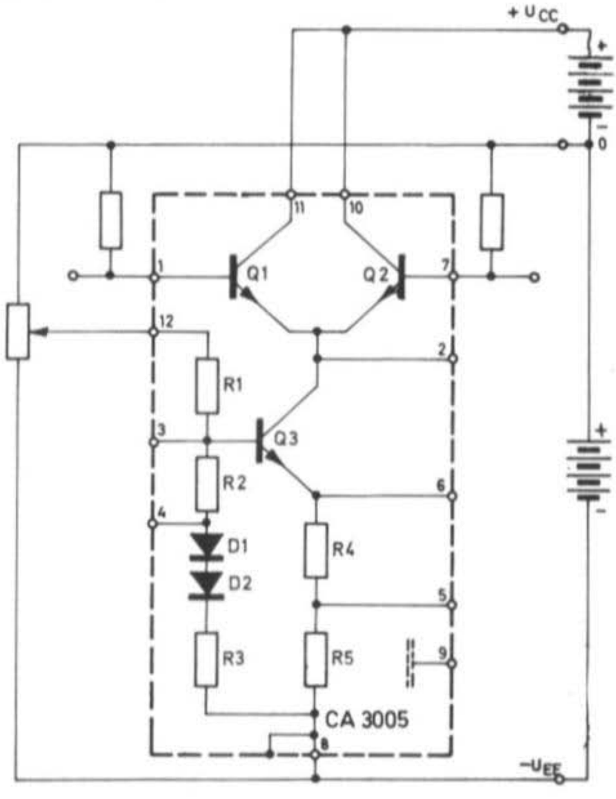


Fig. 8
Integrated circuit CA 3005 with external circuit for measuring the static characteristic

3.2. THE MOST IMPORTANT SPECIFICATIONS

The reference point for the given voltages is the common junction point of the two voltage sources.

The following static characteristics result in a circuit as given in Fig. 8:

with U_{CC}	= + 6 V	I_{10}, I_{11}	= 1 mA
and U_{EE}	= - 6 V	I_1, I_7	= 19 μA

The following gain (g) and noise figures (F) can be achieved in RF amplifier circuits:

with U_{CC} = + 6 V	g = 20 dB	Cascode
and U_{EE} = - 6 V	F = 7.8 dB	
f = 100 MHz	g = 16 dB	Differential
	F = 7.8 dB	

Permissible power dissipation P_{tot} = 300 mW. The voltage difference between the connections 10 and 11 may not exceed 7 V peak-to-peak.

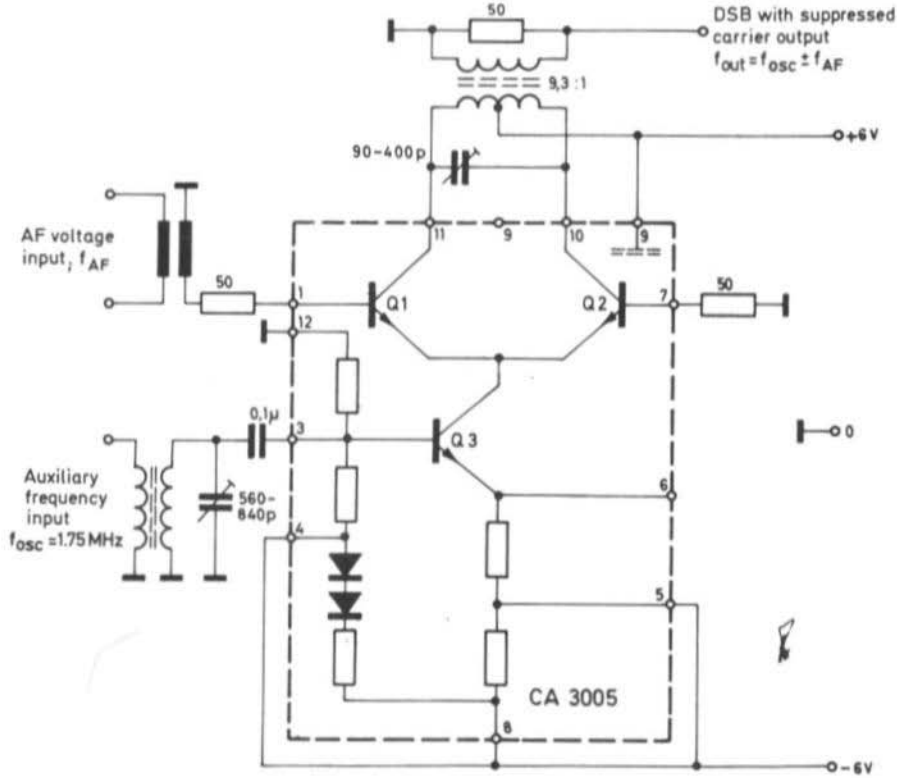


Fig. 9 Integrated circuit CA 3005 as a balanced modulator with carrier suppression at the output

3.3. CIRCUIT DETAILS

The differences between the two integrated circuit types CA 3028 and CA 3005 are not of great importance for most amateur radio applications. If the fact that the CA 3005 has several more circuit elements than the CA 3028 and that the circuit points are fed to different connections is not considered, it will be possible to build up the same circuits as were given for the type CA 3028 (1).

The excellent balance of transistor systems Q 1 and Q 2 makes this integrated circuit suitable for use as a balanced modulator. Figure 9 shows the circuit recommended by the manufacturer (2), (3). The auxiliary frequency is fed to the base of transistor system Q 3, whose collector current feeds the emitters of Q 1 and Q 2 in push-push. The AF voltage is fed to the base of transistor system Q 1. In order to improve the temperature stability, the base of Q 2 is not directly grounded but connected via a resistor whose value is equal to the impedance of the voltage source at the base of Q 1.

The AF drive is therefore not strictly symmetrical but the common feedback of the emitters affects a sufficient balance. The collector resonant circuit is connected in push-pull (double wound primary).

A carrier suppression of approximately 25 dB can be achieved with careful construction.

This value is, of course, not sufficient for a number of applications but reduces the demands made on the shape factor of the subsequent filter. The given circuit (Fig. 9) converts the AF signal by 1.75 MHz. However, since the most important dynamic characteristics of the integrated circuit are practically constant between 1 MHz and 10 MHz, the circuit should be suitable for conversion to 9 MHz providing that the resonant circuits are modified.

The circuit of a matching product detector is given in Figure 10. Since the output signal is available at two outputs with opposite phase position, it is possible to directly drive a push-pull AF amplifier.

The following specifications are given for these two circuits (Fig. 9 and 10).

	Modulator Fig. 9	Demodulator Fig. 10
Between ground and connection 3	approx. 300 mV RMS	approx. 300 mV RMS
Between connection 1 and 7	max. 30 mV RMS	max. 20 mV RMS
Between connection 10 and 11	approx. 250 mV RMS	approx. 220 mV RMS
3rd order distortion (single tone measurement)	-37.5 dB	-47.5 dB

It will be seen when comparing Figure 6 in (1) with Fig. 9 and 10 that the same circuit principle is valid in all cases: The auxiliary carrier is fed (via Q 3) to transistor systems Q 1 and Q 2 in push-push and is suppressed by the push-pull configuration at the output. The input signal feeds Q 1 and Q 2 in push-pull (exact symmetrical feeding or automatic balancing are ensured by the common negative feedback of the emitters). An unbalanced load at the output will only be permissible when the spacing between auxiliary and required output frequency is great enough for the auxiliary frequency to be suppressed by capacitors or a resonant circuit configuration (see Fig. 10).

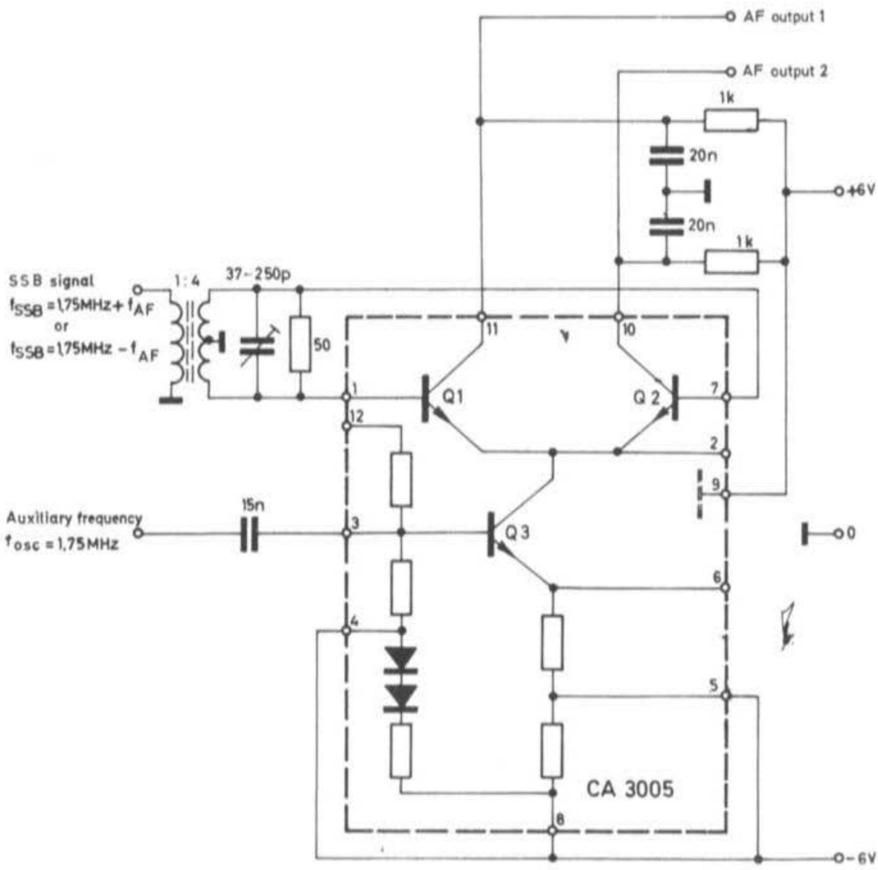


Fig. 10 Integrated circuit CA 3005 as a product detector for SSB signals

4. AF AND RF AMPLIFIER CA 3020
4.1. CIRCUIT DIAGRAM

The circuit diagram of this linear amplifier is given in Figure 11. This integrated circuit is suitable for frequencies in the order of several MHz and possesses a push-pull output. If required, it is possible to use transistor system Q 1 as an emitter-follower to drive Q 2. The common emitter pair Q 2 and Q 3 serves as a phase splitter (phase reversal) stage. The collectors then feed the base electrodes of transistor systems Q 4 and Q 5 in push-pull. These two transistor systems are in an emitter-follower configuration and are used as drivers for the output stage comprising Q 6 and Q 7. In order to linearize the drive characteristic and to compensate the temperature drift, the emitters of Q 4 and Q 5 are connected via resistors R 5 and R 7 to the base of Q 2 and Q 3 respectively. The diodes D 1, D 2 and D 3 match the operating voltages to the operating points over a wide temperature range. This ensures that the integrated circuit is fully operational in the temperature range of -55° C to +15° C. The integrated circuit CA 3020 (A) is suitable for operation as a:

AF/RF preamplifier
AF/RF driver
Low power AF/RF output stage
Wideband amplifier with a bandwidth of 6 MHz
Overloadable amplifier in servo systems.

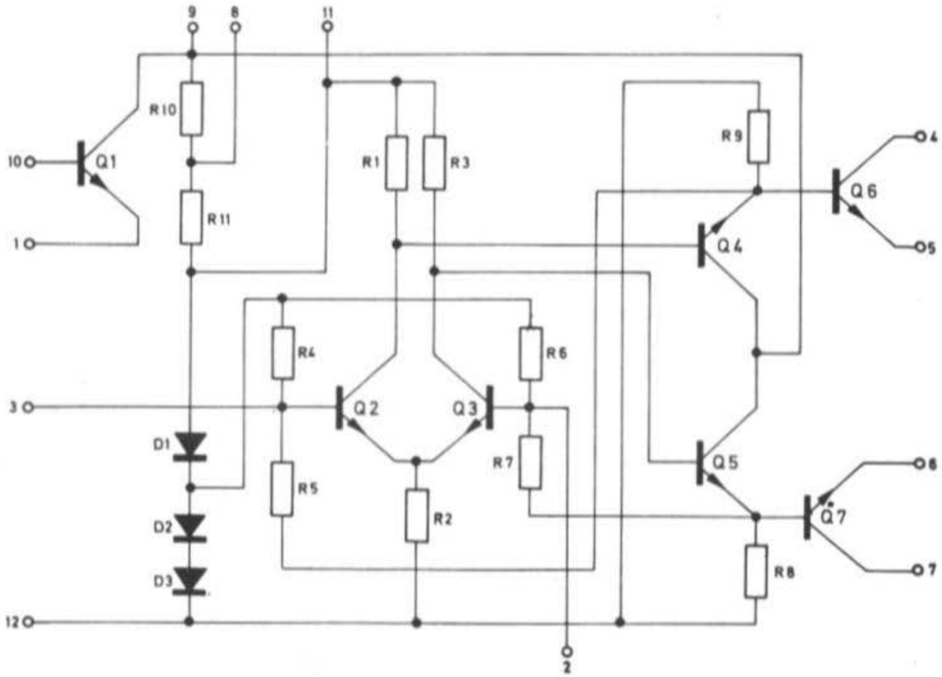


Fig. 11 Circuit of the AF amplifier CA 3020

4.2. THE MOST IMPORTANT SPECIFICATIONS

4.2.1. OPERATING VALUES

Connections 2 and 3 must be DC-free. The following values are valid for an integrated circuit CA 3020 when connected in a circuit according to Fig. 11:

Output power (f = 1 kHz)		at U_{CC} = 6 V	P = 300 mW
		at U_{CC} = 9 V	P = 500 mW
	(CA 3028 A	at U_{CC} = 12 V	P = 1 W)
Power gain (f = 1 kHz)		at U_{CC} = 6 V	P = 58 dB
		at U_{CC} = 9 V	P = 65 dB
Distortion factor k (f = 1 kHz, connection 11 connected via 1 kΩ to U_{CC}).		at U_{CC} = 6 V	1% at P = 150 mW

4.2.2. LIMIT VALUES

Operating voltage for preamplifier (connection 12 = 0)	U_8, U_9	= 12 V
Operating voltage for output stages (connections 5 and 7 = 0)	U_4, U_7	= 20 V
Emitter current for Q 1	I_1	= 20 mA
Emitter current for Q 6 or Q 7	I_5, I_6	= 100 mA
Maximum power dissipation (with heatsink for $T_c \leq 55°$ C)	P_{tot}	= 2 W

4.3. APPLICATIONAL EXAMPLE

Figure 12 shows the standard circuit where the integrated circuit CA 3020 (A) is used as a low-power amplifier. The low-impedance output (max. 100 Ω) of a preamplifier can be connected via a capacitor to connection 3. If the preamplifier has a high-impedance output (max. 5 kΩ), transistor system Q 1 can be used as an emitter-follower. Figure 12 shows the required additional connections and components as dashed lines.

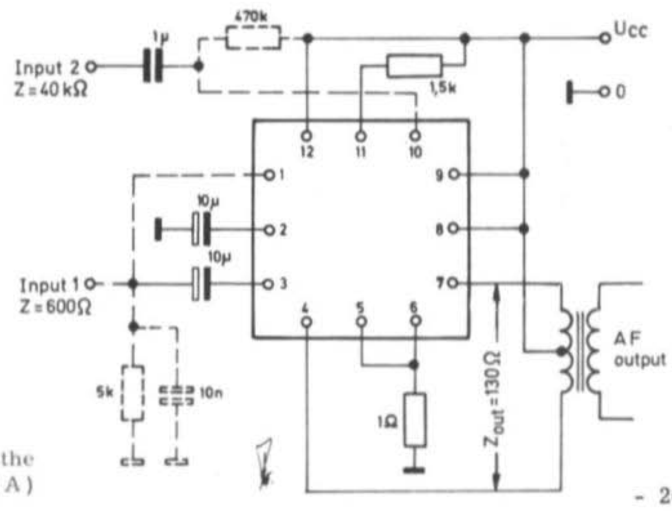


Fig. 12
0,5 W modulator using the AF amplifier CA 3020 (A)

5. TRANSISTOR COMBINATION FOR RF APPLICATIONS CA 3018

5.1. CIRCUIT DIAGRAM

As can be seen in Figure 13, this integrated circuit consists of the two independent transistor systems Q 1 and Q 2, whose specifications very accurately coincide, and a second amplifier where the emitter of Q 3 is directly connected to the base of Q 4. The transistor systems Q 3 and Q 4 (Darlington circuit) may also be independently driven if the common connection 2 is directly grounded or at least grounded with respect to signal voltages. Transistor system Q 3 will then be operated in a common emitter circuit whereas Q 4 will be in a common base configuration. The order of the connections has been chosen so that the capacitance between the base and emitter connections is as low as possible. The result of this is that no neutralization will be required even at frequencies over 100 MHz.

It should be noticed that the collectors operate as diodes with respect to the substrate.

This integrated circuit is suitable for operation as a:
Cascode amplifier for frequencies up to over 100 MHz.
Tuned amplifier, IF amplifier up to approx. 50 MHz.
Wideband amplifier up to 10 MHz.
Modulator or demodulator.

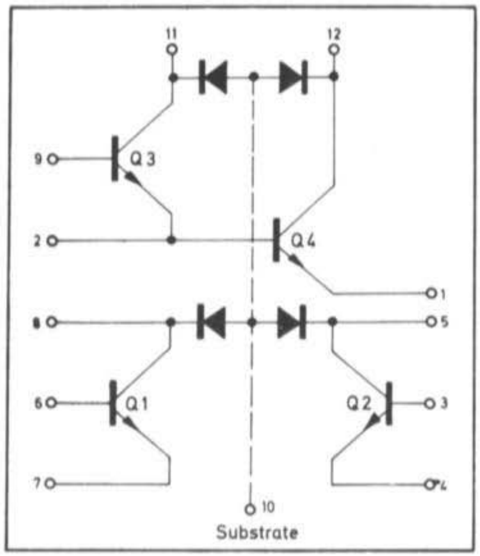


Fig. 13
Circuit diagram of the transistor combination CA 3018

5.2. THE MOST IMPORTANT SPECIFICATIONS

5.2.1. OPERATING VALUES

Each transistor system has the following specifications at:

U_{CE} = 3 V	Current gain	β	=	60
I_C = 1 mA	Transit frequency	f_T	=	400 MHz
f = 1 MHz				

The substrate is connected to the most negative point of the power supply.

5.2.2. LIMIT VALUES

Collector-emitter voltage	U_{CE}	=	15 V
Collector-current	I_C	=	50 mA
Collector-substrate voltage	U_{CS}	=	20 V
Power dissipation (per transistor)	P_{tot}	=	300 mW
Total power dissipation also	P_{tot}	=	300 mW

5.3. APPLICATIONAL EXAMPLES

5.3.1. CASCODE AMPLIFIER

A circuit of a 100 MHz cascode amplifier is shown in Figure 14. Transistor systems Q 1 and Q 2 are used for RF amplification. The gain of the cascode stages can be varied with the aid of the Darlington circuit comprising Q 3 and Q 4. The input impedance for the control voltage (connection 9) is in excess of 2 MΩ. If the control voltage increases, the emitter of transistor system Q 4 will take over some of the emitter current of Q 2 and the gain will be reduced. The following specifications are given for this circuit:

Power gain	26 dB
Control range	70 dB
3 dB bandwidth	4.5 MHz
Noise factor	6.8 dB
DC input power at 6 V	7.7 mW

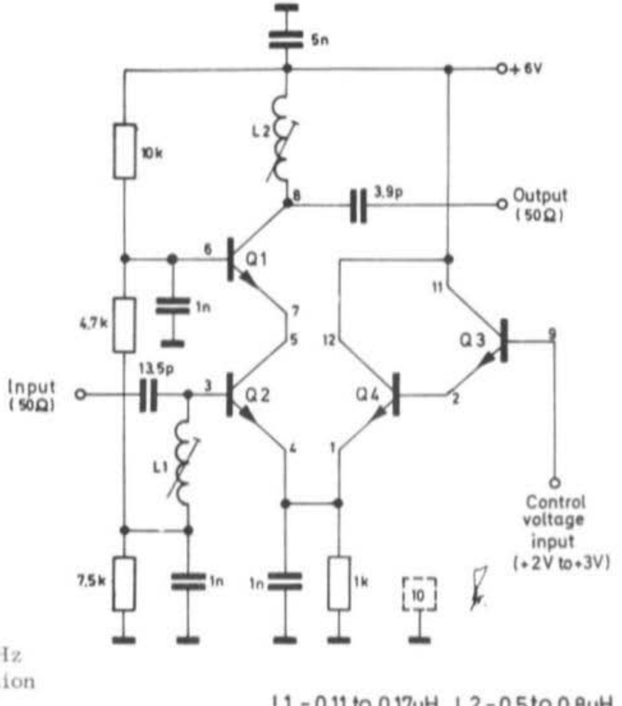


Fig. 14
Cascode amplifier for 100 MHz using the transistor combination CA 3018

L1 = 0.11 to 0.17μH, L2 = 0.5 to 0.8μH

5.3.2. CRYSTAL OSCILLATOR WITH PHASE MODULATION FOR THE 70 cm BAND

The advantages of frequency or phase modulation on the decimetre wavebands, especially with respect to transistorized equipment has been discussed for some time now. The integrated circuit CA 3018 allows a simple FM exciter to be built up using a quartz crystal in the range of 26.875 to 27.5 MHz. The circuit diagram of this exciter is given in Fig. 15. The 16th harmonic of the crystal frequency will fall into the 70 cm band. The two transistor systems Q 3 and Q 4 form a negative feedback link in the diagram given in Fig. 15. The necessary phase shift is caused by internal capacitance and delay time of the charge carriers. The resonant frequency of the feedback link is determined by the quartz crystal. Both branches of the phase bridge are capacitively coupled to the emitter of Q 4. The voltage at output B lags the voltage at output A further and further the more the inductive branch drives the output, i.e. the higher the base voltage of Q 1 is with respect to Q 2. If the base voltage of Q 1 becomes negative with respect to the base voltage of Q 2, the voltage at the output will correspondingly lead. The crest value of the output voltage therefore remains virtually constant. The circuit has a good linearity but only allows a phase deviation of ± 45°. The subsequent frequency multiplication, however, simultaneously increases the phase deviation. The bridge specifications are given for a frequency of 27 MHz.

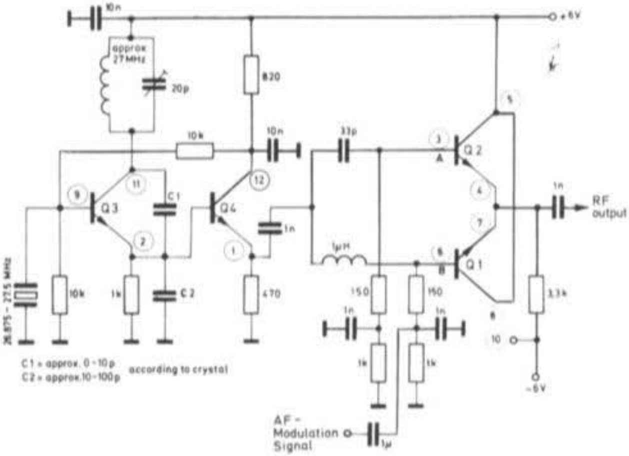


Fig. 15 Crystal oscillator and phase modulator using the transistor combination CA 3018

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(2) RCA Data Sheets:	RF-Amplifier CA 3028 (A) RF-Amplifiers CA 3005, CA 3006 Audio Power Amplifier CA 3020 Transistor Array CA 3018
(3) RCA Publications:	ICAN 5337, ICAN 5022, ICAN 5320, ICAN 5296.

A BANDPASS FILTER FOR 145 MHz

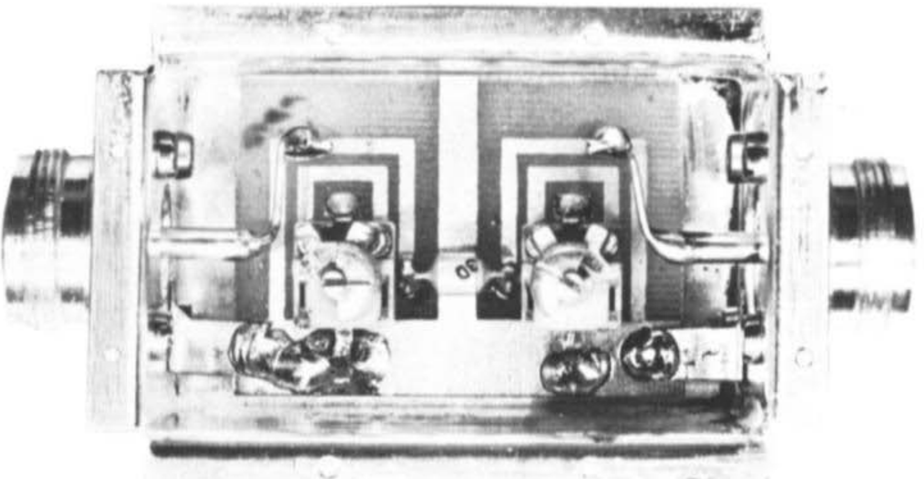
by K. Maiwald, DJ 4 KH

Transistor receivers are, due to the small linear drive range of transistors, often sensitive to cross-modulation interference from commercial, broadcast and television transmitters operating in the VHF region. Strong interference signals can cause spurious reception by mixing themselves with unwanted harmonics of the first auxiliary oscillator.

The performance of such receivers can be improved by the use of field-effect transistors in the input stages which allow these effects to be suppressed. However, this will only be successful if the interfering signals are relatively weak. A simple way of obtaining the required suppression is to use a suitable bandpass filter between the antenna and the receiver input so that all frequencies outside the pass band are attenuated.

A bandpass filter is also suitable for filtering the transmitter output to suppress unwanted spurious and harmonic components.

A bandpass filter suitable for both applications is to be described, whose main feature is the use of printed inductances. This ensures not only a high reproduceability but also a very simple construction.



The photograph Fig. 1 shows a completed filter. It can be seen that the components are mounted on the conductor side of the PC-board.

1. CIRCUIT AND MECHANICAL ASSEMBLY

The circuit diagram Fig. 2 shows the coupling capacitor C which is connected between the tapping points of the two individual resonant circuits. This allows a higher value of coupling capacitance to be used.

The bandpass filter comprises the printed circuit board DJ 4 KH 001 (Fig. 3) with the dimensions 35 mm x 50 mm. The two printed inductances are complemented by two trimmer capacitors with a range (Δ_c) of 3 - 13 pF and a fixed capacitor whose value is determined during the alignment procedure. These components are mounted onto the printed circuit board as shown in Fig. 4.

If the filter is to be operated outside of the receiver or transmitter cabinet, it will be necessary for the PC-board to be enclosed in a metal casing, which should be spaced at least 10 mm from the printed circuit board. If this is not the case, a deterioration of the frequency response will be observed. The cover must have a good connection to the casing to ensure an adequate screening.

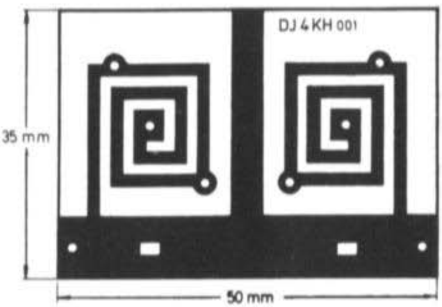


Fig. 3: Conductor side of the bandpass filter

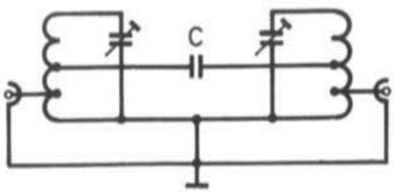


Fig. 2: Circuit diagram of the bandpass filter

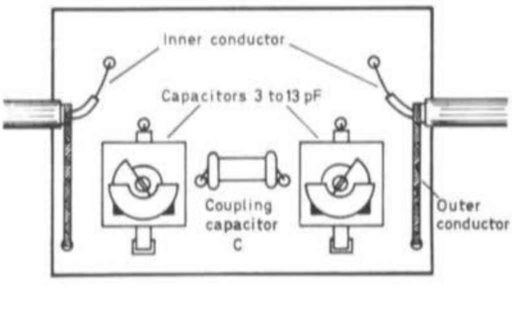


Fig. 4: Component side of the bandpass filter

2. ALIGNMENT OF THE FILTER

The two resonant circuits are roughly aligned to 145 MHz with the aid of a dip meter. During the alignment procedure, the fixed coupling capacitor is replaced by a trimmer of 3 - 13 pF. Of course, the most favourable means of carrying out the alignment would be with the aid of a sweep measuring set and oscilloscope but since few amateurs have such facilities at their disposal, the following alignment procedure is recommended:

a) Connect the bandpassfilter between the antenna and a low power two metre transmitter (< 15 W)

b) Indicate the power output by means of a reflectometer (forward power) or a fieldstrength meter.

c) Align the two resonant circuits for max. output at 145 MHz.

e) Increase the capacitance of the coupling capacitor C step-by-step - realigning the two resonant circuits for resonance at each step - until the output power reaches its maximum value. It should be noted, however, that the <u>lowest</u> value of coupling capacitance should be used that allows the maximum output (minimum insertion loss).

f) The capacitance of the coupling trimmer may now be measured and the trimmer replaced by a fixed capacitor of the same value.

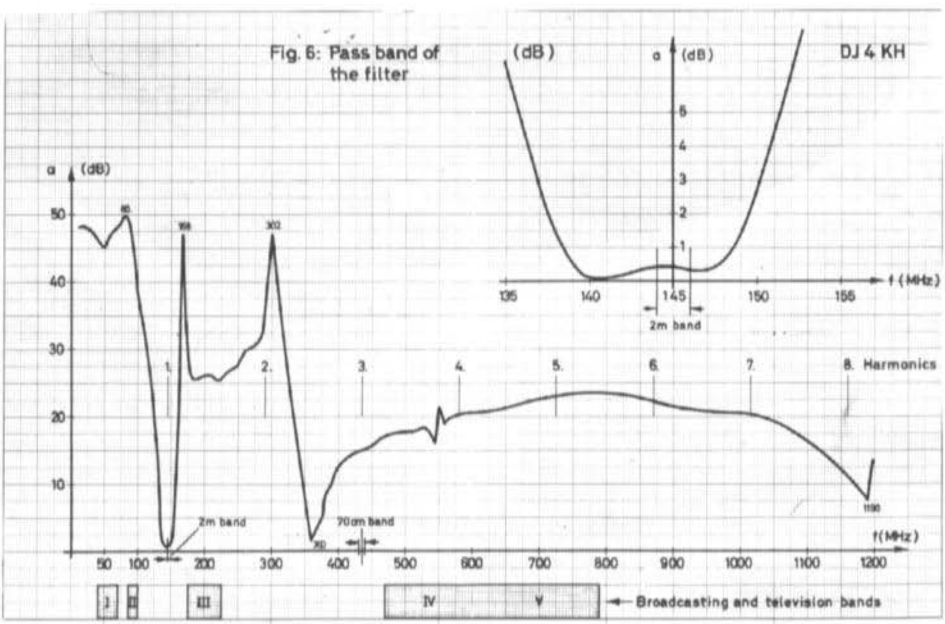


Fig. 5: Measured frequency response (dB) of the 2 metre bandpass filter (Coupling capacit. C = 7 pF)

3. MEASURED VALUES

The frequency response of the described bandpass filter, which was measured using a fixed coupling capacitance of 7 pF, is shown in Fig. 5. The pass band characteristic is illustrated in Fig. 6.

Frequencies of up to 100 MHz are suppressed by more than 40 dB, which means that signals from the television and VHF/FM bands I and II are sufficiently suppressed. The following attenuation values were measured in the pass band range of the filter (see Fig. 6):

135 MHz:	6.5 dB	142 MHz:	0.2 dB	149 MHz:	1.4 dB
136 MHz:	4.5 dB	143 MHz:	0.4 dB	150 MHz:	2.7 dB
137 MHz:	2.9 dB	144 MHz:	0.4 dB	151 MHz:	4.4 dB
138 MHz:	1.4 dB	145 MHz:	0.38 dB	152 MHz:	6.2 dB
139 MHz:	0.5 dB	146 MHz:	0.3 dB	153 MHz:	8.0 dB
140 MHz:	0.1 dB	147 MHz:	0.3 dB	154 MHz:	10.0 dB
141 MHz:	0.1 dB	148 MHz:	0.6 dB	155 MHz:	12.0 dB

The attenuation minimum, which is located between 140 MHz and 142 MHz in the case in question, can be shifted into the two metre band by careful alignment.

Television signals in Band III will be attenuated by approximately 26 dB, Band IV and V signals by approx. 20 dB. This is also true of spurious signals and harmonics originating from the transmitter which could otherwise cause TVI at these frequencies.

4. PRACTICAL EXPERIENCE

It has been found that the frequency response is greatly dependent on the impedance value at the input and output as well as on the ground points and the screening of the bandpass filter.

An effective output power of 15 W was fed through the filter during the alignment procedure without causing any damage. However, the maximum permissible power level was not determined.

The filter has been built up and used successfully in conjunction with several portable stations. In the majority of cases, the previously mentioned interference from VHF/FM and television transmitters was no longer apparent.

A second filter was installed in a cabinet (Fig. 1) and measured. This showed that it is necessary to connect a fixed capacitor of 6.8 pF in parallel to each trimmer. A coupling capacitor of 3 pF resulted in a 3 dB bandwidth of 6 MHz.

5. AVAILABILITY OF COMPONENTS

The printed circuit board DJ 4 KH 001 and the two trimmer capacitors are available from the publishers or our national representatives.

A THREE-STAGE VFO FOR 48.0 - 48.7 MHz

by G. Hoffschildt, DL 9 FX

INTRODUCTION

This variable frequency oscillator (VFO) was designed to replace the crystal oscillator stages in existing transmitters or as the basis of a new VHF transmitter. The oscillator is therefore enclosed in its own cabinet with output socket (SO-239) and offers an output frequency of 48.0 to 48.7 MHz. The control oscillator operates at 12 MHz and is tuned with a variable capacitor. This is followed by two frequency doubler stages. The second frequency doubler provides an output of approximately 500 mV into a load impedance of 60 Ω. This voltage should be sufficient to directly feed a transistorized frequency tripler stage.

The VFO is designed for an operating voltage of 13 to 14 volt which is internally stabilized at approximately 9 V using a pass transistor and zener diode. The overall dimensions of the VFO are 86 by 54 by 29 mm and the weight is approximately 150 grammes.

1. CIRCUIT DESCRIPTION

The variable frequency oscillator shown in Fig. 1 operates at the relatively low frequency of 12 MHz. The circuit is basically a Hartley oscillator with the transistor in a common emitter configuration. In order to obtain a high frequency stability, the collector of the oscillator transistor T 1 is connected to a tapping point on the resonant circuit inductance L 1. The variable tuning capacitor is also connected to a coil tap. The frequency range Δf can be varied by altering this tapping point. The total capacitance of the resonant circuit amounts to approximately 350 pF, of which 15 pF are provided for compensation of the negative temperature coefficient.

In order to obtain the lowest possible reaction on the oscillator stage, the first frequency doubler stage is connected to a tapping point near to the cold end of the resonant circuit inductance L 1. Transistor T 2 is also in a common emitter configuration and operates in class A; the collector current is, however, very low (0.8 mA). Resistor R 4 of the base voltage divider for transistor T 2 is brought out of the VFO cabinet. This facility is provided for transmit receive switching, for keying or for netting; the operating point of transistor T 2 is only maintained when this point is connected to ground (plus) and only then will the signal be amplified and doubled.

The collector of transistor T 2 is connected to a tapping point of the 24 MHz resonant circuit comprising L 2 / C 11. Resistor R 7 damps this circuit so that a bandwidth of at least 350 kHz is achieved.

Transistor T 3 of the second frequency doubler stage is also connected to a tapping point on the previous resonant circuit with L 2. This transistor operates in class B in a common emitter circuit. The collector current will increase from zero in the absence of drive to approximately 5.5 mA after correct alignment of all stages. Since an output voltage of about 500 mV is to be taken from a low impedance tap of inductance L 3, transistor T 3 is not provided with an emitter resistor. Due to this, the drive range will be greater.

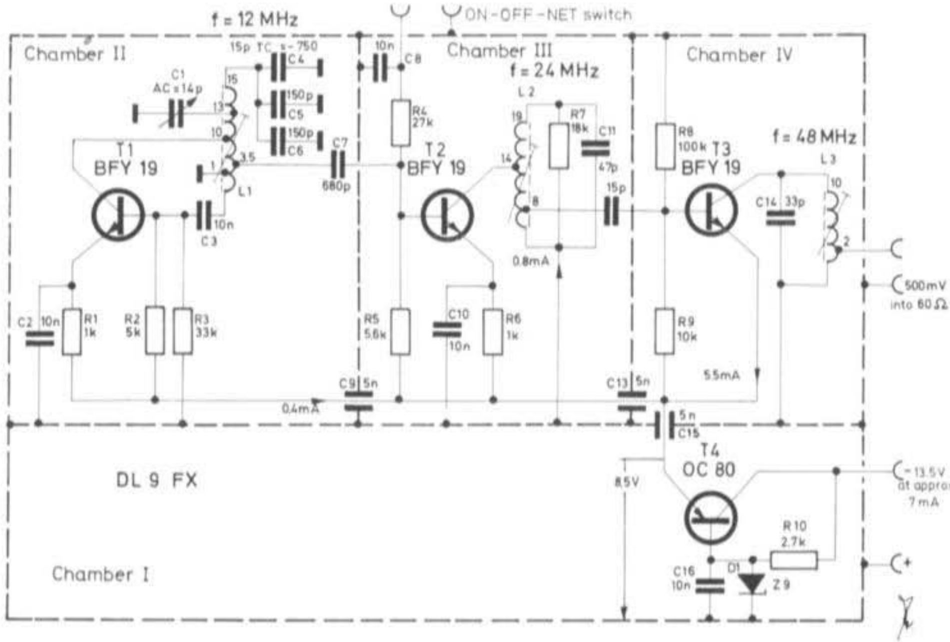


Fig. 1 A 48 MHz VFO for a two metre transmitter

The collector of transistor T 3 is connected to the hot end of the 48 MHz resonant circuit comprising L 3/C 14, so that the required bandwidth of 700 kHz is obtained.

A voltage of approximately 500 mV (into 60 Ω) is available at the output of the second frequency doubler stage. This frequency can be varied in the frequency range of 48.0 to 48.67 MHz and represents the quadruplication of the 12.0 to 12.17 MHz oscillator frequency. In practice, the tuning range is usually arranged so that it is possible to tune 15 to 30 kHz beyond these limits, corresponding to 50 to 100 kHz at 144 MHz.

The oscillator and buffer stages are equipped with the transistor type 2 N 708. A great number of similar types such as 2 N 914, 2 N 918 may also be used.

The operating voltage of the VFO is stabilized by a pass transistor whose base voltage is in turn stabilized by a zener diode. A germanium transistor type OC 80 was used as the pass transistor and complemented with a 9 V zener diode type Z 9.

2. MECHANICAL ASSEMBLY

The whole three-stage VFO is accommodated in (preferably double) copper-coated (unetched printed circuit board) epoxy (see Fig. 2); the material is 1.6 mm thick. The individual pieces are cut out, for instance, with the aid of a fret-saw and formed into a casing with four chambers as shown in Fig. 3.

The longest chamber (I) is provided with the power supply connections and the voltage stabilizing circuit.

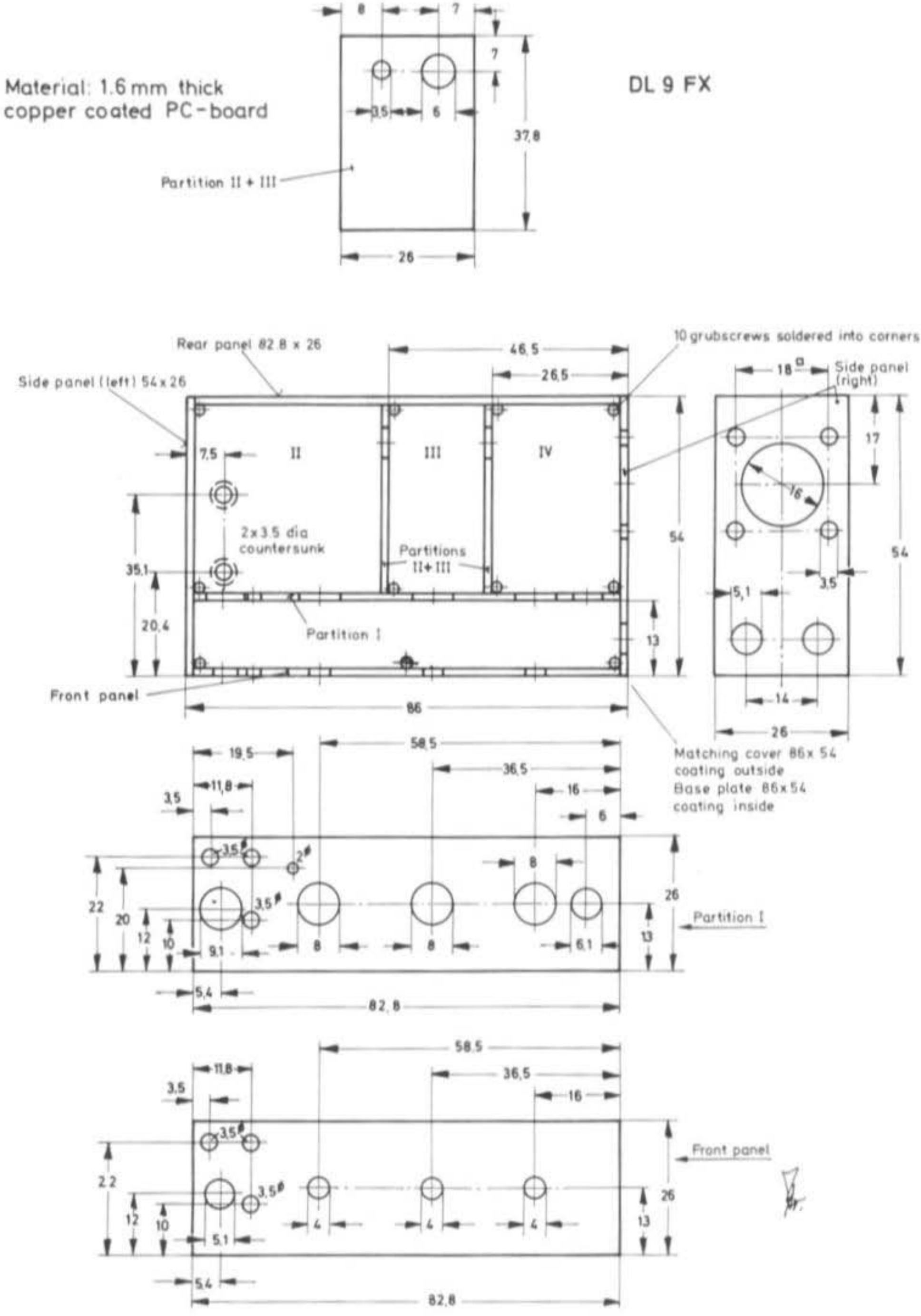


Fig. 2 Mechanical parts of the VFO; dimensions in mm

Chamber II (lower right in Fig. 3) accommodates the actual oscillator. The two-gang variable capacitor, of which only one gang is used, the two 150 pF capacitors C 5 and C 6 of the resonant circuit and the inductance L 1 can be clearly seen.

Chamber III is located to the left and contains the first frequency doubling stage from 12 to 24 MHz. Finally, chamber IV contains the second frequency doubler which provides the required output frequency of 48 MHz. Transistors T 2 and T 3 can be seen together with their resonant circuit elements in Figure 3.

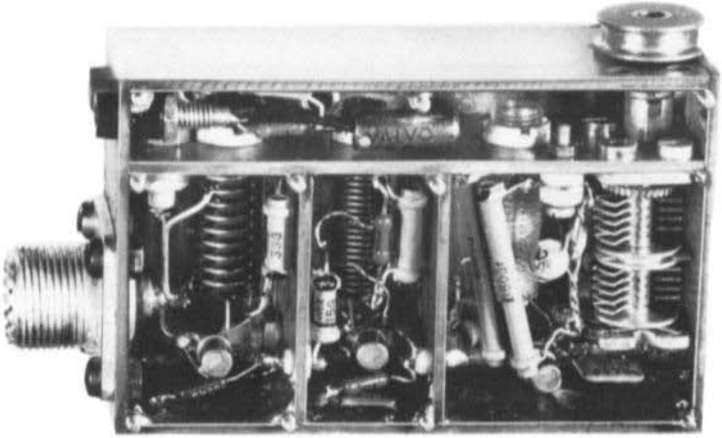


Figure 3

The outer plate of chamber I is provided with three cut-outs, through which the alignment cores of the three inductances are accessible; these cut-outs are covered by plastic insulation tape after completion of the alignment process. The drive drum of the variable capacitor is also mounted on this plate; the variable capacitor possesses a built-in 3 : 1 vernier drive.

The cover is held in place by 10 grub screws which are soldered to various positions of the casing (see Fig. 3). These screws protrude through the corresponding holes in the cover which is then held in place by nuts. The nuts should be provided with spring washers to ensure a good electrical contact between casing and cover. The complete VFO is shown in Fig. 4.

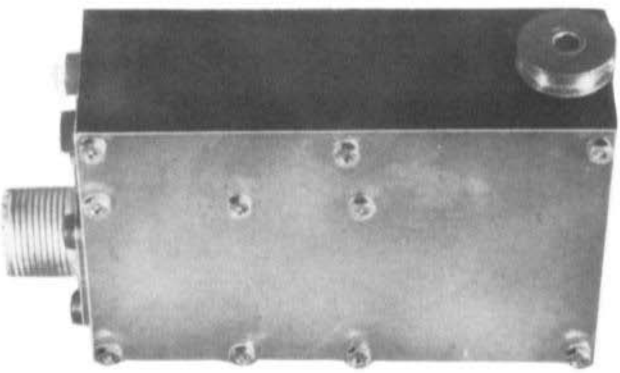


Figure 4

2.1. COIL DATA

L 1 15 turns of 0.4 mm dia. (26 AWG) silver-plated copper wire wound on a 6.4 mm dia. grooved macrolon coil former. Coil length 15 mm. Core shortened to 3 mm. Each turn fixed with a drop of two-component adhesive.

L 2 19 turns of 0.4 mm dia. (26 AWG) silver-plated copper wire wound on a former similar to L 1 but made from molded material. Coil length 15 mm. Core 10 mm long.

L 3 10 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 6.8 mm dia. grooved coil former. Coil length 19 mm. Core 10 mm long.

2.2. COMPONENTS

T 1, T 2, T 3 : BFY 19 or 2 N 708, 2 N 914, 2 N 918
T 4 : OC 80, AC 117
D 1 : Z 9, BZY 85 / D 9, OA 126/9, 1 N 764-3
C 1 : Δ C = 14 pF (two gang) variable capacitor
C 2, C 3, C 8, C 10, C 16 : 10 nF / 63 V disc capacitor
C 4 : 15 pF tubular capacitor (NTC 750 = violet)
C 5, C 6 : 150 pF tubular capacitor (TC approx. 0 = black)
C 7 : 680 pF / 350 V disc capacitor
C 9, C 13, C 15 : 5 nF / 500 V feedthrough capacitor
C 11 : 47 pF tubular capacitor (TC approx. 0 = black)
C 12 : 15 pF tubular capacitor (TC approx. 0 = black)
C 14 : 33 pF tubular capacitor (TC approx. 0 = black)

R 1, R 6 = 1 kΩ	R 4 : 27 kΩ	R 8 : 100 kΩ
R 2 = 5 kΩ	R 5 : 5.6 kΩ	R 9 : 10 kΩ
R 3 = 33 kΩ	R 7 : 18 kΩ	R 10 : 2.7 kΩ

Resistor ratings : 0.1 W

3. ALIGNMENT

With the exception of the exact temperature compensation which requires an accurate frequency meter, the alignment is quite simple.

After checking all DC currents, the oscillator is aligned to the required frequency range. If it is not possible to measure at the fundamental frequency (12 MHz), it will be necessary to roughly align the doubler stages so that the output frequency may be monitored on a two metre receiver. The amplitude of the oscillator voltage is so low that its twelfth harmonic (144 MHz) is not audible (without doublers) on a two metre receiver even when the cover is removed.

The required tuning range plus 50 to 100 kHz reserve beyond the band limits is obtained by aligning the core of inductance L 1 and altering the tapping point for capacitor C 1.

The first doubler stage is now aligned for maximum collector current of transistor T 3.

The 48 MHz resonant circuit is aligned with the transmitter connected for maximum drive or, using a 60 Ω terminating resistor and valve voltmeter (VTVM) for maximum output. It may be necessary to vary the damping of the first frequency doubler circuit until the output voltage of the VFO does not vary more than approximately 10% over the whole frequency band.

The following fine frequency alignment and temperature compensation must be made when the cover is mounted in place.

The procedure depends on the measuring instruments available. In the authors case, a 15 pF capacitor with a negative temperature coefficient of 750 ($\times 10^{-6}$) was found to offer the most favourable temperature compensation.

Finally the whole VFO may be covered by a 1 cm thick layer of styrene foam of other plastic foam material.

4. MEASURED VALUES

When calculated for an output frequency of 144 MHz, the following deviations from the previously measured frequency were determined:

Warm-up test: After 10 minutes 430 Hz
After 30 minutes 580 Hz
After 10 minutes at 65° C : 25 kHz
Switching on after 24 hours: 9 kHz
On reducing the operating voltage by 15% : 108 Hz
Non-load/short-circuit : 900 Hz.

AUTOMATIC SEARCH OSCILLATOR FOR TWO METRE CONVERTERS

by H. Wilhelm DL 8 AT

INTRODUCTION

The following small accessory for a two metre converter has been found very useful for monitoring the two metre band during periods of low or non-activity. It avoids the boring manual tuning over "dead" bands by carrying this out automatically. The search oscillator replaces the crystal oscillator of the two metre converter and thus sweeps the converter over the whole band even though the tuning of the subsequent shortwave receiver remains constant. In order to determine the presence of a signal, it is necessary for the sweep period of the search oscillator to be great enough that a new station will be noticed on the receiver. A sweep period of ten seconds from 144 MHz to 146 MHz has be found to be favourable. After the presence of a station has been determined, it is merely necessary to switch from the swept to the crystal oscillator and tune in the conventional manner.

1. THEORY OF OPERATION

The search oscillator circuit given in Figure 1 consists of an LC-oscillator (T 3) operative at half the required auxiliary frequency (116 ÷ 2 = 58 MHz). This is followed by a frequency doubler (T 4) as well as a multivibrator (T 1 / T 2) whose output pulses sweep the LC-oscillator with the aid of the varactor diode D 2.

The multivibrator with transistors T 1 and T 2 possesses two equally great RC feedback links (10 µF/220 kΩ). This means that a squarewave signal having a duty cycle of 1 : 1 is generated and therefore that the forward and return sweep are symmetrical. The sweep period can be reduced by exchanging the 10 µ F capacitors for ones having a lower capacitance.

The squarewave voltage of the multivibrator is converted into an exponentially increasing and decreasing voltage in the RC combination comprising the 1 MΩ resistor and the 5 µF capacitor. Since the time constant of this integrating link is greater than the pulse duration of the squarewave voltage, only a small portion of the exponential charge and discharge curve is utilized. This means that the sweep voltage for diode D 2 is practically triangular. Figure 2 shows the waveform in the time scale.

In order to ensure that the swept frequency, sweep width and centre frequency of the search oscillator are independent of operating voltage variations, the deflection stage (multivibrator) and thus the varactor diode are provided with a stabilized voltage (zener diode D 1).

The LC oscillator and subsequent doubler stage should be electrically and mechanically constructed in a similar manner to the crystal oscillator circuit used in the converter. The circuit and component values given in Fig. 1 serve only as an example of many possible circuits; it is similar to the oscillator circuit of the DJ 6 ZZ SSB transverter (1). The two 116 MHz coupling capacitors can be connected together at the mixer providing that the RF voltages of the fre-

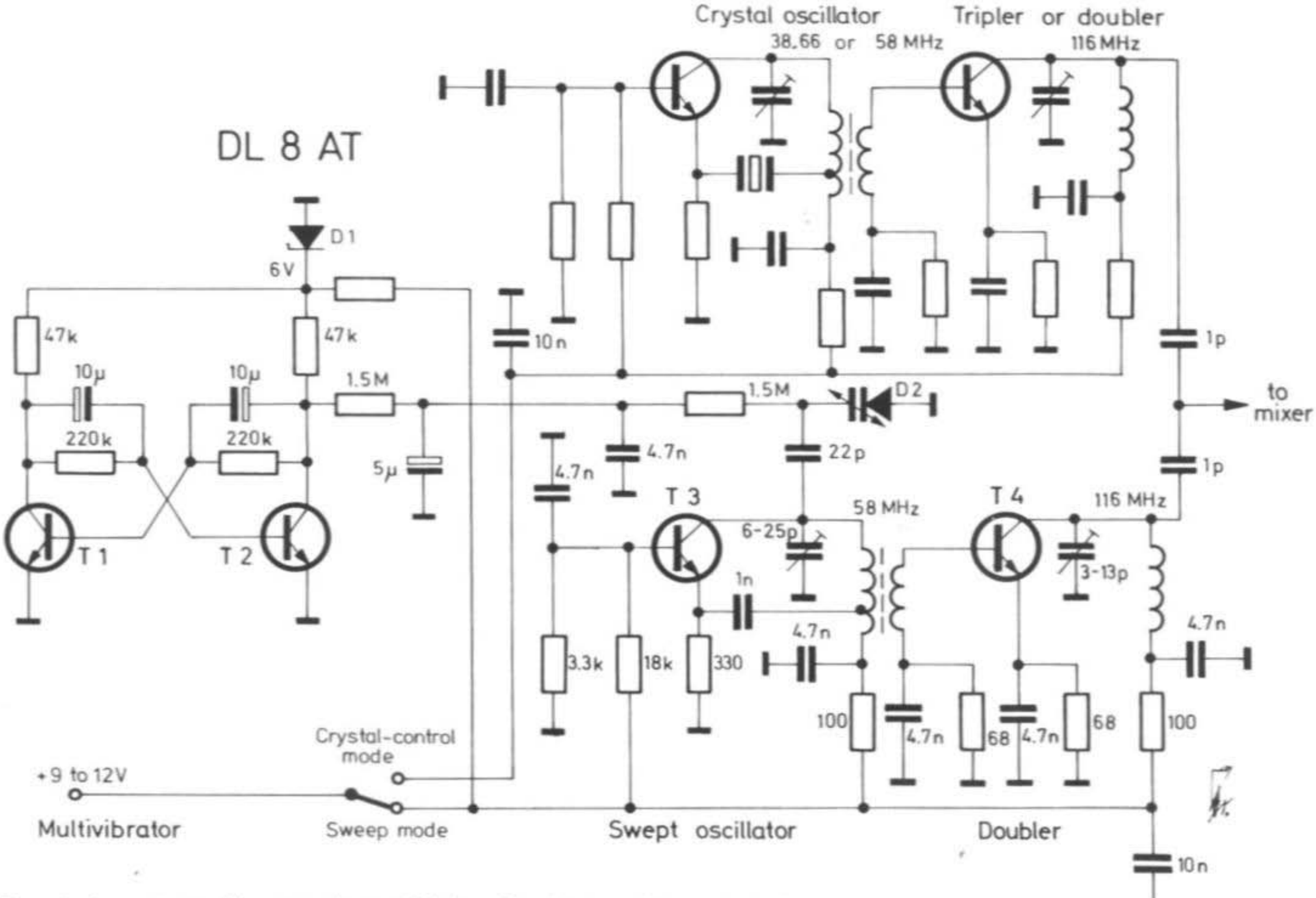


Fig. 1 An automatic search oscillator for two metre converters

quency multiplier stages are sufficiently great as to allow coupling capacitors of max. 1.5 pF. In this case, it will be merely necessary to switch over the operating voltage when switching from the search to manual tuning and vice versa.

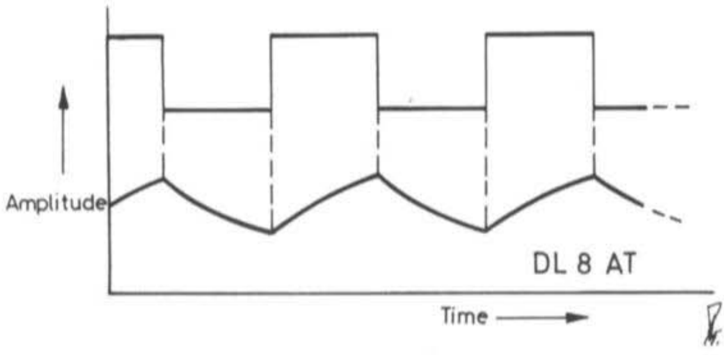


Fig. 2 Output voltage of the multivibrator circuit (upper curve) and sweep voltage across diode D 2 (lower curve)

2. COMPONENT DETAILS

T 1 to T 3	BC 108, BC 183, 2 N 708, 2 N 3903 or similar
T 4	BF 224, BF 173, BF 115, 2 N 918 or similar
D 1	Z 6, OA 126/6, BZY 85/D6V2, 1 N 429, 1 N 821 (U_Z = 6 V)
D 2	BA 101, BA 102, BA 124, BA 125, 1 N 954, (C = 35 pF at 2 V)

3. ALIGNMENT

The LC oscillator and the subsequent frequency doubler stage are aligned for maximum voltage at the mixer. During the alignment process, the 1.5 MΩ resistor and thus diode D 2 are temporarily connected to the stabilized voltage across D 1 and the resonant circuit of the LC oscillator tuned for a frequency of 117 MHz. This means that a frequency of 146 MHz will be received on the shortwave receiver when tuned to 29 MHz.

After this, the 1.5 MΩ resistor is reconnected to the integrating link comprising the 1.5 MΩ resistor and 5 μF capacitor and thus to the multivibrator.

The frequency deviation (sweep width) is now adjusted so that the lower reversal occurs at 115 MHz which means that the receiver tuned to 29 MHz will receive a frequency of 144 MHz. The sweep width is dependent on the coupling capacitor located between diode D 2 and the resonant circuit of the oscillator. The value was 22 pF in the authors prototype but the value is very greatly dependent on the diode characteristics and on the LC ratio of the oscillator.

4. REFERENCES

(1) F. Weingärtner: A Transistorized Transverter for 28 MHz/144 MHz VHF COMMUNICATIONS 1 (1969), Edition 4, Pages 189 - 195

ACTIVE AUDIO FILTERS - PART I

by D. E. Schmitzer, DJ 4 BG

1. INTRODUCTION

Active audio filters are to be described which obtain favourable attenuation curves without using inductances. A principle is given that allows filters to be dimensioned according to prototype measurements. Due to the availability of integrated circuits, the author explains how operational amplifiers can be used for filter configurations which simultaneously offer a noticeable amplification. Practical circuits are not given; they are given in the second part of this description.

It is very advantageous to keep the bandwidth of radio equipment as narrow as possible. This is not only to satisfy often disregarded official requirements but also to keep required transmission bandwidth, and thus interference to other stations, at a minimum. It should be considered that any transmit bandwidth that is greater than that required, represents wasted transmit energy. On the receive side, an excessive bandwidth will cause a reduction of the signal-to-noise ratio and thus a reduction of the receiver efficiency.

It is possible, using some circuits, to build up filter configurations using only resistors and capacitors (RC combinations) instead of inductances and capacitors (LC combinations). Several types of active RC filters are to be described that are especially suitable for tailoring the voice frequency range.

2. CIRCUITS

The four most simple and easily understandable configurations of the many known active filter circuits have been chosen, from which only two will be considered in detail (1), (2).

2.1. BASIC LOW-PASS AND HIGH-PASS CIRCUITS

Figure 1a shows a high-pass filter using shunt resistors, and Fig. 1b the derived low-pass filter. High-pass and low-pass filters using series resistors are shown in Fig. 2a and 2b. Since inductances for filter applications in the audio frequency range are large, heavy and expensive components, only filter circuits based on Figures 1a and 2b, i. e. RC filters, will be explained further.

The amplifiers contained in the basic circuit diagrams are assumed to be ideal impedance transformers having a gain of A = 1, an infinitely high input impedance, an infinitely low output impedance and no phase shift. In practice, a sufficiently good approximation of such amplifiers can be achieved using a common-collector transistor configuration (emitter follower).

2.2. ATTENUATION RESPONSE

The attenuation curves obtainable with these circuits have a maximum skirt slope of 18 dB per octave or 60 dB per decade. The fundamental difference between a high-pass and a low-pass filter is: The attenuation of the high-pass filter using shunt resistors (Fig. 1a) strives towards infinity on decreasing

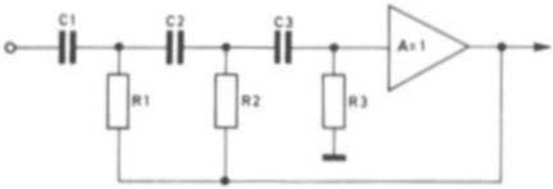


Fig. 1a High-pass filter with parallel resistors

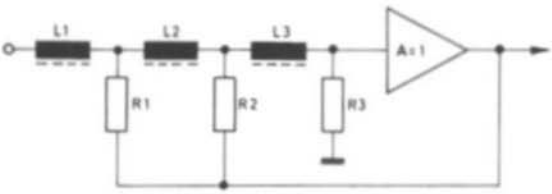


Fig. 1b Low-pass filter with parallel resistors

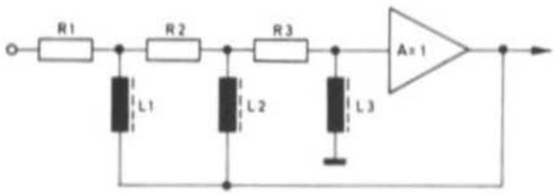


Fig. 2a High-pass filter with series resistors

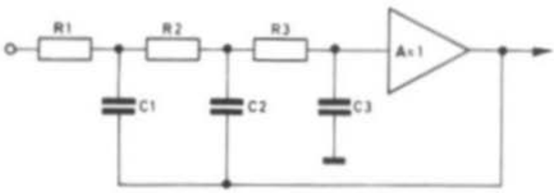


Fig. 2b Low-pass filter with series resistors

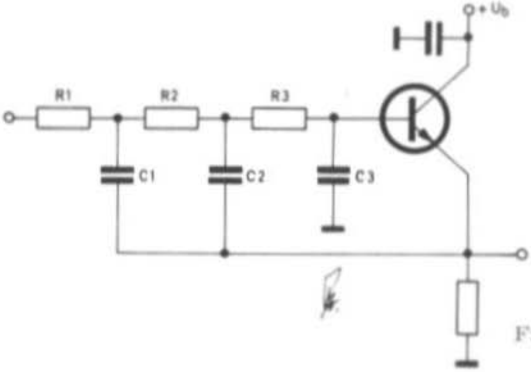


Fig. 3 Operation of the low-pass filter at high frequencies

the frequency, whereas the attenuation of the low-pass filter according to Fig. 2b approaches a finite value on increasing the frequency; this is because a voltage division between resistor R 1 and the output impedance of the amplifier will only occur at high frequencies. With a bipolar transistor in a common collector configuration, this output impedance Z_{out} is equal to the input impedance Z_{in} of the transistor in a common-base configuration since the last shunt capacitor C 3 shorts the base to ground at higher frequencies (Fig. 3). It is thus possible to calculate the maximum stop band attenuation of the low-pass filter whilst dimensioning the circuit since Z_{in} can be calculated according to equation 1a or 1b.

$$Z_{in} \approx \frac{1}{S} \approx \frac{26\ \mathrm{mV}}{I_e} \approx \frac{26\ \mathrm{mV}}{I_c}$$

Equation 1a: Input impedance of a bipolar transistor in a common-base configuration. Where S = transconductance, I_e = emitter current, I_c = collector current.

A transistor having a collector current of 1 mA will have an input impedance of 27 Ω in a common-base configuration. If a resistor of 10 kΩ is selected for R 1, the maximum stop band attenuation a_{max} is:

$$a_{max} = \frac{10\ 000\ \Omega}{26\ \Omega} = 385 \mathrel{\hat{=}} 51.7\ \mathrm{dB}$$

Higher attenuation values can be obtained by increasing the collector current so that the input impedance Z_{in} is decreased or by correspondingly increasing the value of resistor R 1. Both cases can cause difficulties because the DC operating point of the transistor must be adjusted and maintained. An improvement can be achieved by using a Darlington circuit according to Fig. 4 instead of a single transistor. The Darlington circuit allows greater series resistance values to be used since the increased current amplification means that less base current will flow for the same collector current.

When calculating the maximum possible attenuation obtainable with this configuration, it should be noted that the input impedance of the Darlington circuit is somewhat higher than that of a transistor having the same collector current. This is because the base of the second transistor in a Darlington circuit is not directly connected to zero potential in the AC sense but via the input impedance value of the first transistor as given in equation 1a. The collector current of the first transistor must therefore be included in the calculation. However, since this current is very low, the impedance, which is no longer negligible, will also appear at the output reduced by the current amplification factor β_2 of the second transistor.

$$Z_{in}\ (D) \approx \frac{26\ \mathrm{mV}}{I_{c2}} + \frac{26\ \mathrm{mV}\ B_2}{I_{c2}\ \ \beta_2}$$

Since the static current amplification B_2 and the dynamic current amplification β_2 are approximately equal, the following offers a good approximation

$$Z_{in}\ (D) \approx \frac{50\ \mathrm{mV}}{I_{c2}}$$

Equation 1b: Input impedance of a Darlington circuit (Fig. 4).

Very large series resistors are permissible when a field effect transistor is used in a common drain amplifier circuit as given in Fig. 5. This configuration leads to high stop band attenuation values although the output impedance is one or two orders-of-magnitude higher than that of bipolar transistors due to the low transconductance ($Z_{out} \approx 1/S$, is in the same order as for vacuum tubes). However, the attenuation of approximately 60 dB attainable with a single transistor should be sufficient for most applications.

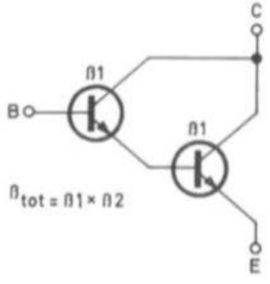


Fig. 4
Darlington Circuit

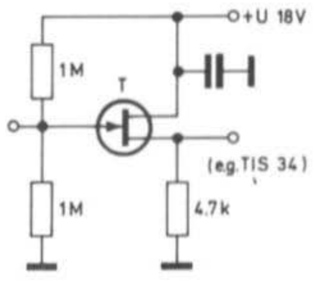


Fig. 5 Field effect transistor in common-drain circuit

2.3. PROTOTYPE MEASUREMENT AND MEANS OF CALCULATION

Since active filter circuits are not easily calculated, some fundamentals were laid down. A prototype measurement was made under these conditions from which any desired cutoff frequency and within limits, any curve form can be derived. The prototype measurement was made on a low-pass filter, as shown in Fig. 2b, according to the following considerations:

The series resistors R 1, R 2 and R 3 are of the same value, the shunt capacitors C 1 and C 2 are also equal and it is only the value of C 3 that is varied (For a high-pass filter as given in Fig. 1a, the following is valid: C 1 = C 2 = C 3, R 1 = R 2; R 3 is variable). By reducing the value of C 3 (or increasing R 3 for the high-pass filter) with respect to C 1 and C 2 (or R 1 and R 2 for the high-pass filter) it is possible for the transition between the pass band and stop band to be varied up to overshoot conditions; at lower values of C 3, the circuit can even break into oscillation, but this is not considered here. Since it has been found that the same principle is valid for both high-pass and low-pass filters, only the low-pass filter need be discussed.

The curves obtained from the prototype circuit according to Fig. 6, are shown in Fig. 7. To redimension the circuit for other cutoff frequencies, it is necessary to base same on the cutoff frequency of the first RC-link comprising R 1 and C 1. This frequency is designated f_1 in our example.

$$f_1 = \frac{1}{2\pi \; R1 \; C1} \qquad \text{Equation 2}$$

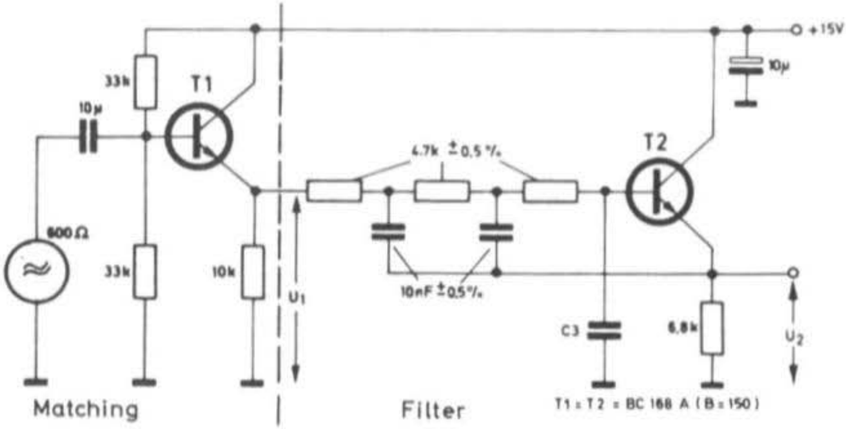


Fig. 6 Measuring arrangement for the prototype measurement

This frequency will only coincide to the -3 dB limit frequency when C 3 = 0.45 x C 1. The amplitude of the overshoot with respect to the level at frequencies well below the cutoff frequency is not considered during this definition. Figure 8 has therefore been provided as an additional aid to display the relationship between f_1 and $f_{-3\ dB}$ as a function of the relationship of C 3/C 1. If a filter for another frequency is to be recalculated from these two prototype curves, this should be made in the following manner: Firstly select the required attenuation curve from Fig. 7 which in turn lays down the relationship C 3/C 1. Fig. 8 then indicates the relationship $f_{-3\ dB}/f_1$.

Resistor R 1 or capacitor C 1 are given and the value of the appropriate second component, i.e. C 1 or R 1, must be determined with the aid of f_1 from equation 2. Capacitor C 3 is determined from C 1 and the relationship C 3/C 1 indicated in Fig. 7. Since it was determined that R 1 = R 2 = R 3 and C 1 = C 2, all frequency-determining components of the circuit are known.

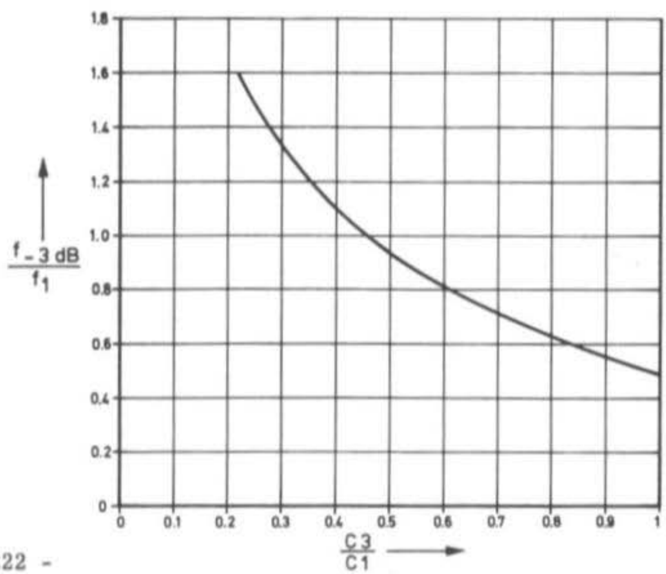


Fig. 8

Relationship between $\frac{C\ 3}{C\ 1}$ and $\frac{f_{-3\ dB}}{f_1}$

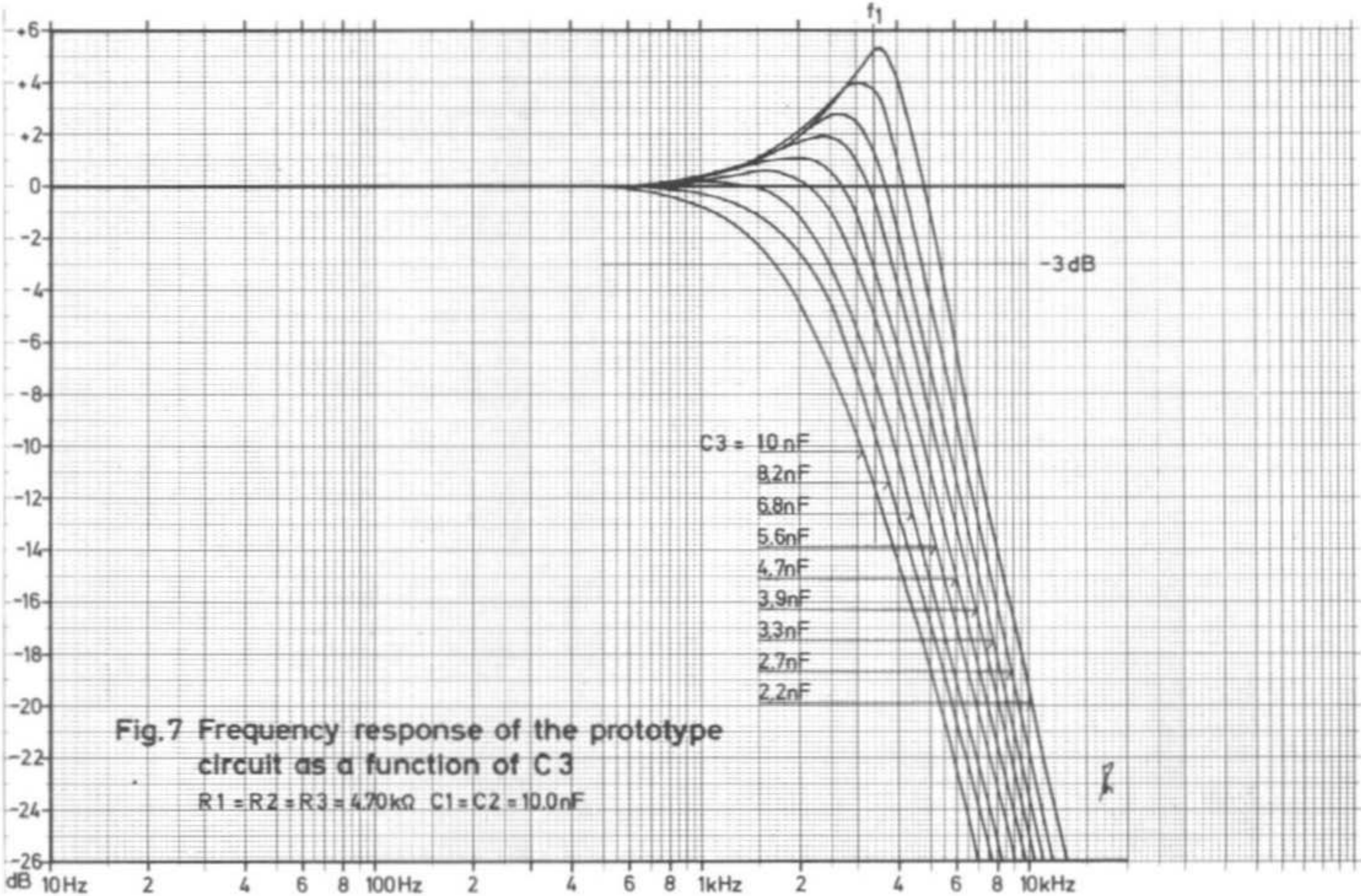


Fig. 7 Frequency response of the prototype circuit as a function of C 3

R1 = R2 = R3 = 4.70 kΩ C1 = C2 = 10.0 nF

2.3.1. EXAMPLE

A low-pass filter with the following specifications is required: $f_{-3\ dB}$ = 2.7 kHz and an overshoot of max. 1 dB. Also given is: R 1 = R 2 = R 3 = 10 kΩ. Required are the values of C 1, C 2 and C 3. Figure 6 shows that C 3 was 5.6 nF in the prototype (C 1 = C 2 = 10 nF) in order to obtain an overshoot of approx. 0.6 dB. This indicates: that C 3/C 1 = 0.56 and that according to Fig. 8 $f_{-3\ dB}/f_1$ = 0.88; f_1 is thus 2.7 kHz/0.88 = 3.07 kHz.

Equation 2 shows that: C 1 = 5.18 nF and C 3/C 1 = 0.56 so that C 3 = 2.9 nF. Normally, conventional standard values must be used (e.g. C 1 = C 2 = 5 nF and C 3 = 2.7 nF) and the series resistors somewhat varied if this should be necessary when deviating greatly from the required value.

3. ACTIVE FILTERS USING OPERATIONAL AMPLIFIERS

An operational amplifier with negative feedback down to a voltage amplification of A = 1 is essentially better with respect to satisfying the demands of an ideal impedance transformer than is a simple transistor. Such amplifier configurations are available as integrated circuits at prices well within reach of amateurs. Since these components have very low input current requirements, they are very similar to the darlington circuit given in Section 3.2. which means that they are suitable for use with filters having rather high series resistance values and low shunt capacitance. This is especially true when an additional resistor (R 4) is used to compensate for the voltage drop across the series resistor (see Fig. 9). This resistor should have a value equal to the sum of the resistance values at the non-inverted input.

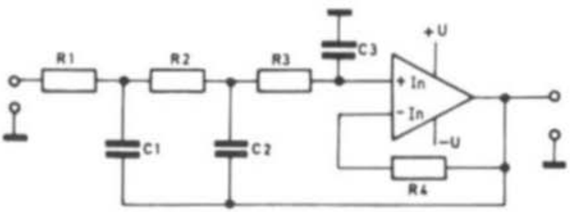


Fig. 9 Filter using an operational amplifier

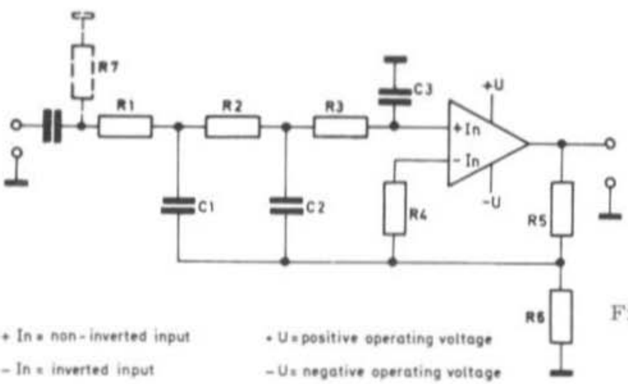


Fig. 10 Filter and amplifier using an operational amplifier

By slightly extending the circuit, it is possible to not only use it as an active filter, but also to obtain an additional amplification. Such a circuit is given in Fig. 10. If the remaining negative feedback is sufficient, the available amplification will only be dependent on the relationship (R 5 + R 6)/R 6. Since conventional operational amplifiers offer non-load amplification values of over A = 1000, it is possible to obtain operational amplification values of up to A = 100, corresponding to 40 dB. The resistance relationship in the negative feedback loop is thus 100 : 1, for instance, R 5 = 100 kΩ, R 6 = 1 kΩ.

Operational amplifiers are usually operated from two operating voltages, i. e. one positive and one negative voltage to ground (zero), for instance ± 6 V. The inputs should be connected to zero in the DC sense. If this occurs using an additional resistor (R 7 in our example), it will be necessary to increase the value of R 4 by the same amount. In actual fact, the following must be fulfilled: R 1 + R 2 + R 3 + R 7 = R 4 + R_{tot}, where R_{tot} represents the parallel configuration of R 5 and R 6 (Both are connected to zero potential, R 6 direct and R 5 to the output where zero is usually present). However, since R 5 is substantially greater than R 6 in the case of high amplification, the following simplification is possible

$$R\,1 + R\,2 + R\,3 + R\,7 = R\,4 + R\,6 \qquad \text{Equation 3}$$

In addition to this condition, it may be necessary to provide one or two small capacitors or RC combinations in order to neutralize any tendency to RF oscillation. The values of these components are dependent on the internal build-up of the operational amplifier which means that it is impossible to give any exact values here. The data sheets and application notes of the manufacturers contain such details. Some of the available integrated operational amplifiers are built-up so that no additional circuitry will be required for most applications.

4. APPLICATIONAL NOTES

It can be seen that the accuracy of the resulting attenuation curves is solely dependent on the accuracy with which the frequency-determining components coincide to the calculated values. It is therefore important that only low tolerance resistors (5% or better) and capacitors are used.

By connecting low-pass and high-pass filters in series, it is possible to obtain a bandpass characteristic.

In order to ensure that the attenuation curve is not distorted, it is necessary to feed the filter from a source impedance that is at least ten times smaller than R 1, whereas the load impedance Z_L at the output must not be less than Z_L = 30 x R 1/B where B is the static current amplification of the transistor.

5. REFERENCES

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Das DL-QTC 33 (1962), H. 3, Pages 104-107

(2) Mc Vey: An Active RC-Filter Using Cathode Followers
Electronic Engineering (1962), July, Pages 458-463

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ACTIVE AUDIO FILTERS
PART II - PRACTICAL CIRCUITS

by D.E. Schmitzer, DJ 4 BG

INTRODUCTION

The introductory article "Active Audio Filters" (1) is now to be extended by a series of proved and measured circuits designed to provide favourable AF response for both transmit and receive applications.

The measured frequency response curves are given for each of the given circuits. A universal printed circuit board has been developed by the author that allows all described circuits to be built up in a simple and space saving manner.

The intelligibility of processed speech is not noticeably less than wideband transmissions under good operating conditions (high signal-to-noise ratio); however, under poor conditions (high noise level, interference etc.) the advantages are very apparent. These advantages were explained in detail in (1) and (2).

1. AF FILTERS IN THE TRANSMIT MODULATOR

The available transmit energy of a voice transmitter is better utilized if the audio frequency range provided by the microphone is limited to that frequency spectrum required for voice transmission. The lower -3 dB frequency limit need not be lower than 300 Hz and the upper limit not greater than 3 kHz. In addition to this, the remaining transmission range should fall by 6 dB per octave in the lower frequency direction (half the frequency = half the gain).

1.1. SIMPLE LOW-PASS FILTER

The easiest manner of suppressing the unwanted higher frequency components above 3 kHz is to use a simple low-pass filter such as that given in Fig. 1a. The disadvantage of this circuit is that it must be driven from a low impedance source ($Z \leq 500\,\Omega$) so that the frequency response curve is not distorted. Since such a source (microphone) is not always available, the high impedance circuit given in Fig. 1b - where the source impedance may be up to $10\,k\Omega$ - could be of advantage. Smaller capacitance values also result that may be somewhat easier to obtain and which therefore tend to compensate for the extra expense of the second transistor.

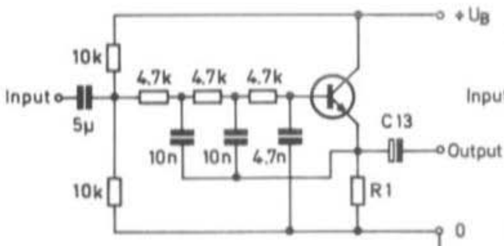


Fig. 1a Simple low impedance low-pass filter

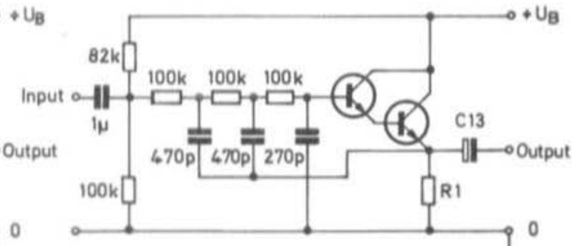


Fig. 1b Simple high impedance low-pass filter

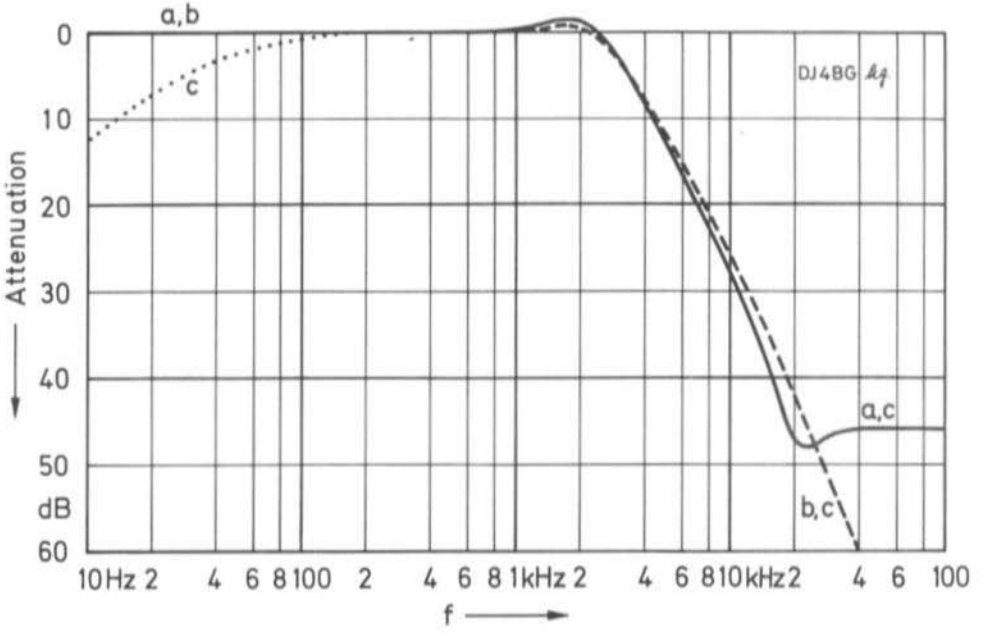


Fig. 1c Circuit 1a (curve a) and 1b (curve b) as well as circuit 2a/2b (curve c)

The measured frequency response curves of both circuits are given in Figure 1c. They were measured at an operating voltage of 9 V using the component values given in Table 2. Since the curves of circuits designed for other operating voltages offered no noticeable deviation, no further curves are given.

It can be seen that the high impedance circuit (Fig. 1b) has a more favourable response at higher frequencies. The reason for this was given in (1). In spite of this, the frequency response does not exhibit any considerable difference between the low impedance and the high impedance configuration. Since this is also valid for the following circuits, only the frequency response curve of the low impedance circuits will be given.

Both versions of the simple low-pass filter can be built up on part "B" of the printed circuit board DJ 4 BG 001 (see Fig. 7). The associated component location plans are given in Figures 1d and 1e. The values of the designated components are listed in Table 2.

1.2. ADDITION OF AN AMPLIFIER STAGE

It is possible with a suitable modification of the circuit to combine an amplifier with a low-pass filter to form a space and component saving combination. This is explained in conjunction with the simple low-pass filter in Figure 2a or 2b. The frequency response curves correspond at medium and high frequencies exactly to those of the simple low-pass filters without amplifier (Fig. 1c). At lower frequencies, the capacitors C 3 and C 4 of the amplifier stage cause a gain reduction (curve "c" in Fig. 1c). This "bass cut-off" is very desireable for speech applications. If the gain reduction may commence at 300 Hz, the capacitance values can be reduced to 1 μF for C 3 and 25 μF for C 4.

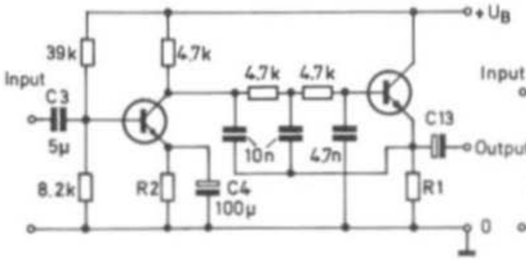


Fig. 2a Low impedance low-pass filter with amplifier

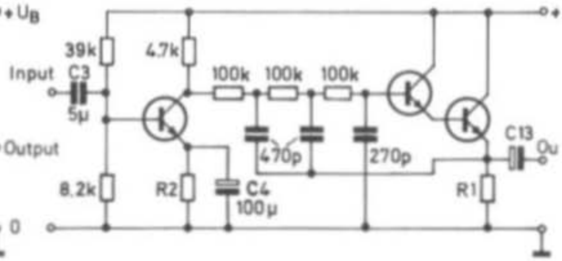


Fig. 2b High impedance low-pass filter with amplifier

The 0 dB point of the frequency response curve corresponds to the voltage amplification A in Table 1 for the circuits given Fig. 2a and 2b.

Operating voltage	6 V	9 V	12 V	18 V
Voltage amplification A for C 4 = 0 μF	7.8 17.8 dB	4.4 12.8 dB	3.9 11.8 dB	3.1 9.8 dB
Voltage amplification A for C 4 = 100 μ F	104 40.2 dB	137 42.7 dB	177 45 dB	230 47.2 dB
Current requirements	approx. 1.5 mA	approx. 2.1 mA	approx. 2.4 mA	approx. 3.2 mA

Table 1 Measured values for circuits given in Fig. 2a and 2b.

The component location plans for the low-pass filter circuits according to Fig. 2a and 2b are given in Fig. 2c and 2d.

The filters described in the following sections can also be combined with an amplifier stage providing that the input circuit is in the form of a low-pass filter. The important point is that the first filter resistor of the low impedance configuration is also used as collector resistor for the preceding transistor.

1.3. A STEEP SKIRTED LOW-PASS FILTER WITH BUFFER

If the principle is taken a little further, it is possible to build up a circuit that allows a steeper treble cut-off. The circuit, which is given in Figures 3a (low Z) and 3b (high Z), consists of a common collector stage followed by two low-pass filter sections. The corresponding frequency response curves are given in Fig. 3c and the component location plans in Fig. 3d and 3e.

If the second low-pass filter is deleted, a simple low-pass filter circuit will result with a frequency response according to Fig. 1c. The additional buffer, however, allows the low impedance circuit to be used with source impedances up to 5 kΩ.

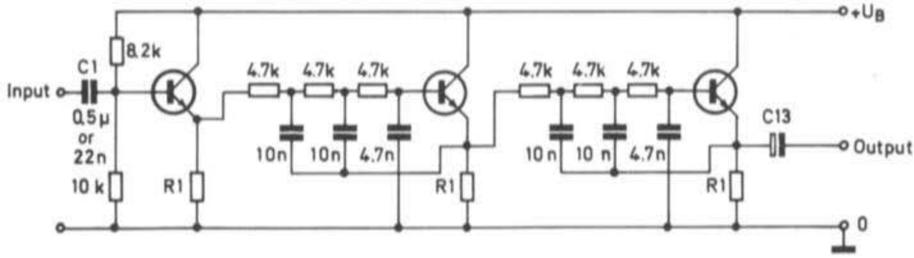


Fig. 3a Steep skirted, low impedance low-pass filter with buffer DJ 4 BG

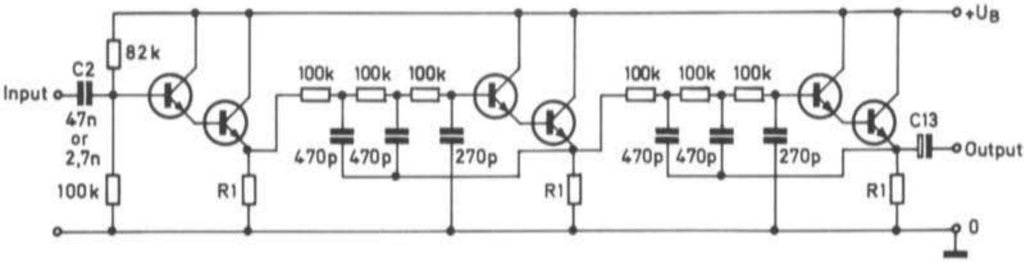


Fig. 3b Steep skirted, high impedance low-pass filter with buffer DJ 4 BG

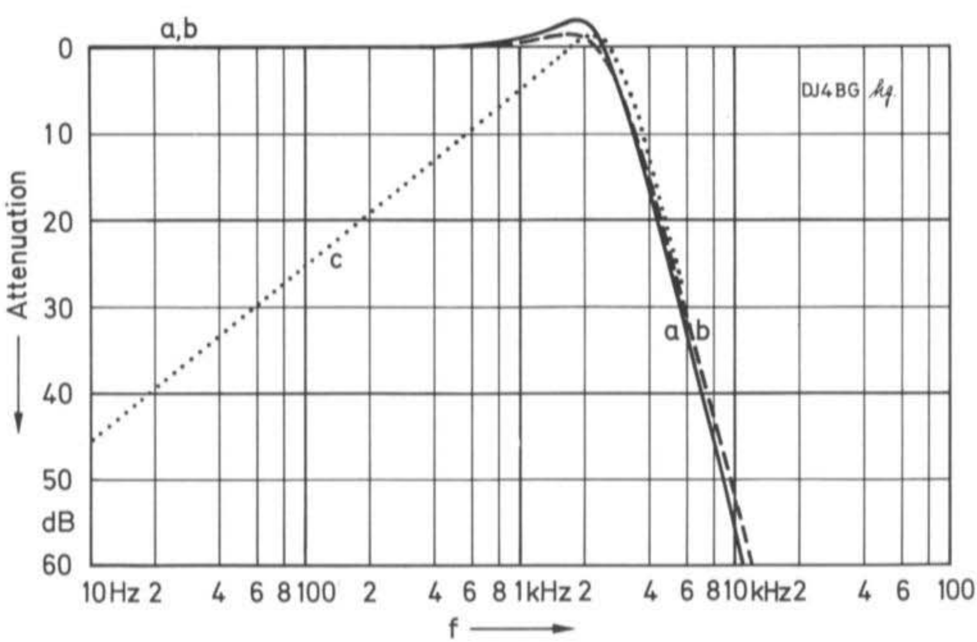


Fig. 3c Circuit 3a and 3b for large values of C 1/C 2; curve c = low values

1.4. SPEECH WEIGHTING FILTER

If the optimum frequency response for voice transmissions with a bass cut-off of 6 dB per octave is to be achieved, a point in the circuit having a defined impedance is required so that the required frequency response can be obtained using a suitable coupling capacitance. A common collector stage is suitable for this purpose since the transistor input impedance is so high that the effective input impedance is only determined by the desired low impedance of the base voltage divider. It is only necessary to use the lower values given in Table 2c for C 1 and C 2. This filter can be combined with a simple low-pass filter as shown in Fig. 4a (low Z) and Fig. 4b (high Z). Figure 4c gives the frequency response curves of the arrangement, which can be built up on part A of the PC-board. The corresponding component location plans are given in Figures 4d and 4e.

The steep skirted low-pass filter given in Fig. 3a and 3b can be used in the same manner. The frequency response curve obtained with such a configuration is also given in Fig. 3c. Both the simple and steep skirted speech weighting filter must be driven from a low impedance source; the source impedance for the low impedance configuration should be less than 1 kΩ, for the high impedance circuit less than 5 kΩ.

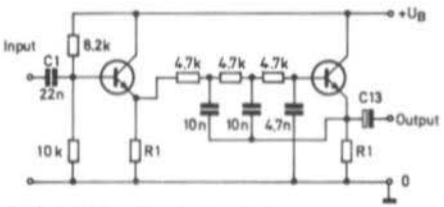


Fig. 4a Simple low impedance speech weighting filter

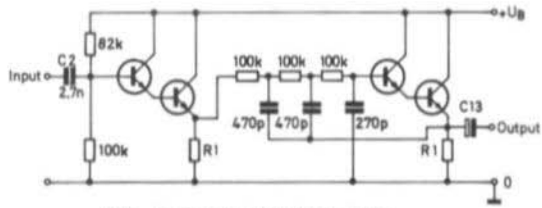


Fig. 4b Simple high impedance speech weighting filter

Fig. 4c Speech weighting filter using a simple low-pass filter

2. AF FILTERS FOR RECEIVER APPLICATIONS

In many cases, it is not necessary to provide additional filters in the AF chain of a receiver since the bandwidth is sufficiently narrow due to the IF selectivity. In spite of this, it is possible that additional filters could provide an improved signal-to-noise ratio. This is the case, for instance, if the IF bandwidth is essentially greater than 5 kHz, or if the overall selectivity of the receiver takes place near the input of the IF chain, perhaps with a crystal filter, and the subsequent IF stages contribute a noticeable noise component.

AF filtering is especially necessary in the receiver during reception of frequency modulated signals since only then is it possible to utilize the advantage of this mode to the full. With frequency modulation, the unwanted IF noise is no longer equally distributed after demodulation but increases its level with frequency deviation. The AF signal-to-noise ratio will therefore be improved if the frequencies above approximately 3 kHz are suppressed. This can be achieved with the filter circuits given for transmit applications in Figures 1a, 1b, 3a and 3b.

2.1. AF BANDPASS FILTERS

Since frequency components under 300 Hz do not contribute to speech intelligibility, they may also be suppressed using the bandpass filter circuits given in Figures 5a or 5b; the frequency response curve of these configurations is given in Fig. 5c. These circuits may be so dimensioned that the remaining bandwidth amounts to only about 500 Hz. This results in a considerable improvement of the signal-to-noise ratio during telegraphy reception using a receiver not equipped with a narrow band CW filter. The affected components are given in the "telegraphy filter" column of Table 2. The frequency response curve of the modified filter is given in Fig. 5c; the component location plans of the bandpass filter circuits are given in Figures 5d and 5e.

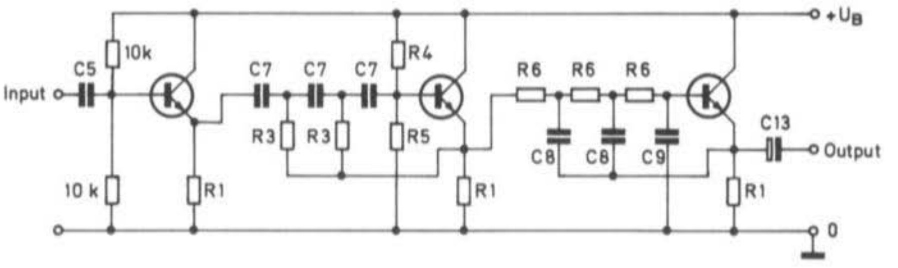


Fig. 5a Low impedance bandpass filter DJ4BG

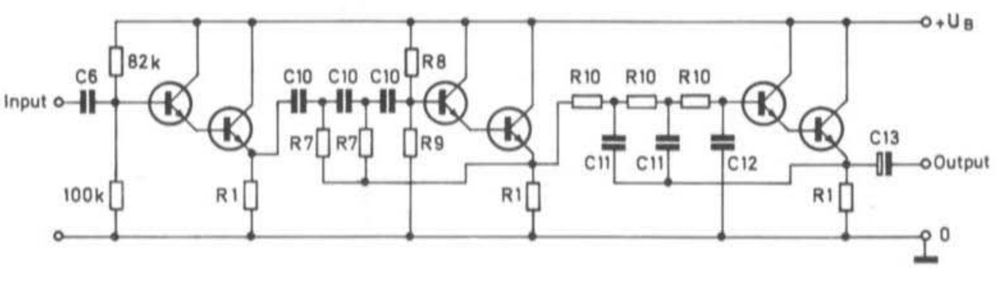


Fig. 5b High impedance bandpass filter DJ4BG

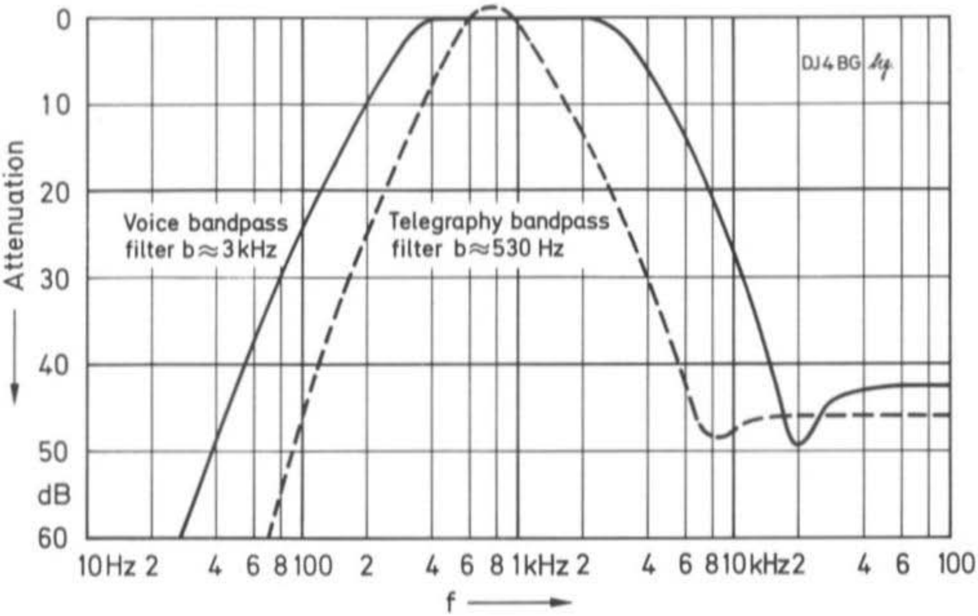


Fig. 5c Bandpass filter according to Fig. 5a or 5b

2.2. BANDWIDTH SELECTION

If the AF bandwidth of a receiver is to be switched, it will be advisable to use separate filters for telephony and telegraphy. The arrangement shown in Fig. 6 is very effective since the telephony filter increases the ultimate selectivity during reception of telegraphy.

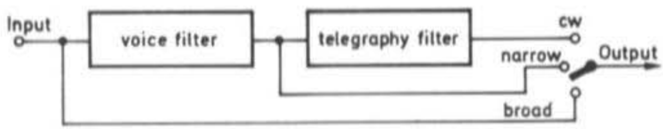


Fig. 6 Universal PC-board for active audio filters DJ 4 BG 001

3. DRIVE CAPABILITY

Generally speaking, the given filter circuits are insensitive to overload. Peak-to-peak levels up to half the operating voltage U_B can be handled without distortion which corresponds to an RMS input voltage of 1/6th of U_B. When using a preamplifier, the permissible input voltage will be reduced by the value of the gain factor.

4. COMPONENTS

The author used the silicon transistor BC 167 for all circuits. The types BC 168, BC 169 and the parallel typec BC 107 to BC 109 as well as any silicon NPN transistors having a high current amplification ($B \geq 100$ at $I_C = 1$ mA) are equally suitable. Especially low-noise transistors will only be required in conjunction with extremely low levels.

All resistors and capacitors should have the lowest tolerance possible (5% or less). Although good frequency response curves can be obtained with components having higher tolerance values, certain deviations from those given in this description will have to be expected.

The following table (Table 2) list the values of the components given in the circuit diagrams:

U_B =	6 V	9 V	12 V	18 V
R 1 =	2.7 kΩ	3.9 kΩ	5.6 kΩ	8.2 kΩ
R 2 =	560 Ω	1 kΩ	1.2 kΩ	1.5 kΩ

Table 2a Component values dependent on the operating voltage U_B

	without bass rejection	with 6 dB/octave
C 1	0.5 μF	22 nF
C 2	47 nF	2.7 nF

Table 2c Input capacitors

	Voice	Telegraphy
R 3	56 kΩ	33 kΩ
R 4	220 kΩ	120 kΩ
R 5	270 kΩ	180 kΩ
R 6	4.7 kΩ	6.8 kΩ
R 7	560 kΩ	330 kΩ
R 8	2.2 MΩ	1.2 MΩ
R 9	2.7 MΩ	1.8 MΩ
R 10	47 kΩ	68 kΩ
C 5	1 μF	0.1 μF
C 6	0.1 μF	10 nF
C 7	10 nF	10 nF
C 8	10 nF	22 nF
C 9	5 nF	10 nF
C 10	1 nF	1 nF
C 11	1 nF	2.2 nF
C 12	500 pF	1 nF

C 13 according to the impedance of the following stage: between 0.1 μF and 10 μF

Table 2b Values of the bandwidth determing components

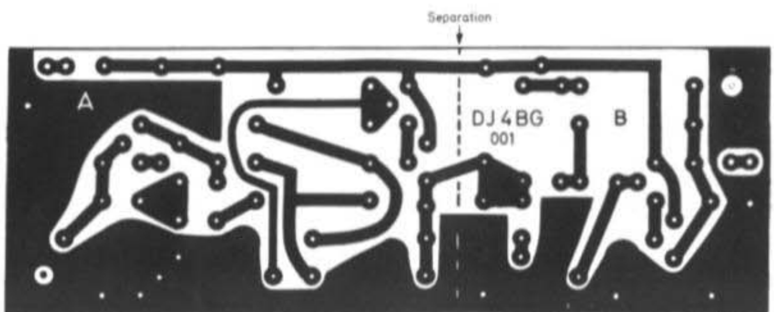


Fig. 7 Universal board for active audio filters

4.1. PRINTED CIRCUIT BOARD DJ 4 BG 001

In order to facilitate assembly, a printed circuit board having the dimensions 100 mm x 35 mm was designed on to which all the described filter circuits can be built. Due to the versatility of the PC-board, it will be necessary to leave some component positions vacant or to make individual wire bridges to suit each of the circuits. The PC-board can be separated at the dotted line (see Fig. 7) for simple configurations so that further variations are possible. The conductor side of this printed circuit board is given in Figure 7; the corresponding component location plans are given in Fig. ...d and ...e of each individual circuit.

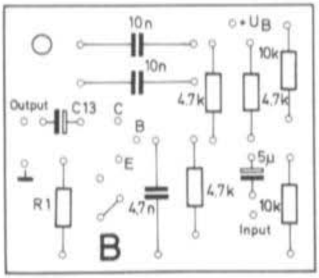


Fig. 1d Component location plan to Fig. 1a

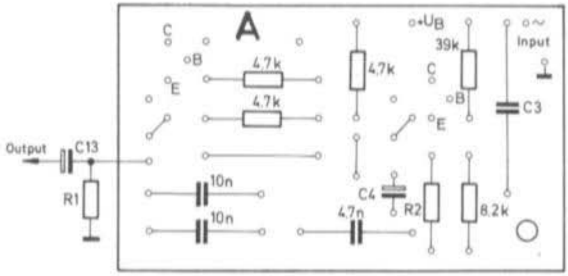


Fig. 2c Component location plan to Fig. 2a

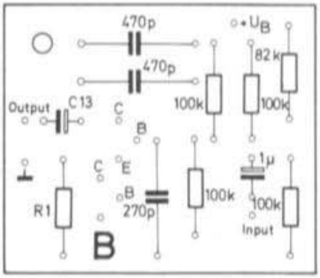


Fig. 1e Component location plan to Fig. 1b

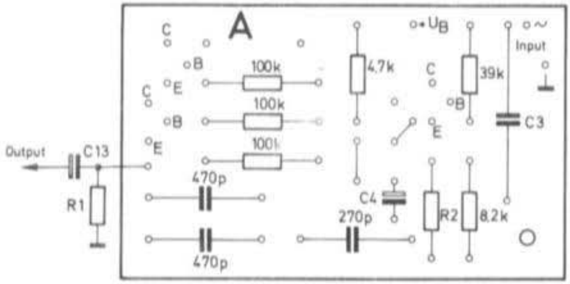


Fig. 2d Component location plan to Fig. 2b

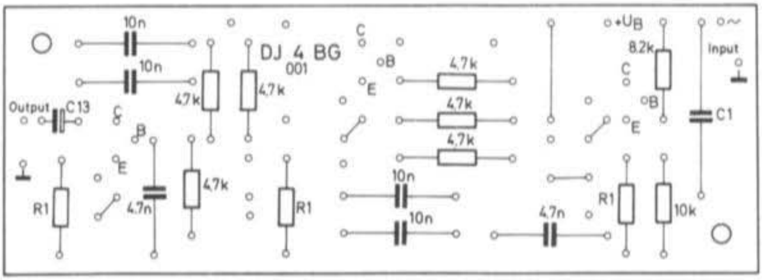


Fig. 3d Component location plan to Fig. 3a

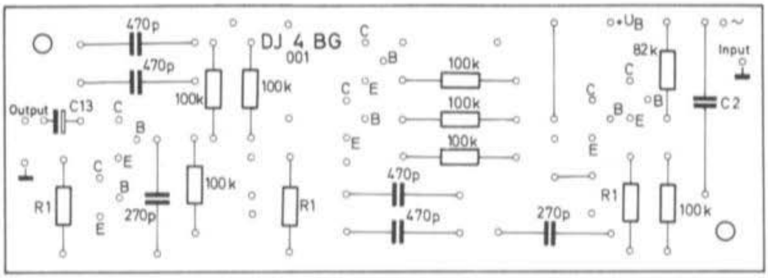


 Fig. 3e Component location plan to Fig. 3b

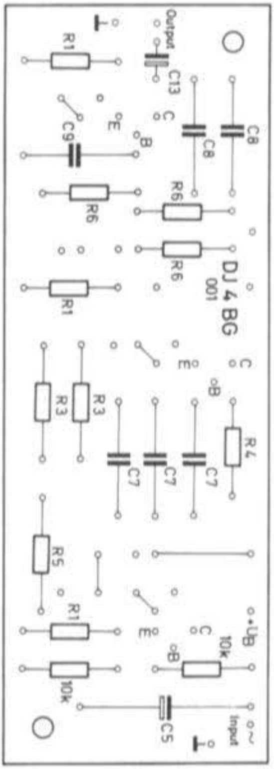


Fig. 5d Component location plan to Fig. 5a

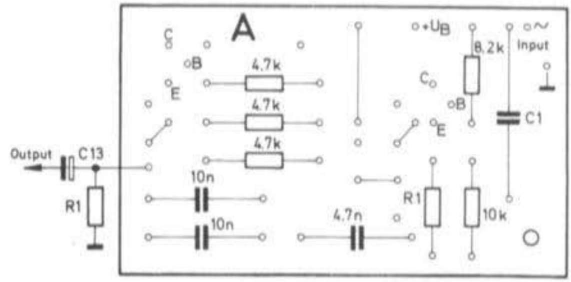


Fig. 4d Component location plan to Fig. 4a

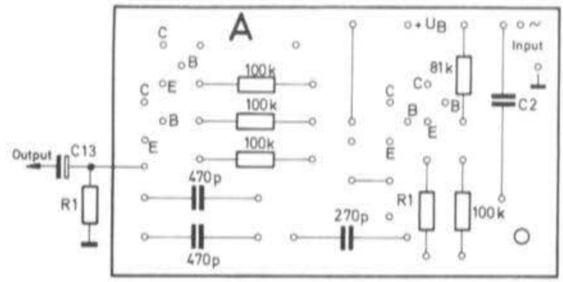


Fig. 4e Component location plan to Fig. 4b

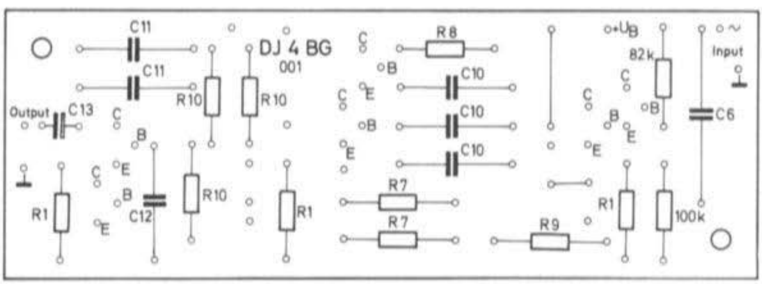


Fig. 5e Component location plan to Fig. 5b

4.2. AVAILABLE COMPONENTS

The printed circuit board DJ 4 BG 001 is available from the publishers or their national representatives. See advertising page.

5. REFERENCES

(1) D. E. Schmitzer: Active Audio Filters Part I
VHF COMMUNICATIONS 1 (1969), Edition 4

(2) D. E. Schmitzer: Preamplifiers to Improve Speech Intelligibility under Poor Operating Conditions
VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 110 to 114.

A BALLOON-CARRIED TRANSLATOR

by K. Meinzer, DJ 4 ZC

INTRODUCTION

A large number of our European readers will have heard about the balloon-carried translators of the German ARTOB and BARTOB projects. Since, however, these projects are virtually unknown outside Germany and the bordering countries, we thought that a description of the principle involved as well as a typical circuit would be of interest, especially to overseas clubs and societies which may like to attempt similar ventures.

Basically, the technical complement consists of a transistorized translator which is brought to a height of approximately 30 km using a balloon.

1. THE BALLOON

The only method for amateurs - not having NASA support - to bring an active translator to a sufficient height is to use a (weather) balloon. Such balloons are manufactured from a highly extensible material which are then filled with hydrogen. In our case, the balloon was filled to a diameter of approximately two metres. Due to the fact that the atmospheric pressure reduces during the ascent, even a slight overpressure at ground level will be sufficient to expand the balloon at higher altitudes. The reduction of the ascent speed caused by the decreasing air pressure is largely compensated for by the lift resulting from the expansion of the balloon. This means that a virtually constant ascent speed of 15 to 20 km/h can be assumed. A diameter of approximately 7 metres - which represents the maximum expansion of the specimen balloon - is obtained at an altitude of approximately 25 to 30 km. At this point, the balloon will burst and some type of parachute will be required to control the descent. The parachute used in our example has a diameter of 2.40 metres.

Figure 1 shows a typical flight plan. It can be seen that the balloon ascends steadily up to an altitude of approximately 30 km. After the balloon has burst, the parachute and translator will firstly descend rapidly and then slower and slower until they reach the ground. The horizontal dashed lines indicate the maximum distance that can be covered between balloon and a surface station; the total distance between two surface stations is therefore approximately twice this distance (see Fig. 2).

The length of these dashed lines indicates the period during which communication over this distance is possible. This, however, only deals with weather type balloons; with a combination of several balloons or using a different type of balloon it would probably be possible to achieve even greater altitudes.

2. TRANSLATOR AND BEACON TRANSMITTER

Both receiver and transmitter of the specimen translator operate on the two metre band. It is dimensioned so that amateur stations having an output power of 100 watts into an antenna with 10 dB gain will be able to fully drive the translator at a distance of 600 km. In order to ensure that weak stations are not prejudiced with respect to stronger stations, no form of gain control is used. It is therefore necessary that the power limits given in Fig. 3 are not exceeded; otherwise, the translator would be overloaded and spurious intermodulation products will appear in the transmit band.

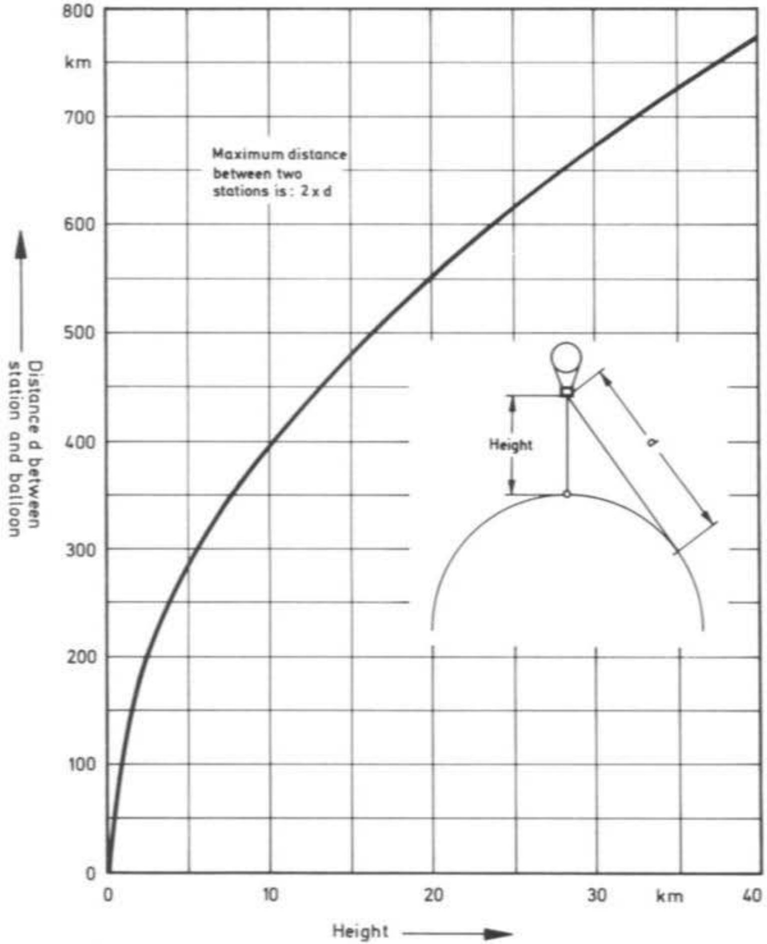


Fig. 2 Maximum distance as a function of the balloon altitude

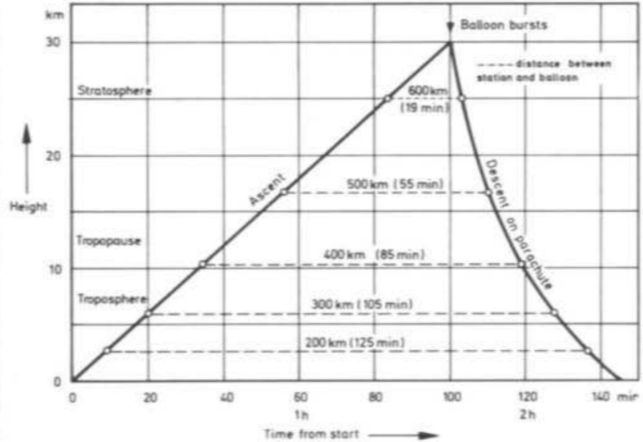


Fig. 1 Typical flight plan of a balloon-carried translator

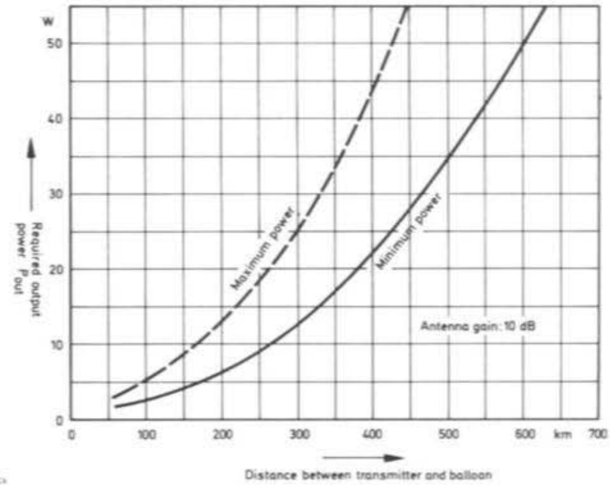


Fig. 3 Required output power as a function of distance

The same frequency ranges have been chosen as were used for the amateur satellite Oscar III; however, the bandwidth has been limited - by cost reasons - to 40 kHz. This means that approximately twenty single sideband channels are available at any one time. The limited bandwidth, however, means that only SSB and CW should be used because the bandwidth required by AM and FM stations is too great.

A beacon transmitter is also included which operates at a frequency of 145.95 MHz. It is provided for plotting purposes - and to find the translator after it has landed. The output power of the beacon is 3 mW; this means that the signal will be about 20 dB to 30 dB above the noise level during line-of-sight conditions when an effective receive bandwidth of 3 kHz is assumed. The signal will fall into the noise as soon as the balloon disappears below the horizon. This means that reception of the beacon transmitter will be a good indication that communication is possible.

The peak output power of the translator is 300 mW. If line-of-sight and a receive antenna having 10 dB gain are assumed, this will correspond to a signal-to-noise ratio of 40 dB to 50 dB at a distance of 200 to 650 km. The transmit energy that runs along the surface of the earth is attenuated by about 30 dB, but will still be audible. This means that the transposed band will be audible even when line-of-sight conditions do not exist, e.g. when the balloon is below the horizon.

3. A TYPICAL TRANSLATOR

The following description deals with the problems involved in the development of a balloon carried translator; the block and circuit diagrams of the translator inquestion are given in Figures 4 and 5.

The receiver is a single-conversion superhet with an intermediate frequency of 27 MHz. This frequency was chosen because it is one of the very few intermediate frequencies where spurious signals and unwanted conversion products cannot appear in the passband range of the receiver. Since the receive band is only spaced about 1.8 MHz from the transmit band, it is necessary to place a filter in the antenna input. This filter must suppress the transmit energy - which appears at the receiver input inspite of the ring filter - to a value that will not cause any cross modulation. A crystal filter having a centre frequency of 144.1 MHz, an insertion loss of 5 dB and a stopband attenuation of approx. 45 dB was chosen in our example.

The transmit oscillator signal must also not be allowed to inject itself into the receive mixer. If this were to occur, the transmit frequency range would be transposed into the IF range and self-oscillation could take place.

The receive signal is passed via the previously mentioned crystal filter from the antenna to the RF-amplifier stage where it is amplified and fed to the subsequent mixer stage. The 144.1 MHz signal frequency is converted to the intermediate frequency of 27 MHz with the aid of the carrier frequency obtained from the local oscillator. The bandwidth of the IF signal is then limited to a bandwidth of 40 kHz in a crystal filter and the signal amplified by 70 dB in a three-stage IF amplifier. This is followed by a further crystal filter which has been provided for a specific purpose: Since the IF amplifier has a bandwidth

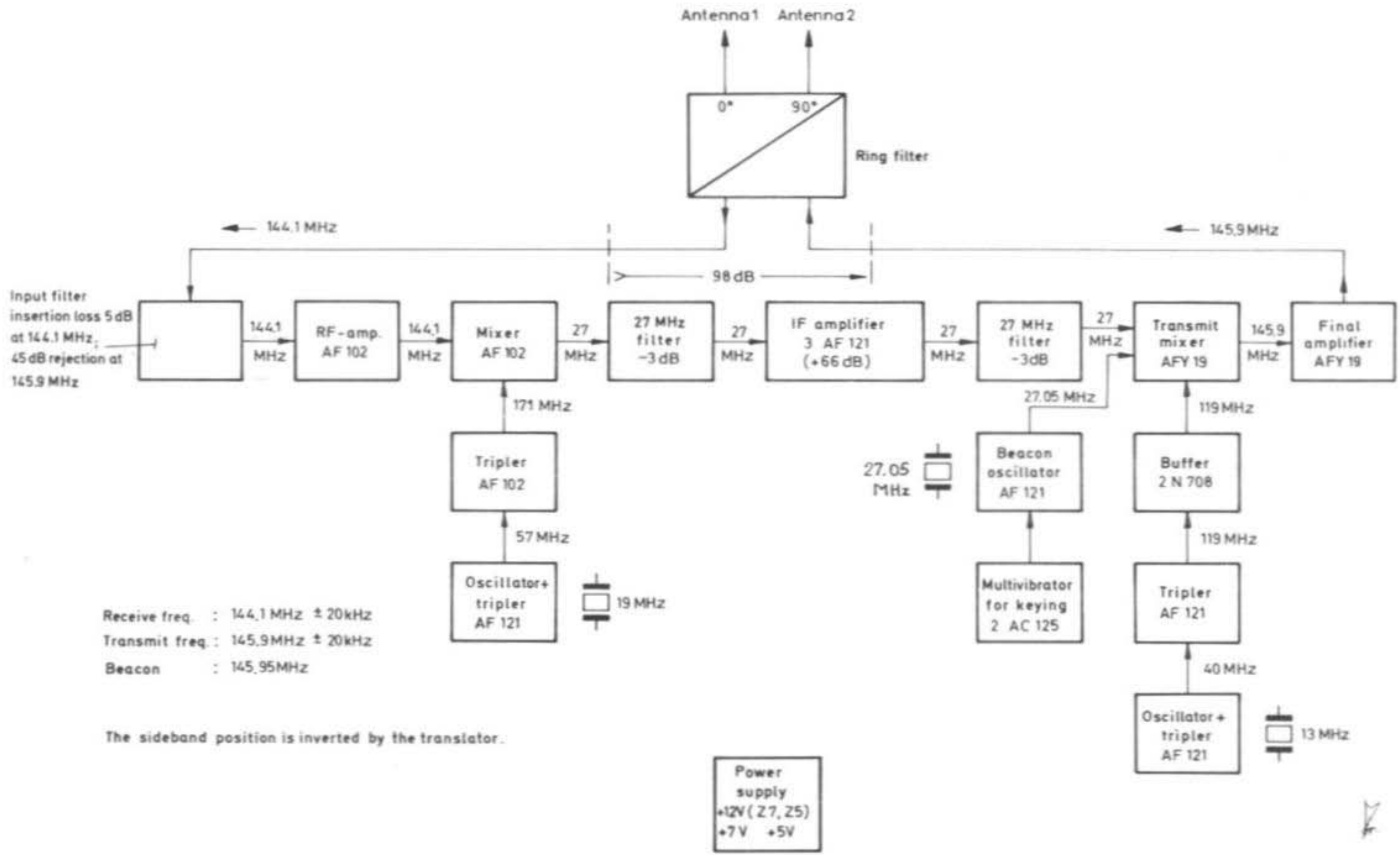


Fig. 4 Block diagram of the translator

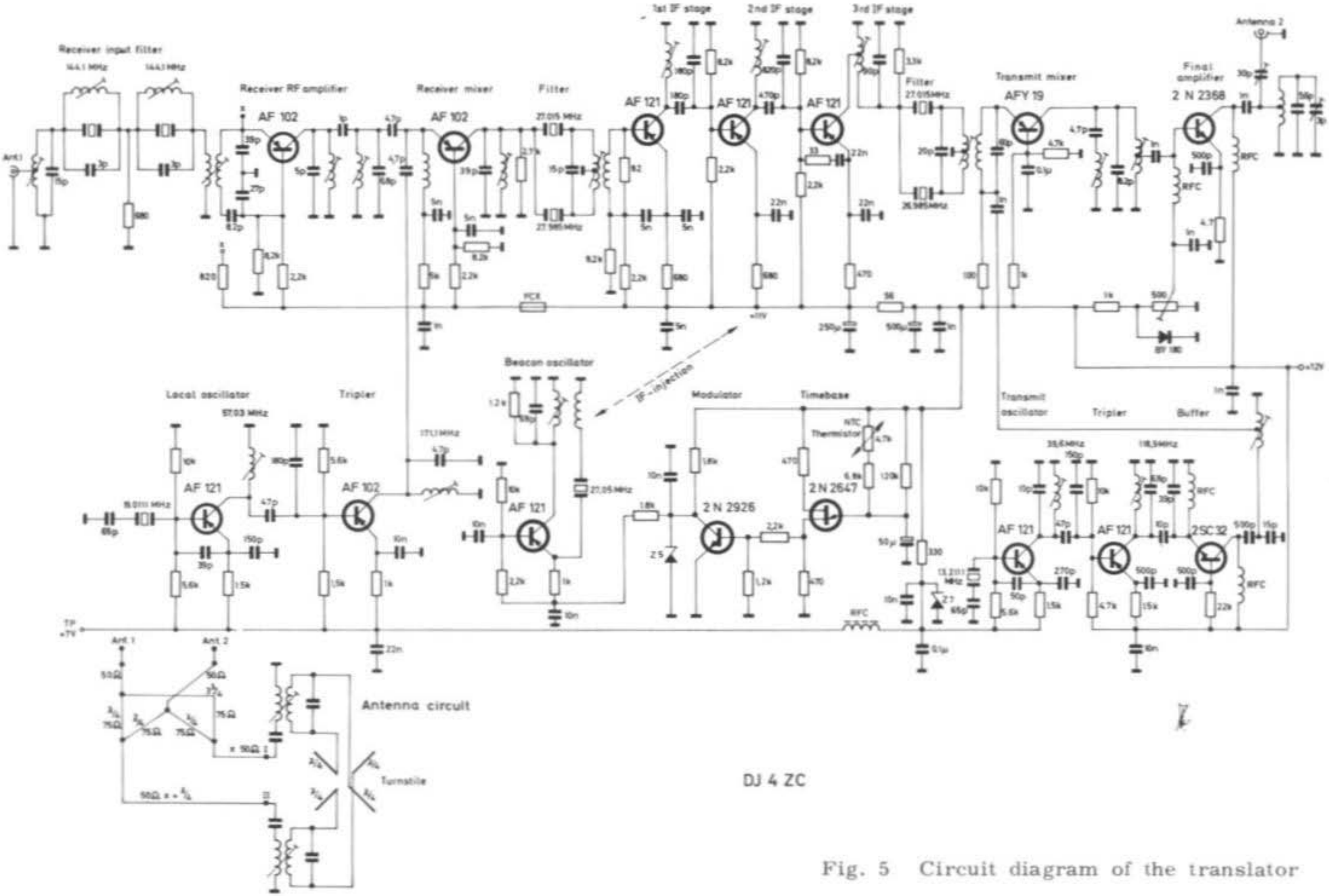


Fig. 5 Circuit diagram of the translator

of 4 MHz, the gain at a frequency spaced 1.8 MHz from the centre frequency will not be considerably less. The IF amplifier - as all amplifiers - possesses a certain intrinsic noise. This means that any noise component spaced 1.8 MHz from the centre frequency will be transposed into the receive band by the transmit mixer. This would result in the receiver being blocked by this noise. The task of the second crystal filter in the IF chain is to suppress these noise components to a level that will not interfere with the receiver.

The IF signal at the output of this crystal filter is converted with the aid of a 119 MHz auxiliary frequency to 145.9 MHz and amplified to an output power of 300 mW in the power amplifier stage. The same fundamental principles were used during the development of this transmit mixer as for a conventional single sideband transmitter. The bandwidth of the system is great enough that no noticeable fall-off of the conversion noise could be noticed at 144.1 MHz, the receive frequency. One of the most difficult problems involved with this translator was to find a suitable transmit mixer - power amplifier combination whose noise power in the receive band was at least 120 dB down on the output power. This problem was solved by suitable selection of the operating point and by use of an extremely low conversion gain.

The oscillator for the beacon signal operates at 27.05 MHz and is fed to the transmit mixer in the same manner as the IF signal.

The antenna is a horizontal turnstile configuration which is fed via a ring filter. When viewed from below, the transmit signal will be found to transmit clockwise circular polarized waves, whereas the receive antenna receives anticlockwise circular polarized signals. This allows an isolation of 20 dB to be obtained between the transmit and receive connection of the antenna array. The antenna arrangement has virtually omnidirection characteristic for horizontally polarized waves.

Further problems that had to be solved were caused by the environmental conditions under which the equipment was to operate. The translator is, for instance, required to operate correctly at temperatures down to -50° C and under the low pressure conditions encountered at such high altitudes. Most of the components are usually cooled by convection. Due to the lower atmospheric pressure at high altitudes, the cooling effect of convection will be greatly reduced and the warmth is only dissipated by conduction and radiation.

The completed translator weighs 600 grammes, the weight of the whole array including translator, antennas, ring filter, batteries, thermo-insulation and parachute is 1.4 kilogrammes.

4. GENERAL

Two different groups are active in Western Germany: These are the ARTOB (Amateur Radio Translator on Balloon) group in the Hannover area and their Bavarian collegues BARTOB.

Several different projects are in use or planned including cross-band translators 432/144 MHz and even 1296/144 MHz. The translator described in Section 3. is only given as an example of many different configurations that could possibly be used. It is hoped that this description will lead to similar ventures in other countries.

Editors Note:
The coordinators of the ARTOB and BARTOB groups are willing to assist other groups contemplating similar ventures. Please address any correspondence to either:

BARTOB Group
Mr. Bernd Helmke, DL 7 HR
8051 Mintrach ing 109/Bavaria
West Germany

ARTOB Group
Mr. Fritz Herbst, DL 3 YBA
3161 Burgdorf/Hannover
Marktstr. 64
West Germany

A TEN WATT TRANSMITTER FOR 70 cm

by H.J. Franke, DK 1 PN

INTRODUCTION

The following 70 cm transmitter was conceived to extend a transistorized two metre exciter such as described in (1) in the simplest possible manner for operation on the 70 cm band. The most favourable concept with respect to simplicity and available output power was found to be a two-stage circuit comprising a varactor tripler and a power amplifier using an EC 8020 tube. This transmitter, which is shown in Figure 1, possesses the following features:

The transistor transmitter (1) provides more than enough power (2 W at 12 V) to fully drive the 70 cm portion.

The output power of the EC 8020 power amplifier amounts to 10 W at a DC input power of only 14 W. This represents an efficiency of over 70%. These values are valid for the peak-power alignment which is used in the telegraphy and frequency modulated mode.

The output power is sufficient to fully-drive a linear amplifier equipped with a 4 x 150.

For amplitude modulation, it is merely necessary to re-align the output coupling link; the output power is then 6 to 7 W.

It is possible for the varactor tripler to be built-up separately so that it may be used for portable operation.

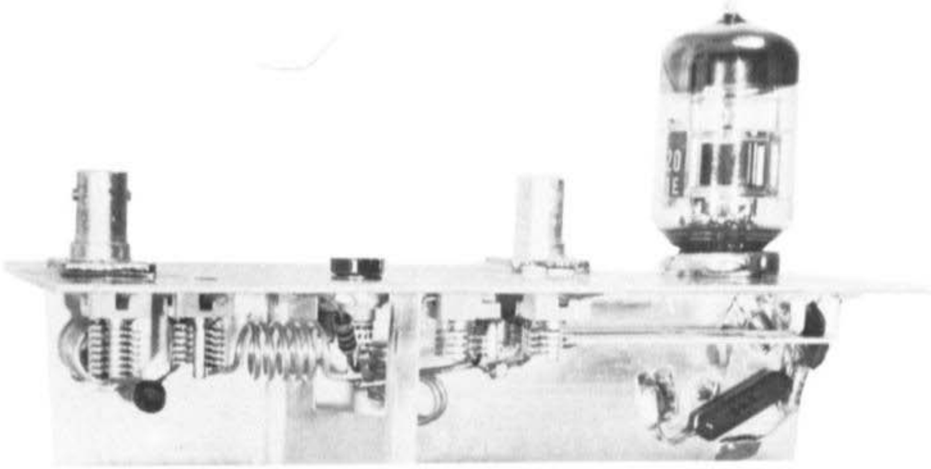


Fig. 1 10 W transmitter for 70 cm

2. CIRCUIT DESCRIPTION

The circuit diagram of the two-stage transmitter is given in Figure 2. The varactor is equipped with a parallel resonant circuit and π-filter for 145 MHz at the input; the output circuit consists of a strip line bandpass filter. The 288 MHz idler-circuit and resistor R 1 for the varactor bias voltage complete the varactor circuit. Whereas the 432 MHz resonant circuits of the varactor output filter are built-up as $\lambda/4$ strip lines, the input and idler circuit comprise coils. The value of R 1 (56 kΩ) represents a compromise between high efficiency and good linearity.

The tubed power amplifier stage operates with the triode EC 8020 in a grounded-grid configuration. The cathode is coupled to the secondary of the 432 MHz bandpass filter. A negative grid bias is generated across the capacitively-bridged cathode resistor R 2 which limits the anode current to approximately 30 mA under non-drive conditions. The anode circuit consists of a $\lambda/2$ resonant line. The output power is coupled out using a coupling link whose inductivity is compensated for with the aid of a trimmer capacitor.

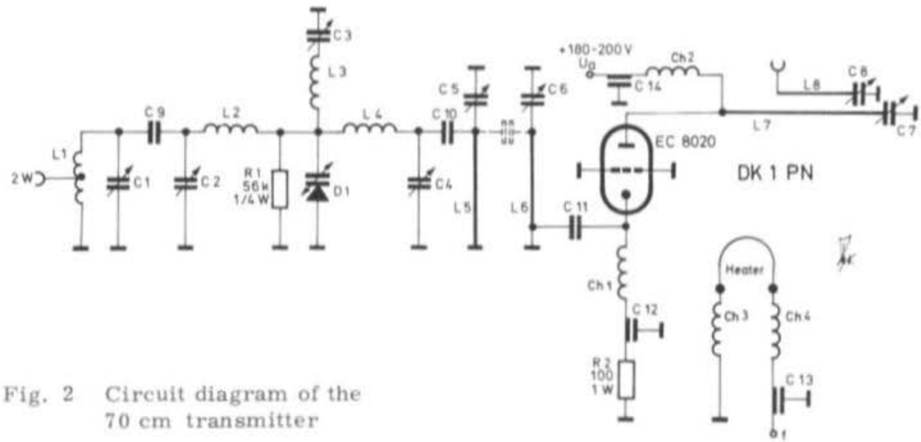


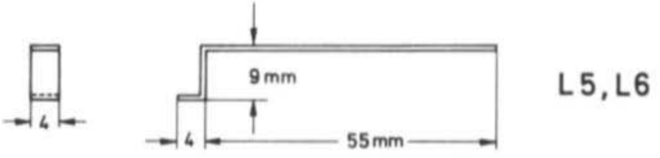
Fig. 2 Circuit diagram of the 70 cm transmitter

2.1. THE TRIODE EC 8020

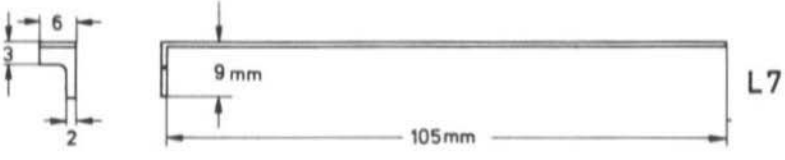
This tube has been especially developed for UHF television transposers. Since the tube exhibits a minimum feedback admittance at approximately 435 MHz, its stability is very good on the 70 cm band. The grid wires have a diameter of only 6 μm and the grid-cathode spacing is only 35 μm (grid-anode spacing: 350 μm). This results in a very high transconductance (60 mA/V at an anode current of 40 mA) and a very short electron path which allows this tube to be used at relatively high frequencies. However, the thin grid leads are not able to cope with high input power levels. The tube should therefore not be driven with much more than the required drive power. For the same reason, the tube should not be keyed by breaking the cathode lead in the grounded-grid mode. The very low power impedance value of this tube (approx. 200 Ω) allows a high output power and high efficiency. To achieve this, the tube must be driven into the grid current region, which required a drive power of approx. 0.5 W.

2.2. SPECIAL COMPONENTS

PA tube :	EC 8020 (AEG-Telefunken)
Varactor :	MA-4061 B (Microwave Associates) or BAX 11 (AEG-Telefunken)
C 1, C 2, C 8 :	2.0 - 13 pF (Tronser type 110 2070 013) air spaced
C 3 - C 7 :	1.7 - 6 pF (Tronser type 110 2070 005) trimmer
C 9 :	2.2 pF ceramic disc capacitor
C 10 :	0.5 pF " " "
C 11 :	40 pF " " " (Value uncritical)
C 12 :	1 nF " " "
C 13, C 14 :	1 nF feedthrough capacitors
R 1 :	56 kΩ, 1/4 W low-inductive composition resistor
R 2 :	100 Ω, 1 W
L 1 :	5 turns of 1.5 mm dia. (15 AWG) silver-plated copper wire wound on an 8 mm former, self-supporting. Coil length 14 mm with centre tap.
L 2 :	6 turns wire and former as L 1, coil length 16 mm.
L 3 :	4 turns wire as L 1, 7 mm former, coil length 11 mm.
L 4 :	2 turns as L 3, coil length 7 mm.
L 5, L 6 :	Brass strip 1 mm thick, 4 mm wide, 70 mm long, bent according to following diagram:



L 7 : Brass strip 1 mm thick, 6 mm wide, 115 mm long, bent according to following diagram. Bent end must be filed down to 2 mm width.



L 8 :	Coupling loop 50 mm long with a 2 mm spacing to L 7. 1 mm dia. (18 AWG) silver-plated copper wire.
Ch. :	10 turns of 0.5 mm dia. (24 AWG) enamelled copper wire close-wound on a 5 mm former, self supporting.

3. MECHANICAL ASSEMBLY

The brass chassis is drilled according to Fig. 3, bent and soldered together, after which the tube socket is soldered into place. A special UHF socket should be used which is characterized by its very thin contact plates and correspondingly short contact springs (The contact plates can be pertinax). The use of the

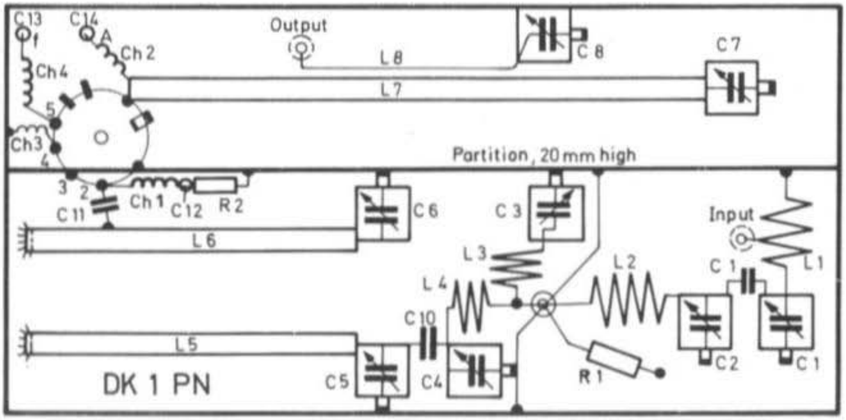


Fig. 3 Mechanical assembly of the 70 cm transmitter (from below)

UHF socket allows the screening plate between cathode and anode circuits to be mounted practically on the tube base. The grid connections are soldered to the chassis whilst the latter is still warm from the assembly.

Since the EC 8020 tube is wider than the tapped collar for the screening can, the collar is sawn off 1 mm over the socket base. The remaining 1 mm should be pressed inwards so that the various parts of the tube socket are held together.

The next process is to solder the screening plate and the feedthrough elements into place.

This completes the main mechanical assembly and it is now possible to install the trimmer capacitors. The author recommends that the trimmer capacitors are glued into place using a dual-component adhesive (such as Araldit or UHU-plus) since the ceramic portion of the trimmer breaks easily when using screws. Before gluing, both surfaces should be grease-free; the plates can be slightly roughend. Of course, one can also use tubular trimmers of the same capacity, which should be soldered into place.

After the adhesive has hardend, it is only necessary to solder the resonant lines, inductances and remaining components. The transmitter is now ready for alignment. A photograph showing the completed transmitter from below is shown in Fig. 4.

4. ALIGNMENT

Besides the requirements for a two metre exciter with an output power of 2 W (carrier), a selective power meter at 432 MHz is needed. It is very advantageous to use a reflectometer and a 70 cm antenna. Of course, the reflectometer and antenna should, in the amateur sense, operate correctly at 432 MHz. As long as the standing-wave ratio remains at its normal low level, one can assume that the indicated power is at the correct frequency. However, a 432 MHz low-pass or bandpass filter would simplify the alignment procedure.

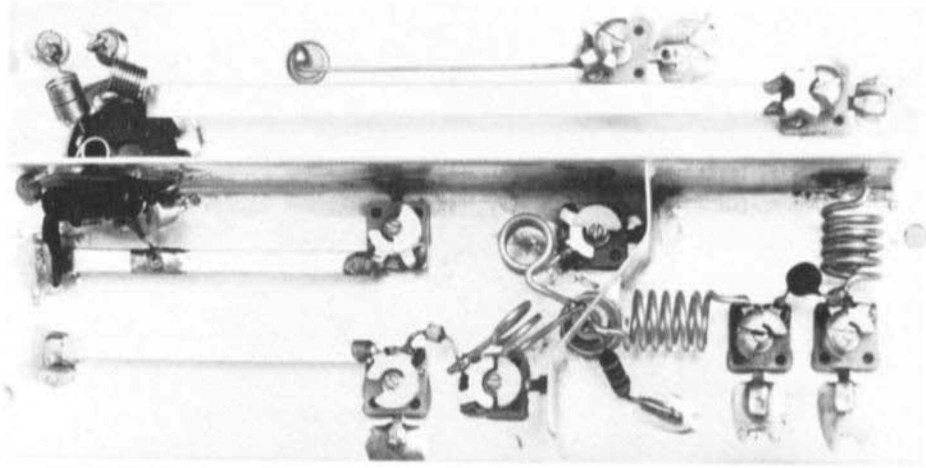


Fig. 4 Photograph of the 70 cm transmitter from below

4.1. ALIGNMENT OF THE VARACTOR TRIPLER

Unsolder trimmer capacitor C 11 from the resonant line L 6 and connect the coaxial cable of the UHF power meter in its place. The tripler is now fed with 2 watts of RF power and aligned commencing at the output. The trimmers are aligned in the following order for maximum power reading: C 5, C 6, C 3, C 4 (nearly minimum capacity), C 1 and C 2. This alignment should be repeated several times and optimized so that all trimmers can be varied slightly without causing the output power to jump suddenly. After this, the coaxial cable is removed and trimmer capacitor C 11 re-connected.

4.2. ALIGNMENT OF THE POWER AMPLIFIER

The transmitter output is now connected to the power meter after which the heater voltage and anode voltage of 200 V can be connected. An anode current of approximately 30 mA should now flow. Under full-drive conditions, the anode current will increase to 90 or 100 mA. Trimmer capacitor C 8 is now adjusted for maximum capacity and C 7 adjusted for maximum output power.

This preliminary alignment ensures that the varactor tripler is loaded; trimmer capacitors C 6 and C 5 can now be finally adjusted for maximum anode current.

Here a note regarding the meter: It has been found that a number of the "inexpensive" multimeters are sensitive to RF injection and thus indicate values that are too high. The meter can be checked by touching the connection leads and the meter itself to see if the indicated value alters. If this is the case, RF injection will be present.

The power amplifier is now adjusted for maximum output power by increasing or reducing the spacing between the coupling loop L 8 and the resonant line L 7, aligning at the same time trimmer capacitors C 7 and C 8. After completing this alignment, an output power of 9 to 10 watts should be measured at an anode current of approx. 70 mA.

4.3. AM ADJUSTMENT

If the transmitter is to be anode modulated, it will be necessary to provide the PA tube with a higher load impedance. This is achieved by varying the coupling loop L 8. A measure of the correct adjustment is when the anode current of the fully aligned transmitter is 30% less than at the peak-power setting, i.e. approx. 45 to 50 mA.

An audio power of about 5 W is required to modulate the transmitter which exhibits an impedance of approx. 4 kΩ to the modulator.

4.4. NOTES

Since the drive power is passed through the power amplifier in a grounded grid configuration and only the actual power generated in the PA is modulated, it will be impossible to achieve 100% modulation (this would require modulating the driver). However, since the power passed through the PA is only 0.5 W (7 to 8% of the output power), a modulation depth of about 92% will be achieved.

Finally a note regarding the drive power:
The transmitter was designed to operate with a power of 2 W at 144 MHz. At an efficiency of 50%, the varactor multiplier will provide 1 W of drive power for the PA tube. This means that sufficient power is available to allow an uncritical assembly. If less than 2 W of drive power are available at 144 MHz, it will be necessary to find the most favourable connection point on the resonant line L 6 for capacitor C 11. If necessary, a trimmer capacitor can be used for C 11.

5. REFERENCES

E. Flügel: The 2 metre Transmitter UTS 5 with 2 Watts mean output at an Operating Voltage of 12 V
VHF COMMUNICATIONS 1 (1969), Edition 3, Pages 179-187.

REVALUATION
The revaluation of the German Mark will unfortunately cause most of the national subscription rates and material prices to be increased in spite of the fact that the prices remain at their old DM value. However, we hope that our publications and material will be worth the extra burden thus caused. We profit in no way whatsoever from the revaluation.

DELAYS
Unfortunately delays are still being encountered, especially with material. This is out of our control since it is mainly caused by the post and customs authorities. Especially the British customs are causing us considerable difficulties and have seized a number of parcels.

PRICE LIST OF EQUIPMENT DESCRIBED IN VHF COMMUNICATIONS

Edition 1, 2, 3, and 4/1969

DM 3,60 = US$ 1.00 / DM 8,70 = £ 1.

All prices are given in West German Marks. The following prices do not contain post and packing for which an extra charge will be made: . . . . . . DM 1.50

The prices do not include any customs duty where applicable. All supplies having a value of over DM 80,-- (or less when requested) will be dispatched per registered mail and charged with: . . . . DM 1.--

Some delays may be caused at the moment due to delivery difficulties at the manufacturers. Equivalent semiconductor types will be supplied if original types are not available.

Semiconductors, quartz crystals and crystal filters can not be exchanged.

It is not possible for us to dispatch orders per C.O.D. All orders should be made cash with order including the extra charges for post and packing, registered mail etc. A transfer to one of our accounts or via our representatives is also possible.

Any items (such as handbooks) which include post and packing are correspondingly annotated.

PRINTED CIRCUIT BOARDS
(Glass fibre epoxy with silver-plated conductors)

2 metre FET converter	DL 6 SW 004	DM	6.--
Phase-locked oscillator	DJ 7 ZV 001 + 002	DM	11.--
24 cm converter, IF preamp.	DL 3 WR 001	DM	2.--
70 cm converter	DL 9 GU 001	DM	6.--
5W SSB transmitter	DJ 9 ZR 001	DM	15.--
V X O	DJ 9 ZR 002	DM	3.--
SSB receiver, IF/AF portion	DJ 9 ZR 005 +	DM	42.--
SSB receiver, VHF portion	DJ 9 ZR 006 ++	DM	43.50
2W AM transmitter	DJ 1 NB 004	DM	10.--
Electronic Fuse	DL 3 WR 002	DM	1.50
Bandpass filter	DJ 4 KH 001	DM	1.--
Active Audio Filters	DJ 4 BG 001	DM	3.--
28/144 MHz transverter	DJ 6 ZZ 001	DM	15.--

\+ Double coated with through contacts

++ Teflon base material with printed inductances

COIL FORMER SETS for the above mentioned PC-boards:

DL 6 SW 004	DM	1.70
DJ 7 ZV 001 + 002	DM	3.20
DL 9 GU 001	DM	-.70
DJ 9 ZR 001	DM	2.50
DJ 9 ZR 002	DM	1.40
DJ 9 ZR 005	DM	-.35
DJ 9 ZR 006	DM	-.35
DJ 6 ZZ 001	DM	2.10

TRIMMER SETS for the above mentioned PC-boards:

DL 6 SW 004	DM	3.50
DL 9 GU 001	DM	1.40
DJ 9 ZR 006	DM	10.40
DJ 1 NB 004	DM	3.10
DJ 6 ZZ 001	DM	6.30
DJ 4 KH 001	DM	6.--

1.3 mm diameter special drill for PC-boards	DM 3.--
This special drill has been specially developed for epoxy PC-boards. It will be supplied free-of-charge on orders amounting to more than DM 30,--.	

Potted core with former and holder for DJ 9 ZR DC/DC-converter	DM 15.--

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METAL CASINGS (Resonant lines), silver plated

70 cm converter DL 9 GU 001, complete with three feed-through capacitors, three trimmers, inner conductors soldered into place and cover	DM 65.--
24 cm converter DL 3 WR 001, partitions, trimmers and inner conductors installed	DM 127.40

COMPLETED EQUIPMENT (ready to operate)

70 cm converter DL 9 GU 001	DM 196.50

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145 MHz HB 9 CV ANTENNA, chroma-plated, detachable including postage	DM 40.--

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K I T S

2 m FET converter DL 6 SW 004	DM 42.60
(PC-board, coil formers, trimmers, transistors, 38.6667 MHz quartz crystal and 1.3 mm special drill)	
Transistors only	DM 17.90
Quartz crystal 38.6667 MHz (HC-6/U) only	DM 13.70
70 MHz FET converter DL 6 SW / G 3 JHM	DM 44.50
(PC-board, coil formers, trimmers, transistors 42.0000 MHz qaurtz crystal and 1.3 mm special drill)	
Transistors only	DM 17.90
Quartz crystal 42.0000 MHz (HC-6/U) only	DM 15.60
Phase-locked oscillator DJ 7 ZV 001 + 002	DM 103.95
(2 PC-boards, coil formers, transistors, diodes, 1 quartz crystal and 1.3 mm special drill)	
Transistors and diodes only	DM 40.75
2 quartz crystals (46.0000 and 46.3333 MHz) (HC-18/U)	DM 49.--
24 cm converter DL 3 WR 001	Price on reques
70 cm converter DL 9 GU 001	DM 130.--
(Silver-plated metal casing as above, PC-board, coil formers, trimmers, transistors, 96.0000 MHz quartz crystal and 1.3 mm special drill)	
Transistors only	DM 37.40
Quartz crystal 96.0000 MHz (HC-6/U) only	DM 19.50
5 W SSB transmitter DJ 9 ZR 001	DM 138.40
(PC-board, coil formers, trimmers, Ferroxcube-Drosseln, transistors, diodes and 1.3 mm special drill)	
Crystal filter XF-9A with both sideband crystals	DM 96.--
Crystal filter XF-9B with both sideband crystals	DM 125.50
Transistors and diodes only	DM 116,90

VXO DJ 9 ZR 002	DM 18.45
(PC-board, coil formers, transistors, diodes)	
Transistors and diodes only	DM 14.05
Quartz crystal (45.478 MHz for SSB spot frequency 145.416 or other crystal in same range) (HC-18/U)	DM 24.50
DJ 9 ZR 005 SSB Receiver, IF/AF Portion for 9 MHz	DM 122.60
Double coated PC-board, coil formers, and semiconductors (3 x CA 3028A, 2 x CA 3020, 1 x BC 182A, 1 x 2 N 706, 6 x 1 N 914, 2 x 1 N 277) without crystal filter	
Semiconductors only	DM 80,15
Crystal filter XF-9A with both sideband crystals	DM 96,--
Crystal filter XF-9B with both sideband crystals	DM 125.50
DJ 9 ZR 006 SSB Receiver, VHF Portion 145/9 MHz	DM 90.--
Teflon (PTFE) PC-board, disc trimmers, air spaced trimmers, coil formers, 1.3 mm drill and semiconductors (1 x TA 7153, 1 x 3 N 140, 1 x BC 182A)	
Semiconductors only	DM 33.30
DL 3 WR 002 Electronic Fuse (24 V)	DM 11.85
(PC-board, semiconductors (1 x AC 117, 2 x AC 122, 1 x BZY 87)	
Semiconductors only	DM 10.35
DJ 6 ZZ 001 28/144 MHz Transverter	DM 80.20
Comprising PC-board, coil former set, trimmer capacitor set and 38.6667 MHz crystal in HC-6/U holder as well as 1.3 mm special drill and semiconductors	
Semiconductors only	DM 43.50
Crystal only	DM 13.70
DJ 1 NB 004 2W AM transmitter	
Comprising PC-board, trimmer capacitor set, Ferrox cube chokes and semiconductors	
Semiconductors only	DM 60.20
Crystal to order. Please give frequency (Delivery 4 weeks)	DM 19.50
9 MHz Crystal Filters from KVG (see rear cover)	
XF-9A with both sideband crystals	DM 96.--
XF-9B with both sideband crystals	DM 125.50
XF-9E	DM 125.--
RADIO COMMUNICATION HANDBOOK (including postage)	DM 35.--
VHF-UHF-MANUAL (including postage)	DM 13.40

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INDEX TO VOLUME 1 OF VHF COMMUNICATIONS 1969

1. TECHNICAL DESCRIPTIONS AND ASSEMBLY INSTRUCTIONS

1.1. VHF RECEIVERS AND CONVERTERS

A 2 Metre Converter with Field Effect Transistors	W. v. Schimmelmann DL 6 SW	Ed. 1	P. 2-10
A Modern Concept for Portable 2 Metre Receivers	D. E. Schmitzer DJ 4 BG	Ed. 2	P. 115-122
A Field Effect Transistorized Converter for 70 MHz	D. Hayter G 3 JHM	Ed. 2	P. 123
A 145 MHz/9 MHz Receive Converter using Printed Inductances	K. P. Timmann DJ 9 ZR	Ed. 3	P. 129-135
A 9 MHz IF-AF Portion using Integrated Circuits	K. P. Timmann DJ 9 ZR	Ed. 3	P. 136-150 158-159
An Automatic Search Oscillator for Two Metre Converters	G. Hoffschildt DL 9 FX	Ed. 4	P. 215-217

1.2. VHF EXCITERS, TRANSMITTERS AND POWER AMPLIFIERS

A 144 MHz Adapter for use with a 10 Metre Transceiver	E. Krahe DL 9 GU	Ed. 1	P. 54-61
A 5 Watt Transistorized SSB Transmitter for 145 MHz	K. P. Timmann DJ 9 ZR	Ed. 2	P. 73-82
The Two Metre Transmitter UTS 5 with 2 Watts Mean Output at an Operating Voltage of 12 V	E. Flügel DJ 1 NB	Ed. 3	P. 179-187

1.3. UHF RECEIVERS AND CONVERTERS

A Solid-State Converter for 24 cm	R. Lentz DL 3 WR	Ed. 1	P. 36-53
144 MHz/432 MHz Transverter for Low Power and Field Day Applications	L. Wagner DL 9 JU	Ed. 1	P. 31-35
A 432 MHz/144 MHz Converter with Silicon Transistor Complement	E. Krahe DL 9 GU	Ed. 2	P. 65-72

1.4. UHF EXCITERS AND TRANSMITTERS

144 MHz/432 MHz Transverter for Low Power and Field Day Applications	L. Wagner DL 9 JU	Ed. 2	P. 31-35
A Ten Watt Transmitter for 70 cm	H. J. Franke DK 1 PN	Ed. 4	P. 243-248

1.5. COMPLETE VHF TRANSCEIVERS

A 144 MHz SSB Transceiver comprising:

144 MHz/9 MHz Receive Converter using Printed Inductances	K. P. Timmann DJ 9 ZR	Ed. 3	P. 129-135
A 9 MHz IF-AF Portion using Integrated Circuits	K. P. Timmann DJ 9 ZR	Ed. 3	P. 136-150 158-159
A 5 Watt Transistorized SSB Transceiver for 145 MHz	K. P. Timmann DJ 9 ZR	Ed. 2	P. 73-82

1.6. TRANSMIT-RECEIVE CONVERTERS (TRANSVERTERS) FOR VHF AND UHF

144 MHz/432 MHz Transverter for Low Power and Field Day Applications	L. Wagner DL 9 JU	Ed. 1	P. 31-35
A 144 MHz Adapter for use with a 10 Metre SSB Transceiver	E. Krahe DL 9 GU	Ed. 1	P. 54-61
A 28 MHz/144 MHz Transistorized Transverter	F. Weingärtner DJ 6 ZZ	Ed. 4	P. 189-195
A Balloon-Carried Translator	K. Meinzer DJ 4 ZC	Ed. 4	P. 236-242

1.7. VARIABLE FREQUENCY OSCILLATORS

A Phase-Locked Oscillator for Transmit and Receive Mixers in Amateur Radio Equipment	K. P. Timmann, DJ 9 ZR and V. Thun, DJ 7 ZV	Ed. 1	P. 11-25
Regarding the Phase-Locked Oscillator	G. Loebell DJ 6 AH	Ed. 2	P. 85-86
A Variable Frequency Crystal Oscillator (VXO)	K. P. Timmann DJ 9 ZR	Ed. 2	P. 87-94
A Three-Stage VFO for 48.0 to 48.7 MHz	G. Hoffschildt DL 9 FX	Ed. 4	P. 209-214

2. ANTENNAS AND ANTENNA ACCESSORIES
2.1. ANTENNAS

The HB 9 CV Antenna for VHF and UHF	H.J. Franke DK 1 PN	Ed. 1	P. 26-30
Determining the Impedance of Rod Antennas in the VHF Range	H.J. Dohlus DJ 3 QC	Ed. 2	P. 98-109
Determining the Impedance of Quarter Wave Groundplate Antennas	H.J. Dohlus DJ 3 QC	Ed. 3	P. 160-168

2.2. ANTENNA ACCESSORIES

A Coaxial Relay with a High Coupling Attenuation and Good SWR	E. Beberrich DL 8 ZX	Ed. 2	P. 124-125
A Bandpass Filter for 145 MHz using Printed Inductances	K. Maiwald DJ 4 KH	Ed. 4	P. 205-208

3. MODULATION

Preamplifiers to improve Speech Intelligibility under Poor Operating Conditions	D.E. Schmitzer DJ 4 BG	Ed. 2	P. 110-114

4. FILTERS

A Bandpass Filter for 145 MHz using Printed Inductances	K. Maiwald DJ 4 KH	Ed. 4	P. 205-208
Active Audio Filters Part I (Theoretical)	D.E. Schmitzer DJ 4 BG	Ed. 4	P. 218-225
Active Audio Filters Part II (Practical)	D.E. Schmitzer DJ 4 BG	Ed. 4	P. 226-235

5. MEASURING TECHNOLOGY

A Calibrated Attenuator	J. Wasmus, DJ 4 AU and G. Laufs, DL 6 HA	Ed. 3	P. 169-173

6. OTHER DESCRIPTIONS

A 12 V/24 V DC-DC Converter	K.P. Timmann DJ 9 ZR	Ed. 2	P. 83-84
The Effect of the Printed Circuit Base Material on the Q of Printed Inductances	K.P. Timmann DJ 9 ZR	Ed. 3	P. 127-128
Linear Integrated Circuits for Amateur Applications Part I	D.E. Schmitzer DJ 4 BG	Ed. 3	P. 151-157
Linear Integrated Circuits for Amateur Applications Part II	D.E. Schmitzer DJ 4 BG	Ed. 4	P. 196-204
A Simple Electronic Fuse	R. Lentz DL 3 WR	Ed. 3	P. 174-178

All Editions contained Advertisements, Material Lists and Short Notices

VHF COMMUNICATIONS

A PUBLICATION FOR THE RADIO AMATEUR
ESPECIALLY COVERING VHF, UHF AND MICROWAVES

VERLAG UKW-BERICHTE, HANS J. DOHLUS oHG, DJ 3 QC, D-8520 ERLANGEN
Gleiwitzer Strasse 45 - Telefon (0 91 31) 3 33 23 + 6 33 88

Konten: Postscheckkonto Nürnberg 30 455
Deutsche Bank Erlangen 476 325
Stadtsparkasse Erlangen 314-620 956
Österreichisches Postsparkassenamt Wien 116 914