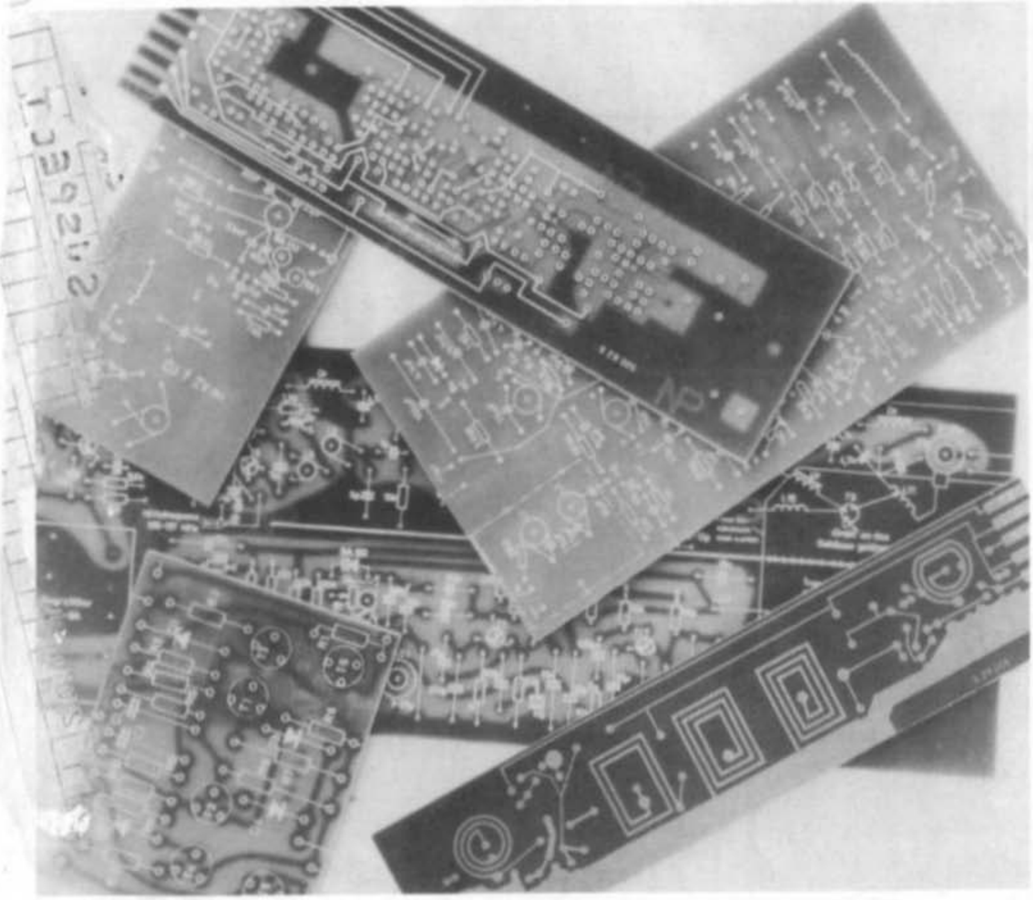
J 20 419 F

VHF COMMUNICATIONS

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

VOLUME NO. 1 EDITION 2 MAY 1969 US $ 1.00



CONTENTS:

E. Krahé, DL 9 GU	432/144 MHz CONVERTER WITH SILICON TRANSISTOR COMPLEMENT	page 65—72 and 96—97
K.P. Timmann, DJ9ZR	A 5 WATT TRANSISTORIZED SSB TRANSMITTER FOR 145 MHz	page 73—82
K.P. Timmann, DJ9ZR	A 12 V / 24 V DC-DC CONVERTER	page 83—84
G. Loebell, DJ 6 AH	REGARDING THE DJ 7 ZV / DJ 9 ZR PHASE-LOCKED OSCILLATOR	page 85—86
K.P. Timmann, DJ9ZR	VARIABLE FREQUENCY CRYSTALL OSCILLATOR (VXO)	page 87—94
H. J. Dohlus, DJ 3 QC	DETERMINING THE IMPEDANCE OF ROD ANTENNAS IN THE VHF RANGE	page 98—109
E. Schmitzer, DJ 4 BG	PREAMPLIFIERS TO IMPROVE SPEECH INTELLIGIBILITY UNDER POOR OPERATING CONDITIONS	page 110—114
E. Schmitzer, DJ 4 BG	A MODERN CONCEPT FOR PORTABLE 2 METRE RECEIVERS	page 115—122
D. T. Hayter, G 3 JHM	THE 70 MHz DL 6 SW FET CONVERTER	page 123
E. Berberich, DL 8 ZX	A COAXIAL RELAY WITH A HIGH COUPLING ATTENUATION AN GOOD SWR	page 124—125

VHF COMMUNICATIONS, the international edition of the well-established German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. Published in February, May, August and November.

The subscription price is US $ 3.00 or national equivalent per year. Individual copies are available at $ 1.00, or equivalent, each.

Subscriptions, orders of individual copies, purchase of printed circuit boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representatives. It is important that any change of address be reported as soon as possible to ensure the correct and punctual arrival of the publication. Please give your address in block letters and make sure to place sufficient postage.

Publisher: VERLAG UKW-Berichte, Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Str. 45, Federal Republic of Germany, Tel. (0 91 31) 3 33 23

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

Printed in the Federal Republic of Germany by Richard Reichenbach KG, D-8500 Nuernberg, Krelingstraße 39.

We would be grateful if you would address your orders and queries to your national representative:

VERTRETUNGEN: **REPRESENTATIVES:**

Austria	Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Straße 45, see Germany
Australia	WIA, PO Box 67, East Melbourne 3002, Victoria
Belgium	E. Drieghe, ON 5 JK, B-9160 HAMME, Kapellestr. 10, Telefon (052) 49457, PCR 588111
Canada	see USA
Denmark	Sven Jacobson, SM 7 DTT, Ornbogatan 1, S-21232 MALMÖ, Tel. 49 16 93, Postkto. København 14985
France	Christiane Michel, F 5 SM, F-89 PARLY, les Pillés, CCP PARIS 16.219-66
Finland	Eero Valio, OH 2 NX, 04740 SÄLINKAA, Postgiro 4363 39-0 und Telefon 915/86 265
Germany	Verlag UKW-BERICHTE H. Dohlus oHG, D-8520 ERLANGEN, Gleiwitzer Str. 45, Telefon (09131) 3 33 23 + 6 33 88, Deutsche Bank Erlangen Kto. 476 325, Postscheck 304 55 Nürnberg
Holland	S. Hoogstraal, PA ø MSH, ALMELO, Oranjestraat 40, giro 137 2 282, telefon (05490) — 1 26 87
Italy	STE s.r.l. (I 2 GM) via maniago 15, I-20134 MILANO, Tel. 21 78 91, Conto Corrente Postale 3/44968
Luxembourg	P. Wantz, LX 1 CW, Télévision, DUDELANGE, Postscheckkonto 170 05
Norway	H. Theg, LA 4 YG, Ph. Pedersensv. 15, N 1324 LYSAKER pr. Oslo, Postgiro 31 6000
South Africa	Arthur Hemsley ZS 5 D, P.O. Box 64, POINT, Durban Tel. 31 27 27
New Zealand	E. M. Zimmermann, ZL 1 AGQ, P.O. Box 56, WELLSFORD, Tel: 80 24
Spain+Portugal	Julio A. Prieto Alonso, EA 4 CJ, MADRID-15, Donoso Cortès 58 5°-B, Tel. 243.83.84
Sweden	Sven Jacobson, SM 7 DTT, S-21232 MALMÖ, Ornbogatan 1, Tel. 491693, Postgiro 43 09 65
Switzerland	Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Straße 45, see Germany
United Kingdom	MICROWAVE MODULES Ltd., 4 Newling Way, WORTHING/SSX. Tel. 0903-64301
USA-East Coast	VHF COMMUNICATIONS Russ Pillsbury, K 2 TXB, & Gary Anderson, W 2 UCZ, 915 North Main St. JAMESTOWN, NY 14701, Tel. 716-664-6345
USA-Central	Bob Eide, WØENC, 53 St. Andrew, RAPID CITY, SD 57701, Tel. 605-342-4143
USA-West Coast	Darrel Thorpe, Circuit Specialists Co. Box 3047, SCOTTSDALE AZ 85257, Tel. 602-945-5437

UKW - BERICHTE
VHF
COMMUNICATIONS

VOLUME 1 MAY 1969 EDITION 2

PUBLISHER: VERLAG UKW-BERICHTE
Hans J. Dohlus, DJ 3 QC
D-8520 ERLANGEN
Gleiwitzer Strasse 45
Fed. Republic of Germany

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

432/144 MHz CONVERTER WITH SILICON TRANSISTOR COMPLEMENT

by E. Krahe, DL 9 GU

1. INTRODUCTION

The described 70 cm converter is equipped entirely with silicon transistors. It consists of two preamplifier stages with the transistor 2N 3478, a crystal oscillator followed by a tripler stage, as well as a mixer stage using a junction field-effect transistor (JFET). The two preamplifier stages are accommodated in a metal casing in a similar manner to the 70 cm preamplifier described in (1). The oscillator and the mixer stage are built up on a printed circuit board which is firmly connected to the metal casing (Fig. 1). This PC-board is a modified version of the DL 6 SW 2 metre converter (2). The converter operates from a 9 V supply (negative pole to ground); the current drain is approximately 18 mA.

The manufacturer of the 2N 3478 (RCA) lists a typical noise figure of 5 dB (at 470 MHz) for this RF-amplifier transistor. This means that a noise figure in this order can also be assumed for the converter.

2. DESCRIPTION OF THE CIRCUIT

The circuit diagram of the converter is given in Fig. 2. The two preamplifier stages operate in a common-base circuit. The emitter and collector are connected to the capacitively shortened quarter wave ($\lambda/4$) coaxial circuits. The DC operating voltages are fed to the transistors via 1 kΩ resistors. The tuning of the preamplifier stages is not critical and the bandwidth is wide enough to enable television signals to be received.

The signal is amplified by approximately 18 dB and is tapped off by means of a coupling loop from the last resonant line circuit. It is then fed via connection-point (4) in the shortest possible manner to the gate of the mixer transistor T 5. A resonant circuit (L 5, C 23) in the drain circuit filters out the 2 metre difference frequency which is fed via a coupling link at a low source impedance to the output connection point (2) and from there to the two metre receiver.

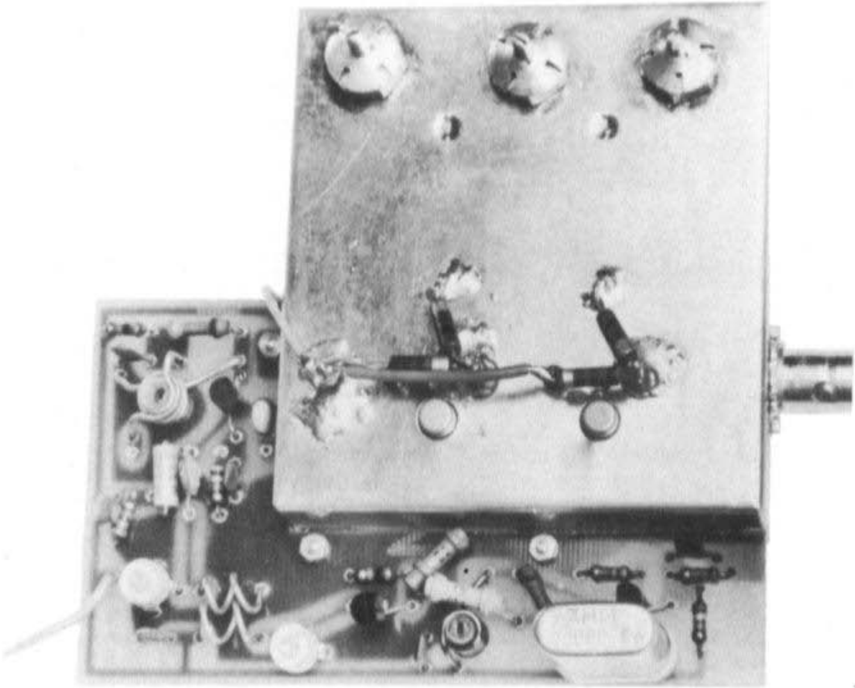


Fig. 1

The popular overtone oscillator circuit is equipped with the bipolar transistor T 3 in a common-base circuit and uses a 96 MHz quartz crystal. The capacitance of the crystal holder is neutralized by the inductivity of L 1.

The tripler transistor T 4 also operates in a common-base circuit. The emitter of this transistor is connected to a capacitive tapping point of the oscillator resonant circuit (L 2, C 13, C 14, C 15). The tripled oscillator frequency (288 MHz) is then fed via a capacitively tuned, unductively coupled bandpass filter (L 3/C 16, L 4/C 18) and an isolating capacitor (C 19) to the gate of the mixer transistor T 5.

2.1. COIL DATA

L 1 and L 7	=	0.3 μH choke or 15 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound on a 3 mm former, self-supporting.
L 2	=	7 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 4.3 mm dia. coil former with VHF core.
L 3 and L 4	=	2 turns, wire as for L 3, wound on to a 5 mm former, self-supporting, coil length approx. 10 mm.
L 5	=	6 turns, wire and coil former as for L 2, coil length approx. 8 mm.
L 6	=	1.75 turn of 0.5 mm dia. (24 AWG) enamelled cotton-covered wire placed between the last turns at the cold end of L 5.

Note: The choke L 1 is directly soldered on the conductor side of the board to the connection leads of the crystal, or crystal socket.

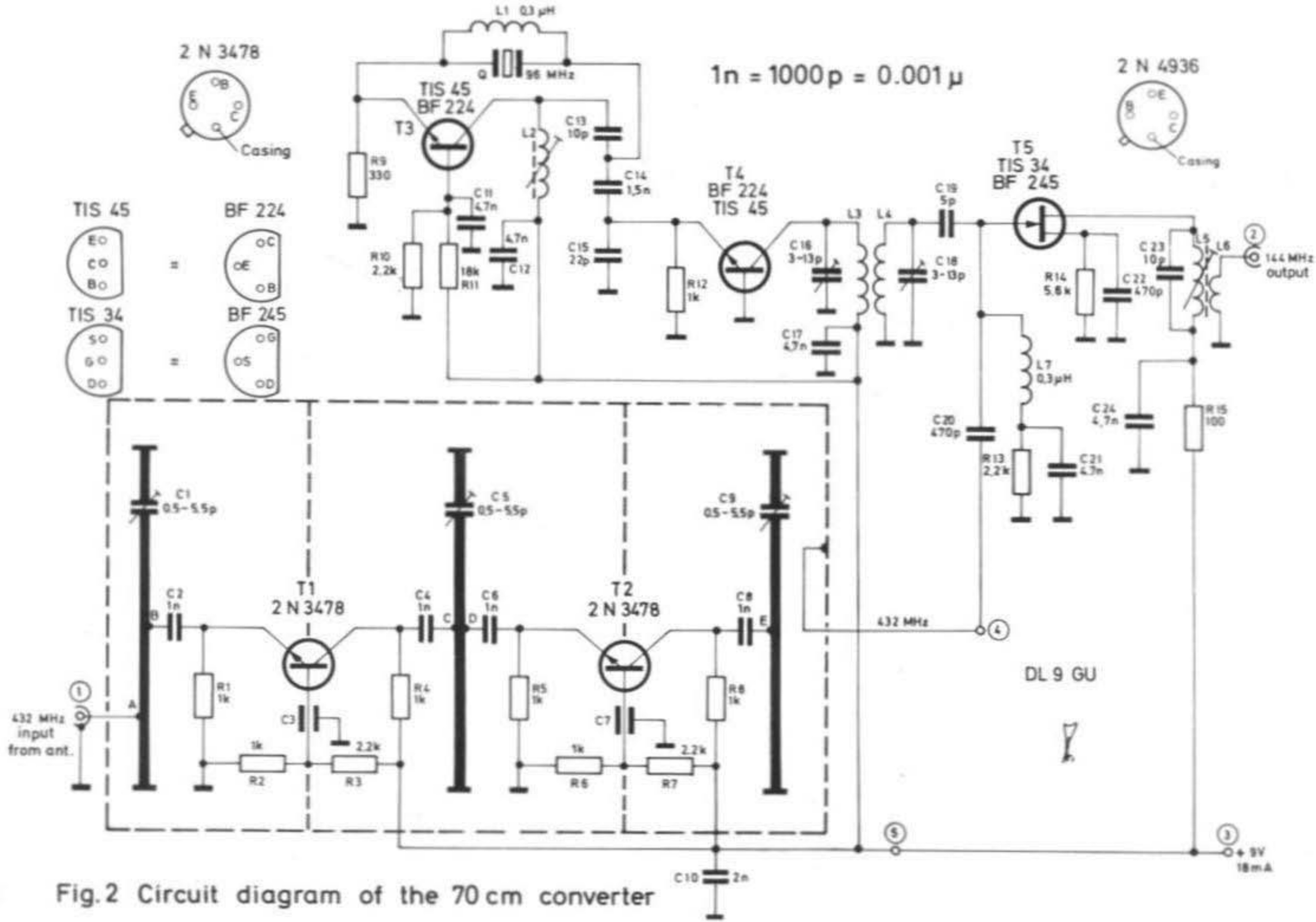


Fig. 2 Circuit diagram of the 70 cm converter

2.2. SPECIAL COMPONENTS

C 1, C 5, C 9	= tubular trimmers 0.5 to 5.5 pF e.g. JFD type PT 903 or Valvo (Philips) type C 004 EB / 6E
T 1, T 2	= 2 N 3478 (RCA) 2 N 4936
T 3, T 4	= BF 224 or TIS 45 (Texas Instruments)
T 5	= BF 245, TIS 34 or TIS 88 (Texas Instruments)

3. MECHANICAL ASSEMBLY

As already mentioned, the converter consists of two sub-assemblies. The metal casing for the two preamplifier stages and the L-shaped printed circuit board for the oscillator and mixer stage.

3.1. PREAMPLIFIER CASING

The assembly of the metal casing is made according to Fig. 3, where all essential dimensions are given. Before mounting the intermediate partitions, it is necessary for the 6 mm diameter cut-outs required for the accommodation of transistor T 1 and T 2 to be drilled into the base plate directly under each partition. Semi-circle cut-outs of approximately 7 mm dia are now filed into the partitions directly over each of the transistor positions; these enable the transistor connections to be made.

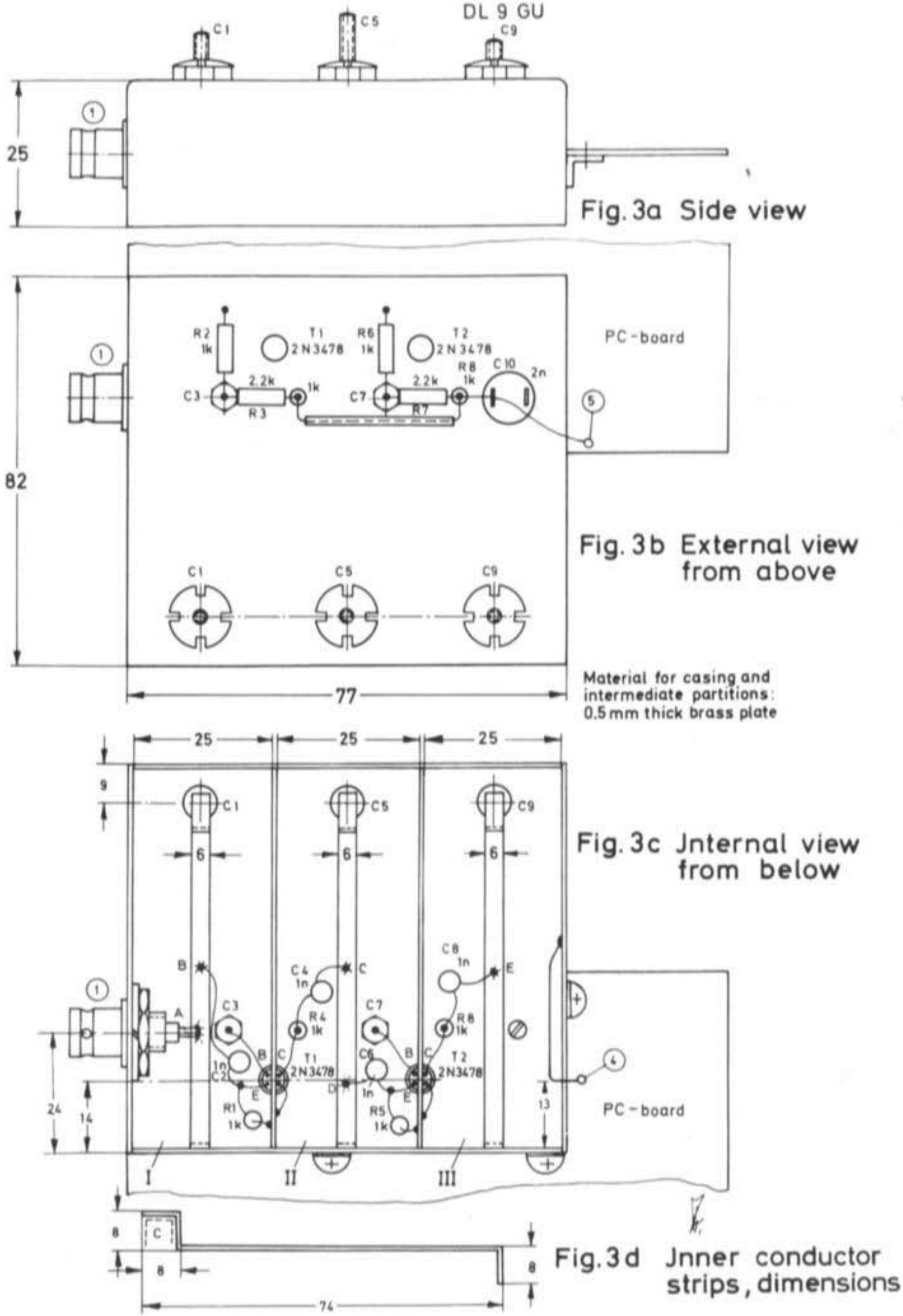
It is also necessary to drill holes in suitable positions (see Fig. 3b and 3c) for the two feed-through capacitors C 3 and C 7, as well as for the two 1/4 watt resistors R 4 and R 8. After the casing has been soldered and silver-plated, these resistors are vertically mounted into the prepared holes and are glued into place with a two-component glue (e.g. UHU-Plus) so that a third profrudes into the inner chamber. Transistors T 1 and T 2 are also glued in such a manner into the holes as to allow the socket to touch the base plate. The by-pass capacitor C 10 is to be found on the outside of the cabinet. This capacitor is either soldered or screwed into place according to its design.
The inner conductors of the three coaxial circuits are made by bending the 0.3 to 0.5 mm thick copper or brass plate as shown in Fig. 3d and 3c.

The tubular trimmer capacitors C 1, C 5 and C 9, which in the authors case protrude 22 mm, from the base plate of the converter, also serve as support for one end of the associated inner conductor; the other end being firmly soldered to the inside of the lower wall as shown in Fig. 3c. The inner conductors run in the centre axis of the chamber.

The mounting of the input socket , transistors, feed-throughs and inner conductors is followed by soldering the remaining resistors (R 1, R 2, R 3, R 5, R 6, R 7) and capacitors (C 2, C 4, C 6, C 8) into the positions shown in Figures 1 and 3. The spacing from connection points A to E to the lower wall shown in Fig. 3c is given in the following table:

Connection point	A (Input)	:	24 mm
" "	B (for C 2)	:	approx. 38 mm
" "	C (for C 4)	:	38 mm
" "	D (for C 6)	:	14 mm
" "	E (for C 8)	:	37 mm

Fig. 3 Mechanical assembly of the 70 cm converter



All dimensions in mm

The decoupling loop is made up from 1 mm diameter (18 AWG) silver-plated copper wire. The ground point for this loop is spaced 50 mm from the lower wall.

The open casing is covered after the alignment procedure with a 0.5 mm brass plate which is then soldered onto the casing at various points.

It is, of course, possible for the preamplifier described in (1) to be used in conjunction with this converter. If this is desired, it is merely necessary for the output socket to be removed and a decoupling loop to be inserted in its place.

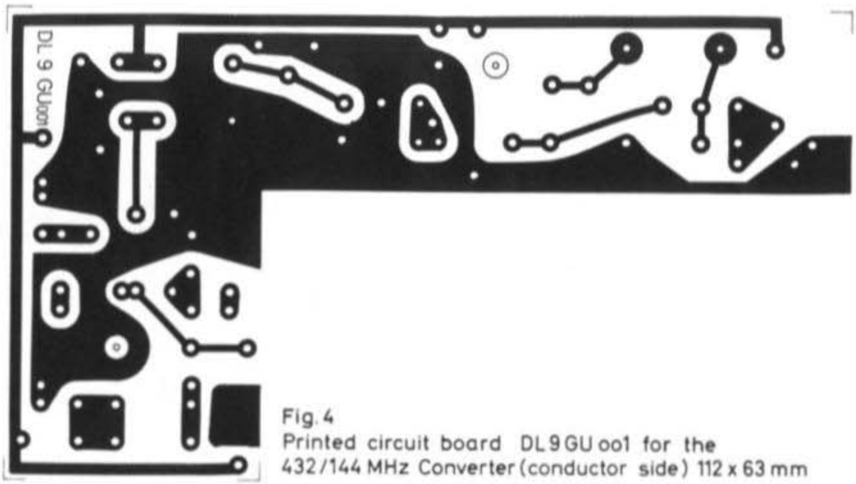


Fig. 4
Printed circuit board DL9GU oo1 for the 432/144 MHz Converter (conductor side) 112 x 63 mm

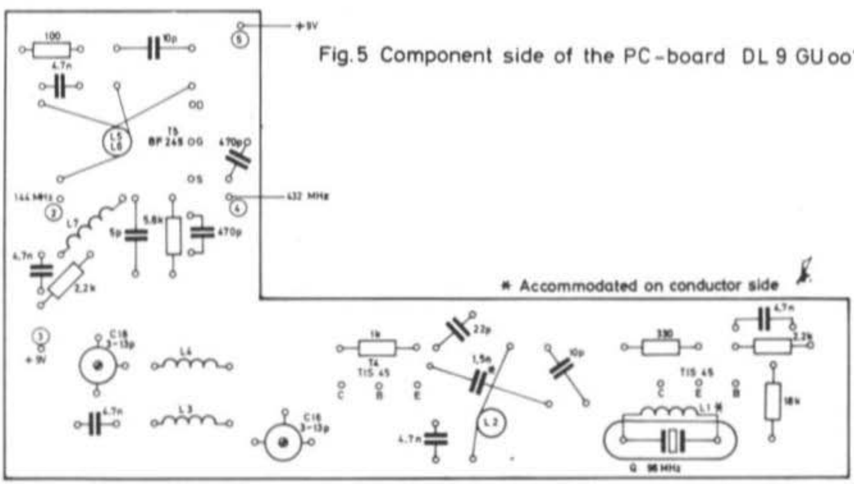


Fig. 5 Component side of the PC-board DL 9 GUoo1

3.2. PRINTED CIRCUIT BOARD DL9GU 001

The two oscillator stages and the mixer stage are to be found on printed circuit board DL9GU 001 which is shown in Fig. 4. The location of the components is given in Fig. 5. The printed circuit board is screwed onto the preamplifier casing by means of three metal brackets which have already been soldered onto the casing. These brackets can be clearly seen in Figures 3 and 6. The latter views the open converter from below. The ground connection between the printed circuit board and the casing is made via the brackets. This means that it is only necessary to make the connections (4) and (5), shown in the circuit diagram Fig. 2, in order to operate the converter.

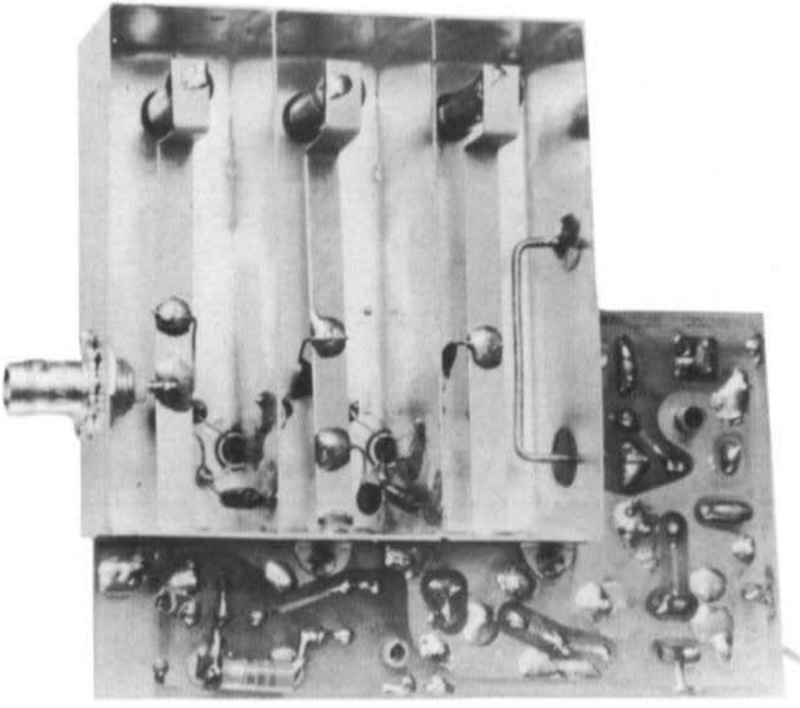


Fig. 6

4. ALIGNMENT

The crystal oscillator can be monitored by means of a VHF/FM broadcast receiver tuned to 96.0 MHz. The oscillation should cease on detuning the core of inductance L 2. If any spurious oscillations, such as unstable signals, noise, etc., are observed above the crystal frequency, this will mean that the inductance of L 1 must be decreased, or increased if the opposite is found to be the case. The tripler stage is aligned for maximum voltage drop across the source resistor R 14 of the mixer stage (in actual fact: it is aligned for the maximum voltage difference between no-signal and full signal conditions). If no 70 cm signals are audible after completing the alignment, it should be established if the correct harmonic of the crystal frequency has been selected.

Finally the preamplifier stages are aligned by means of trimmers C 1, C 5 and C 9 for maximum noise indication on the S-meter.

The decoupling loop is also adjusted for maximum noise.

The most favourable sensitivity position can only be achieved with the aid of a noise generator. This is carried out by varying the position of tapping point B and aligning C 1. It may also be advisable to solder a short wire onto the input socket so that the most favourable position of tapping point A can be found.

Editorial notes:
A noise figure of 6.5 dB was measured in a professional laboratory on a preamplifier according to reference (1). The swept frequency measurement indicated that the 3 dB bandwidth of the preamplifier was approximately 20 MHz.

5. REFERENCES

(1) E. Krahe: 70-cm-Vorverstärker mit Transistoren
UKW-BERICHTE 6 (1966), H. 1, pages 18 to 21

(2) W. v. Schimmelmann: A 2 Metre Converter With Field Effect Transistors
VHF COMMUNICATIONS 1 (1969), Edition 1, pages 2 to 10.

6. AVAILABLE PARTS

A variety of parts are available from the publishers or our national representatives: The printed circuit board DL 9 GU 001; the coil formers; the metal casing complete with feed-through capacitors, inner-conductors and trimmers; or metal casing complete with PC-board, coil formers and trimmers (See price list).

A 5 WATT TRANSISTORIZED SSB TRANSMITTER FOR 145 MHz

by K. P. Timmann, DJ 9 ZR

1. INTRODUCTION

Following the development of the phase-locked oscillator described in the last edition of VHF COMMUNICATIONS (1), several other units were built in order to form a modern SSB transceiver for fixed and mobile operation. The next piece of equipment to be assembled by the author was a fully transistorized SSB transmitter with an output power of 5 watts at 145 MHz. This unit is accommodated on a printed circuit board with the dimensions of 85 by 255 mm. A matching 2 metre receiver was also developed and will be described in the following edition of VHF COMMUNICATIONS. A variable auxiliary frequency of 135 MHz to 137 MHz is required for both units, which can be supplied from the already described phase-locked oscillator (1). It is, of course, possible for a VXO to be used instead of the phase-locked oscillator. The disadvantage of this, however, is that the frequency range of the transceiver is limited to a range of $\pm$ 20 kHz from the crystal controlled frequency of, for instance,145.41 MHz. Such a fully transistorized VXO is described in this edition.

The above mentioned transmitter and receiver units have been sufficiently well tested. They are easy to construct and are uncritical to align. The measured values at the end of each description offer information regarding the characteristics of each unit.

Up to RF power levels of approximately 10 W, transistor power amplifier stages have the advantage over tubed types because the heater power of tubes in this power category is in the same order as the RF output power. At higher output power levels, tubed power amplifiers are more economical since their heater power is negligible with respect to the output power and because their power amplification is greater whereas the plate efficiency of such modern tubes (i.e. ceramic types) is roughly equal to the efficiency of transistors. The disadvantage of having to provide cooling is normally common for both PA types.

The output power of the described 5 W SSB transmitter is sufficient to fully drive a linear amplifier equipped with the tubes 4X150A, D, G (drive power approx. 2 W), 4CX250B (drive power approx. 4 W) or 4CX350B (drive power approx. 6 W).

2. THEORY OF OPERATION

The block diagram of the 145 MHz SSB transmitter is shown in Fig. 1. Further details regarding the transmitter are given in the circuit diagram Fig. 2.

2.1. BLOCK DIAGRAM

The operation of the SSB transmitter is shown in Fig. 1. The voice frequency is amplified in a AF preamplifier and fed to a balanced modulator. The auxiliary frequency of 8.9980 MHz (USB) is generated in a 9 MHz carrier oscillator. The 9 MHz DSB signal is fed from the balanced modulator via the first

9 MHz amplifier to the crystal filter. The single sideband signal at the output of the crystal filter is fed via the 9 MHz buffer to the push-pull mixer. The 9 MHz SSB signal is then mixed with the variable auxiliary frequency of 135 MHz to 137 MHz obtained from the phase-locked oscillator (1) or the auxiliary frequency of 136.4 MHz ± 20 kHz from the already mentioned VXO. One obtains, in this manner, a 2 m SSB-signal which is amplified in the buffer and the driver stage and is fed to the PA-stage equipped with the overlay transistor 2 N 3375. The output power of the PA-stage amounts to 5.8 W.

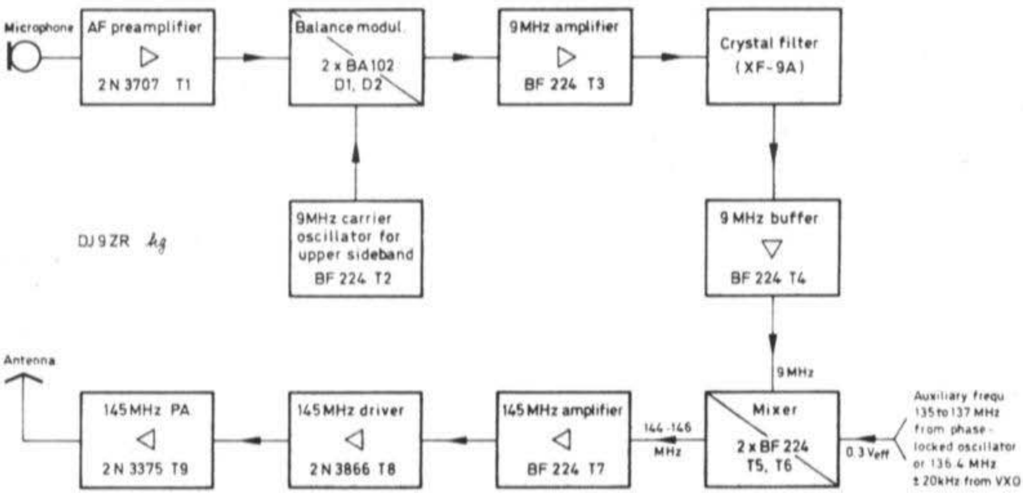


Fig.1: Block diagram of the 5watt SSB transmitter for 145 MHz

2.2. CIRCUIT OF THE TRANSMITTER

A detailed circuit diagram of the 2 m SSB transmitter is shown in Fig. 2. The input of the AF preamplifier is designed for the connection of a dynamic microphone. It is desirable for such a microphone to have a frequency response limited to the voice frequency range so that the frequency response favourable for single sideband transmission is achieved.

The voice frequency is fed via a resistor of 470 Ω and a 2.2 μF capacitor to the base of the AF preamplifier comprising the transistor T 1.

It is necessary, that a low-noise silicon planar transistor be used for T 1. The base of transistor T 1 is decoupled with a 10 nF capacitor to ground which, together with the resistor of 470 Ω, ensures that any RF-voltages are grounded. A current of 1 mA flows via transistor T 1, which causes a voltage drop of approx. 10 V to appear across the 10 Ω collector resistor. This ensures that T 1 is not overloaded even at the maximum permissible operating voltage of 28 V. The positive voltage from feed point A is filtered by the 390 Ω resistor and the 20 μF capacitor.

The AF preamplifier is coupled to one diagonal of the subsequent balanced modulator via a 50 nF capacitor which is series-connected to a 10 kΩ resistor. The carrier of 8.9980 MHz obtained from the 9 MHz carrier oscillator is fed via two 62 pF capacitors to the other diagonal of the balanced modulator. This principle was explained in (3). Due to the AF-modulating voltages, the capacity of the two varactor diodes D 1 and D 2 (BA 102) will be sometimes increased and sometimes reduced with respect to another. The RF-voltage at the centre point of the bridge is correspondingly varied and is tapped-off as a double sideband signal with suppressed carrier by potentiometer P 1. The carrier suppression in the balanced modulator amounted to approx. 20 dB. Further suppression of the carrier frequency is made in the crystal filter. The operating points of the two varactor diodes D 1 and D 2 are stabilized by means of a 8 V Zener diode Z 8. If the carrier is to be re-injected, the wiper of the potentiometer P 1 can be shorted to one of its ends by means of, for instance, a reed contact.

The 9 MHz carrier oscillator is equipped with a silicon planar transistor BF 224 (T 2). The load impedance of the oscillator transistor T 2 is very low to ensure that the AF voltage at the bridge cannot frequency modulate the oscillator.

The oscillator is built-up in the form of a capacitive Hartley-circuit. The crystal is aligned to the nominal frequency with the aid of a series capacitor of 300 pF. The nominal frequency is 2 kHz below the filter centre frequency of 9.000 MHz. The operating voltage of the oscillator is also stabilized by the Zener diode Z 8.

The 9 MHz signal from the bridge modulator is inductively coupled via inductances L 3 and L 2 to the base of the subsequent amplifier comprising transistor T 3 (BF 224). The output of this stage is connected via a resistor of 560 Ω to the operating voltage.

The 9 MHz DSB signal is coupled to the input of the crystal filter (XF-9A) via a 1 nF capacitor. The parallel capacity of 20 pF prescribed by the manufacturer is to be found at the input and output of the filter. The required load resistor of 680 Ω is also connected to the output of the filter. The crystal filter is not driven at the usual level of up to 500 mV but only at approximately 20-30 mV. This, together with the dimensioning of the subsequent buffer stage, is the main reason for the extraordinary good speech quality of the SSB signal.

The single sideband signal is fed via a 680 Ω resistor and a 30 pF series capacitor to the base of the 9 MHz buffer comprising transistor T 4 (BF 224). This stage provides virtually no amplification and is merely used to avoid any reaction on the filter from the subsequent mixer or 145 MHz amplifier chain. A relatively high resistor of 5.6 kΩ is to be found in the emitter lead of T 4. The voltage divider is dimensioned with two 33 kΩ resistors to obtain a ratio of 1:1. This means that the emitter voltage is very high (about 12 V or half the operating voltage). The amplification is reduced by the 1.8 kΩ resistor in the collector circuit of transistor T 4. The output circuit of this stage is aligned to 9 MHz by adjusting inductance L 3.

The 9 MHz signal is fed via inductances L 3 and L 4 to the push-pull mixer comprising transistors T 5 and T 6 (BF 224). A 100 Ω resistor is to be found in the base leads of both transistors to ensure that the auxiliary frequency of

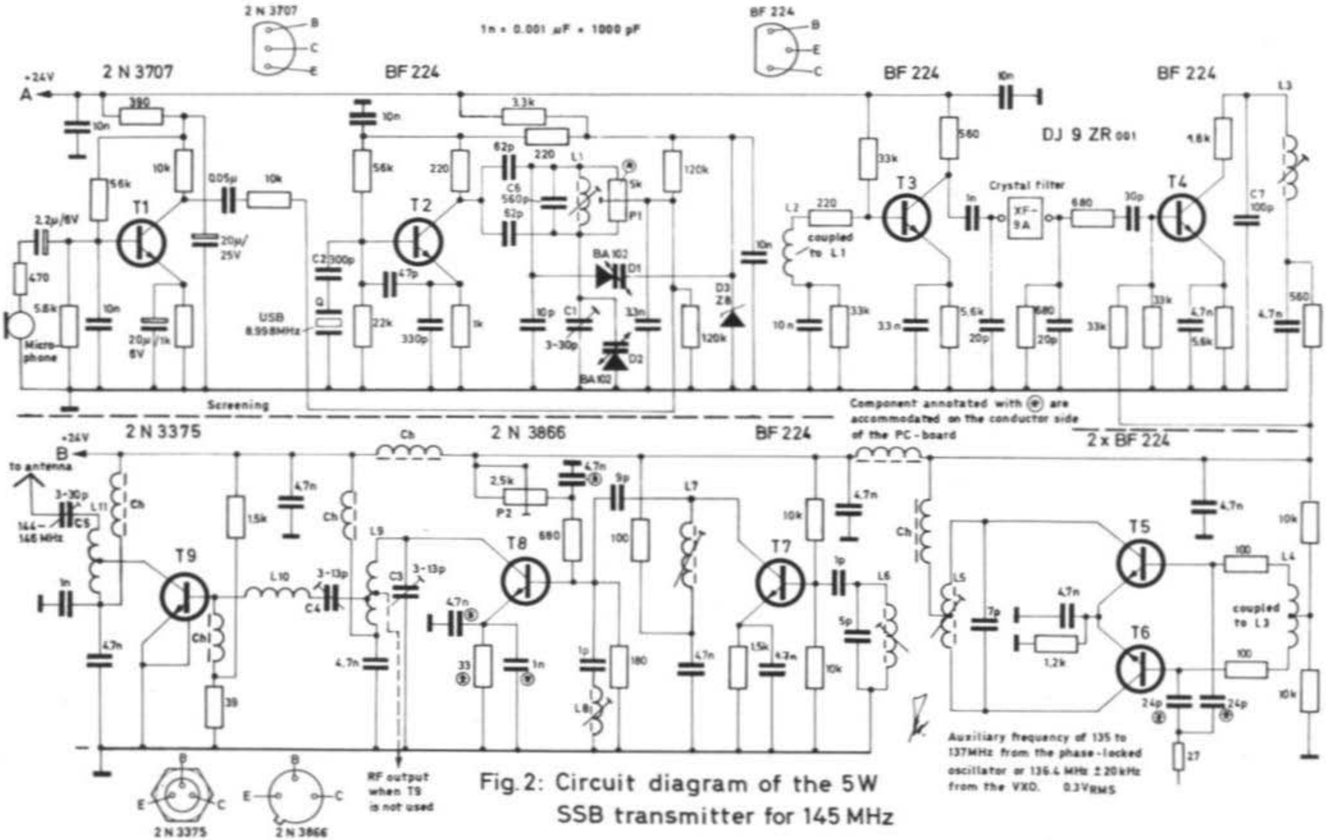


Fig. 2: Circuit diagram of the 5W SSB transmitter for 145 MHz

135 to 137 MHz (or 136.4 $\pm$ 20 kHz from the VXO) is not affected by inductance L 4. The auxiliary frequency input should be low-inductive and low-capacitive. The bases of T 5 and T 6 are fed in push-push with the auxiliary frequency voltage.

This voltage should have an amplitude of approximately 0.3 V in order to achieve the most favourable gain in the mixer. When using the VXO, a 27 Ω resistor should be connected directly previous to the two 24 pF capacitors to suppress any tendency of the mixer to self-oscillation. The emitters of transistors T 5 and T 6 are grounded. It is not possible to vary the operating point (balance). The auxiliary frequency rejection of the mixer is 10 - 15 dB which could be improved by approximately 5 dB if the mixer were fully balanced. However, the author did not consider this to be worth while. A push-pull output is obtained from the mixer. The positive operating voltage is fed via the centre tap of inductance L 5 and a choke.

The 145 MHz SSB signal is now fed via L 6 from the output of the mixer to the first 145 MHz linear amplifier comprising the transistor T 7 (BF 224). Small coil formers are used for inductances L 5 and L 6 to avoid any possible stray coupling. The DC input power of transistor T 7 is 70 to 80 mW; the RF output power approximately 40 to 50 mW.
This 50 mW signal is now coupled via a capacitor of 9 pF to the base of the driver stage comprising the overlay transistor T 8 (2 N 3866). A voltage divider consisting of a 2.5 kΩ trimmer resistor, a 680 Ω resistor and a 180 Ω resistor is to be found at the base of transistor T 8. In addition to this, a series-resonant circuit tuned to the auxiliary frequency (135 to 137 MHz) is connected to this point providing a further 10 dB suppression. The emitter of T 8 is grounded via a 4.7 nF capacitor as well as via several 1 nF capacitors to various positions on the PC-board. The 33 Ω resistor in the emitter circuit of T 8 is provided for temperature stabilization of the operating point. Approximately 0.6 W of RF power are available at the output of the driver T 8. If this power level is sufficient or if, for cost reasons, the power amplifier transistor is not to be used, it is possible to tap off the RF signal at the centre of L 9 (2.5 turns from either end) and feed it directly to the antenna relay.

Transistor T 8 reaches a temperature of 70° C at a DC input power of 1.5 W. It is provided with a heat sink in this transmitter although it operates correctly at temperatures even up to 120° C.

The 145 MHz SSB signal is fed via a series trimmer of 3 - 13 pF and the series-connected inductance L 10 to the base of the power amplifier transistor T 9 (2 N 3375). This means of coupling allows the input impedance of the overlay transistor (1 - 3 Ω) to be matched to the output of the driver. This series circuit, as well as the tank circuit (L 11), has a very sharp resonance which is of advantage for the suppression of harmonics. The transistor T 9 is screwed, together with intermediate copper or aluminium plates, to the printed circuit board. During normal SSB operation, transistor T 9 will become slightly warm; under continuous tone conditions at an operating voltage of 24 V, the heat sink of transistor T 9 will reach a temperature of approximately 70° C. The positive operating voltage is fed via a choke from point "B" to the power amplifier stage.

The common positive operating voltage of 24 V is fed to two different points of the printed circuit board which have been annotated points "A" and "B". This has been carried out to achieve an isolation between the operating voltages of the radio frequency and voice frequency circuits.

With the aid of the DC-DC converter also described in this edition of VHF COMMUNICATIONS, it is possible to operate this transmitter from a 12 V supply, e.g. from a car battery.

Transistors T 1 - T 8 operate in class A; the power amplifier transistor T 9 operates in class AB 2 which means that the transistor has a low quiescent current. The transistor type BF 224 (TIS 45) has a transit frequency of 850 MHz and possesses a very low intrinsic distortion when operated in class A up to a power of approximately 50 mW.

3. MECHANICAL ASSEMBLY OF THE 145 MHz SSB TRANSMITTER

The complete 145 MHz transmitter is built-up on a printed circuit board with the dimensions 85 mm by 225 mm. A photograph of the completed PC-board is shown in Fig. 3. The printed circuit, which has been designated DJ 9 ZR 001, is shown in Fig. 4. The component location plan is given in Fig. 5.

(on the centre pages 96 - 97)

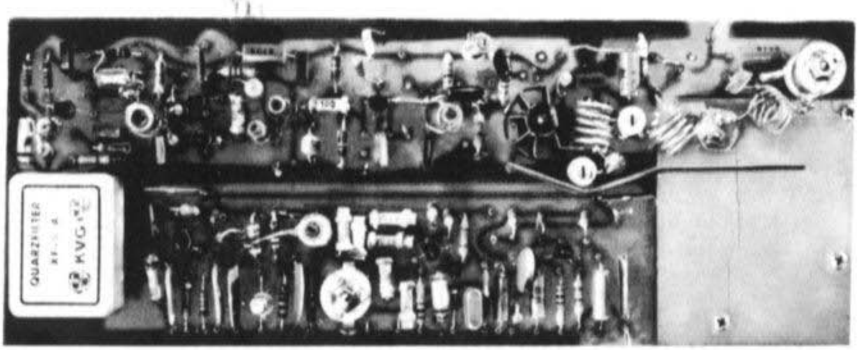


Fig. 3: The completed printed circuit board of the 5 W SSB transmitter

3.1. ASSEMBLY INSTRUCTIONS

After the required holes have been drilled into the printed circuit board, the components are mounted as shown in Fig. 5. The trimmer potentiometer P 1 of the balanced modulator and the two 24 pF capacitors at the auxiliary frequency input are mounted on the conductor side of the printed circuit board. In addition to this, the auxiliary frequency connection and the emitter combination of transistor T 8 are also located below the PC-board. These components are annotated by ⊛ in the circuit diagram (Fig. 2). Transistor T 8 is equipped with cooling fins (heat sink). The emitter of this transistor must be grounded via the previously mentioned combination in the shortest possible manner; this is not possible on the PC-board itself. The two 100 Ω resistors between inductance L 4 and the mixer transistors T 5 and T 6 stand vertically on the printed circuit board and are directly soldered to the coil.

The printed circuit board possesses a large, continuous copper coating in the corner where the power amplifier is located. A 1 mm thick copper plate with the dimensions 54 mm by 64 mm is mounted above and below this area with the aid of several screws. Transistor T 9 (2 N 3375) possesses a stud which is placed through the plates and the printed circuit board and screwed into place. These two plates ensure sufficient cooling for the transistor. A 30 mm high copper screening plate (approximately 0.5 mm thick) is placed on the component side of the printed circuit board to separate the RF and the AF portions. This screening runs from the heat sink nearly to the crystal filter and serves both as screening and for additional cooling (see Fig. 3). Heat-conductive paste should be used to improve the heat transfer.

3.2. COMPONENTS

3.2.1. SEMICONDUCTORS

T 1	2 N 3707 (yellow), BC 109 B, BC 131 B
T 2 - T 7	BF 224, TIS 45 (Texas Instruments)
T 8	2 N 3866 (RCA)
T 9	2 N 3375 (RCA) and possibly XB 404 (TI)
D 1, D 2	BA 102 (ITT-Intermetall), BA 101 (Telefunken) 1 N 954 (Varactor: 15 pF at 10 V)
D 3	Z 8 (ITT-Intermetall), BZY 85/D 8 V 2 (Telefunken) (Zener diode, 8 V/250 mW).

3.2.2. TRIMMER CAPACITORS

C 1, C 5	3 - 30 pF Philips tubular trimmers
C 3, C 4	3.5 - 13 pF ceramic trimmers as used in the DL 6 SW FET converter

3.2.3. CAPACITORS WHOSE VALUES MAY HAVE TO BE VARIED:

C 2	300 pF; may have to be reduced to 82 pF if, on monitoring the SSB signal, the impression is gained that the sideband is spaced too far from the carrier.
C 6	560 pF; may have to be reduced to 470 pF if resonance at 9 MHz cannot be achieved
C 7	100 pF; may have to be reduced to 56 pF if resonance at 9 MHz cannot be achieved.

All decoupling capacitors with values between 1 nF and 10 nF can, for simplification, be modified to 4.7 pF ceramic disc types.

3.2.4. RESISTORS

Resistors with a rating of 0.1 watt may be used; exceptions to this are: the 1.2 kΩ resistor in the emitter circuit of T 5 and T 6 and the 1.5 kΩ resistor in the emitter circuit of T 7, which have a rating of 0.2 watt. The 33 Ω resistor in the emitter circuit of T 8 should be a 0.5 W type. The resistors for the base voltage divider of T 9 have a rating of 0.2 watt.

3.3. COIL DATA

L 1	10 turns of 0.1 mm dia. (38 AWG) enamelled copper wire wound onto a 5 mm coil former with SW iron-dust core (brown); coil length 10 mm.
L 2	2 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound onto the coil former of L 1. Separated from L 1 by two layers of sellotape or plastic foil.
L 3	30 turns of 0.1 mm dia. (38 AWG) enamelled copper wire wound onto a 5 mm coil former with SW iron-dust core.
L 4	2 x 5 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound onto the upper portion of the L 3 coil former near to L 3. Centre tap.
L 5	5 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound onto a 4 mm coil former. Coil length spread over a width of 10 mm. Centre tap. SW core.
L 6	5 turns of 0.3 mm dia. (29 AWG) enamelled copper wire close wound; coil former and core as for L 5.
L 7	4.75 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound onto a 5 mm coil former; coil length approximately 10 - 12 mm, VHF core.
L 8	10 turns of 0.3 mm dia. (29 AWG) enamelled copper wire close wound; coil former and core as for L 5.
L 9	5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 9 mm former; self-supporting. Coil tap two turns from cold end. Coil length approximately 10 mm.
L 10	3.75 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 9 mm former; self-supporting coil length approximately 10 mm.
L 11	4.5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 9 mm former; self-supporting. Coil tap two turns from the cold end. Coil length approximately 10 mm.
Ch	all chokes are Ferroxcubes 4312 020 36691 manufactured by Philips ($Z \approx 450\,\Omega$).

3.4. INSTALLING THE PRINTED CIRCUIT BOARD INTO A CABINET

The printed circuit board should have a spacing of at least 25 mm from the cabinet. Large RF currents flow on the printed circuit board which can induce RF voltages into the neighbouring metal work. This could cause an unwanted stray coupling to other stages.

Only certain points of the printed circuit board should be grounded to the cabinet. These points are located in the vicinity of the antenna socket, near L 7, just before transistor T 8 and at both corners of the PC-board near the crystal filter.

4. ALIGNMENT

The alignment of the transmitter, which is not critical, should be carried out as follows:

Connect the operating voltage of 24 V to points "A" and "B" of the printed circuit board. Trimmer capacitor C 1 of the balance modulator is adjusted for maximum capacitance (fully in). This means that the modulator is unbalanced and that the carrier signal is not suppressed. All resonant circuits are now adjusted so that the maximum output power is achieved from the transmitter.

Trimmer C 1 of the balance modulator is then aligned for minimum output power. This is followed by further alignment of inductance L 1, trimmer C 1 and potentiometer P 1 of the balance modulator to carefully find the maximum carrier suppression. It is advisable to monitor the output signal on a receiver and align the balance modulator for minimum reading on the S meter.

The sensitive VHF voltmeter is now connected to the transmitter output. Inductance L 8 which forms, in conjunction with the 1 pF capacitor, an absorption circuit for the auxiliary frequency (135 to 137 MHz), is aligned for minimum RF voltage at the output socket.

The microphone is now connected to the AF input of the transmitter board and a VHF power meter (wattmeter with a large time constant) connected to the output. The VHF wattmeter will only indicate roughly a third of the maximum output ($5.8 \div 3 = 1.9$ W) when speaking into the microphone at normal volume. If more or less is indicated, it will be necessary to increase or reduce the value of the 680 Ω resistor between the crystal filter and transistor T 4. The resistor should not be reduced to less than 220 Ω since this could cause a self-oscillation of the buffer stage T 4. However, this resistance value is usually in the range of 680 Ω to 1 kΩ.

The driver transistor T 8 is aligned for class A operation by adjusting trimmer potentiometer P 2. The collector current of T 8 should not vary on speaking into the microphone. The collector current can be determined by measuring the voltage drop across the emitter resistor (33 Ω) which should be constant value between 2 and 3 V.

The quiescent current of the power amplifier transistor T 9 should be between 20 mA and 50 mA. If this is not the case, the 1.5 kΩ resistor between the base circuit of T 9 and point "B" should be changed.

Since most of the present 2 metre SSB contacts are carried out around the SSB frequency of 145.41 MHz, the transmitter should be aligned for maximum output at this frequency. If operation is desired over the whole two metre band, it is advisable to align the transmitter for maximum at 145 MHz where it should provide a mean output power of 5.8 W. The output power will drop by approximately 20% at the band limits (144 MHz and 146 MHz) which corresponds to an output of 5 W.

5. MEASURED VALUES

The overall efficiency of the described transmitter was measured to be nearly 50%. A mean output power of 5.8 W was obtained at 145 MHz with an overall DC input power of 12 W (24 V/0.5 A). The dependence on the operating voltage U_{op} was:

U_{op} = 21 V: 4.8 W mean output power
U_{op} = 24 V: 5.8 W mean output power
U_{op} = 28 V: 7.0 W mean output power

This transmitter can, with the described cooling, be operated continuously at 5.8 W mean output power.

A carrier suppression of 40 dB down when referred to the maximum output power of 5.8 W (U_{op} = 24 V) was measured.

The intermodulation distortion was 24 dB down with respect to 5.8 W. This corresponds to an overall harmonic distortion factor of approximately 5%. This measurement was made via all stages of the transmitter.

The first harmonic of the transmitter (288 MHz) was measured to be 45 dB down referred to 5.8 W mean output power. Only two spurious frequencies were found when operating the transmitter in conjunction with the phase-locked oscillator. These were the oscillator frequency itself (135 MHz to 137 MHz) and the corresponding image frequency (126 MHz to 128 MHz). The auxiliary oscillator is 40 dB down (ref. to 5.8 W) when injected at a voltage of 0.3 V. If this voltage is reduced from 0.3 V to 0.1 V, the break through will be approximately 50 dB down. However, this will also reduce the output power. The oscillator break through is essentially dependent on the absorption circuit at the emitter of T 8 and on the ground points of the printed circuit board.

Rejection of the image frequency (126 MHz to 128 MHz) was more than 60 dB down.

Measuring equipment used:

Hewlett Packard Spectrum Analyzer 851 B 8551 B
Hewlett Packard Signal Generator
Hewlett Packard Wattmeter and Milliwattmeter

6. REFERENCES

(1) K.P. Timmann and V. Thun: A Phase-locked Oscillator for Transmit and Receive Mixers in Amateur Radio Equipment.
VHF COMMUNICATIONS 1 (1969), Edition 1, pages 11 - 25.

7. AVAILABLE PARTS

The following components are available from the publishers or our national representatives: Printed circuit board DJ 9 ZR 001 as shown in Fig. 4; coil former set including cores; ceramic trimmer capacitors 3.5 - 13 pF. We are also able to provide the crystal filter XF-9A.

A 12 V / 24 V DC-DC CONVERTER

by K.P. Timmann, DJ 9 ZR

1. INTRODUCTION

This DC-DC converter was designed by the author to operate in conjunction with the 5 W SSB transmitter described in this edition of VHF COMMUNICATIONS. It is possible to operate the described transmitter from a 12 V supply; the output power is, in this case, approximately 1.4 W. It was only necessary to adjust the base resistors of transistors T 8 and T 9 for maximum output. However, in order to run this transmitter at full-power from a 12 V battery, a DC-DC converter with an output of 24 V at 0.8 A will be required. Since the transmitter only requires a current of 0.5 A, this converter provides a sufficient reserve of power.

2. CIRCUIT OF THE DC-DC CONVERTER

The circuit diagram is given in Fig. 1. The converter is a push-pull chopper using two silicon NPN transistors (T 1 and T 2) and a ferrite transformer (pot core). The voltage at the secondary of the transformer is rectified in a bridge rectifier comprising four silicon diodes (D 1 - D 4). The special feature of this circuit is that the rectified secondary voltage is added to the primary battery voltage. This means that it is not necessary for the whole power to be passed through the transformer and that the efficiency is higher than would be possible without this feature.

The quiescent current I_r and the maximum operating current I_{max} of the DC-DC converter are determined by the combination of R 1 / C 1. At R 1 = 68 Ω, I_r will be approximately 0.1 A and I_{max} approx. 1.05 A. At R 1 = 15 Ω, I_r will be approximately 0.4 A and I_{max} approx. 1.40 A. The most favourable value of R 1 is dependent on the load current and the current amplification of the transistors in question. A resistor of 100 Ω is often sufficient for the operation of the described transmitter.

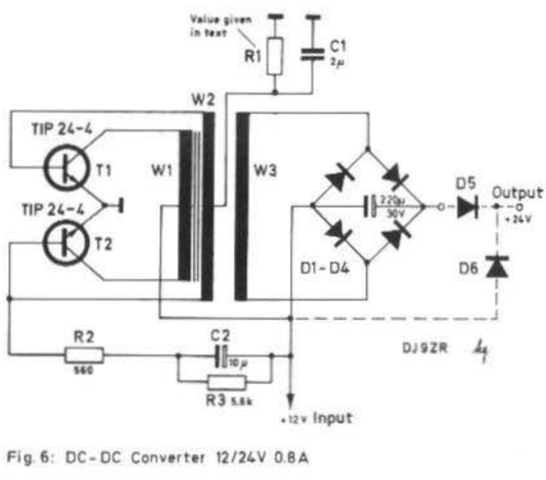


Fig. 6: DC-DC Converter 12/24V 0.8A

The RC network comprising R 2, R 3 and C 2 ensures that the chopper commences oscillation. This network may remain connected during operation because the current flowing via R 3 is limited.

3. PRACTICAL EXPERIENCE

In spite of this starting aid, it is advisable not to load the 24 V output of the converter before oscillation has commenced. One way of doing this is to connect the energizing coil of the transmit-receive relay between the 24 V and 12 V connections. This means that the relay (12 V type) is energized when oscillation commences and that one of the relay contacts may be used to connect the transmitter to the 24 V output of the converter.

If this method is not desired, two diodes (D 5 and D 6) can be used to enable the load to be continuously connected to the DC-DC converter.

If neither of these measures were to be used, it is possible that the converter would not commence oscillation or would oscillate at too high a frequency. This is because the secondary of the chopper transformer is short circuited by diodes D 1 - D 4. It will be noticed that until the transformer produces a secondary voltage, a current path exists from the + 12 V terminal via diodes D 1 - D 4 and the load resistance to ground which causes the diodes to conduct. Diode D 6 by-passes the current drain of the load resistance from the bridge rectifier until the higher voltage from the now oscillating converter blocks D 6.

The author and the editors would like to thank the Dutch amateurs PA Ø HAL and PA Ø JPB for the described modification of the original circuit.

4. COMPONENT DETAILS

T 1 and T 2 (matched pair)	TIP 24-4 (Texas Instruments) BD 106 A or BD 107 A (ITT-INTERMETALL) 40347V2 or 40250 (RCA) or other silicon NPN transistors with a transit frequency f_T = 100 kHz and a dissipation of P_{tot} = 10 W.
D 1 - D 6	Silicon diodes 100 V / 1 A
Transformer:	Ferrite pot core, 34 mm dia. 28 mm high. Manufacturer Siemens; core type 59 T 4 - 2000 T 26, W 0,5 o.L., A_L value = 5900 nH / turn^2 . Available from the publishers or from your national representative. Coil data: Primary: 22 turns of 0.6 mm dia.(23 AWG) enamelled copper wire with centre tap. Feedback: 8 turns of 0.3 mm dia. (29 AWG) enamelled copper wire with centre tap. Secondary: 14 turns of 0.6 mm dia. (23 AWG) enamelled copper wire.

REGARDING THE DJ 7 ZV / DJ 9 ZR PHASE-LOCKED OSCILLATOR

by G. Loebell, DJ 6 AH

1. EXPERIENCE

A phase-locked oscillator assembled as described in (1), operated correctly after careful alignment according to the given instructions. However, it was felt that the alignment was rather complicated for less experienced amateurs. Further instructions are given in Section 3.

Furthermore, it is necessary to suitably screen the whole oscillator against RF injection from the transmitter, which could cause an unsynchronized condition or an RF feedback into the SSB modulation.

2. MODIFICATIONS

2.1. It is advisable to mount a screening plate near to choke Ch 2 on printed-circuit board DJ 7 ZV 002 of the 135 MHz to 137 MHz oscillator (Fig. 5 and Fig. 9a in (1)). The slight tendency of the buffer stage T 11 to self-oscillation is completely suppressed by this measure and allows the stage to be neutralized even without 60 Ω termination.

2.2. A trimmer capacitor of 2 to 6 pF was installed and used, instead of the iron-dust core of L 10, to align the output circuit of the oscillator (Fig. 5 in (1)). Sufficient space is available for this trimmer. This measure has the further advantage of increasing the RF output voltage by approximately 30%.

2.3. Inductances L 1 and L 2 (Fig. 4 and Fig. 8b in (1)) are too tightly coupled. The minimum bandwidth obtainable with this degree of coupling is 6 MHz. Since the stages from inductance L 1 to transistor T 5 and the subsequent diode D 1 represent a very sensitive receiver, the input should possess the narrowest possible bandwidth so that unwanted signals are rejected. The coupling is reduced by means of a screening plate in the screening can. The metal plate with the dimensions 36 mm x 26 mm is bent in a right-angle at 18 mm along the longest side and mounted so that the inductance L 2 is screened as shown in Fig. 1.

2.4. The 30 μH choke and the series capacitor of 125 pF at connection point (4) (Fig. 4 and Fig. 8b in (1)) were found to be unnecessary.

3. TECHNOLOGICAL CONSIDERATIONS

In the first paragraph of Section 6 (1) it was stated that a high resistance voltmeter with an FSD of 10 V could be connected to connection point (3). It is, however, necessary to connect a low resistance μA meter to this point so that the discriminator diode is sufficiently loaded. The most favourable method is given in Fig. 10 in (1).

3.1. The amplifier can be frequency swept and aligned in the following manner: Connect a sweep frequency generator to connection point (1). Connect an oscilloscope via a 100 kΩ resistor to the cathode of the discriminator diode AA 112. Disconnect the 0.05 μF capacitor between the 1 kΩ resistor and ground.

Remove the 1 pF capacitor between the tripler stage T 6 and the mixer transistor T 2, and replace same with a coupling link.

The alignment is made whilst sweeping as follows:

Inductance L 1 allows the whole curve to be shifted in the frequency sense.

Align the bandwidth (dip) of the curve with inductance L 2. The amplification (crest slope) is aligned with inductance L 3.

Adjust the link between inductances L 3 and L 4 for maximum mixer amplification. Inductances L 4 and L 5 are, at the same time, aligned so that the pass band curve is retained, both in amplitude and form, on switching to crystal Q 1 or Q 2.

The frequencies 135 MHz and 137 MHz are located on the humps of the slightly dipped pass band curve.

4. REFERENCES

K.P. Timmann, DJ 9 ZR and V. Thun, DJ 7 ZV:
Phase-locked Oscillator for Transmit and Receive Mixers in Amateur Radio Equipment.
VHF COMMUNICATIONS 1 (1969), Edition 1, pages 11 to 25.

VARIABLE FREQUENCY CRYSTAL OSCILLATOR (VXO)
FOR 136 MHz

by K.P. Timmann, DJ 9 ZR

This variable crystal oscillator was designed by the author for operation with the 2 metre SSB transmitter (1) (as well as with the matching receiver which will be described in a later edition). The task was to find a simple means of obtaining the 136 MHz auxiliary frequency required for mixing both the 9 MHz SSB signal to 145 MHz and the 145 MHz input frequency to an intermediate frequency of 9 MHz. This VXO, which may be used to replace the phase-locked oscillator (2), is suitable for operation at frequencies up to 20 kHz above and below the nominal crystal frequency. It is especially suitable for SSB operation on or near the SSB spot frequency of 145.41 MHz.

1. BASIC PRINCIPLES

In the vicinity of its fundamental resonance, the impedance of a quartz crystal can, with good approximation, be simulated by a capacitance C_o parallel-connected to a "lossy" series-resonant circuit comprising L, C and R (Fig. 1 and Fig. 2). Prerequisite of this equivalent circuit is that spurious resonances are sufficiently spaced from the fundamental resonance. The capacitance C_o represents the capacitance of the crystal holder (capacitance between the electrodes and connection leads of the crystal). This is in the order of a few picofarads. The quantities of L, C and R are solely dependent on the individual characteristics of the quartz crystal, which represents a mechanical resonator electrically activated and firmly coupled to C_o. The inductance L (in the order of 0.1 H is very high compared to the inductances used in conventional LC circuits whereas the capacity is very low (1/400 th to 1/130 th of C_o, according to crystal type and plate cut). The equivalent resistance R, which represents the mechanical damping of the quartz crystal plate, including dielectric and connection loss, is in the order of 100 Ω.

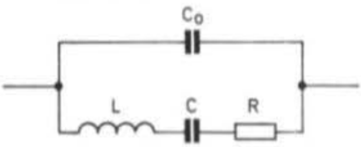


Fig. 1
Equivalent circuit of a quartz crystal in the vicinity of its fundamental resonance

The equivalent circuit also allows a good approximation of the impedance characteristics at high-order resonances to be shown as a function of frequency. However, this results in different values for L, C and R. The high-order resonant frequencies are not exactly multiples of the fundamental resonant frequency. This means that the crystal behaves in a similar manner to a shortened resonant line.

The series resonant frequency f_s of an unmounted crystal is solely determined by its oscillation characteristics.

$$f_s = 1/2\pi\sqrt{L \times C}$$

The following is valid for the parallel resonant frequency (anti-resonance) f_p:

$$f_p \approx f_s \times \left(1 + \frac{C}{2\,C_o}\right)$$

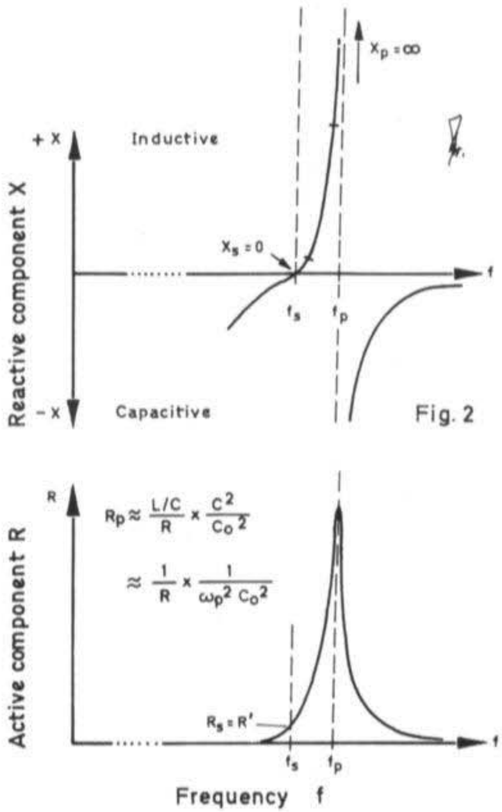


Fig. 2

Active and reactive components of the crystal impedance in the vicinity of its fundamental resonance shown as a function of frequency.

The accentuated curve portion in the upper diagram indicates the operating range of inductive crystals.

Especially stable oscillators usually utilize the series resonance of a quartz crystal. The majority of oscillators constructed by radio amateurs, however, operate at frequencies where the crystal represents an inductive impedance. This is the case for all circuits that seemingly operate at the parallel resonant frequency of the crystal, e.g. the Pierce circuit. The frequency of oscillation is in fact between the series and parallel resonant frequency; it is possible to slightly vary the frequency in this range by altering the phase of the feedback or of the amplification. It is possible to partially compensate C_o by connecting an inductance in parallel with the crystal. The parallel resonant frequency of this arrangement will be higher than the original f_p. If the inductance is connected in series with the crystal, a new series resonant frequency will result which is lower than the original f_s. These two arrangements therefore allow the spacing between the series and parallel resonance of the crystal to be increased and enable the pull-range of such "inductive" crystal oscillators to be slightly expanded. A pull-range approximately 1×10^{-3} of the fundamental frequency is possible; in some cases, however, it is even difficult to achieve a pull-range of 2×10^{-4}. It should be noted when using an inductance that an additional parallel resonant frequency exists below f_s in the parallel configuration or an additional series resonant frequency above f_p when series-connected.
The pull-range of overtone crystal oscillators is generally lower than that obtained from oscillators operating at their fundamental resonance. It can happen that the necessary detuning has a detrimental effect on the transient qualities

of the oscillator. The additional resonance points caused by the series or parallel-connection of inductances, are often located so near to the required resonant frequency, and are so easily activated, that it is possible for the frequency to jump.

F.W. Noble (3) described how a greater pull-range, together with sufficiently good frequency stability, could be achieved. Nobles argument was that the capacitive reactance represented by the crystal in the frequency range of a few percent below the resonant frequency f_s, is determined essentially by the resonance characteristics of the crystal plate (L and C) and not by C_0. This range is therefore also suitable for oscillator control purposes and allows, at the same, a greater pull-range. It is this method that is utilized in the described oscillator circuit.

The oscillator operates at a crystal frequency of approximately 45.5 MHz which is subsequently tripled to provide an output frequency of 136.5 MHz. The output frequency, which can be pulled approximately ± 20 kHz, is designed to produce the auxiliary frequency required to provide an intermediate frequency of 9 MHz in 2 metre receivers, or to mix 9 MHz (SSB) signals to a frequency of 145 MHz in 2 metre transmitters.

2. THEORY OF OPERATION

In the described circuit (Fig. 3), the quartz crystal is connected in series with a series configuration of capacitor C 2 and inductance L 2. This configuration forms the feedback link between the coil tap in the collector circuit (L 1 , C 1) and the emitter of transistor T 1 and, in this manner, determines the frequency of oscillation.

The oscillator frequency is varied by adjusting the forward voltage of the two varactor diodes D 1 and D 2. These diodes are connected in parallel with inductance L 2 and partially compensate its inductivity. The frequency stability of the complete oscillator is not only dependent on the stability of the crystal characteristics but also on the stability of the series inductance L 2. This means that special care must be paid during the manufacture of this coil. If a capacitor having a negative temperature coefficient is used (approximately $-750 \times 10^{-6}/^{\circ}C$), this will compensate for the positive temperature coefficient of the coil inductance and other circuit elements. It is, of course, important to ensure that the voltage across the varactor diodes does not fluctuate from the selected value. A capacitor of 100 pF is connected between the base and emitter of transistor T 1 (2 N 3707) in order to suppress any tendency to self-oscillation. Oscillation must cease completely when the crystal is disconnected. The damping resistor of 2.2 kΩ parallel to the collector circuit (L 1 , C 1) also has the same effect and, in addition, provides a noticeable extension of the pull-range by increasing the bandwidth of the collector circuit.

The subsequent frequency tripler with the transistor T 2 (BF 224) operates in class B and provides a high harmonic content. The output circuit with inductance L 3 is tuned to the third harmonic of the oscillator frequency. This is followed by a capacitively-coupled amplifier stage, equipped with the transistor T 3 (BF 224). The output voltage is then inductively coupled from the collector circuit of T 3 (L 4, L 5). A mean RF output voltage of 0.3 V is available at the output. Spurious signal rejection is better than 30 dB. The spurious signals are spaced more than 45 MHz from the nominal frequency and do not cause any interference points in the receive or transmit range.

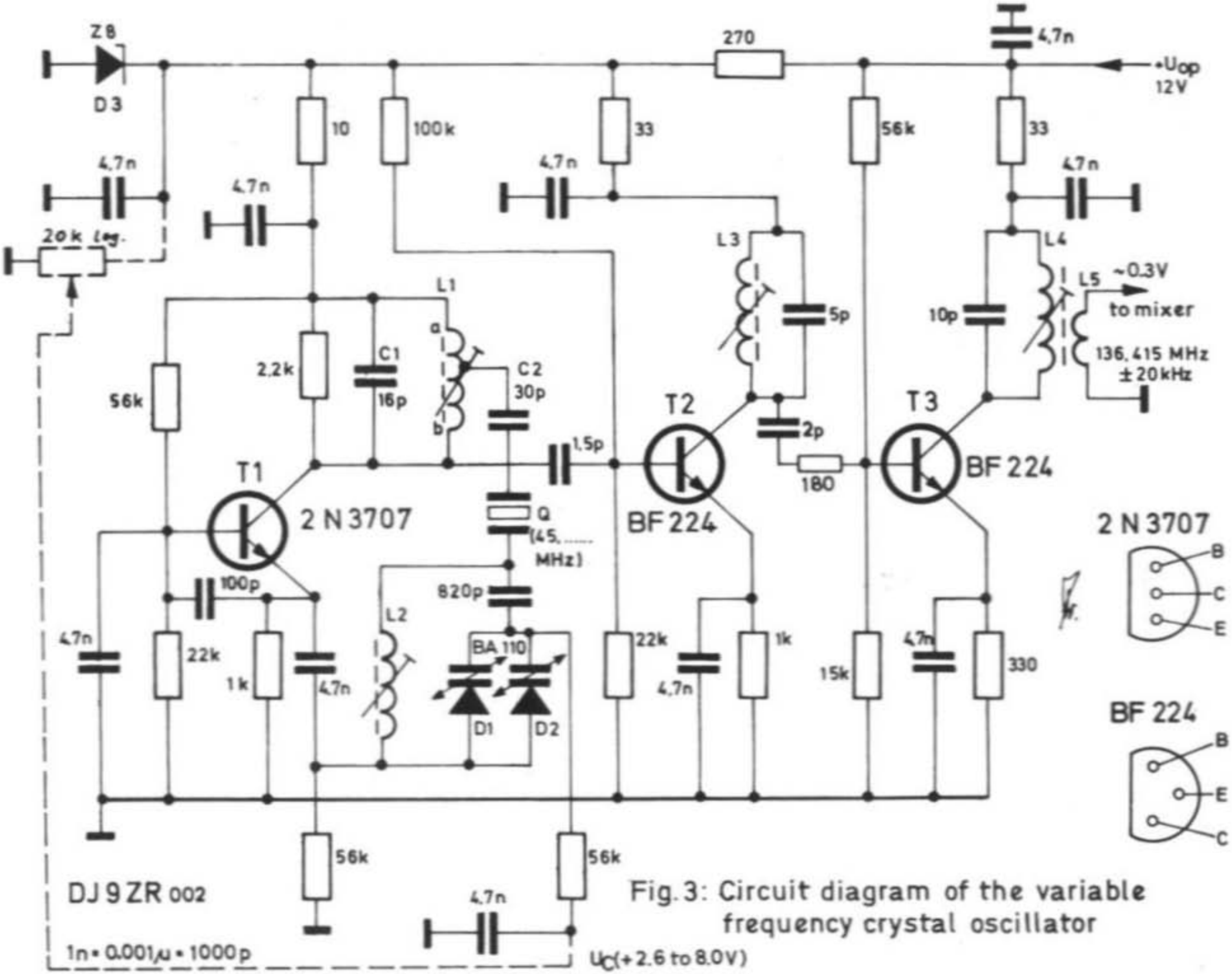


Fig. 3: Circuit diagram of the variable frequency crystal oscillator

3. MECHANICAL ASSEMBLY OF THE OSCILLATOR

The variable frequency crystal oscillator is accommodated on an epoxy printed circuit board with the dimensions 47 mm by 88 mm (see Fig. 4). The component location plan is given in Fig. 5 and a prototype shown in the photograph Fig. 6.

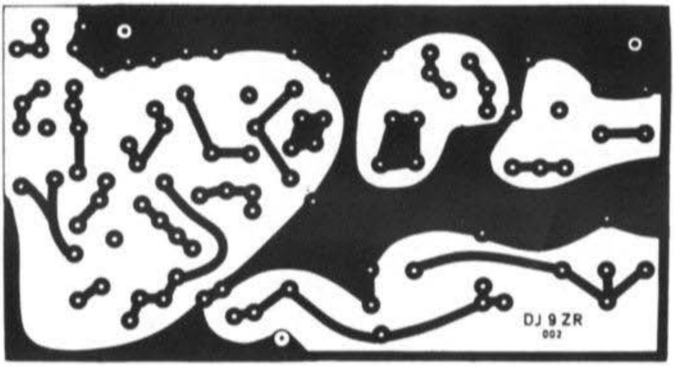


Fig. 4: Printed circuit board of the variable frequency crystal oscillator DJ 9 ZR 002

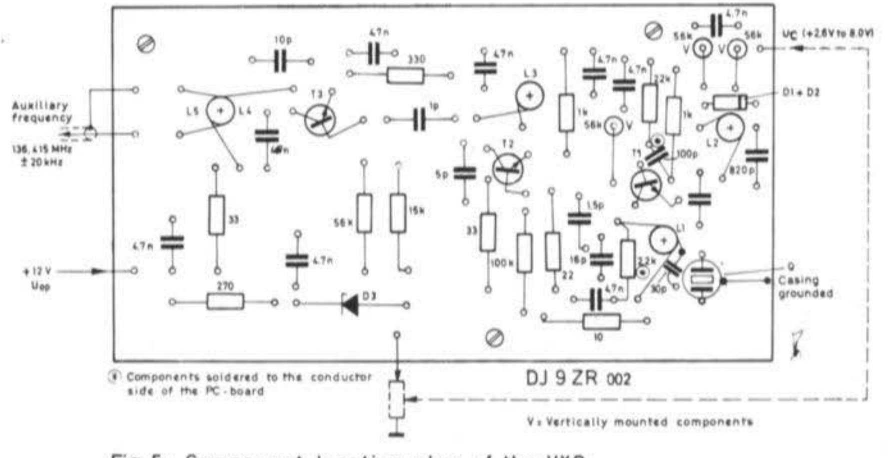


Fig. 5: Component location plan of the VXO

ERROR CORRECTION

A SOLID-STATE CONVERTER FOR 24 cm, Edition 1 of 1969, page 42, Fig. 2: Width should read 110 mm (not 100 mm). Chamber IV is 35 mm wide.

3.1. ASSEMBLY INSTRUCTIONS

The tubular ceramic capacitor C 2, located between the coil tap of inductance L 1 and the quartz crystal, is mounted vertically on the printed circuit board. It is important that the connection leads of this capacitor are kept as short as possible.

All resistors that are mounted vertically on the printed circuit board are designated by V, the components soldered onto the conductor side by ⊛

It is advisable to mould the whole printed circuit board including components into a solid block using a two-component adhesive or suitable epoxy resin. The top of the coil formers should protrude from the moulded block to allow alignment of the cores. The cores themselves must fit tightly in the coil formers and be provided with a slight padding.

The prototype oscillator was moulded into an epoxy resin block and protected against air currents by a small cabinet. It is advisable for further thermo-isolation to be made by covering the oscillator with a thin layer (5 mm to 10 mm) of styrene foam.

3.2. COMPONENTS

The nominal frequency f_q of the quartz crystal can be determined according to the following equation :

$$\text{Crystal frequency } f_q \text{ (MHz)} = \frac{\text{Required centre frequency (MHz)} + 20\text{ kHz} - 9\text{ MHz}}{3}$$

Example : Required centre frequency for the SSB transceiver is 145.416 MHz

$$f_q = (145.416 \text{ MHz} + 0.02 \text{ MHz} - 9 \text{ MHz}) \div 3$$

$$f_q = 136.436 \div 3 = 45.478 \text{ MHz}$$

Sub-miniature crystals, type HC-18/U are suitable for direct soldering to the PC-board.

T 1	2 N 3707, BC 149 or BC 109
T 2, T 3	BF 224, TIS 45 (Texas Instruments)
D 1, D 2	BA 110 (ITT-Intermetall), BA 121 (AEG-Telefunken) Varactor diode: 10 pF at 2 V.
D 3	Z 8 (ITT-Intermetall), BZY 85/C8V2 (AEG-Telefunken), 1 N 1931 (Zener diode: 8 V, 250 mW).
C 2	30 pF tubular ceramic capacitor with a negative temperature coefficient of 750 (violet colour point).

3.3. COIL DATA

L 1	11 turns of 0.3 mm dia. (28 AWG) enamelled copper wire wound along the whole length of the coil former (9 mm). Coil tap 2 turns from point "a". Coil former 4.3 mm dia. with VHF core and core padding.
L 2	6.5 turns; wire, former and core as for L 1. Coil length: 6 mm.
L 3	5 turns of 0.5 mm dia. (24 AWG) enamelled copper wire. Coil former and core as for L 1. Coil length: 6 mm.

L 4	5 turns; wire and coil length as for L 3. Coil former and core as for L 1.
L 5	1.5 turns wound between the turns of L 5.

All coil formers and cores are available either from our national representatives or directly from the publishers.

4. ALIGNMENT

The alignment is commenced by adjusting the resonant circuits comprising L 3 and L 4 to 136 MHz with the aid of a dip meter. This is followed by aligning the core of inductance L 2 to approximately 48 MHz. The operating voltage of + 12 V and the control voltage for the varactor diodes (+ 2.6 V to + 8 V) are connected and inductance L 2 aligned for maximum output. At the lowest control voltage of + 2.6 V, inductance L 2 is adjusted to the lower frequency limit of the pull-range. Inductance L 2, therefore, determines the lower frequency limit of the VXO. The upper frequency limit, which is determined by inductance L 1, is not influenced greatly by L 2 (Not more than 100 Hz). The transient oscillation characteristics, however, depend on the alignment of inductance L 1.

5. MEASURED VALUES

After correct alignment, the crystal oscillator will have a pull-range of ± 20 kHz with respect to the selected centre frequency. If a centre frequency of 145.416 MHz is assumed, the pull-range will be from 145.396 MHz to 145.436 MHz. However, the extent of the pull-range, is somewhat dependent on the manufacturer of the crystal. The frequency stability within the nominal pull-range of ± 20 kHz was found to be good. The author measured the short-term drift to be approximately 10 Hz in 15 minutes. It is possible to exceed the pull-range, but only at the cost of reduced stability. The frequency drift in the range 20 kHz to 40 kHz above or below the centre frequency was measured to be between 20 Hz and 30 Hz in 15 minutes and approximately 100 Hz to 200 Hz in the range of 50 kHz to 100 kHz from the centre frequency. These values were measured on a Hewlett-Packard frequency counter over a considerable period.

6. REMARKS

A stabilized operating voltage of 12 V was used to power the whole VXO. A stabilization of ± 0.1 V was obtained using a zener diode. The control voltage U_c for the two varactor diodes is stabilized with a further zener diode (diode D 3 in Fig. 3). A high-quality zener diode must be used at this point to avoid a variation of the stabilization between the transmit and receive mode. If this were the case, a frequency shift of 100 Hz to 200 Hz could be present between the transmit and receive frequency. This can be avoided by isolating the zener diode D 3 from ground (see Fig. 7) as well as by adding the transistor T 4 (2 N 3707) and a 560 Ω resistor. The stabilizing effect of the zener diode is then improved by the current gain of transistor T 4. No frequency shift could be observed after carrying out this modification. The additional components can be easily accommodated on the printed circuit board. The author used a conventional potentiometer which could be rotated over approximately 270°; a helical type potentiometer would be better. It should be noted that oscillation will cease if the control voltage drops below + 2.6 V. It is necessary for high-Q varactor diodes to be used.

The frequency range of ± 20 kHz from the SSB spot frequency of 145.41 MHz has been found to be fully sufficient for normal operation, even during contests. In many cases, even ± 10 kHz would have been enough.

It is not advisable to switch the quartz crystal by means of a ceramic or diode switch, since even the slightest shock of the contacts or variation of the transient resistance can cause a deterioration of the oscillator stability. If two ranges are required, the most favourable method would be to build up a further oscillator board for the second frequency range.

The printed circuit board DJ 9 ZR 002 is available from our national representatives or directly from the publishers.

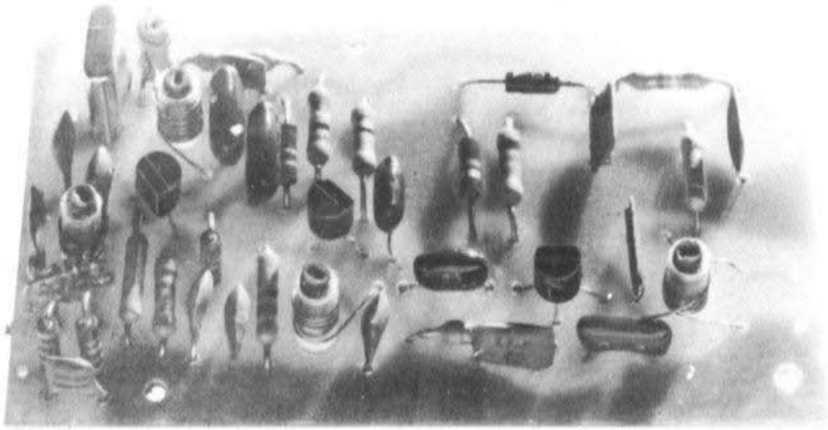


Fig. 6

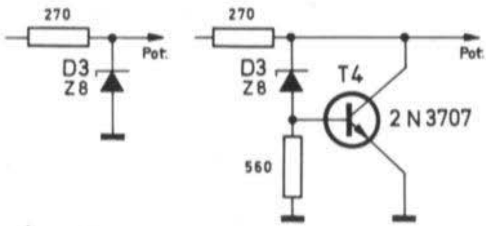


Fig. 7: Improvement of the stabilization using transistor T4

7. REFERENCES

(1) K.P. Timmann: A 5 watt Transistorized SSB Transmitter for 145 MHz VHF COMMUNICATIONS 1 (1969), Edition 2, pages 73 - 82

(2) K.P. Timmann: A Phase-locked Oscillator for Transmit and Receive Mixers in Amateur Radio Equipment.
VHF COMMUNICATIONS 1 (1969), Edition 1, pages 11 - 25

(3) F.W. Noble: Inductor ups crystal bandwidth to 2%
Electronic Design 15 (1957), Edition 6, pages 144 - 147

Centimetre / inches

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

Millimetre / inches

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

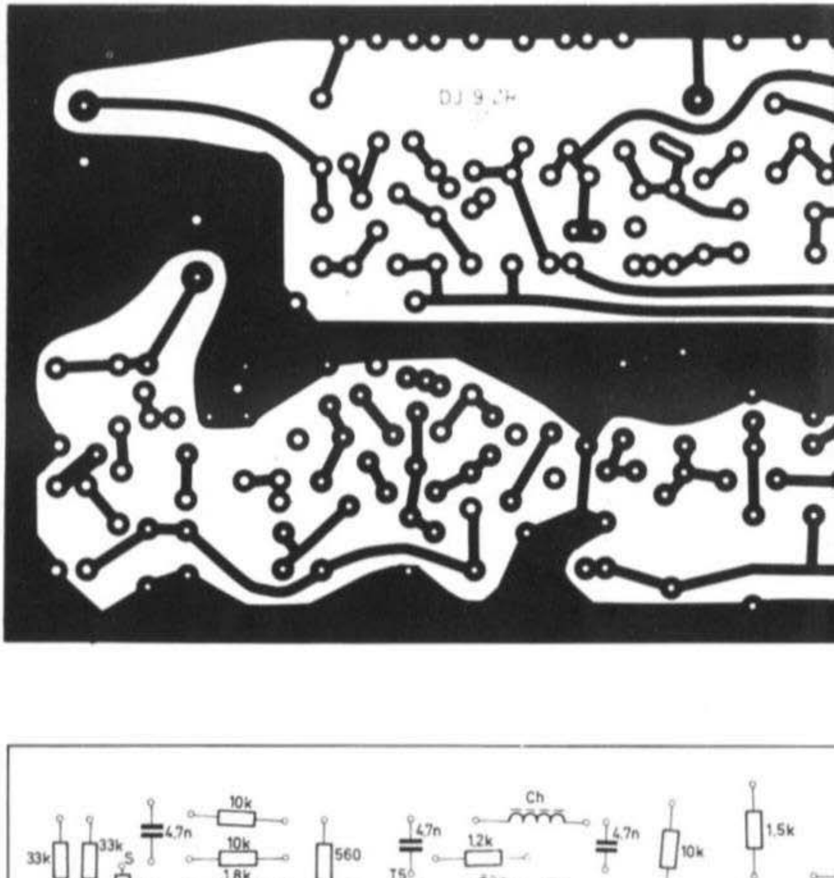
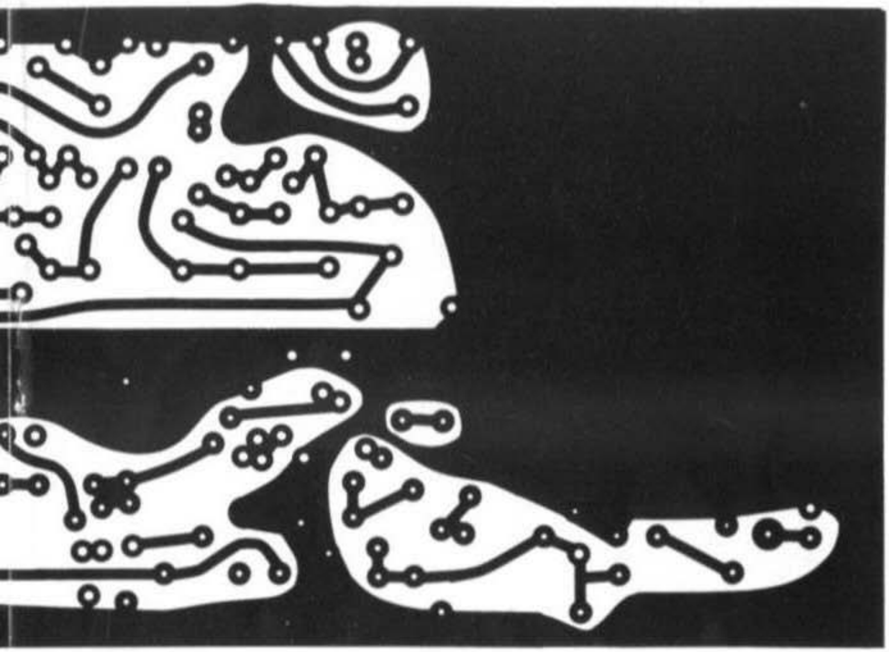
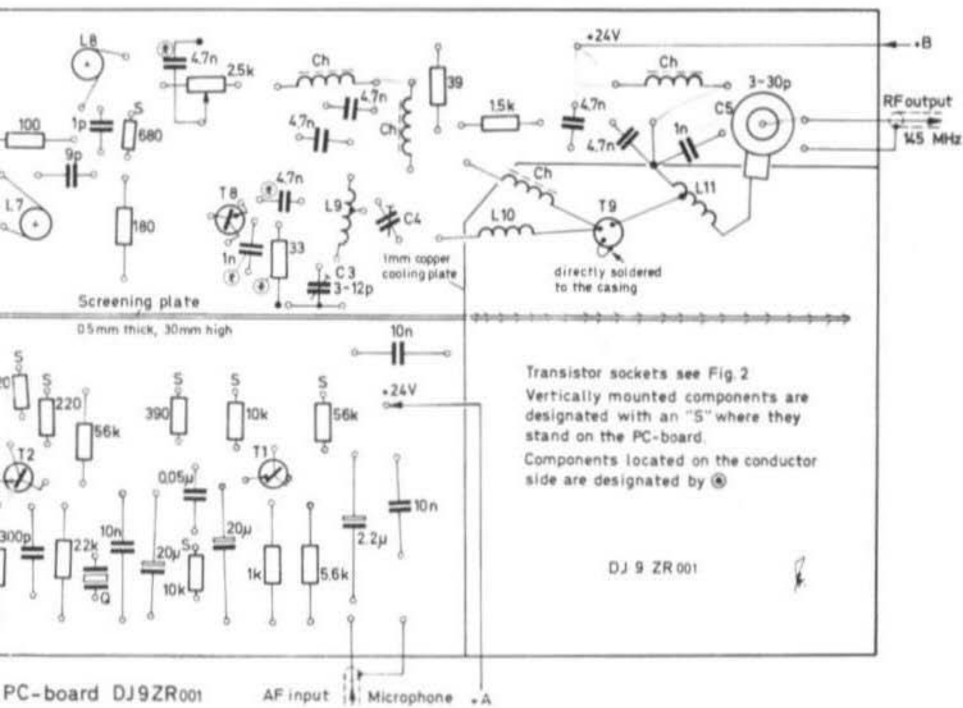


Fig. 4: Conductor side (above) and component side of the 2m SSB transmitter





PC-board DJ9ZR001

DETERMINING THE IMPEDANCE OF ROD ANTENNAS IN THE VHF RANGE

by H.J. Dohlus, DJ 3 QC

1. ROD ANTENNAS

Vertical rod antennas are popular for mobile and portable operation on the VHF bands; they are inconspicuous and are similar to antennas used for car radio, taxi communications etc. Further advantages are that they may remain mounted on the car or are easily connected and removed during portable operation.

In the majority of cases, amateur VHF transmissions are horizontally polarized, and rod antennas represent a certain compromise. However, horizontal antennas are more conspicuous, heavier and often directional. For this reason they are normally only used in contests, mobile rallies etc. A number of fixed stations are equipped with a vertically polarized antenna in addition to the normal, horizontally polarized array so that they are better prepared for communication with mobile stations.

Under ideal conditions, vertical rod antennas exhibit a constant, omni-directional characteristic with vertical polarization. However, due to the form and dimensions of the counterpoise (the car bodywork of mobile stations, or the body of the operator, the cabinet and the intermediate resistance between the two with portable stations), large deviations from the omni-directional characteristic, the vertical polarization and the expected impedance are possible.

In order to determine how the matching of such antennas is influenced, the impedance of a number of rod antennas ($\lambda/4$; $\lambda/2$; $5\lambda/8$; $3\lambda/4$; λ) was measured with a quadratic counterpoise, whose side-length was $3\lambda/4$; $5\lambda/8$; $\lambda/2$; $\lambda/4$; $\lambda/8$ and 0.

The directional characteristics and the gain were not measured, but some conclusions are given in Section 4.

2. MEASURING ARRANGEMENT AND PROCEDURE

A four-probe slotted line and an oscilloscope were used so that the impedance values at a predetermined frequency could be displayed as a direct Smith Chart representation.

The antenna test configuration (Fig. 1) consisted of a square aluminium plate with the side-length l and a rod antenna with the length L. The rod itself was built-up from a 6 mm dia. lower section designated L_o and a variable 3 mm dia. top section. The values of l and L were varied from measurement to measurement in order to determine the effect on the impedance. A coaxial cable with a characteristic impedance of 60 Ω was used to connect the slotted line to the antenna test arrangement.

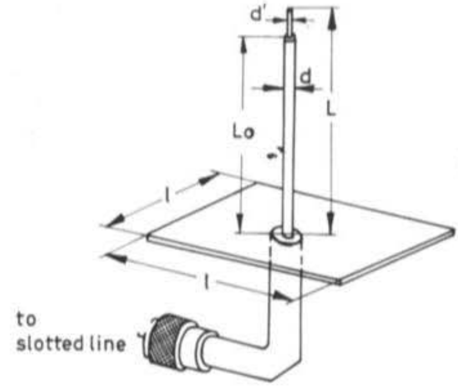


Fig. 1: Arrangement for measuring the impedance of vertically-polarized unsymmetrical rod antennas

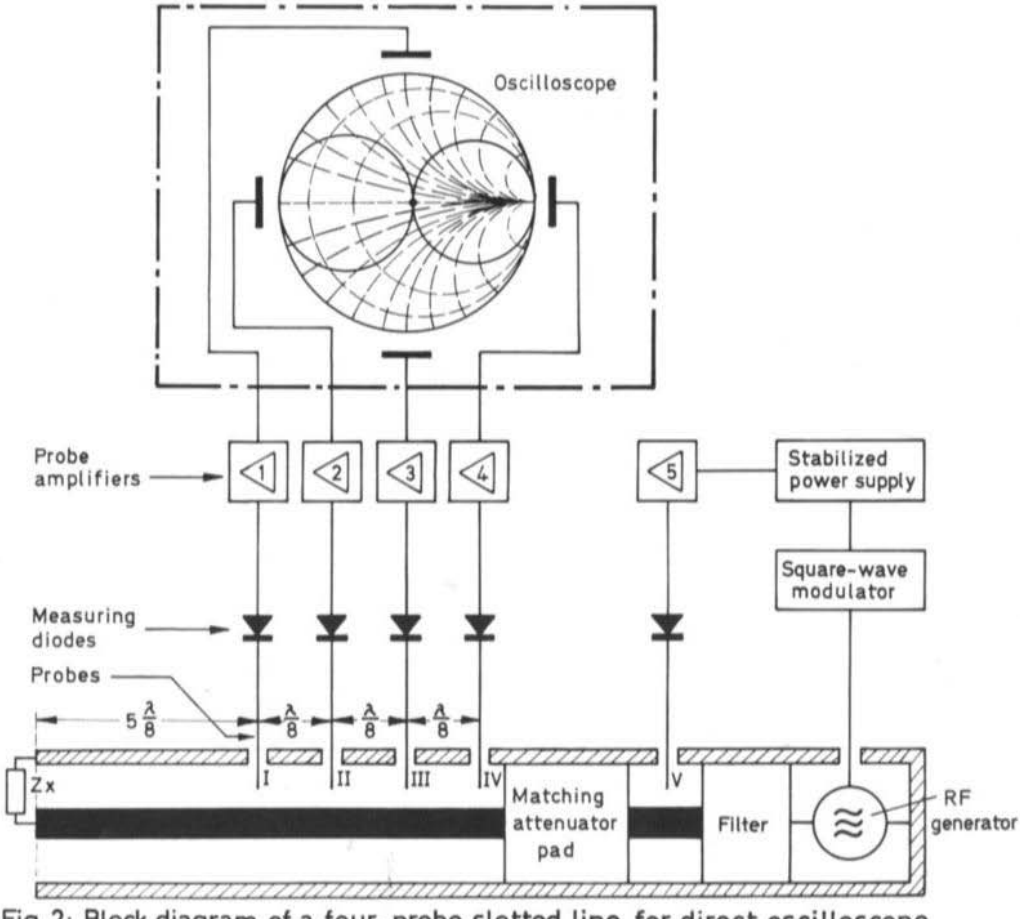


Fig. 2: Block diagram of a four-probe slotted line for direct oscilloscope display in Smith Chart representation. Z_x represents the impedance to be measured.

A block diagram showing the four-probe slotted line in conjunction with a Smith Chart oscilloscope is given in Fig. 2. The input of the slotted line was fed with a 145 MHz voltage from a signal generator. The other end was terminated by the unknown impedance Z_x of the antenna. A harmonic filter and a matching attenuator pad ($Z = 60\ \Omega$) were connected between the generator and the actual input of the slotted line. The generator output voltage was stabilized with the aid of probe V which was to be found between the filter and the attenuator pad. The other four probes (I - IV) were used to scan the voltage distribution caused by the unknown impedance Z_x at four points spaced $\lambda/8$ along the slotted line. These four voltages were then rectified by diodes having a square-law characteristic, amplified in subsequent amplifiers and fed to the plates of a cathode ray tube. The generator was square-wave modulated (10 kHz) so that AC voltage amplifiers (1 to 5) could be used to amplify the rectified voltages. The manufacture of an accurate four-probe slotted line is a very time consuming process and places great demands on the mechanical and electrical precision. Since it is felt that very few amateurs would wish to assemble such a device, a more detailed description was considered to be superfluous.

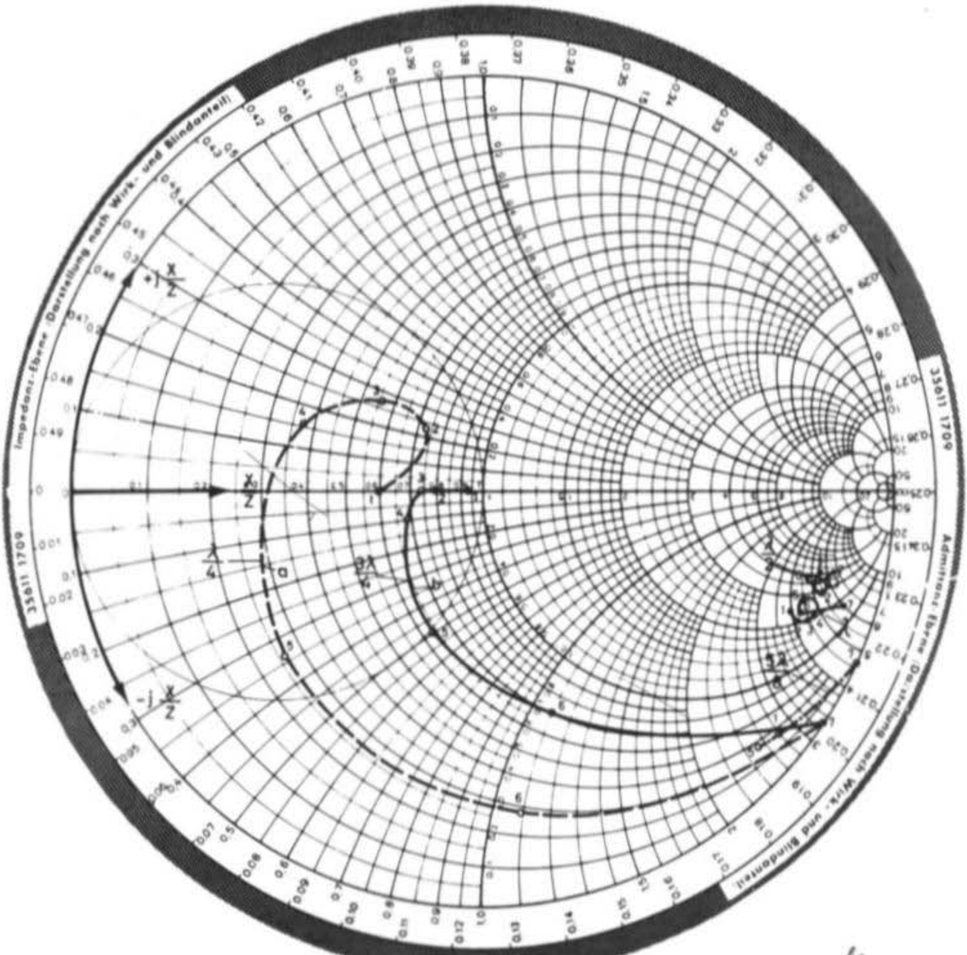
The co-ordinate grid usually used to display the magnitude and phase of the complex reflection coefficient may be covered with a transparent Smith Chart. The electron beam of the oscilloscope will then display the active and reactive component or the magnitude and phase of the unknown impedance Z_x. The impedance values indicated on the Smith Chart are referred to the corresponding characteristic impedance of the slotted line. The active component scale is to be found on the diameter line. The inner scale at the circumference is valid for the reactive components. In order to find the active component of a given measuring point, the corresponding circular arc crossing the diameter line is followed and read off. The reactive component is determined by following the nearest arc leading to the circumference and noting the value. The upper half of the circumference indicates inductive reactive impedances, the lower capacitive. The indicated values must be multiplied by the characteristic impedance of the test arrangement which was $60\ \Omega$ in the case in question.

The measurements were carried out in a large enclosed room. The walls and ceiling should not have affected the feedpoint impedance to any degree. In order to avoid any sheath current between the measuring set-up and the antenna, a quarter-wave coaxial sleath (balun) was added. This is connected to the $\lambda/2$ cable section just before the counterpoise connection (see Fig. 3). The other end of the coaxial cable is connected to the slotted line. If the coaxial sleath were not to be used, large measuring errors would be introduced, as can be seen by comparing the two results shown in Fig. 3 and Fig. 4.

3. ANTENNA MEASUREMENTS

3.1. IMPEDANCE OF THE $\lambda/4$ ROD ANTENNA

The following results are displayed in Fig. 3. for the $\lambda/4$ vertically polarized, unsymmetrical rod antenna:



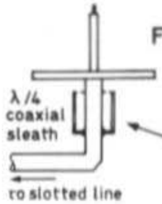


Fig. 3: Feedpoint impedance

a) a vertically polarized λ/4 rod antenna

b) a vertically polarized 3λ/4 rod antenna

as a function of the counterpoise area (see Fig.1)

Measurement carried out with coaxial sleath.

Side-length of the counterpoise for the measuring points:

1= λ; 2= 3λ/4; 3= 5λ/8; 4= λ/2; 5= λ/4; 6= λ/8; 7=0

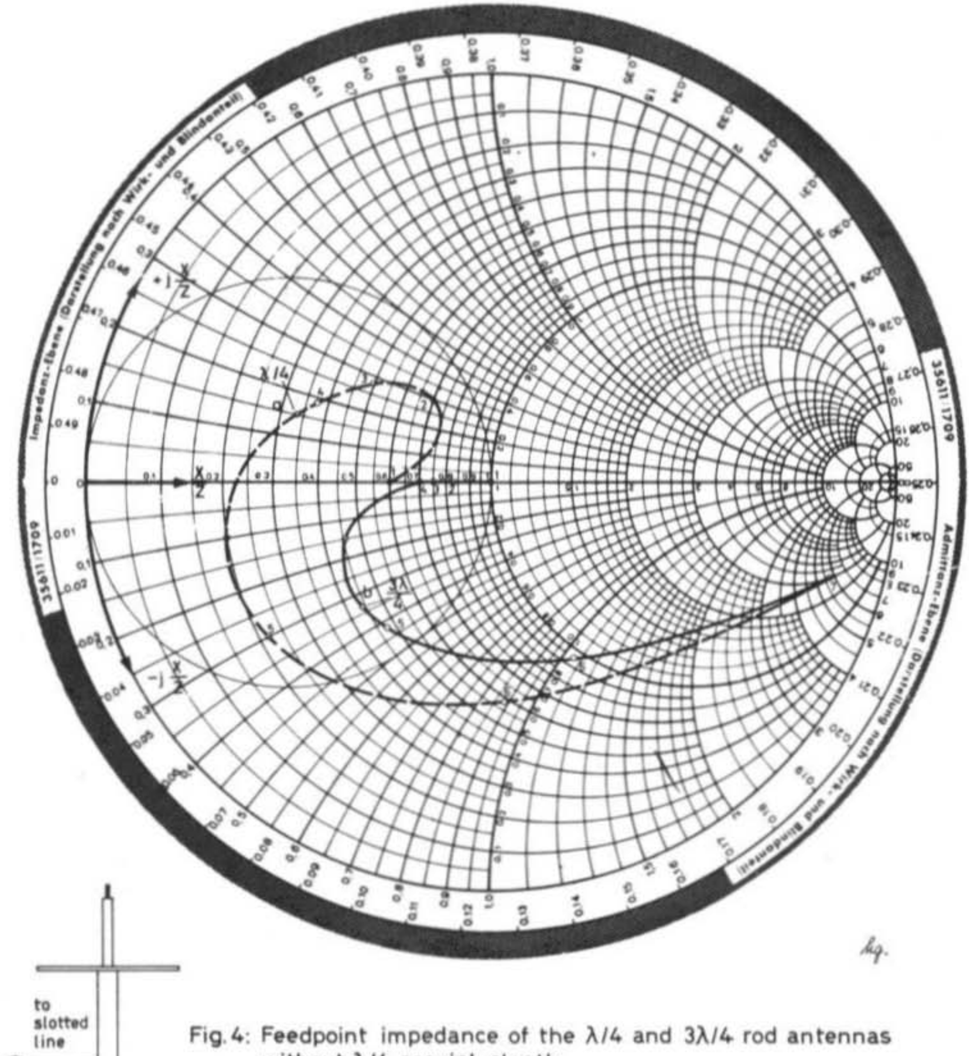


Fig. 4: Feedpoint impedance of the λ/4 and 3λ/4 rod antennas without λ/4 coaxial sleath.

Measuring point	Side-length of the counterpoise	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_x in Ohm
1	λ	0.6	36
2	3λ/4	0.73 + j 0.23	43.8 + j 13.8
3	5λ/8	0.56 + j 0.27	23.6 + j 16.2
4	λ/2	0.38 + j 0.16	22.8 + j 9.6
5	λ/4	0.27 - j 0.34	16.2 - j 20.0
6	λ/8	0.29 - j 1.09	17.4 - j 65.4
7	∅	- j 3.60	- j 216

Measuring point 2 is now considered as an example to illustrate how the measuring values are read off. The real reference value is obtained by following the arc via point 2 to the real axis (X/Z) which is crossed at the value 0.73. The complex reference value is to be found on the arc + j X/Z at the value 0.23. By multiplying X/Z + j X/Z by Z = 60 Ω, a value of 43.8 + j 13.8 Ω results for the impedance of the λ/4 rod antenna shown in Fig. 1; the side-length of the counterpoise was 3λ/4.

The impedance curve given in Fig. 3 only crosses the real axis X/Z at two points, at 0.32 and 0.6. A λ/4 antenna having a counterpoise with a side-length of approximately λ/3 exhibited an impedance of 20 Ω. At side-lengths greater than λ, the feedpoint impedance remained real and amounted to approximately 36 Ω.

This means that λ/4 rod antennas only have real, low-impedance feedpoint under the two previously mentioned conditions. At 145 MHz, a real impedance of 36 Ω can only obtained in the centre of a metal car roof having the dimensions of about 2 x 2 metres.

Unter normal conditions such a large, flat area is not available for 145 MHz rod antennas. This is especially true of walkie-talkie type stations where the counterpoise is partially replaced by the person holding the equipment.

3.2. IMPEDANCE OF THE 3λ/4 ROD ANTENNA

As can be seen in Fig. 3, the feedpoint impedance of the vertically polarized 3λ/4 rod antenna is higher than that of the λ/4 rod:

Measuring point	Side-length of the counterpoise	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_x in Ohm
1	λ	0.94	56.4
2	3λ/4	0.80	48.0
3	5λ/8	0.75 - j 0.01	45.0 - j 0.6
4	λ/2	0.68 - j 0.10	40.8 - j 6.0
5	λ/4	0.64 - j 0.50	38.4 - j 30.0
6	λ/8	0.70 - j 1.10	42.0 - j 66.0
7	∅	- j 3.60	- j 216

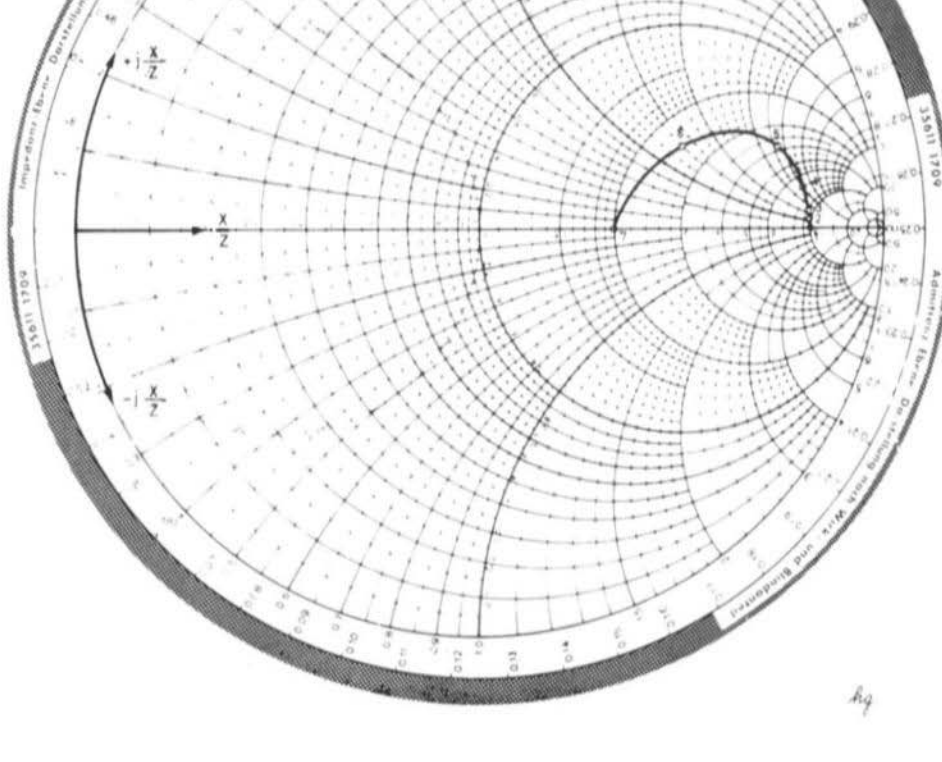


Fig. 5: Feedpoint impedance of a vertically polarized 0.4λ rod antenna as a function of the counterpoise area. Measuring points are given in Fig. 3

$3\lambda/4$ rod antennas exhibit a real impedance down to counterpoise side-lengths of λ, the rod had an impedance of 56.4 Ω. The impedance values at $3\lambda/4$ and $5\lambda/8$ were 48 Ω and 45 Ω respectively. The feedpoint impedance will exhibit a capacitive component if the counterpoise dimensions are further reduced; the behaviour, however, is still more favourable than that of $\lambda/4$ rods.

It should be noted that $3\lambda/4$ antenna rods radiate some of the energy at high angles, even under ideal conditions. This means that somewhat poorer radiation characteristics than $\lambda/4$ antennas must be expected when the antenna is mounted on the car roof or used during portable operation. The only way of suppressing this steep omni-directional lobe is when the lower quarter of the antenna is screened by the car body, as is the case with boot (fender) mounted antennas. This, however, also effects the feedpoint impedance.

3.3. IMPEDANCE OF THE $5\lambda/8$ ROD ANTENNA

As can be seen in Fig. 5, the $5\lambda/8$ rod antenna exhibited the following impedance values:

Measuring point	Side-length of the counterpoise	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_X in Ohm
1	λ	0.35 - j 2.80	21.0 - j 168
2	$3\lambda/4$	0.40 - j 2.75	24.0 - j 165
3	$5\lambda/8$	0.42 - j 2.65	25.2 - j 159
4	$\lambda/2$	0.42 - j 2.55	25.2 - j 153
5	$\lambda/4$	0.38 - j 2.50	22.8 - j 150
6	$\lambda/8$	0.35 - j 3.00	21.0 - j 180
7	ø	- j 4.60	- j 276

The feedpoint impedance of this antenna is virtually independent of the counterpoise dimensions and possesses a very strong capacitive component, which requires a certain amount of compensation.

3.4. IMPEDANCE OF A 0.4λ ROD ANTENNA

The feedpoint impedance of a vertically polarized 0.4λ rod antenna as a function of the counterpoise area is given in Fig. 5.

The impedance is real at a counterpoise side-length of λ and amounts to approximately 600 Ω. This value remains roughly constant down to a side-length of $\lambda/2$ and then becomes more and more inductive. The feedpoint impedance without counterpoise is approximately 120 Ω and is once more real.

3.5. IMPEDANCE OF VERTICALLY POLARIZED λ AND $\lambda/2$ ANTENNAS

As can be seen in Fig. 6, the $\lambda/2$ and λ antennas possess nearly equal feedpoint impedances which are virtually independent of the counterpoise dimensions. The measuring point in the diagram moves spirally in approximately the same area on varying the counterpoise area. The following values were measured in conjunction with the $\lambda/2$ antenna:

Measuring point	Side-length of the counterpoise	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_x in Ohm
1	λ	4.0 - j 5.5	240 - j 330
2	$3\lambda/4$	2.9 - j 5.6	174 - j 336
3	$5\lambda/8$	3.5 - j 6.0	210 - j 360
4	$\lambda/2$	3.7 - j 5.8	222 - j 348
5	$\lambda/4$	3.8 - j 5.3	228 - j 318
6	$\lambda/8$	3.8 - j 4.9	228 - j 294
7	∅	4.0 - j 7.5	240 - j 450

The high capacitive component of the impedance can be easily compensated by slightly shorting the antenna rod. This results in a real impedance of approximately 600 Ω (see Section 3.6.). The radiation characteristic of the $\lambda/2$ antenna is more favourable than that of a $\lambda/4$ rod.

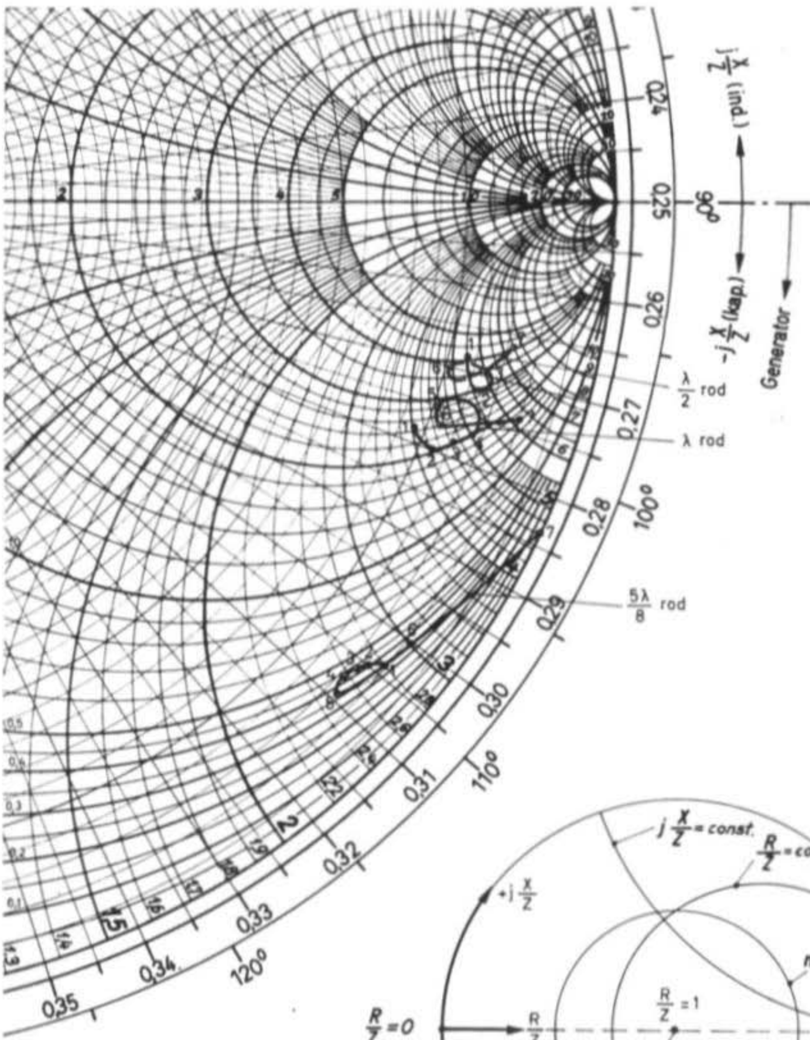
Measurements on the λ antenna resulted in the following values:

Measuring point	Side-length of the counterpoise	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_x in Ohm
1	λ	2.5 - j 4.0	150 - j 240
2	$3\lambda/4$	2.0 - j 4.2	120 - j 252
3	$5\lambda/8$	2.0 - j 4.6	120 - j 276
4	$\lambda/2$	2.0 - j 5.0	120 - j 300
5	$\lambda/4$	2.8 - j 4.6	168 - j 276
6	$\lambda/8$	2.3 - j 4.5	138 - j 270
7	∅	1.6 - j 6.4	96 - j 372

The measurements on the λ antenna were only made to complete the measuring series. The radiation characteristics of such antennas are very poor and are not even suitable for rear-mounted car antennas.

3.6. VARIATION OF THE FEEDPOINT IMPEDANCE AS A FUNCTION OF THE ANTENNA LENGTH

In this measurement, the side-length of the counterpoise is kept at a constant value of λ. The rod length L is reduced from λ to approximately $0.2\,\lambda$ in order to determine the dependence of the feedpoint impedance on the antenna length. The measuring curve commences with test point 1 of Section 3.5. which dealt with the λ antenna. As can be seen in Fig. 7, the resulting curve is an anticlockwise spiral which crosses the real axis on four occasions: At $0.9\,\lambda$, where the feedpoint impedance is approximately 450 Ω; 56.4 Ω at $0.75\,\lambda$ ($3\lambda/4$); approx. 600 Ω at $0.4\,\lambda$ and 36 Ω at $0.25\,\lambda$ ($\lambda/4$). The measurement was completed at $0.23\,\lambda$. The expected run at lower lengths is shown as a dashed line in Fig. 7. The impedance value ∞ is reached at $\lambda = 0$.



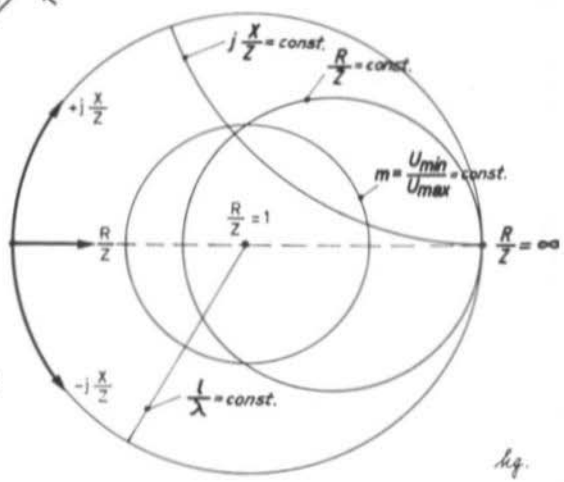


Fig. 6: Feedpoint impedance of vertically polarized rod antennas of λ/2, 5λ/8 and λ as a function of the counterpoise area. See Fig. 3 for measuring points.

hq.

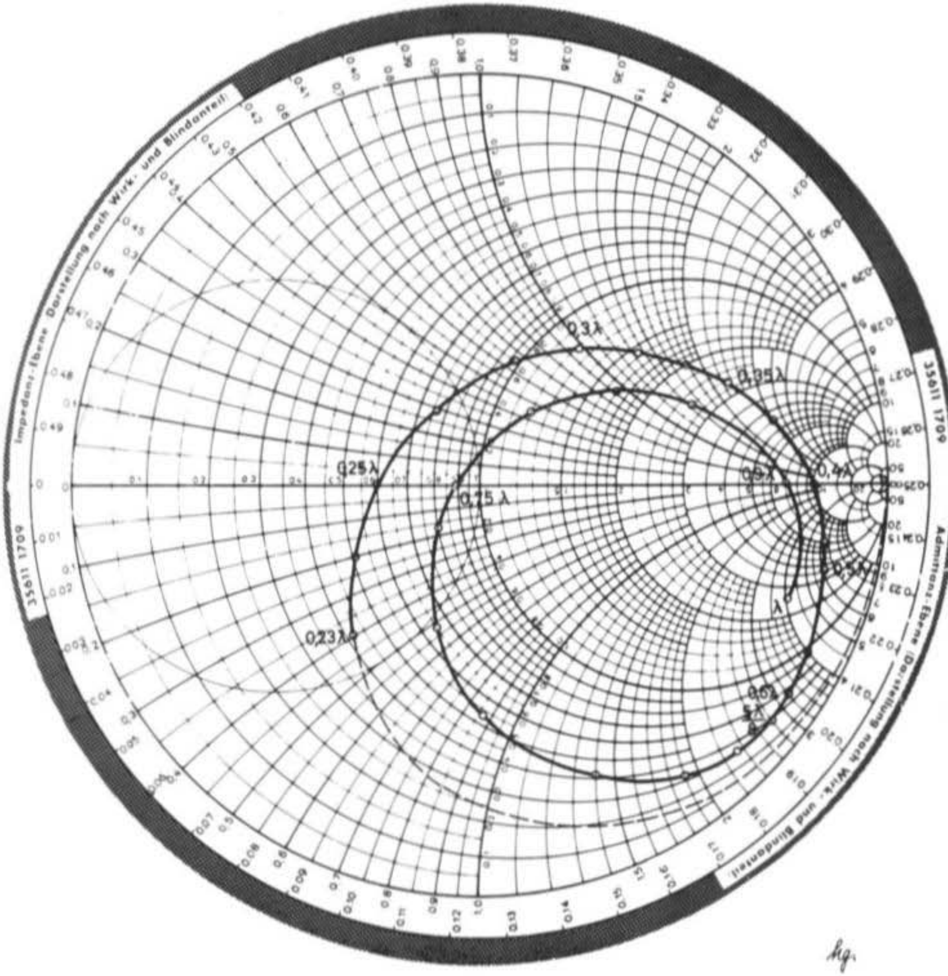


Fig. 7: Feedpoint impedance of vertically polarized rod antennas of various lengths over a constant counterpoise having a side-length of λ.

4. RESULTS OF THE MEASUREMENT

The feedpoint impedance of the rod antenna used with walkie-talkie stations must be high in addition to having a low reactive component. This is to ensure that the capacitive influence caused by the proximity of the operator and the intermediate resistance between the operator and the cabinet will be as small as possible. The measurements and subsequent considerations have shown that the $0.4\ \lambda$ antenna is most suitable for such applications. This is further underlined by the fact that the radiation characteristic is also slightly more favourable than that of $\lambda/4$ antennas.

The $3\lambda/4$ ($0.75\ \lambda$) antenna is suitable for rear-mounted mobile antennas where the lower portion of the antenna runs in the vicinity of the car body. However, such antennas will radiate at too high an angle when located on the car roof or when used with portable stations.

The $\lambda/4$ antenna is suitable for roof mounting and will exhibit a real feedpoint impedance of approximately $36\ \Omega$ as long as the counterpoise side-length is at least λ. In addition, the very low radiation angle makes this antenna popular for fixed (club) stations requiring an inexpensive, omni-directional antenna.

PREAMPLIFIERS TO IMPROVE SPEECH INTELLIGIBILITY UNDER POOR OPERATING CONDITIONS

by E. Schmitzer, DJ 4 BG

1. INTRODUCTION

The author would like to describe two preamplifier circuits that allow a virtually linear increase of the frequency response in the range of 300 to 2500 Hz. This represents an essential improvement of the speech intelligibility under poor operating conditions.

2. BASIC PRINCIPLES

The intelligibility is the most important consideration for both professional and amateur radio communications and not the "broadcast quality" that was often popular in the past.

This can be achieved at the transmitter by suitable tailoring (preemphasis) of the frequency response of the modulator preamplifier. If, for instance, the treble response within the AF bandwidth were to be accentuated to the value of 6 dB per octave, the speech would, it is true, seem unnaturally high but the individual speech characteristics would still be easily recognized. During normal conditions, the tailored speech has not been found to adversely affect the intelligibility. It is, however, under poor conditions where the great advantages are noticeable. The tailored speech is much more insensitive to subsequent non-linear distortion (by overloading or limiting devices) than linearly amplified speech. Speech tailoring is not only limited to AM but can be used with advantage for frequency or phase modulation as well as in the single sideband mode.

3. A TRANSISTORIZED PREAMPLIFIER

The following circuit shows how the desired effect of increasing the amplification proportionally with the frequency, namely by 6 dB per octave, can be achieved using simple circuitry. The amplification falls off steeply above 3 kHz so that the audio frequency bandwidth is not too broad. As the circuit is fully-transistorized, it is just as suitable as a microphone preamplifier for mobile and portable operation as it is for fixed station use.

A disadvantage of the described circuit is that only microphones with impedances lower than 2 kΩ can be used to ensure that the input low-pass filter is not detuned. But it is especially these low-impedance dynamic microphones that are preferable for mobile and portable stations since they are less sensitive to humidity and temperature than crystal microphones. A dynamic tape recorder microphone with an impedance of 500 Ω was used by the author.

3.1. CIRCUIT DETAILS

The circuit of the transistorized preamplifier is shown in Fig. 1. A two stage low-pass filter is to be found between the microphone input socket and the actual amplifier. The task of this RC filter is firstly to suppress undesired frequencies in excess of 3 kHz, and secondly to reject RF voltages. The latter case is especially true if a 10 nF feed-through capacitor is used for C 1.

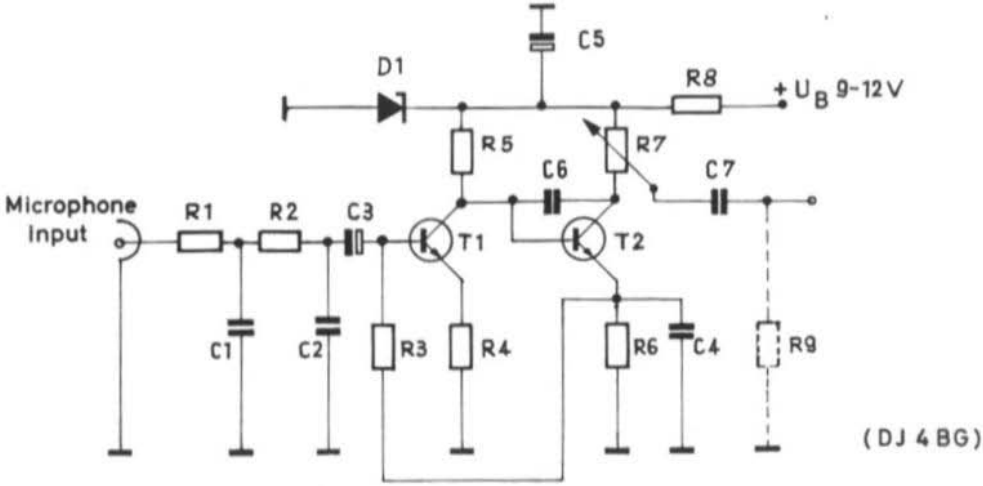


Fig. 1 Circuit diagram of a transistorized preamplifier with tailored speech response

The frequency response of this low-pass filter is shown graphically in Fig. 2. The insertion loss is less than 3 dB for microphones with impedances of 200 Ω or 2 kΩ. The curves corresponding to these two impedances do not differ greatly up to a frequency of 1 kHZ, the difference at 3 kHz is less than 2 dB. The 3 dB limits are to be found at 3 kHz for the higher impedance value and at 5 kHz for the lower.

The low-pass filter is followed by a two-stage amplifier which is stabilized by a large negative feedback. The author would now like to explain the operation of the circuit: The capacitor C 4 at the emitter of transistor T 2 has been dimensioned so that a frequency dependent feed-back results at low frequencies.

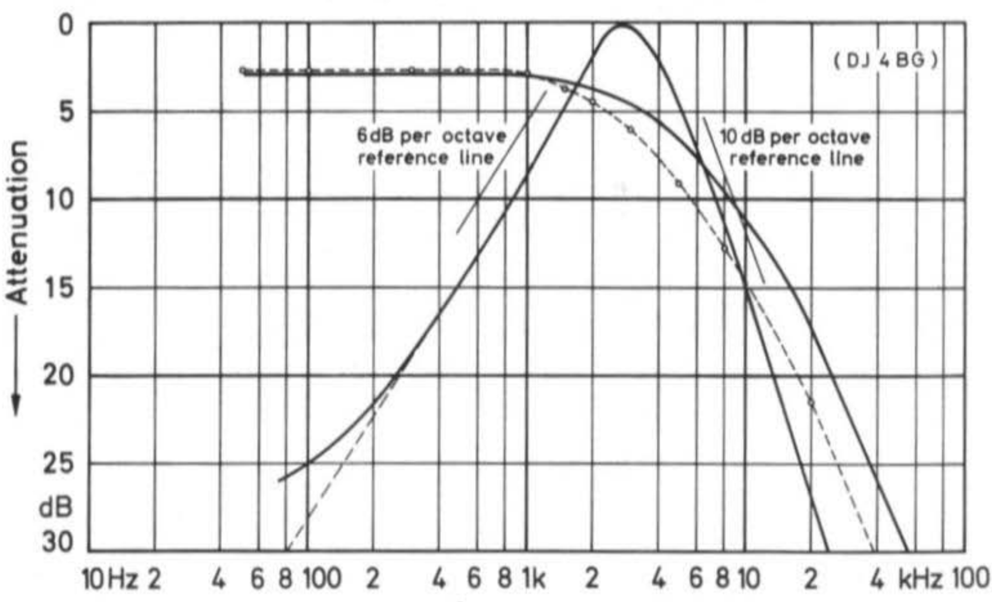


Fig. 2 Frequency response of the transistor preamplifier (0 dB corresponds to the maximum gain of the amplifier)

The voltage across C 4 / R 6, which represents an antiphase component to the voltage at the base of transistor T 1 at low frequencies, reduces the input impedance via R 3 (negative feed-back) and increases the voltage drop across R 1 and R 2. This combination of negative feed-back and voltage division results in a frequency response that increases by 6 dB per octave in the range 200 Hz to 2.5 kHz and decreases by 10 dB per octave over 4 kHz. This frequency response is also given in Fig. 2.

At frequencies lower than 300 Hz the reactive impedance of C 4 is so great that the negative feed-back is only determined by resistor R 6. This is the reason why the curve flattens out below 200 Hz. This range is, however, of little importance to the intelligibility and may be better suppressed by using a suitable combination of R 9 and C 7 for the coupling to the subsequent amplifier. The cut-off frequency of this RC-combination should be 100 Hz. If the output of the preamplifier is to be fed to a tubed modulator with an input impedance of 500 kΩ (Grid leak resistance of the first tube), this will result in a value of 3 nF for capacitor C 7 (Dashed line in Fig. 2). If on the other hand, the preamplifier is to be followed by a transistor amplifier, the correspondingly lower impedance must be taken into consideration.

A stabilized operating voltage is obtained using the zener diode D 1 and the resistor R 8. This ensures that frequency response variations will not occur due to operating voltage fluctuations. Since the circuit is equipped with NPN transistor types, it is possible to obtain the operating voltage by a suitable dropper resistor from the plate supply of the subsequent tubed modulator. An 80 kΩ resistor with a rating of 1 W will be required at a plate voltage of 250 V.

3.2. SELECTION OF SUITABLE TRANSISTORS

The author used transistor type 2 N 708 for T 1 and T 2. However, due to the great-negative feed-back, any transistor having similar AF characteristics can be used i.e. BFY 37, BFY 39, or even low-noise types such as the BC 107, BC 108. The economical Japanese transistor types 2 SC 182 or 2 SC 185 can be used if a miniaturized assembly is desired.

The selected transistor should have a DC amplification of B = 30 to 100.

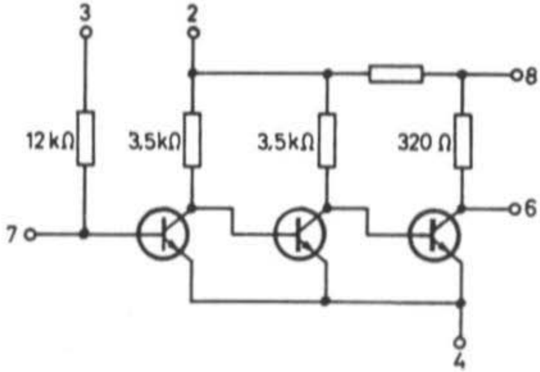


Fig. 3 Circuit of the integrated circuit TAA 111 (Siemens)

(DJ 4 BG)

H 0167

4. AN INTEGRATED CIRCUIT PREAMPLIFIER

Since integrated circuits are now available at economical prices, the author decided to examine their suitability for amateur applications. An integrated circuit with the designation TAA 111 (Siemens) was used in these experiments.

The attempt was made to use this integrated circuit in a preamplifier circuit having similar frequency response characteristics to the transistorized preamplifier described in Section 3. The circuit of the integrated circuit is given in Fig. 3; this is complemented by Fig. 4 which shows the connection diagram of the additional components. A microphone with an impedance of 500 Ω was again used in conjunction with this circuit.

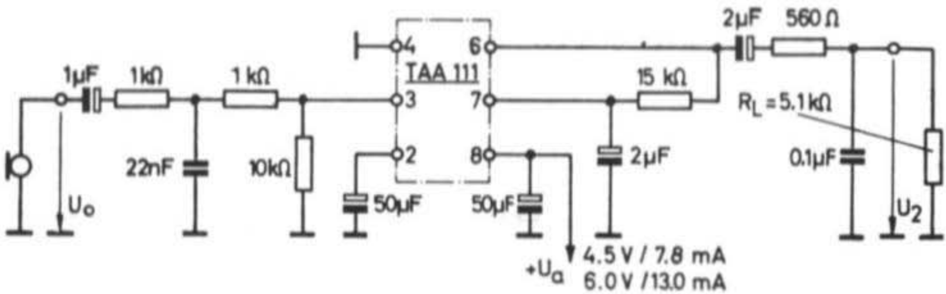


Fig.4 Connection diagram of the integrated circuit TAA 111 (DJ 4 BG)

4.1. CIRCUIT DETAILS

The microphone output is fed via an isolating capacitor to an RC-link which suppresses the treble component. This is followed by a resistor which reduces the danger of non-linear distortions. The parallel resistor determines, together with the dropper resistor between points 7 and 3 and the negative feed-back resistance between points 6 and 7 of the integrated circuit, the operating point of the first amplifier stage.

The negative feed-back of the lower frequencies is carried out via all three amplifier stages; the RC combination 15 kΩ / 2 µF allows a sufficiently great bass-rejection of approximately 6 dB per octave to be achieved.

If a negative feed-back network for the suppression of unwanted frequencies in excess of 3 kHz were to be placed between connection points 6 and 3 of the IC, this would lead to a self-oscillation of the amplifier. The treble rejection is, therefore, carried out with the aid of RC-combinations at the input and output. This has the disadvantage of requiring more components and of having a somewhat wider transition range. However, one saves the analysis of the complicated transfer loci (frequency and phase response) of the amplifier itself and the calculation of the negative feed-back network. Of course, a steeper rejection of frequencies in excess of 3 kHz and a smaller transition range would be possible if this calculation process were to be carried out.

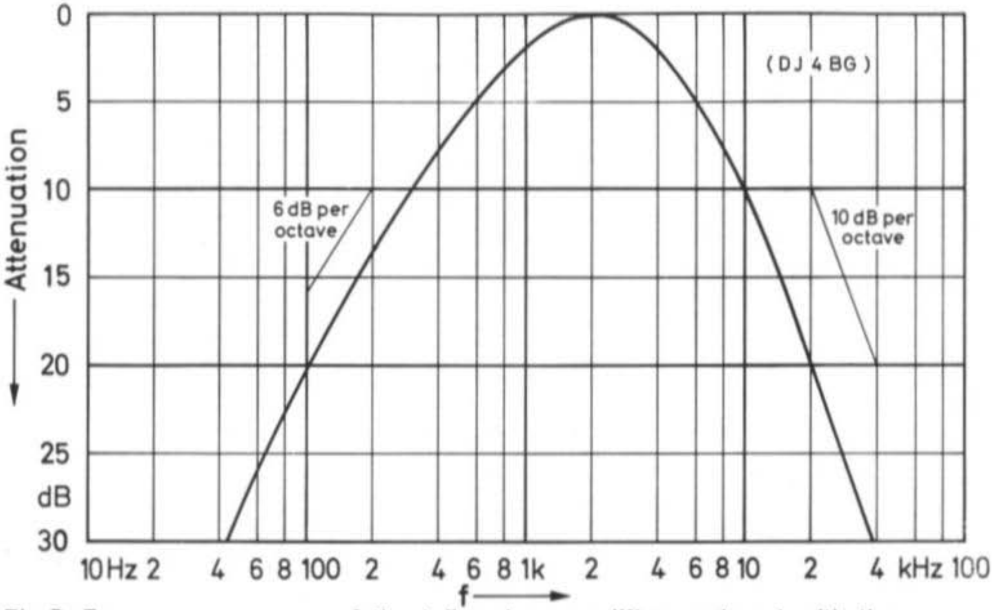
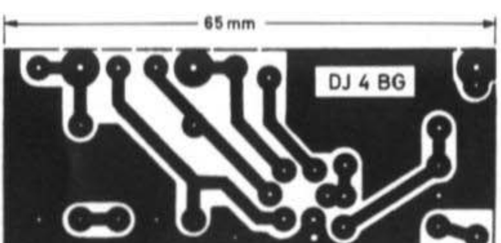


Fig. 5 Frequency response of the tailored preamplifier equipped with the integrated circuit TAA 111 (0dB corresponds to an amplification of 51 dB)

The maximum voltage amplification of 51 dB is very high. This means that the output voltage is usually sufficient to feed the driver stage of a modulator. A value of 100 mV was measured at the output when speaking into the previously mentioned microphone.

The frequency response (Fig. 5) and amplification remain very constant in the operating voltage range of 4.5 V to 7.0 V. The amplification falls off rapidly under 4 V; a voltage in excess of 7 V should not be applied since this could lead to destruction of the integrated circuit. It is therefore advisable to limit the operating voltage with the aid of a 5 V zener diode.

The circuit was assembled using conventional components on a PC-board having the dimensions 27.5 mm by 65 mm (see Fig. 6 and 7). However, a further miniaturization of the circuit would be possible.



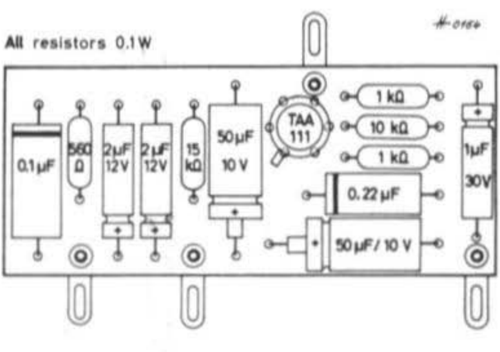


Fig.6 Conductor side of the integrated circuit preamplifier board.

Fig.7 Component side

A MODERN CONCEPT FOR PORTABLE 2 METRE RECEIVERS

by E. D. Schmitzer, DJ 4 BG

INTRODUCTION

The major disadvantage of the double and triple conversion superheterodyne principle as used at present is that most spurious signals are not suppressed until after the third mixer stage. Since these signals - together with the required signal - will have been amplified by 30 to 60 dB at this point, even very weak signals can cause a considerable cross-modulation.

It is thus advantageous to return to the single conversion superhet principle and to suppress all spurious signals after the first mixer by using a 9 or 10.7 MHz crystal filter. This means that cross-modulation can virtually only occur in the first mixer; however, the amplification is relatively low up to this stage.

Instructions with respect to the design of the RF amplifier, mixer and the coupling to the IF transformer are given. A simple home-made crystal filter is also described.

The following recommendations essentially deal with portable VHF receivers. However, if the IF section is extended to allow suitable switching of the IF bandwidth and if special attention is made to the variable frequency oscillator, there is no reason why the recommended method should not be used to construct a high-quality station receiver for 144 MHz. This concept requires, of course, that all UHF converters (for 432 MHz, 1296 MHz etc.) use an intermediate frequency of 144 to 146 MHz.

1. CONVENTIONAL RECEIVERS

The majority of todays efficient VHF receivers operate according to the principle shown in Fig. 1. As can be seen in this block diagram, the 144 MHz signal is transposed in a crystal controlled converter to the first IF which is normally in the order of 30 MHz. This is often followed by a further conversion to approximately 3 MHz, from where the final conversion to around 455 kHz is made. The advantages of this rather extensive principle are firstly that a good selectivity can be easily achieved at the low final IF and secondly that a good stability may be simply obtained because the tuning is made at approximately 30 MHz.

The main disadvantage of this method is that unwanted signals will sometimes not be suppressed until after the final conversion. Since these interfering signals will have been amplified by approximately 30 to 60 dB at this point, they can cause severe cross-modulation, especially in the third mixer stage. This shows clearly that the mere provision of field effect transistors in the VHF converter cannot affect a great improvement in the behaviour under large-signal conditions.

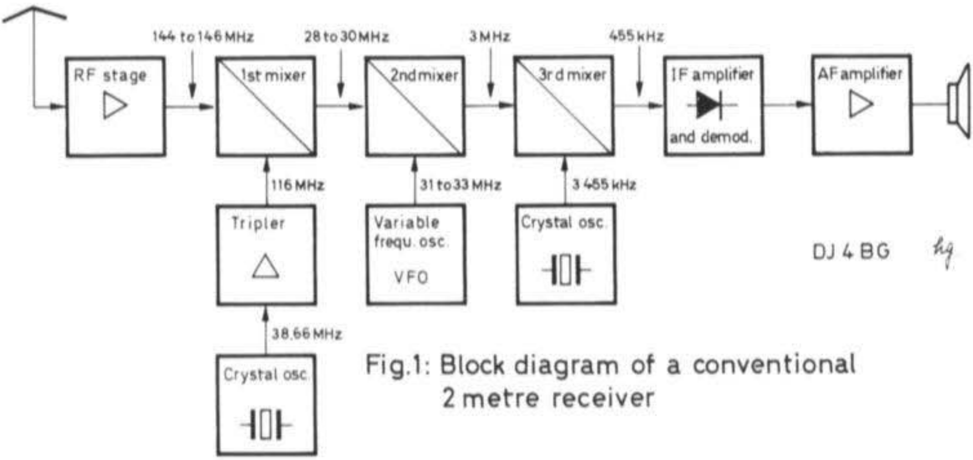


Fig. 1: Block diagram of a conventional 2 metre receiver

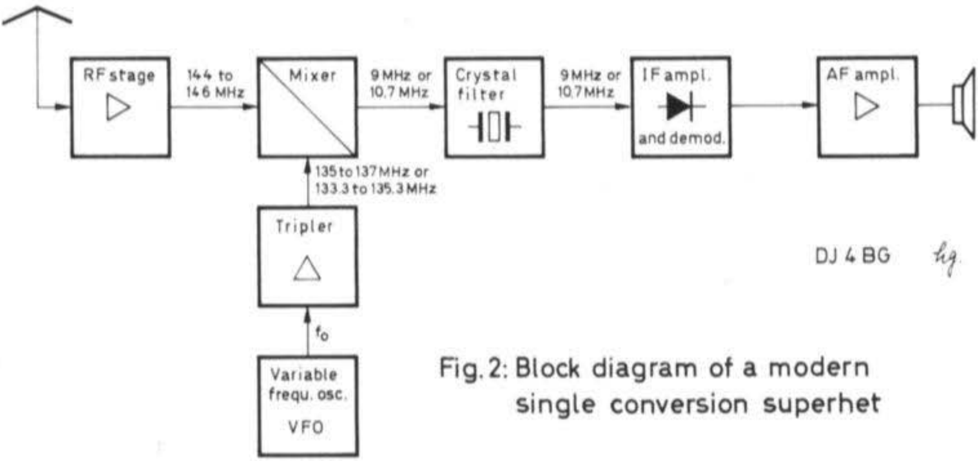


Fig. 2: Block diagram of a modern single conversion superhet

2. THE RECOMMENDED PRINCIPLE

Since it has been determined that cross-modulation only occurs when the unwanted signals have reached a relatively high level, it can be seen that these signals may not be amplified with the required signal. This means that the whole selectivity must be made directly after the first mixer. Further conversion is then superfluous. The use of this "obsolete" single conversion principle is greatly assisted by modern crystal filters in the 9 MHz or 10.7 MHz range. Of course, new developments should also include the advantages of modern silicon and field effect transistors.

As can be seen in Fig. 2, the principle of the single-conversion VHF receiver is to transpose the required two metre signal with the aid of a stable, variable oscillator frequency to a fixed intermediate frequency of 10.7 or 9 MHz. The mixer is followed by a narrow band filter (crystal filter). The crystal filter provides a skirt selectivity of approximately 80 to 100 dB, according to type. Subsequent IF transformers are used more for coupling than for reasons of selectivity and only affect the ultimate attenuation. The bandwidth of the crystal filter should not be too narrow - 10 to 15 kHz is sufficient - since this could lead to some difficulties when tuning during mobile and portable operation. Such a bandwidth should also be chosen in anticipation of the up-and-coming narrow band frequency modulation (NBFM) on 70 cm and higher bands.

Only the required signal will be present subsequent to the crystal filter, which means that no cross-modulation can occur in the IF amplifier. It is now advantageous to equip the receiver with field effect transistors in the RF and mixer stages so that the overall cross-modulation characteristics are good.

2.1. THE MIXER

The mixer is the most critical stage with respect to overloading. Many different ideas have been brought forward regarding this stage. One of these is the push-pull type arrangement using two field effect transistors which is said to be extremely favourable. Another is the push-pull diode mixer with which good results have been experienced. In professional electronics, the ring modulator type mixer has proved itself to be very favourable. Since double gate MOSFET's are now available at reasonable prices, it is considered that they will be very popular in mixer circuits in the near future. These transistors possess an especially high cross modulation rejection, conversion gain and sensitivity as well as allowing an isolation between the input and oscillator voltage.

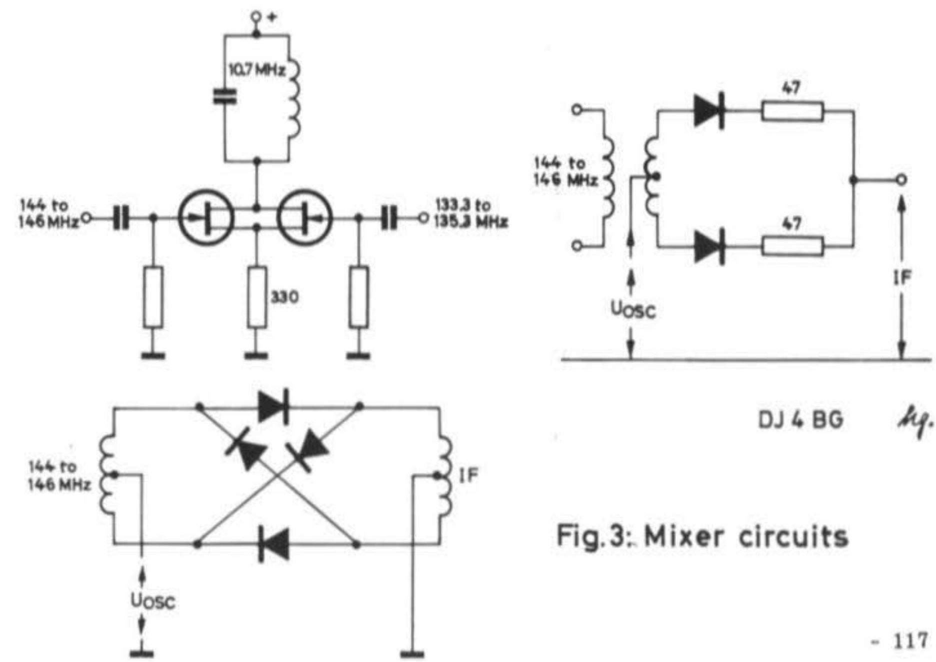


Fig.3: Mixer circuits

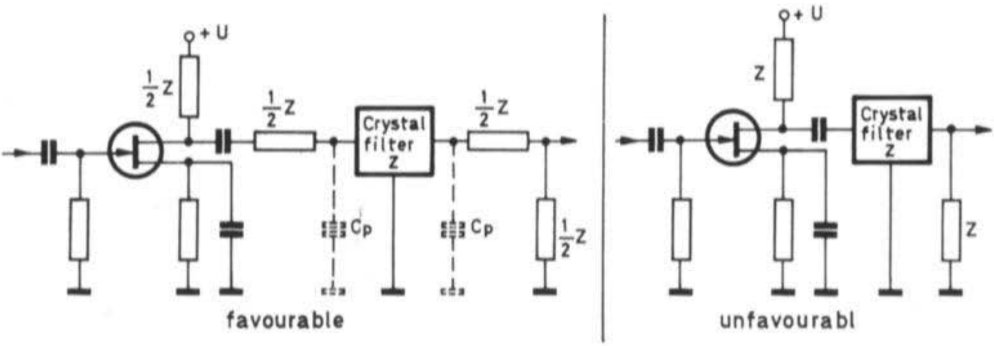
All three of the circuits shown in Fig. 3 allow unwanted signals of even 100 mV at the antenna to be processed as long as the gain of the RF amplifier stage is not too great and if the mixer stage is directly followed by a narrow band crystal filter to ensure that only the required signal is amplified.

2. 2. THE IF FILTER

Economical crystal filters are available to provide the necessary selectivity at IF level. Japanese types are even cheaper and even though their specifications are slightly poorer, they are still suitable for such applications.

It may be advantageous when using industrial crystal filters to place an intermediate stage between mixer and filter which must be able to handle the wide-band signal. Such a common-collector stage should be equipped with an RF transistor having a wide drive range, or better still a field effect transistor. The intermediate stage must be adjusted for the widest drive range and for most symmetrical slope, i. e. the voltage drop across the emitter resistor should be equal to half the operating voltage. Figure 4 gives examples of such circuits. C_p is the parallel capacitance prescribed by the manufacturer of the crystal filter. The intermediate stage, equipped with a transistor in common collector configuration, must be driven from a low impedance source.

If an industrial crystal filter is not to be used, it is possible to build up a simple crystal filter using a single quartz crystal. The 3 dB bandwidth of the filter configuration shown in Fig. 5 is approximately 5 to 10 kHz. As many coil filters (IF transformers) as possible whould follow to ensure that the image frequency is sufficiently suppressed.



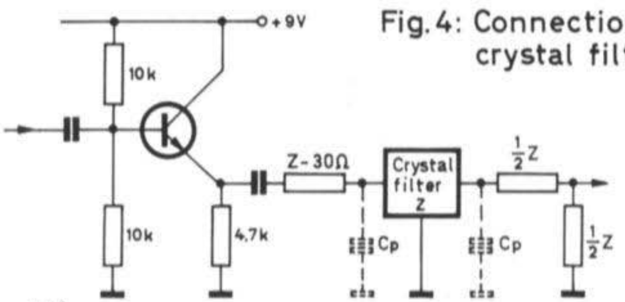


Fig. 4: Connection of manufactured crystal filters

DJ 4 BG hq.

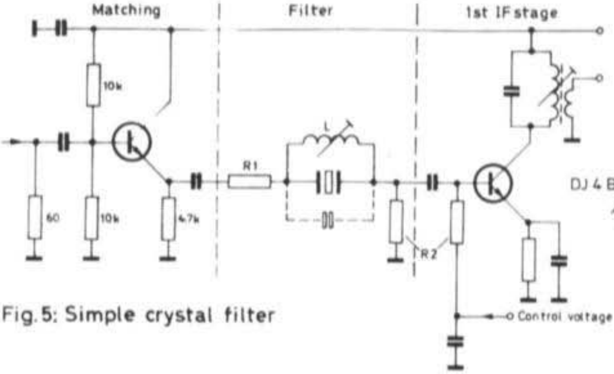


Fig. 5: Simple crystal filter

It is, of course, possible to make a second conversion to a lower intermediate frequency, for instance, to 455 kHz. This leads, it is true, once again to a double conversion superhet but the design is far more favourable. Due to the crystal being placed directly at the input of the IF chain, all unwanted signals will be attenuated by 10 to 30 dB with respect to the required signal, according to their spacing from the centre frequency (see Fig. 6). This means that the cross modulation rejection is increased. However, this configuration is a compromise. Unfortunately, it is not possible to give a detailed description of this simple filter because the source and load impedance required for a certain bandwidth sometimes deviates considerably between the various crystal manufacturers. In order to align the bandwidth of such a crystal filter, it is necessary to have a signal source with a tuning range of ± 30 kHz around the centre frequency. A suitable signal generator can be easily constructed.

More extensive filter configurations, i.e. bridge filters, require an alignment which cannot be carried out using the measuring equipment normally available to amateurs. The described inductively-neutralized crystal filter requires, on the other hand, hardly any alignment. As long as the source and load impedance are known, it is only necessary to align the filter to the centre frequency (for instance, 10.7 MHz) by adjusting the parallel inductance. The only measuring equipment that is required will be a dip meter.

A great improvement in the selectivity can be achieved by using a second filter of the same type which is isolated from the first by an intermediate IF stage. This allows any unwanted pass bands caused by spurious resonances of the crystal to be greatly attenuated. The selectivity curves calculated for various crystal filter designs are shown in Fig. 6. The spurious resonances are only shown as an example since they vary greatly both in amplitude and spacing from the centre frequency according to crystal type and load impedance. It can be seen that the increase of selectivity obtained from a narrow bandwidth is not so great as one might assume, whereas the configuration using two crystals offers more favourable results.

Regarding other parts of the IF chain, it should only be stated that field effect transistors have proved themselves to be very favourable since they can be voltage-controlled (practically no gate current). Several economical integrated circuits are now available which are suitable for applications at IF level. These integrated circuits have the advantage that their low reaction does not require neutralization. Favourable results can also be obtained using IF amplifiers in a cascode configuration.

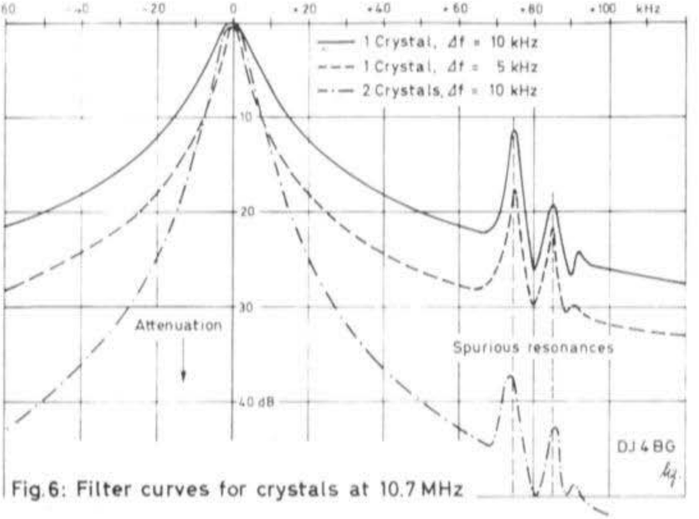


Fig.6: Filter curves for crystals at 10.7 MHz

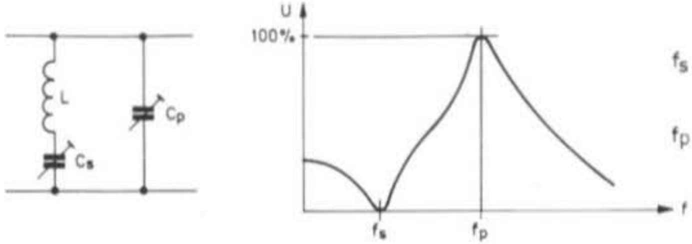
2.3. THE RF AMPLIFIER

The high cross modulation rejection of field effect transistors has already been discussed. This means that they are also suitable for use in RF amplifier stages. However, since the amplification of most junction field effect transistors is too small in the gate configuration, it is usually necessary to use a cascode or neutralized common source circuit. The experts are still not sure which circuit offers the best results. Metal oxide silicon field effect transistors (MOSFET) are now available on the market which are very suitable for VHF applications. The transconductance of such transistors (e.g. 3N140, TA 7153) is far greater than junction field effect transistor types (JFET) such as BF 245 etc. and allows the RF amplifier to comprise a single MOSFET in a common source circuit.

It is adviseable to provide the 144 MHz input circuit with a trap having series resonance at the image frequency. The series resonance f_s will appear at the frequency determined by L and C_s in Fig. 7, whereas the parallel resonance f_p is determined by L and the series configuration of C_s and C_p. For alignment, the series resonance is firstly selected by adjusting the minimum to correspond to the image frequency. This is followed by aligning C_p to the parallel resonance, that is for maximum in the two metre band.

2.4. THE VARIABLE FREQUENCY OSCILLATOR

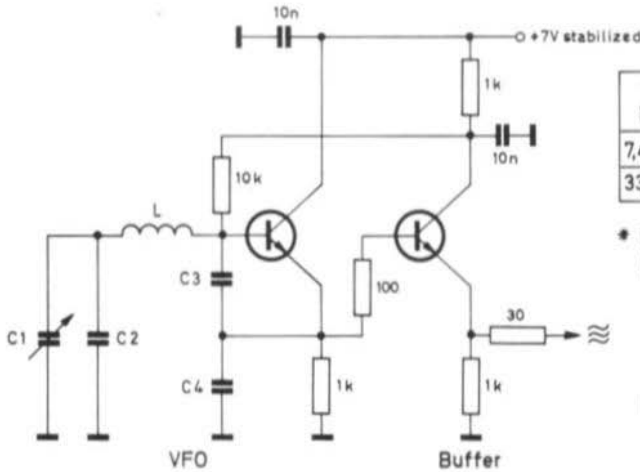
The heart of the receiver is the VFO, on which especially high demands are placed. The author recommends an oscillator having a fundamental frequency in the range of 7 to 40 MHz. The oscillator signal can then be multiplied to the required final frequency. A clapp circuit as shown in Fig. 8 has been found satisfactory. The inductance should be made from enamelled copper wire which is tightly wound onto a glass or ceramic rod, fixed into place with an epoxy resin adhesive (Araldit, UHU-Plus etc.) and subsequently aged artifically. A good VHF variable capacitor should be used for tuning, preferably a dual stator-type not having a sliding contact on the axle.



$$f_s = \frac{1}{\sqrt{L \times C_s}}$$

$$f_p = \frac{1}{\sqrt{L \times \frac{C_p \cdot C_s}{C_p + C_s}}}$$

Fig. 7: Resonant circuit with image frequency trap

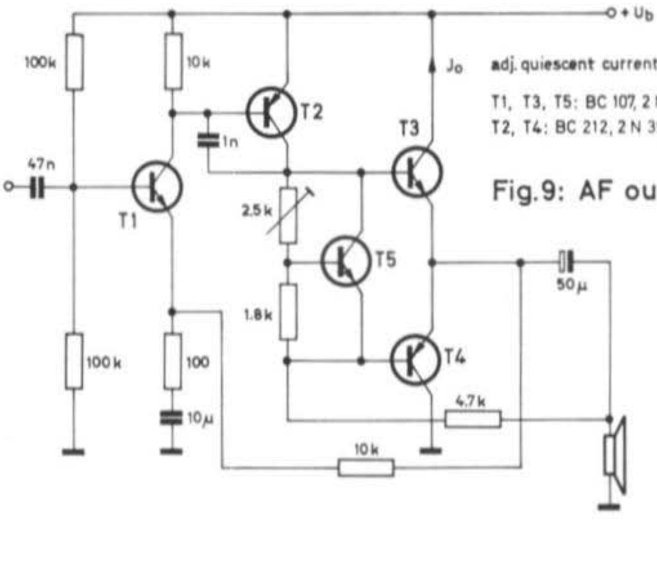


f MHz	C1 pF	C2* pF	C3 pF	C4 pF	L µH
7,4-7,52	2/14	230	1000	2000	2,5
33,3-33,8	2/14	120	500	200	0,34

* Form C2 by 2 or 3 capacitors so that the temperature compensation can be more easily carried out

DJ 4 BG

Fig. 8: Clapp-VFO



adj. quiescent current to 0.5 mA

T1, T3, T5: BC 107, 2 N 3903 or similar NPN
T2, T4: BC 212, 2 N 3905 or similar PNP

Fig. 9: AF output stage

DJ 4 BG

The resonant circuit must be carefully temperature compensated. It is also necessary for the operating voltage to be well stabilized using a transistor control circuit or with zener diodes (preferably temperature compensated). If possible, it is adviseable to enclose the VFO in a small casing made out of at least 2 mm thick aluminium plate so that any temperature variations have a slower and more equal effect on the individual resonant circuit elements. The silicon transistors used in the oscillator circuit should have a transit frequency at least 10 to 20 times higher than the oscillator frequency. In addition they should have low intrinsic capacity. Transistor types 2SC185, 2N918 (BFY 66), BF 173 and BF 224 are very suitable for such applications. Taking this into consideration, the author built up a variable frequency oscillator with a frequency range of 7.4 MHz to 7.52 MHz whose stability is equal to that of a crystal controlled oscillator.

Of course, it would be possible to improve the frequency stability still further by using a sythesis or phase-locked oscillator (1), however, this will not be required for reception of AM, FM or CW signals. The advantages of the slight improvement in frequency stability are often outweighed by the production of spurious signals.

A very simple and room saving method of tuning the oscillator is to use a varactor diode. It is, however, necessary to take the high temperature dependence of the junction capacity into consideration. Several diodes can be connected to the same forward voltage supply so that the RF amplifier tuning can track the oscillator. This allows the bandwidth of the RF stages to be kept far narrower which also allows an improvement of the image and cross modulation rejection. The bias voltage for the varactor diodes must be carefully stabilized against supply voltage fluctuations and temperature variations. In addition to this, it is adviseable to let the receive oscillator run all the time, even during transmit.

2.5. THE AF AMPLIFIER

A small transformerless class B amplifier is especially suitable for the AF output stage. Figure 9 shows a circuit suitable for an intermediate impedance (150 Ω) loudspeaker. Since no output or driver transformers are required, an extremely small and light-weight output stage is obtained. The temperature compensation for the small quiescent current is especially effective due to the auxiliary transistor T 5. The output power into a 150 Ω loudspeaker is 100 mW at an operating voltage of 12 V; at a voltage of 18 V an output power of up to 250 mW is obtained. In the experience of the author, this power output is completely sufficient for portable VHF receivers.

This paper was brought forward by the author at a meeting of the Czechoslovakian radio amateurs in Joachimstal (CSSR).

3. REFERENCES

K.P. Timmann and V. Thun: Phase-locked Oscillator for Transmit and Receive Mixers in Amateur Radio Equipment.
VHF COMMUNICATIONS 1 (1969), Edition 1, pages 11 - 25.

THE 70 MHz DL 6 SW FET CONVERTER

by D. T. Hayter, G 3 JHM

The original idea of converting a DL 6 SW converter (1) for operation on the four metre band (70.025 to 70.7 MHz) originated from Mr. Heinz Stellberg of Königswinter, West Germany. The author decided to attempt this modification and the result was a small, efficient 70 MHz converter with an IF output of 28 to 28.7 MHz.

It was not necessary to vary any component values but only to use differently dimensioned inductances and a 42.000 MHz third overtone crystal.

Field effect transistors type TIS 88 can be used instead of the TIS 34 types if the neutralizing coils are slightly modified to compensate for the reduced reverse capacitance.

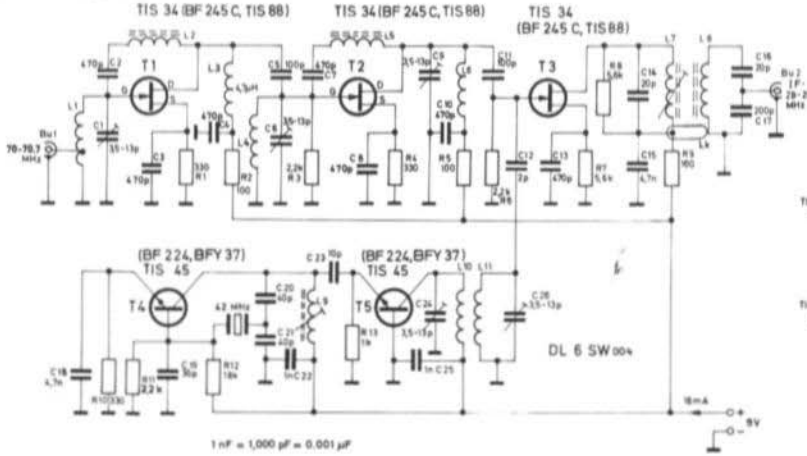


Fig. 1 Circuit Diagram of the 70 MHz Field Effect Transistor Converter

The required component modifications are as follows:

L 1, L 4, L 6	8 turns of 1 mm dia. (18 AWG) enamelled copper wire. Self-supporting, 10 mm inner diameter; close wound.
L 2, L 5, L 7, L 8	20 turns of 0.3 mm dia. (29 AWG) enamelled copper wire close wound on a 4 mm coil former with VHF core.
L 3	4.7 µH RF choke
L 9	15 turns of 0.3 mm dia. (29 AWG) enamelled copper wire close wound on a 4 mm coil former with VHF core.
L 10, L 11	12 turns of 0.8 mm (20 AWG) enamelled copper wire. Self-supporting, 10 mm inner diameter; close wound.

Quartz crystal 42.000 MHz, third overtone.

Editorial note

We would like to congratulate Mr. D. T. Hayter on his second place in a recent 70 MHz contest where the modified prototype converter was used to good effect. We hope to bring a modification for the 50 MHz band in the near future.

REFERENCES

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

A COAXIAL RELAY WITH A HIGH COUPLING ATTENUATION AND GOOD SWR

by E. Berberich, DL 8 ZX

The coaxial relays used by the majority of amateurs have such a low coupling attenuation at VHF that it is possible to measure up to 50 mW of RF energy at the receiver when transmitting a 10 watt signal at 145 MHz. This example, which has been calculated at a coupling attenuation of 23 dB, shows that even the low power of 10 watts can endanger the input transistor of the receiver. The author describes a coaxial relay assembled using reed contacts. The characteristics are very good.

1. REED CONTACTS

Reed contacts are available from a great number of manufacturers of telecommunications equipment. These modern components possess a capacitance of approximately 0.2 pF. Working contacts have been found to be more suitable for RF applications since change-over contacts have two parallel tongues which increase the coupling capacitance and thus reduce the coupling attenuation.
It is, of course, also necessary to actuate the transmit-receive relay during reception if two working contacts are used.

The contact tongues are made from a soft magnetic material which means that they can be actuated by either an electro-magnetic field (Fig. 2) or by a slideable permanent magnet (Fig. 3) mechanically coupled to the transmit-receive switch. The latter is especially suitable for low-power battery driven equipment.

2. MECHANICAL ASSEMBLY

In order to obtain the highest possible coupling attenuation and the lowest possible standing wave ratio, the reed contacts are enclosed in a brass tube, which also serves as a support for the coil former.

The inner diameter of the tube should be selected to obtain the lowest possible SWR; the outer diameter is selected to suit the coil former.
However the tubing should not be too thick, otherwise the air gap will be too great. This means that the number of ampere turns would have to be increased to ensure correct switching. Coil formers of transformers found in miniature transistor driver stages were used.

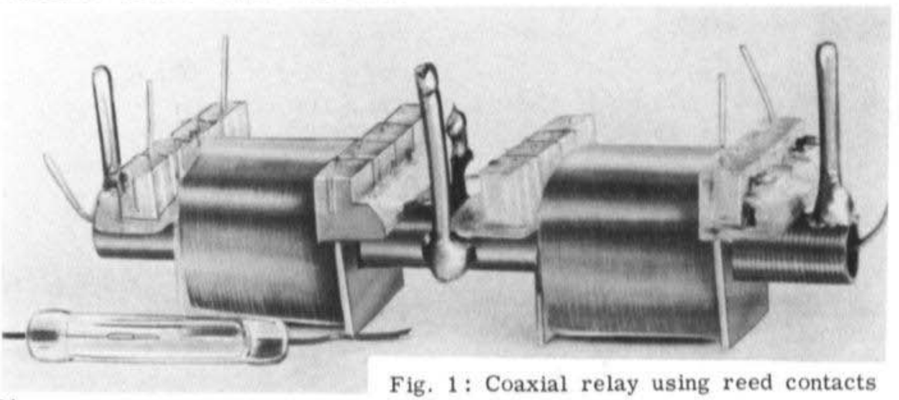


Fig. 1: Coaxial relay using reed contacts

The authors prototype and an individual reed contact are shown in Fig. 1. The dimensions are given in Fig. 4 to allow the transmit-receive relay to be copied and assembled. If printed-circuit coil formers are used, it is possible for instance, to mount the relay as an integral part of the transmitter or receiver board.

Coil data: Coil former type M 22. 6000 turns of 0.07 mm dia. (40 AWG) enamelled copper wire. Operating voltage 12 V Current 12 mA

or 12,000 turns of 0.063 mm dia (42 AWG) enamelled copper wire wound on a coil former suitable for PC-boards. Current flow per winding amounted to 6.5 mA at 12 V; or 3.5 mA at 6 V where the relay still continued to operate correctly.

Reed contacts: Types MRR 2 or MRMF 2 manufactured by Hamlin or type DCR from Hathaway
Required energization is approximately 50 ampere turns.

3. MEASURED VALUES

The relay was tested with regard to its service life by connecting it to a multivibrator. The change-over frequency was 4 Hz and the load was 10 W into 60Ω at 145 MHz. The test was ceased after 20,000 operations. No deterioration of the contacts could be determined.

The coupling attenuation between the transmitter and receiver socket was 43 dB at 145 MHz and 33 dB at 432 MHz. The receiver connection was terminated with 60 Ω during this measurement. No deterioration of the standing wave ratio was noticed even at 432 MHz.

These values could possibly be improved by using reed contacts possessing a lower contact capacitance, such as the Hathaway type DCR with less than 0.1 pF.

This coaxial relay was used in the Bavarian Winter Field Day at a height of approximately 5000 ft. No difficulties were encountered whatsoever.

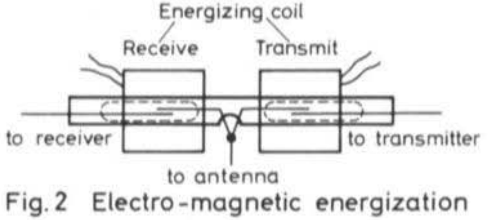
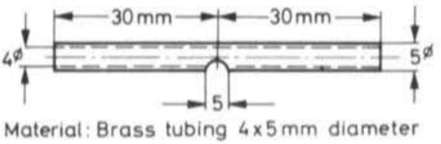


Fig. 2 Electro-magnetic energization



Material: Brass tubing 4x5 mm diameter

Fig. 4 Dimensions of the brass tubing

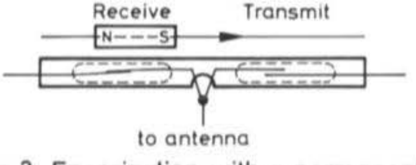


Fig. 3 Energization with a permanent magnet

DL 8 ZX

MATERIAL PRICE LIST OF KITS and COMPONENTS available from the publishers of VHF COMMUNICATIONS or their national representatives

This price list is valid for all orders reaching us after December 1st 1970. All previous price lists are invalid after this date.

We were able to considerably reduce a great number of prices due to:

Standardization of components (especially semiconductors)
resulting in a more favourable purchase rate
Rationalization of the kits and partial kits
Some (unfortunately few) price reductions of the manufacturers.

A number of new components (marked +) may not be available until January 1971. Kits including these parts will be dispatched immediately so that you can commence construction and the missing parts will be supplied without re-ordering as soon as available.

Please read the Sales Conditions on the last page of the price list.

If you are not a reader of VHF COMMUNICATIONS please request full advertising material.

CRYSTALS and CRYSTAL FILTERS for equipment described in VHF COMMUNICATIONS that are available ex stock. Only highest quality crystals are offered.

Crystal filter	XF-9A	(for SSB) with both sideband crystals	DM 106.--
Crystal filter	XF-9B	(for SSB) with both sideband crystals	DM 137.--
Crystal filter	XF-9C	(for AM; 3.75 kHz)	DM 137.--
Crystal filter	XF-9D	(for AM; 5.00 kHz)	DM 137.--
Crystal filter	XF-9E	(for FM; 12.00 kHz)	DM 137.--
Crystal filter	XF-9M	(for CW; 0.5 kHz) with carrier crystal ...	DM 106.--
Crystal discriminator	XD 09-03	matching XF-9E	DM 78.--

Crystal	96.0000 MHz	(HC- 6/U)	for 70 cm converters	(DL 9 GU, DL 9 JU)	DM 21.50
Crystal	96.0000 MHz	(HC-25/U)	for 70 cm converters	(DL 9 GU, DL 9 JU)	DM 28.--
Crystal	95.8333 MHz	(HC-18/U)	for 70 cm converters	(DC 6 HY)	DM 28.--
Crystal	84.5333 MHz	(HC- 6/U)	for 24 cm converters	(DL 3 WR)	DM 21.50
Crystal	65.5000 MHz	(HC- 6/U)	for 2 m converters	(DL 6 HA)	DM 16.50
Crystal	65.0000 MHz	(HC- 6/U)	for 2 m converters	(DL 6 HA)	DM 16.50
Crystal	46.3333 MHz	(HC-18/U)	for phase-locked oscillator	) set	DM 49.--
Crystal	46.0000 MHz	(HC-18/U)	(DJ 7 ZV / DJ 9 ZR)	)	
Crystal	45.4780 MHz	(HC-18/U)	for VXO	(DJ 9 ZR)	DM 24.50
Crystal	42.0000 MHz	(HC- 6/U)	for 70 MHz converters	(G 3 JHM)	DM 15.60
Crystal	38.6667 MHz	(HC- 6/U)	for 2 m converters	(DL 6 SW, DL 6 HA)	DM 13.70
Crystal	27.8000 MHz	(HC-18/U)	for 24 MHz synthesis VFOs	(DL 3 WR 007)	DM 25.--

Standard frequency crystals

5.0000 MHz	(HC- 6/U)	for calibration spectrum generators (DC 6 HY)	DM 25.--
1.0000 MHz	(HC- 6/U)	for calibration spectrum generators (DJ 4 BG, DC 6 HY)	DM 20.50
100 kHz	(HC-13/U)	for calibration spectrum generators (DC 6 HY)	DM 28.--

Crystals 72.... MHz (HC-6/U) for 2 metre transmitters (DJ 1 NB, DL 3 WR)
Please state required frequency on ordering
(Delivery 4 to 6 weeks) .... DM 21.50
Crystals also available ex stock please see list with DJ 1 NB kit.

Crystals other frequencies available (Please state frequency and type)
(Delivery 6 to 8 weeks)

KITS

2 m FET converter DL 6 SW 004 (PC-board, coil formers, trimmers, transistors, 38.6667 MHz quartz crystal and 1.3 mm special drill)	DM 42,60
Transistors for DL 6 SW 004	DM 17,90
Quartz crystal 38.6667 MHz (HC-6/U)	DM 13,70
70 MHz FET converter DL 6 SW / G 3 JHM (PC-board, coil formers, trimmers, transistors, 42.0000 MHz quartz crystal and 1.3 mm special drill)	DM 44,50
Transistors for 70 MHz converter	DM 17,90
Quartz crystal 42.0000 MHz (HC-6/U)	DM 15,60
Phase-locked oscillator DJ 7 ZV 001 + 002 (2 PC-boards, coil formers, transistors, diodes, 2 quartz crystals and 1.3 mm special drill)	DM 103,95
Transistors and diodes for DJ 7 ZV 001 + 002	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 request
70 cm converter DL 9 GU 001 (Silver-plated metal casing as above, PC-board, coil formers, trimmers, transistors, 96.000 MHz quartz crystal and 1.3 mm special drill)	DM 130,--
Transistors for DL 9 GU 001	DM 37,40
96.000 MHz quartz crystal (HC-6/U)	DM 19,50
5 W SSB transmitter DJ 9 ZR 001 (PC-board, coil formers, trimmers, Ferroxcube-Drosseln, transistors, diodes and 1.3 mm special drill)	DM 138,40
Crystal filter XF-9A with both sideband crystals	DM 96,--
Crystal filter XF-9B with both sideband crystals	DM 125,50
Transistors and diodes for DJ 9 ZR 001	DM 116,90
VXO DJ 9 ZR 002 (PC-board, coil formers, transistors, diodes)	DM 18,45
Transistors and diodes for DJ 9 ZR 002	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

RADIO COMMUNICATION HANDBOOK	(including postage)	DM 38,--
VHF-UHF-MANUAL	(including postage)	DM 15,20

All orders must be prepaid by cheque (check) or money order made payable to the bank account of the national representative, or to the publishers

UKW-BERICHTE, Hans J.Dohlus, DJ 3 QC - D-8520 ERLANGEN,
Gleiwitzer Strasse 45 - Western Germany

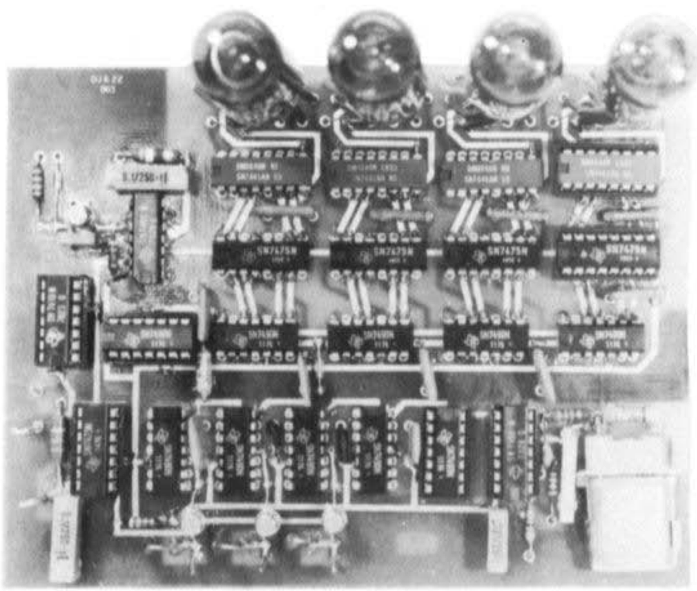
Bankaccount: Deutsche Bank Erlangen, No.476 325
Postscheck: Nürnberg 30 4 55

PLASTIC BINDERS FOR VHF COMMUNICATIONS

- Attractive plastic covered in VHF blue
- Accepts up to 12 editions (three volumes)
- Allows any required copy to be found easily
- Keeps the XYL happy and contented
- Will be sent anywhere in the world for DM 5.50 including surface mail.

Order your binder via the national representative or direct from the publishers of VHF COMMUNICATIONS:

VERLAG UKW-BERICHTE, H.J.Dohlus oHG
D-8520 ERLANGEN (West Germany)
Account: Deutsche Bank Erlangen 476 325



PRICE REDUCTION!

Due to the popularity of the DJ 6 ZZ 003 frequency counter we have been able to reduce the price of the printed circuit board and kit to DM 48.-- and DM 340.-- respectively.

Now is your chance of building a completely fool-proof 30 MHz counter for an extremely favourable price. Double coated PC-board with through contacts, no drilling or bridges necessary. First counter IC selected for a minimum of 30 MHz.