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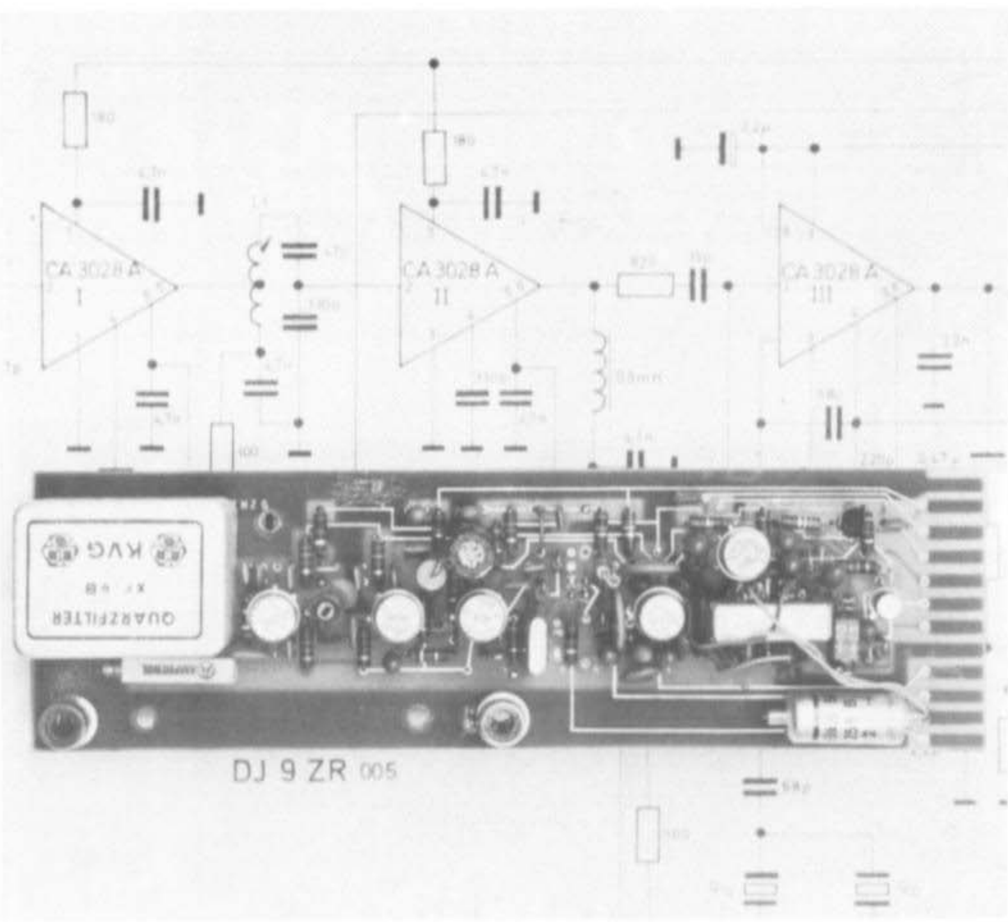

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THE EFFECT OF THE PRINTED CIRCUIT BASE MATERIAL ON THE Q OF PRINTED INDUCTANCES

by K.P. Timmann, DJ 9 ZR

Information allowing the Q and inductivity of printed inductances is given in (1) and (2). When manufacturing such inductances, it should be noted that the dielectric constant ε, the dielectric loss factor tg δ and the thickness of the material essentially influence the Q. If the dielectric constant is increased, this not only causes the conductor lane to be loaded by the capacitive current, but also means that the dielectric loss of the inductances will be increased; in the case in question as the product of the dielectric constant ε and loss factor tg δ.

The results of several experiments made by the author are now given to show the influence of loss effects.

1. A spiral printed inductance with three turns according to Fig. 1 will have an inductivity of 0.15 μH if the following specifications are maintained: Outside radius 1.3 cm, conductor lane width 0.8 mm, turn spacing 1.6 mm, conductor material copper, length of the conductor lane 11 cm.

1.1. According to (1), page 11, Fig. 3, point 5, a conductor resistance of 4 Ω per metre will result at a frequency of 150 MHz, including the skin effect. As a result of the current displacement towards the conductor edges, this value will be increased to 5 Ω per metre.

A radiation loss of 20% must additionally be assumed. The following calculations will result:

Loss impedance	$R = r \times 1.2 \times 0.11\ m = 0.66\ \Omega$
Reactive impedance	$X = \omega_L = 140\ \Omega$
Q	$Q = X/R = 140/0.66 = 215$

1.2. An inductance according to the above specifications was built up on an epoxy printed circuit board. The most important values of this material are:

Dielectric constant	= 5.5
Dielectric loss factor tg δ	= 10^{-2}
Measured Q of the inductance	= <u>120</u>

1.3. A further inductance having the same dimensions was built up on copper-coated teflon (P.T.F.E.). The characteristics were:

Dielectric constant	=	2.1
Dielectric loss factor tg δ	=	10^{-4}
Measured Q	=	<u>200</u>

2. A rectangular printed inductance with three turns was manufactured according to Fig. 2. The conductor lane width is 0.8 mm, conductor spacing is 1.6 mm and the conductor length is 16 cm; the inductance of this coil was measured to be 0.19 μH. According to (1), the following data can be calculated at 150 MHz.

Loss impedance	R = 0.96 Ω
Reactive impedance	X = ω_L = 170 Ω
Q	Q = <u>175</u>

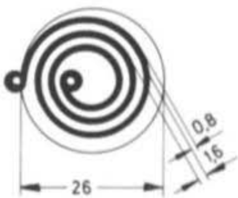


Fig. 1 : Spiral printed inductance with 3 turns

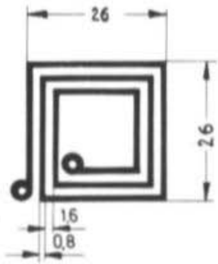


Fig. 2 : Rectangular printed inductance with 3 turns

A coil having the given dimensions was etched on a copper-plated teflon (P.T.F.E.) board. The Q was measured and found to be approximately 175 which practically corresponds to the calculated value.

The values given in (1) show that the spiral printed inductance will have a somewhat higher Q than the rectangular type assuming that the inductivity, conductor width and spacing are identical. The measured values show that the values given in the references are maintained in practice as long as a sufficiently low-loss base material is used.

REFERENCES

(1) Radio Engineering Handbook, 1968 Edition, Chapter 5-22, Fig. 38, 39.

(2) Meinke/Gundlach, Taschenbuch der Hochfrequenztechnik
2. Auflage, Berlin 1962, Page 19

A 145 MHz/9 MHz RECEIVE CONVERTER USING PRINTED INDUCTANCES

by K.P. Timmann, DJ 9 ZR

This 2 m converter was developed for use in a single conversion superhet having an intermediate frequency of 9 MHz. The advantages of this concept are the excellent large-signal characteristics (cross-modulation, intermodulation rejection) and the use of popular and inexpensive 9 MHz crystal filters. A detailed description of this concept was given in (1).

A variable frequency of 135 to 137 MHz is needed to convert the 2 metre signal to a fixed intermediate frequency of 9 MHz. The phase-locked oscillator described in (2) and the VXO given in (3) are especially suitable for this purpose. This receive converter forms, together with the 9 MHz IF portion, the matching receiver for the 5 W SSB transmitter described in Edition 2/1969. The various sub-assemblies: phase-locked oscillator (or VXO), 5 W SSB transmitter, receive converter 145 MHz/9 MHz and 9 MHz IF portion can be combined to form a very modern 2 metre transceiver for both mobile and fixed operation.

1. DESCRIPTION OF THE CIRCUIT

The circuit diagram of the converter is given in Fig. 1. Since its operation is extremely simple, it is only necessary to explain certain details of the circuit.

1.1. MATCHING TRANSISTOR T 1 TO THE ANTENNA INPUT

The input impedance at gate 1 of transistor T 1 (RCA types TA 7153 or 40600) is approximately 3 kΩ at 150 MHz. In order to achieve power matching, it is necessary that the input circuit is also of this impedance. For best sensitivity, however, it is necessary for the transformation ratio of the input circuit to be made smaller. A feed impedance of 800 Ω is necessary for the noise matching.

A simple parallel resonant circuit, consisting of printed inductance L 1 (approx. 2 1/4 turns) and a ceramic trimmer, is used for transformation. The hot end of the resonant circuit is connected to gate 1 of transistor T 1. The tapping point for the antenna connection was found by experiment using a calibrated and matched noise generator. Since the tapping point is already provided on the PC-board, it is not necessary for the constructor to seek the most favourable tapping point.

1.2. DIMENSIONING OF THE TRIPLE-CIRCUIT FILTER

The calculation of the most favourable filter specifications, expecially the couple inductivities, is very time-consuming and does not indicate the coil dimensions and spacings. For this reason, three sample inductances possessing favourable characteristics and dimensions were made and arranged to obtain the required passband characteristics on the oscilloscope.

The loaded Q of the circuits comprising L 2 and L 4 is approximately $Q \approx 50$ in their non-coupled condition (non-load resonant impedance 34 kΩ, L 2 damped with 10 kΩ by the impedance of transistor T 1, L 4 damped at the capacitive voltage divider by the input impedance of transistor T 2. At the most favourable

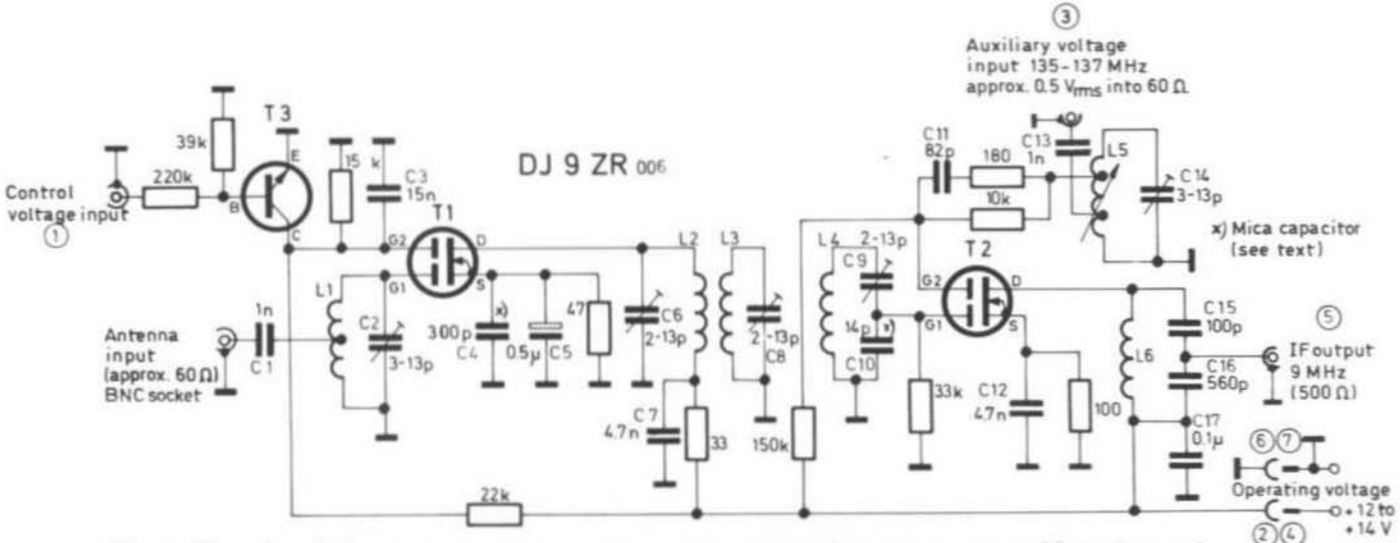
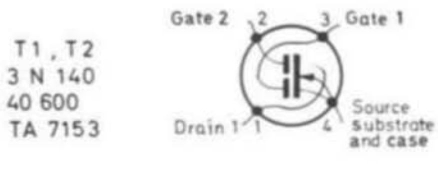
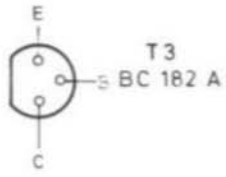


Fig. 1: Circuit of the receive converter with printed inductances (DJ 9 ZR 006)





mixing level, this impedance will be approximately 2.5 kΩ). Air spaced trimmers are used for tuning at this point to ensure a high stability and to maintain the high Q, especially of the intermediate circuit. Further details regarding the advantages of the intermediate circuit are given in (2).

The coupling of the circuits is selected so that the passband ripple is not greater than ± 1 dB in the passband range of 144 to 146 MHz, that an image rejection of over 55 dB is obtained and that the 9 MHz IF rejection is in the order of 110 dB.

The most important consideration allowing the high stopband attenuation of the filter to be achieved is the low capacitive interaction of the inductances as well as the inductive and capacitive coupling between the inductances L 2 and L 4. This is achieved firstly by the selected arrangement of the coils and secondly by the assembly of the board with a spacing of 20 mm parallel to a chassis plate. The ground conductor of the PC-board must additionally be low-inductively connected at 3 specific points (see Fig. 2) to the chassis plate, so that no spurious resonances can occur.

1.3. VFO-VOLTAGE INJECTION

A transformation is also required here because the input impedance of gate 2 of transistor T 2 is relatively high and, on the other hand, the feed cable for the oscillator voltage must be terminated with 60 Ω. This is also achieved with the aid of a resonant circuit equipped with the printed inductance L 5, which possesses two tapping points. The bias voltage for gate 2 is obtained with the aid of the voltage divider 10 kΩ - 150 kΩ. The oscillator voltage is obtained via the series circuit 180 Ω - 82 pF; the 180 Ω resistor is provided to avoid spurious mixing of the receive voltage and the oscillator voltage.

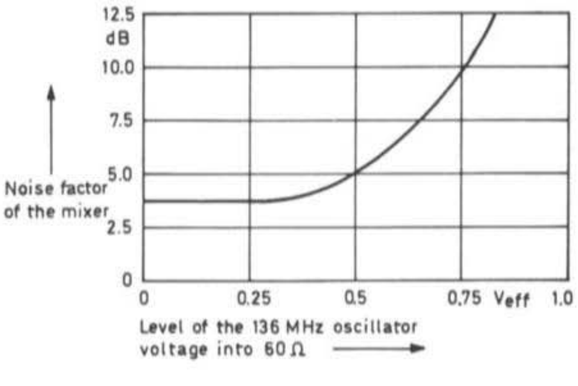


Fig. 4:
Noise factor of the mixer transistor (3N 140 or 40 600) as a function of the auxillary oscillator voltage at the 60Ω input

hg

It should be noted, that the conversion noise of transistor T 2 will increase together with the oscillator voltage, as is shown in Fig. 4. The most favourable value is obtained at approximately 0.5 V_{rms} across the 60 Ω impedance of the PC-board; at lower values, the noise will be reduced, but the conversion gain will simultaneously drop.

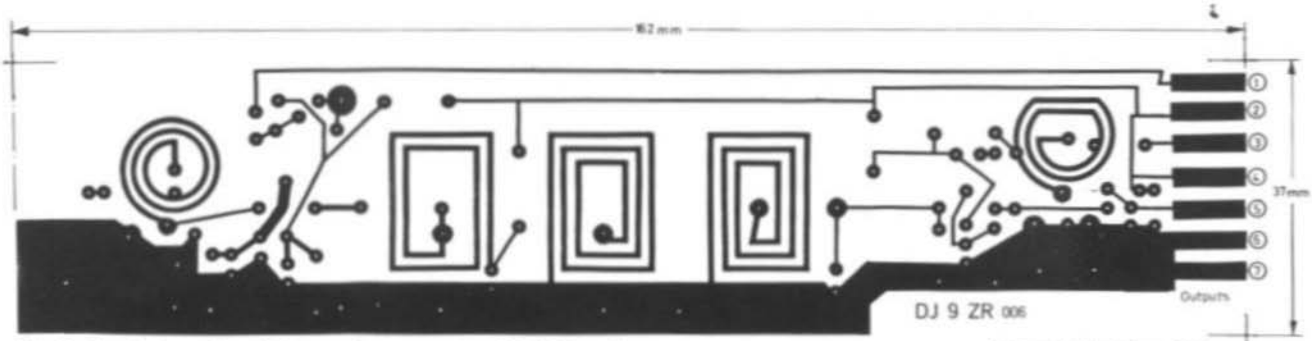


Fig. 3: Conductor side of the receive converter (DJ 9 ZR 006)

Connection points: (1) Control voltage input
(2)(4) Operating voltage +12 V,
(3) Aux voltage 135 - 137 MHz
(5) IF 9 MHz
(6)(7) Ground

37 x 162 mm
See Fig. 1 and 6

Base material: 1.5 mm thick Teflon (P.T.F.E.)

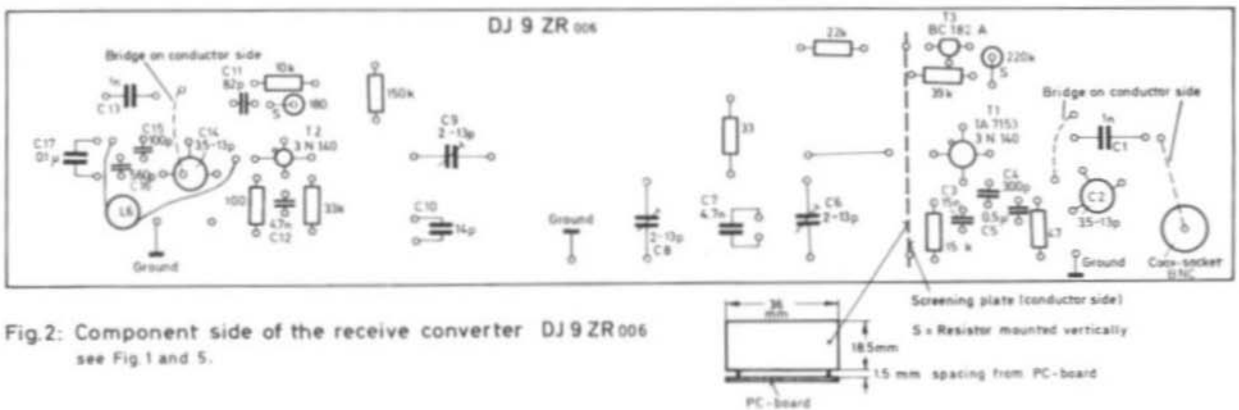


Fig. 2: Component side of the receive converter DJ 9 ZR 006
see Fig. 1 and 5.

1.4. MATCHING OF THE 9 MHz OUTPUT

The most favourable load impedance for the mixer transistor is approximately 3 kΩ; the nominal value for the IF output of the PC-board should amount to 500 Ω in the case in question, since the input impedance of the 9 MHz IF portion (crystal filter input) has this impedance value. The matching is made by transformation with a capacitive voltage division (100 pF - 560 pF) at the resonant circuit. The inductance L 6 comprises 25 turns of 0.1 mm (38 AWG) enamelled copper wire wound on a 4 mm coil former with core.

1.5. CONTROL RANGE OF THE CONVERTER

The overall gain of the converter amounts to approximately 18 dB; it can be reduced by approximately 15 dB by altering the voltage at gate 2 of transistor T 1. This variation range is completely sufficient for control purposes if the converter is used together with the described IF board. Transistor T 3 and the two voltage divider circuits 220 kΩ - 39 kΩ and 22 kΩ - 15 kΩ are provided to match the voltage range of gate 2 to the voltage range at the control amplifier output of the IF board. The circuit can be correspondingly modified for other control voltage ranges.

2. MECHANICAL ASSEMBLY

The component location plan and a diagram of the printed circuit board DJ 9 ZR 006 are given in Fig. 2 and 3. Since the constructor often encounters difficulties in maintaining the self-inductivity and coupling of the required inductances, the input inductors L 1 (145 MHz), the auxiliary frequency inductance L 5 (136 MHz) and the inductances for the triple-circuit filter L 2, L 3, L 4 (145 MHz) have been printed onto the printed circuit board. In order to guarantee a high Q, teflon has been selected as base material. The PC-board can be plugged in and removed if a matching connector is used. Connection points 1 to 7 are given in Fig. 3.

2.1. SPECIAL COMPONENTS

C 2, C 14 :	3.5 - 13 pF, ceramic disc trimmer 7 S - Triko 0.2
C 6, C 8, C 19 :	2 - 13 pF, air spaced trimmer 11 LJ 11 - 13/0.25
T 1 :	TA 7153, 40 600 or 3 N 140 manufactured by RCA
T 2 :	3 N 140 manufactured by RCA
T 3 :	BC 182 A manufactured by Texas Instruments or BC 108, 2 N 3904, 2 N 2926
L 6 :	see Section 1.4.

High quality ceramic disc capacitors having a very low intrinsic inductivity should be used for all other capacitors. Resistors having a rating of 1/10 W can be used.

BNC socket for the 145 MHz input.

2.2. MOUNTING THE COMPONENTS

The components given in Fig. 2 are mounted on the PC-board (Fig. 3) with the dimensions of 37 mm x 162 mm. The 14 pF capacitor C 10 at the capacitive voltage divider of the filter circuit comprising L 3 should be a mica capacitor so that the circuit Q is not unnecessarily reduced. Special care should be taken to ensure that the 300 pF capacitor C 4 at the source connection of transistor T 1 has a low inductivity. A mica capacitor or a ceramic multilayer capacitor is most suitable. Figures 5 and 6 show photographs of the printed circuit board from both sides. A brass screening plate (0.5 mm thick) is soldered between transistor T 1 and L 2 so that a stable operation is guaranteed. Between the soldering points, the screening plate must have a spacing of 1.5 mm to the PC-board and must protrude 18.5 mm past these points. It is possible, after this, to commence the alignment process.

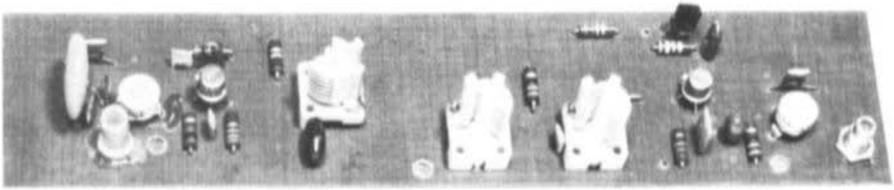


Fig. 5

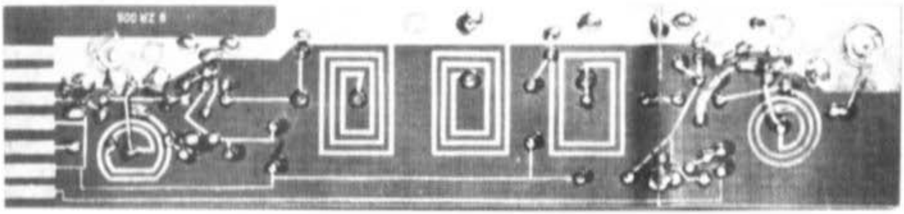


Fig. 6

3. ALIGNMENT

The auxiliary oscillator voltage (136 MHz) (3), a signal generator (145 MHz), the operating voltage (12 - 14 V) and the 9 MHz IF portion are connected to the corresponding points of the printed circuit board. All 6 resonant circuits are aligned for maximum 9 MHz signal. During this alignment, it is not necessary to damp the triple filter or to carry out the alignment at alternative frequencies.

4. TEST RESULTS

Four of these receive converters have been operated for some time now. They were built-up by various amateurs (in DL, G 3 and OK) according to the given assembly instructions and were all found to have the following measured values. The same measuring instruments were used for this measurement, as are given under (1).

a) Noise factor: 3.0 dB typical, 3.3 dB maximum (T 1 = TA 7153)
4.0 dB typical, 4.5 dB maximum (T 1 = 3 N 140)
Transistor type 3 N 140 was chosen as mixer transistor T 2 during both measurements. The R-series of transistor type 40 600 is not recommended, since the specifications vary greatly between individual transistors.

b) Overall gain; 18 dB typical, at least 15 dB; this will be sufficient when using a good IF amplifier.

c) Passband ripple: ± 1 dB.

d) Image rejection: 57 dB.
IF rejection of 9 MHz signals: 110 dB.

e) Input impedance: approx. 60 Ω.
The converter is also stable when no antenna is connected.

f) Output impedance: 500 Ω ± 10%

g) Cross-modulation: 20 mV typical, at least 18 mV of an interfering signal to obtain 10% cross-modulation on a wanted signal of 0.2 μV. Frequency spacing between the wanted and the interfering signal: 100 kHz.

h) Desensitization (reduction of the wanted signal by 3 dB): approximately the same levels as given under g).

If the subsequent 9 MHz IF portion is equipped with a steep-skirted filter at the input, the values given under g) and h) will be valid for the whole receiver.

5. AVAILABLE PARTS

The printed circuit board, various individual components and the whole kit of parts are available from the publishers or their national representatives (see our advertising page).

6. REFERENCES

(1) E. D. Schmitzer: A Modern Concept for Portable 2 Metre Receivers.
VHF COMMUNICATIONS 1 (1969), Edition 2, pages 115-122

(2) 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

(3) K. P. Timmann: Variable Frequency Crystal Oscillator (VXO) for 136 MHz
VHF COMMUNICATIONS 1 (1969), Edition 2, pages 87-94

(4) Radio Engineering Handbook, 1968, chapter 5-22, Fig. 38, 39.

A 9 MHz IF-AF PORTION USING INTEGRATED CIRCUITS

by K. P. Timmann, DJ 9 ZR

The following IF-AF portion is designed to convert SSB signals from 9 MHz into the AF range. It comprises a 9 MHz crystal filter, a local oscillator for reception of the sideband, a product detector, a carefully dimensioned gain control and an audio amplifier with an output power of 350 mW.

1. CONCEPT

The most important considerations during the conception of radio receivers for the metre and decimetre wavebands are the demands for low noise and a high cross-modulation and intermodulation rejection of the equipment. These demands are not only placed on the input circuit but also on the mixer and IF stages.

In order to meet these demands, the IF strip consists of:

a) A crystal filter direct at the input so that practically no cross-modulation interference can occur. This, however, will have no effect on any intermodulation caused in the IF strip or demodulator.

b) A dynamic control range of more than 90 dB in two amplifier stages and an additional manual control of 20 dB in both the 9 MHz and AF range, which ensures that overload and intermodulation distortion is kept at a minimum.

c) A low-noise cascode circuit in the first amplifier stage, which means that an overall gain of the receive mixer of 20 dB will be sufficient.

Practical experience has shown these considerations to be true, namely that it is advisable to keep the overall gain of the receive converter at about 20 dB and to carry out this amplification in the linear input amplifier stages so that the conversion gain can be kept low. This means that the noise figure of the mixer stage will only cause approximately 1% of the total noise figure. In addition to this, the low conversion gain allows the use of mixer circuits that are insensitive to overload and which possess good cross-modulation characteristics. A converter designed according to these considerations is described in this edition of VHF COMMUNICATIONS.

2. REGARDING THE USE OF INTEGRATED CIRCUITS

Type CA 3028 A integrated circuits have been used for the IF amplifier stages, the oscillator and demodulator stage. The type CA 3020 is used in the AF amplifier and for the control circuit. These components are manufactured by RCA. See Figures 1 and 2 for further details.

Circuit CA 3028 A consists of the two transistor systems Q 1 and Q 2, which operate as a differential amplifier. The common emitters are connected to the collector of transistor system Q 3. Due to the high impedance of Q 3, the sum of the emitter currents of Q 1 and Q 2 are practially independent of the operating point of Q 1 and Q 2 (impressed sum currents).

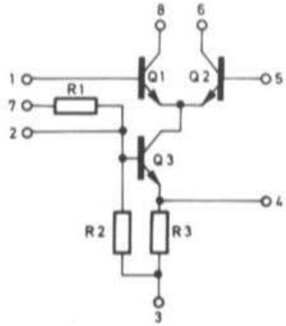


Fig. 1: Circuit of the integrated RF Amplifier CA 3028 A

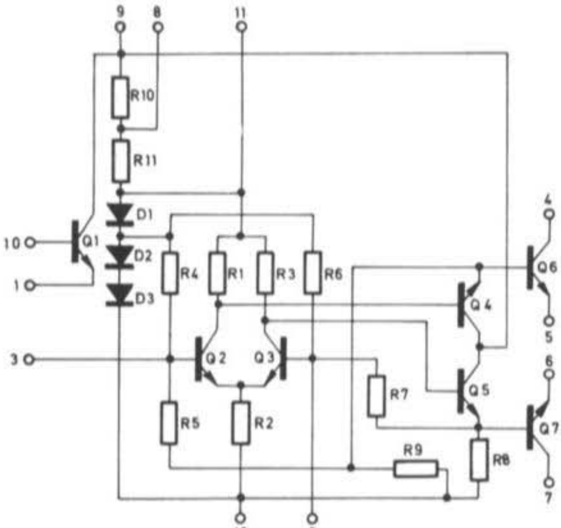
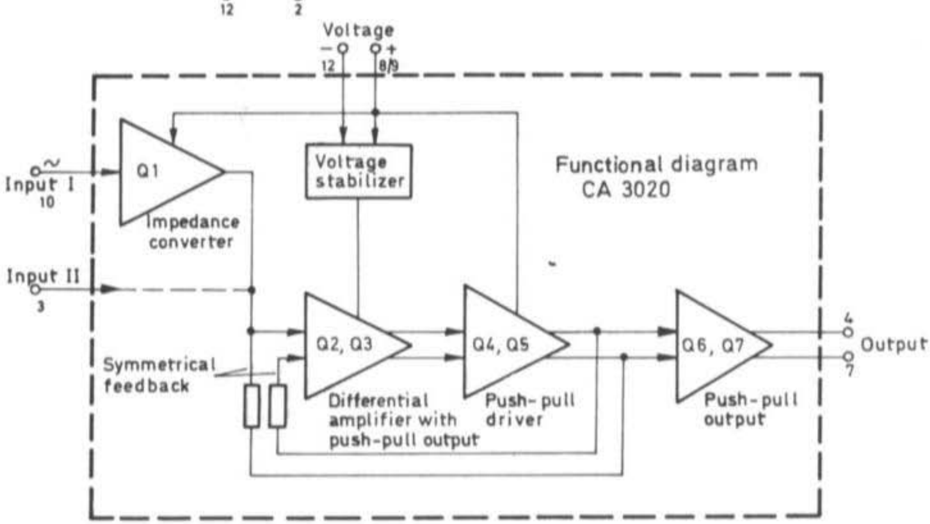


Fig. 2:
Circuit and functional diagram of the Universal Integrated Wideband Amplifier CA 3020



The integrated circuit CA 3020 comprises the transistor system Q 1 (which can be used as an emitter follower), a common emitter transistor pair Q 2 and Q 3 as symmetrical amplifier feeding the driver transistor systems Q 4 and Q 5 which in turn feed the base of the two final amplifier systems Q 6 and Q 7, whose emitter and collector must be connected to an external network.

The operating voltage of + 9 V must be maintained for all integrated circuits. If the operating voltage were to sink below 5 V or increase above 12 V, this would alter the characteristics of the IF strip; in the latter case, the integrated circuits type CA 3020 are endangered.

The use of integrated circuits allows both space and component savings as well as saving the time-consuming process of assorting transistor pairs etc. This means that the somewhat more expensive integrated circuits can in fact be more favourable, even when all systems are not used.

3. DESCRIPTION OF THE CIRCUIT

The circuit diagram of the whole 9 MHz IF-AF portion is shown in Fig. 3.

3.1. IF AMPLIFIER

As previously mentioned, a narrow band crystal filter (type XF-9B manufactured by KVG) is to be found at the input of the 9 MHz IF strip.

If the given filter characteristics are to be maintained, it is necessary to terminate the filter with the input and output impedance values given in the data sheet. They amount to 500 Ω parallel 30 pF in both cases. The required source impedance is formed by the output impedance of the converter. This means that the coupling inductance at the output of the converter must be adjusted accordingly (tune for resonance; non-load voltage must be reduced by half by connecting 500 Ω). The load capacitance is made by using a screened cable (RG - 58 / U : 30 pF per metre; RG - 274 / U: 100 pF per metre). At the output of the filter, the input impedance of the first integrated circuit IC 1 together with the parallel connected 3.9 kΩ and 7 pF are used as a suitable terminating impedance.

After having passed through the crystal filter, the signal is fed to connection 2 of IC 1, that is to the base of transistor system Q 3. This system forms, together with system Q 2, a low-noise cascode amplifier. System Q 1 is not used, the connections 1 and 8 of IC 1 should be removed.

The control voltage U_{C2} is fed to connection 7, where it varies the base bias of Q 3. The control voltage is in the range of + 7.3 V (maximum gain) and + 1.7 V (minimum gain). This voltage range is sufficient to vary the gain of IC 1 by 45 - 50 dB. Connection 4 of IC 1 is connected to ground via the series-connection of a 330 pF capacitor with a 500 Ω trimmer potentiometer P 1. In the radio-frequency sense, this series circuit bridges the resistor R 3 of the integrated circuit. The trimmer potentiometer allows the negative RF feedback of transistor system Q 3, and thus the gain of the integrated circuit to be varied manually by an additional 20 dB; a precision potentiometer manufactured by Amphenol was used for the exact adjustment; it is mounted directly beside the crystal filter. It has been found that potentiometer P 1 will have a value of

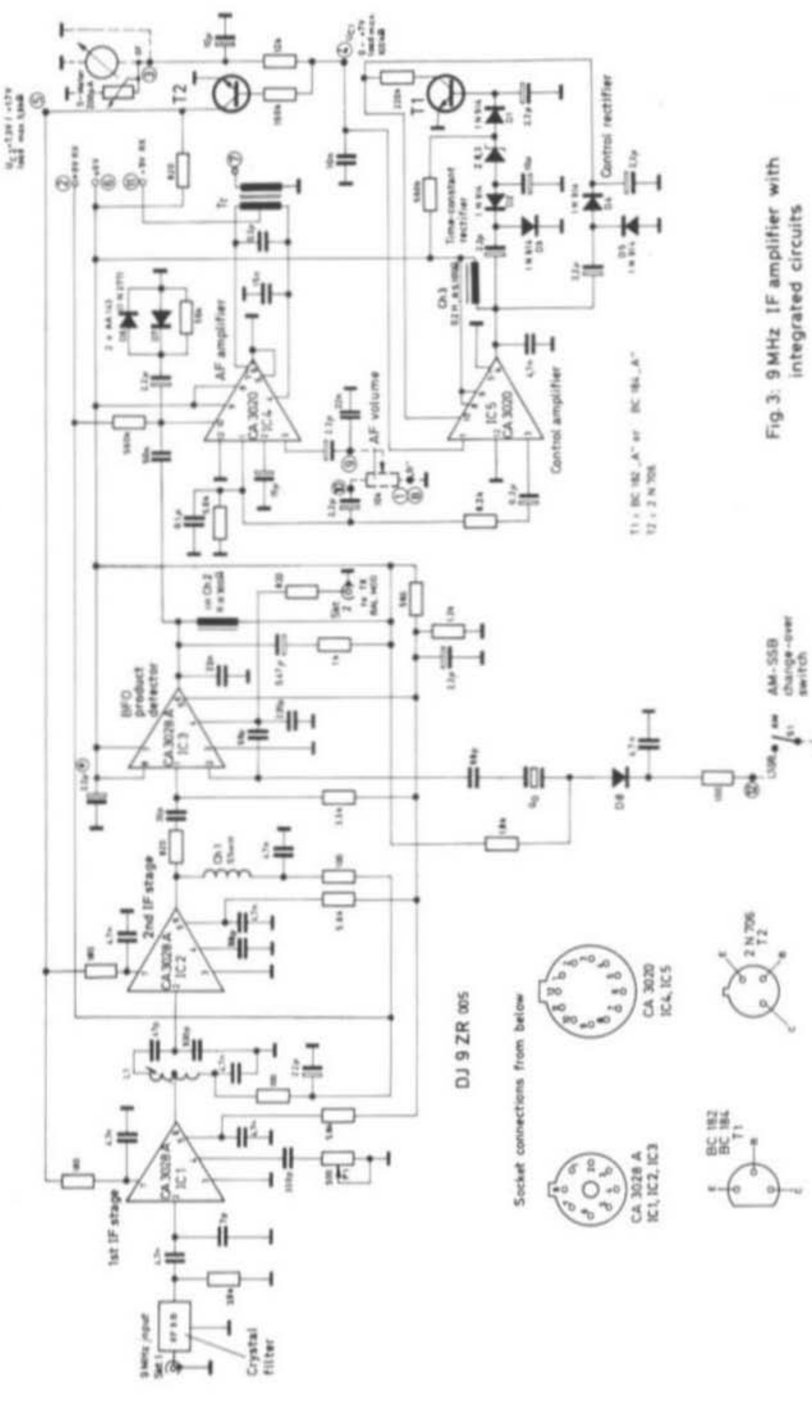


Fig.3: 9MHz IF amplifier with integrated circuits

approximately 100 Ω at a converter gain of 18 dB. This means that the potentiometer can be replaced by a fixed resistor of 100 Ω or a simple 100 Ω potentiometer.

The manual adjustment of the RF gain at this point means that the gain can be reduced to such a degree that the IF strip operates in the most favourable range of the time-constant control (see Section 4). It avoids, at the same time, unwanted distortions which could be caused by the mixer (product detector) when the voltages are too high.

Connection 5 of IC 1 is provided with a fixed bias voltage of approximately 6.2 V which is obtained by voltage division from the 9 V operating voltage by the series-connected resistors of 560 Ω and 1.2 kΩ. The bias voltage is filtered by a 2.2 μF capacitor and is also used as the supply voltage for integrated circuits IC 2 and IC 3. The RC-links (5.6 kΩ, 4.7 nF) block the base of Q 2 of IC 1 and IC 2 and ensure that no coupling is made via the common operating voltage line.

The collector connection 6 of IC 1 is connected to a tapping point of inductance L 1, which, together with the series-circuit 47 pF - 330 pF forms a 9 MHz parallel resonant circuit. It is used to improve the ultimate selectivity of the filter and is essentially damped by the input impedance of IC 2 (connection 2 at the interconnection point of the capacitors).

The second amplifier stage IC 2 is essentially connected in the same manner as IC 1. At the output of IC 2, the operating voltage is not fed via a resonant circuit but via a RF choke Ch 1 (0.5 mH) for means of simplicity.

3.2. LOCAL OSCILLATOR

In the subsequent stage IC 3, transistor system Q 3 operates as an Colpitts oscillator with an "inductive" crystal. This arrangement was also used on printed circuit boards DJ 9 ZR 002 and DJ 9 ZR 003.

The auxiliary frequency required for the demodulation of the 9 MHz single sideband signal can be switched off for AM via switching diode D 8 (1 N 914) by switch S 1. If point 12 is grounded by S 1, a current will flow via diode D 8, so that the differential resistance will drop to approximately 20 Ω. By blocking the diode D 8 with a 4.7 nF capacitor, the change-over lead connected to point 12 is kept RF-free and can therefore not cause any interference effect to other circuits. If the ground connection is switched off at point 12, the oscillator (BFO) will not oscillate and it will be possible to receive amplitude modulated signals at an attenuation of approximately 25 dB.

The emitter of transistor system Q 3 (connection 4) is connected via a 820 Ω resistor to the output (Bu 2). The auxiliary frequency generated in the oscillator is available at this point for the control of a bridge modulator (in the transmitter). This means that exact transceive operation is possible. The operating voltage for the oscillator is continuously connected (+ 9 V, connection point 6) as is also the case with the AF amplifier. The other stages are only switched on (+ 9 V, connection point 2) during reception.

3.3. DEMODULATION

The demodulation is made with the emitter-coupled systems Q 1 and Q 2 of integrated circuit IC 3. The SSB-signal is fed to the base of Q 1 (connection 1) via the series circuit of 820 Ω + 33 pF, in order to avoid any reaction from the product detector on IC 2 and to isolate the bias voltages. The base of transistor system Q 2 (connection 5) is "cold" in the RF sense. The collector current i_{Q2} of transistor system Q 2 also includes the product from the signal voltage u_{Q1} at the base of Q 1 and the collector current i_{Q3} of transistor system Q 3. For the multiplicative mixer circuit used, the following will give a good approximation:

$$i_{Q2} \approx 1/2 + 1/4 \times i_{Q3} \times \frac{u_{Q1}}{26\ \mathrm{mV}} + 1/8 \times i_{Q3} \times \left(\frac{u_{Q1}}{26\ \mathrm{mV}}\right)^2 + \ldots$$

With 9 MHz signal voltages u_{Q1} of up to approximately 4 mV, the third and following terms, on which the intermodulation distortion is mainly dependent, are negligible. i_{Q3} is built-up from the DC component and the superimposed local oscillator component. This means that the required AF signal is also included as a conversion product in i_{Q2}. The required signal is filtered out at connection 6 of IC 3. The 9 MHz components from the signal and local oscillator, as well as higher frequency conversion products are shorted out by the 22 nF capacitor. The DC voltage for the collector is fed via an AF choke Ch 2 (Z = 10 kΩ at 100 Hz, R = 300 Ω); 0.47 μF + 1 kΩ are placed across this choke to compensate for the impedance variation as a function of frequency, which would cause an emphasis of the higher frequencies above 1 kHz. If a signal of 0.5 μV is available at the input of IC 1, an AF voltage of approximately 10 mV will be available at the output of IC 3 if the gain is at maximum.

The gain range of the integrated circuits IC 1 and IC 2 and the operating point of IC 3 must be selected so that no noticeable distortions are caused. Integrated circuits IC 1 and IC 2 would allow output voltages of up to 1 V, however, this could not be utilized because the product detector IC 3 can only be driven at a maximum of 4 mV if the intermodulation distortion is to remain less than -40 dB. The control circuit must be correspondingly dimensioned. In the case in question, the signal voltage at the input of IC 3 amounts to approximately 600 μV.

3.4. AF AMPLIFIER

An integrated circuit type CA 3020 (IC 4) is used for the AF amplification. As can be seen in Fig. 3, the AF signal is fed from the output of integrated circuit IC 3 via a capacitor of 50 nF to connection 10 (base of Q 1) of IC 4. A limiter circuit with two anti-parallel diodes D 6, D 7, is connected via 2.2 μF to this connection. This circuit is designed to suppress all pulse type interference voltage peaks that are greater than the envelope of the AF signal. Such a circuit was also used on PC-board DJ 9 ZR 003. In addition to this, it suppresses the pulses which appear at the commencement of the control process, for instance, when the level of the receive signal rapidly increases. Diode type AA 143 (manufactured by ITT Intermetall) is used for diodes D 6, D 7, due to their very low forward voltage (approx. 0.3 V).

Bias current is fed to the base via a 560 kΩ resistor. The input impedance at connection 10 of IC 4 amounts to approximately 50 kΩ due to the strong negative feedback by the emitter load impedance. This represents approximately the same high load as the limiter circuit at low AF voltages. The emitter of transistor system Q 1 (connection 1) is grounded via 5.6 kΩ and 0.1 μF. This connection is connected via 2.2 μF to connection point 10 of the PC-board. The end connections of the AF volume potentiometer are connected between connection point 10 and the ground. The wiper is connected to connection point 9. The tapped off voltage is fed from connection point 9 via a 2.2 μF capacitor to the base of transistor system Q 2 (connection 3). The common emitter transistor systems, Q 2, Q 3 generate and amplify the antiphase voltages of the push-pull driver Q 4, Q 5. These in turn drive the power amplifier transistors Q 6 and Q 7.

The most favourable load impedance at the collector output of Q 6 and Q 7 amounts to 2 x 65 Ω. A small miniature transformer Tr (2 x 70 Ω/5 Ω; 0.5 W) was used in the prototype. The maximum output power of IC 4 amounts to approximately 0.5 W at an operating voltage of 9 V; however, it is only driven to a value of 350 mW, so that the distortion remains below 5% and that a cooling fin is no longer necessary. A power output of 250 mW is usually sufficient for amateur operation.

4. GAIN CONTROL

4.1. REGARDING THE GAIN CONTROL OF SINGLE SIDEBAND RECEIVERS

The gain control represents certain problems in the single sideband mode. Many commercial system users transmit a residual carrier, whose level is sufficient to control the fading (and the automatic frequency control) even under difficult receive conditions. However, the level of the residual carrier permissible for amateur radio transmissions in the transmit mode A3J is usually too low for such purposes.

The control voltage in the receiver must therefore be generated from the envelope or the mean value of the RF, IF or AF signal in the receiver. Such circuits must have a certain time behaviour, dependent both on the level amplitude and the level variation. The gain control should not cause a noticeable dynamic compression or other dynamic distortion and the interference peaks should be limited but should not affect the control characteristic. On the other hand, fading on the transmission path must be compensated for. In addition to this, the change from one station to another, which may have greatly differing signal strengths at the antenna, must also be compensated for. In the described IF portion, a very satisfactory compromise has been achieved by the fact that a variation of the time constant occurs in the control amplifier which means that small time constants are obtained at low AF levels and great time constants at high AF levels. This virtually compensates for the opposite effect that the control characteristic of the input stages IC 1 and IC 2 have on the control time constant. In addition to this, the time control variation operates with level dependent delay times.

Radio amateurs have usually greatly differing ideas regarding the gain control of single sideband receivers. Since the circuit used in this receiver is very easy to follow, the control process and the most important quantities are to be considered.

4.2. GENERATING THE CONTROL VOLTAGE

Connection 1 (base of the emitter follower system Q 1) of integrated circuit IC 4 is not only connected to connection point 10 and from there to the volume potentiometer, but also via 8.2 kΩ - 0.2 μF to connection 3 of integrated circuit IC 5. The collector (connection 4) of transistor system Q 6 represents the output of the circuit. This system obtains its operating voltage via a choke of 0.2 H (Ch 3). This output feeds the control rectifier D 4 / D 5 (lower rectifier network in the circuit diagram), which provides a positive voltage to ground and the rectifier D 2 / D 3, which is chosen for the time constant variation (upper network). The latter provides a negative voltage to ground.

The control rectifier D 4 / D 5 operates in a voltage doubler circuit using equally great charging capacitors of 2.2 μF. The output is connected to ground via the 220 kΩ resistor and the collector-emitter path of transistor T 1 - this transistor is used as time constant switch. According to the amount of base drive, the following current conditions will be observed in the transistor:

a) Zero current when T 1 is blocked. The load resistor for the control rectifier will then be formed by the input impedance at connection 10 of IC 5 and will amount to approximately 1.2 MΩ (= negative feedback impedance at emitter 12 kΩ x current amplification 100).

b) Current proportional to the base current of T 1.

e) Current corresponding to the difference between the control rectifier voltage and the residual voltage of transistor T 1 divided by 220 kΩ, when transistor T 1 conducts.

The rectifier network consisting of D 3 and D 4, which are used for the time constant variation, are also built-up in a voltage doubler circuit. The capacity of the pass capacitor also amounts to 2.2 μF. The charger capacitor is, however, far greater, having 10 μF, which means that it has a far greater charge time constant with respect to the control rectifier.

In order to see how the change-over switching of T 1 occurs, it is first assumed that the changeover rectifier is discharged.

The base of the switching transistor T 1 (BC 184 A, BC 108, 2 N 3904, with low residual current) is fed with a bias current of approximately 15 μA via the dropper resistor of 560 kΩ and diode D 1 (1 N 914) when it is in its rest position.

A zener diode with a nominal voltage of 6.3 V is connected to the connection point of the dropper resistor and the diode. Diodes having such a low reverse voltage do not normally show a sharp breakdown, but have a somewhat more continuous current-voltage characteristic. At a current of approximately 15 μA, the voltage drop across the zener diode will amount to approximately 4 V. If it is assumed that a voltage of 0.9 V appears at the given connection point when the transistor is conducting, the current flowing via the 560 kΩ resistor will not be noticeably taken over by the zener diode until the rectifier for the time constant variation supplies a voltage which is lower (that is more negative) than - 3.0 to - 3.2 V. If the zener diode takes over the total current of 15 μA, transistor T 1 will be blocked. The combination of the diode and the capacitor at the base of T 1 has the effect that the blocking process is delayed by approximately 30 ms in addition to the charging process of the rectifier. The 2.2 μF capacitor discharges itself via the base, where the discharge current falls on decreasing the emitterbase voltage thus delaying the discharge process.

4.3. THE TIME BEHAVIOUR OF THE CONTROL AMPLIFIER

The time behaviour of the whole control amplifier is essentially determined by the delay times caused by the time constant of the control rectifier and that of the rectifier for the time constant variation. These can be classed as open control loops (the control line is disconnected, the input stages operate at a fixed operating point. Observe the output voltage of the control rectifier on varying the IF signal level). If the subordinate effects are not considered, for instance if the two rectifier branches influence each other at the input, then the following two time constants and two delay times must be considered (the given AF voltages are peak values):

4.3.1. CHARGE TIME CONSTANT OF THE CONTROL RECTIFIER t_1

This time constant is infinitely great with AF voltages of less than 300 mV at the control rectifier input; the rectifier will not be actuated since the diodes are only driven slightly into their forward voltage range. The time constant will decrease on increasing the AF voltage; it amounts to 100 - 50 ms at U_{AF} = 1 V and will finally be decreased to 10 - 20 ms at higher voltages.

4.3.2. DISCHARGE TIME CONSTANT OF THE CONTROL RECTIFIER t_2

This time constant amounts to: approx. 2.2 μF x (220 kΩ/1.2 MΩ) $\approx$ 0.5 s when transistor T 1 conducts and approx. 2.2 μF x 1.25 MΩ $\approx$ 2.6 s when transistor T 1 is blocked.

If transistor T 1 operates in the transition range between its blocked and conducting condition, the voltage drop as a function of time can be approximated by interpolation between these two values. The transition time from blocked to conducting condition is relatively short (see Section 4.3.4.).

4.3.3. RECOVERY TIME t_{r1}

This is the time required between the injection of an AF voltage (at output 4 of IC 5) and the transition of transistor T 1 from conducting to blocked condition. This time is dependent on the duration and amplitude of the AF voltage at the input of the switching rectifier. AF voltages, which are less than approximately 1.4 V, will not be sufficient to switch transistor T 1 from conducting to blocked condition, since the output voltage of the switching rectifier must amount to at least 3.2 V before T 1 is switched (t_{r1} infinitely great).

If a higher AF voltage appears at output 4 of IC 5, this recovery time will fall from a few seconds at an AF voltage of 1.8 V to 50 to 100 ms at 3 V. This is mainly caused by the large charge time of this rectifier circuit (charge capacitance 10 μF is great compared with the pass capacitance of 2.2 μF). In addition to this, the 2.2 μF capacitor at the base of T 1 causes an additional slight delay.

4.3.4. RECOVERY TIME t_{r2}

This is the time which passes between ceasing the AF voltage (at output 4 of IC 5) and the switching of transistor T 1 from the blocked into the conducting condition.

This time is dependent on the charge of the 10 μF capacitor. It amounts to approximately 3.5 s at full charge (approx. 7 V AF voltage at output 4 of IC 5) and will be reduced to zero if the AF voltage is less than 2.4 V.

The transition of transistor T 1 from conducting to blocked condition occurs in the range of U_4 = 2.4 to 2.5 V, assuming that U_4 is varied slowly. If U_4 is quickly varied, the transition will occur within 0.2 s after completion of the recovery time t_{r2} (U_4 = voltage at output 4 of IC 5).

4.4. S-METER CONNECTION AND REVERSAL STAGE FOR THE CONTROL VOLTAGE

It is not possible to use the output voltage of the control rectifier to control the two input stages: the variation direction must be firstly reversed. To achieve this, the emitter of the emitter follower Q 1 in IC 5 is connected via 150 kΩ to the base of T 2 (2 N 706). A further 12 kΩ resistor is connected to ground. A 200 μA meter can be connected in this branch and used as S-meter. The control voltage U_{C2} for the two IF stages IC 1 and IC 2 is available at the collector of transistor T 2; in addition to this, further stages can be controlled from connection point 10, such as a converter. If the input stages are connected to the collector of T 2, a voltage drop will appear across the collector resistor of 820 Ω due to the current requirements of IC 1 and IC 2, even when T 2 is blocked. A voltage of 7.3 V remains at the collector. In the other limit condition - with 7 V AF voltage at output 4 of IC 5 - the collector voltage will fall to 1.7 V. This means that a voltage variation of 1.7 V to 7.3 V is available for control purposes (U_{C2}).

4.5. THE TIME BEHAVIOUR OF THE WHOLE CIRCUIT

The time behaviour of the whole circuit, that is the closed control loop, is not only influenced by the time behaviour of the control rectifier but also by the amplification-control characteristic (amplification as a function of the control voltage): The time constants of the control rectifier are reduced. This is understandable when one considers that a rapid increase of the IF level will cause a rapid increase of the AF level which must, however, be compensated for when the voltage at the control rectifier output increases. The more the amplification of the stages IC 1 and IC 2 varies with the control voltage, the more rapid will be the IF voltage be reduced. This is the reason why the time constant, which indicates the transition into the final condition, is reduced with respect to the control rectifier time constant when the control loop is closed. This is valid for the time constants when the IF signal level is rapidly increased or decreased.

The behaviour of the IF portion subsequent to rapid variations of the IF signal level can be characterized as follows:

a) Since the control voltage cannot have an immediate effect - due to the time constant t_1 and t_2 - the rapid variations of the dynamic speech appear at the speaker at full strength. Slow variations, however, will be controlled in a similar manner to fading.

b) Strong and rapid level variations at the antenna input will - as long as the limiter circuit does not suppress them - therefore be fully operative at the loudspeaker output for a short period of time since the voltage at the output of the control rectifier cannot immediately obtain its final value, but will vary exponentially according to the time constants in this direction. This means that short individual interference pulses will not desensitize the IF amplifier.

c) The time constants having an effect on the AF level are shorter than those of the control rectifier. The greater tha level of the IF signal, the greater is the factor by which the time constant is reduced. The cause of this is that the gain of the controlled stages IC 1 and IC 2 increases linearly with the controlled voltage (and not exponentially, as is the case with control tubes). The time constant variation essentially aids the compensation of this effect.

d) This means that the mean AF level remains more constant the greater the level of the IF signal is. The gain characteristics of the input stages IC 1 and IC 2 are also important in this respect.

e) The control can be adjusted with trimmer potentiometer P 1 (IC 1) so that the switching rectifier commences reduction of the T 1 base voltage at an IF signal-to-noise ratio of approximately 30 to 35 dB. The variation of the discharge time constants from a very low to the very high value occurs in a very small range of the signal level (approx. 5 dB).

f) The recovery times t_{r1} and t_{r2}, which are the times passing between a signal variation and the switching of T 1, possess approximately the same values in a closed as in an open control loop. However, the variation of the IF level must be large if the larger values of t_{r1} are to be reached. This is caused by the fact that the recovery times are dependent on the mean AF level, which varies far less than the IF signal level due to the control.

The circuit used in the described IF amplifier represents a simple means of providing a good gain control during single sideband reception. The circuit has been found to offer very good results under greatly differing operating conditions, such as when switching between stations having greatly differing signal strengths, or when tuning over the band, with pulse type interference and strong fading. The selected combination of time constants, recovery times and control ranges have been found very favourable.

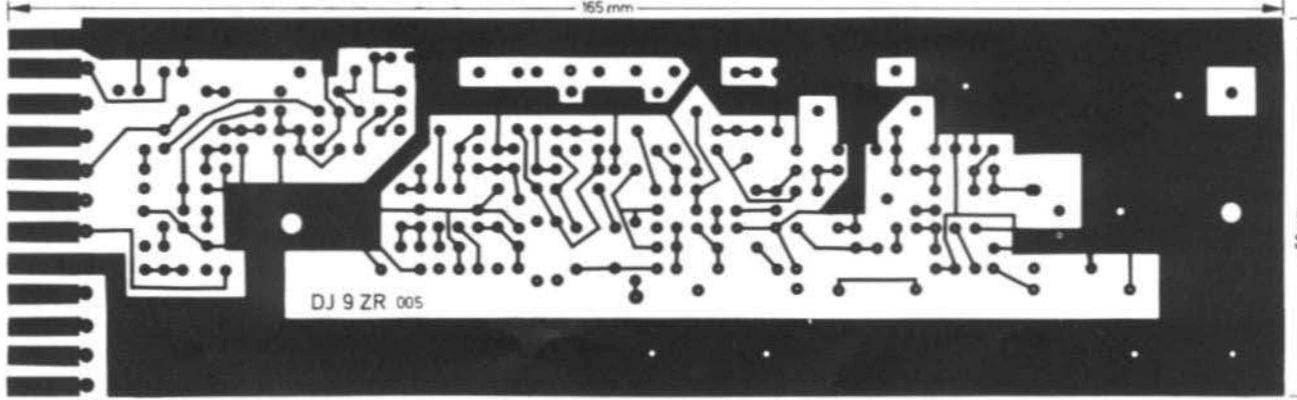


Fig. 4a: Conductor side "a" of the IF - AF amplifier DJ 9 ZR 005

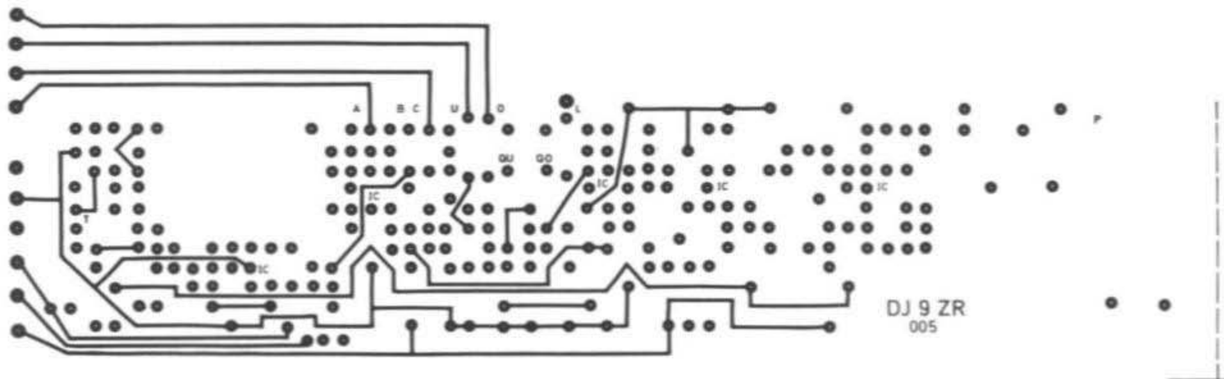


Fig. 4b: Conductor side "b" of the IF - AF amplifier DJ 9 ZR 005

Intermodulation:	< 30 dB at IF input levels between 0.3 μV and 60 mV
AF distortion factor:	< 10% at 350 mW output power and IF input signals between 0.3 μV and 60 mV
Stability of the oscillator frequency:	< 35 Hz between 0° and 65° C when using temperature compensating capacitors
AF frequency response:	see Fig. 5 The frequency response curve has a maximum at approximately 1.2 kHz, the characteristic curve falls above and below this frequency to approximately 500 Hz and 2400 Hz.
Ultimate attenuation:	The ultimate attenuation of the crystal filter is improved by use of the IF resonant circuit comprising L 1. Attenuation ± 50 kHz from centre frequency: > 90 dB ± 500 kHz from centre frequency: > 120 dB.
Noise figure:	$F \approx 3$ A 9 MHz signal voltage of approximately 0.3 μV is required into 500 Ω for a signal-to-noise ratio of 10 dB (corresponding to 0.1 μV into 50 Ω).
Control point:	At IF signal levels where the signal-to-noise ratio of approximately 15 dB is exceeded or according to the IF gain adjustment.
Level compensation:	The AF level will be varied by 10 dB on altering the IF signal level between 0.5 μV and 60 mV (100 dB). The static curve showing the AF signal as a function of the 9 MHz input signal is given in Fig. 6.
Overload rejection:	The IF input voltage must remain below 60 mV in the passband of the filter so that an intermodulation of - 30 dB will not be exceeded. A maximum of 1 V is permissible outside the passband of the crystal filter. It has been found that even 4 V can be fed in without damaging the filter or causing the signal to break through.
Gain stability as a function of varying the operating voltage:	The overall gain variation will be less than 5 dB with battery voltages between + 6 V and + 10 V.

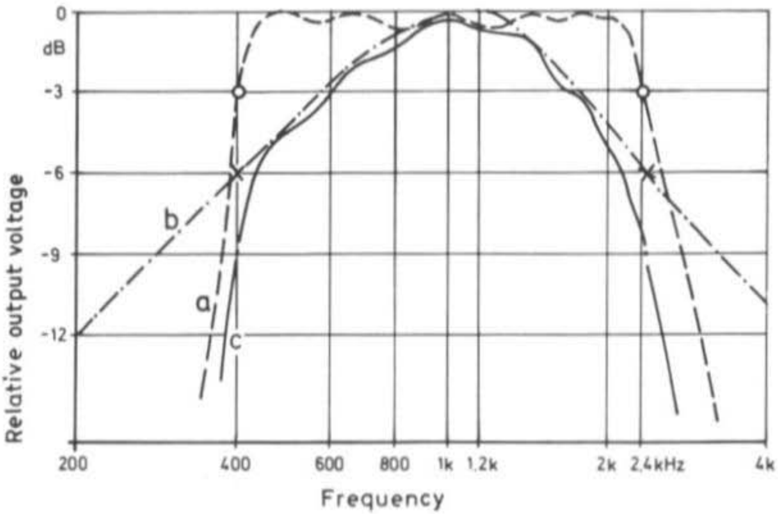


Fig. 5: Frequency response curve of the IF-AF amplifier
0 dB = max. output power

a. 9MHz crystal filter component

b. Component obtained with the RC-link and the AF output transformer

c. Total frequency response

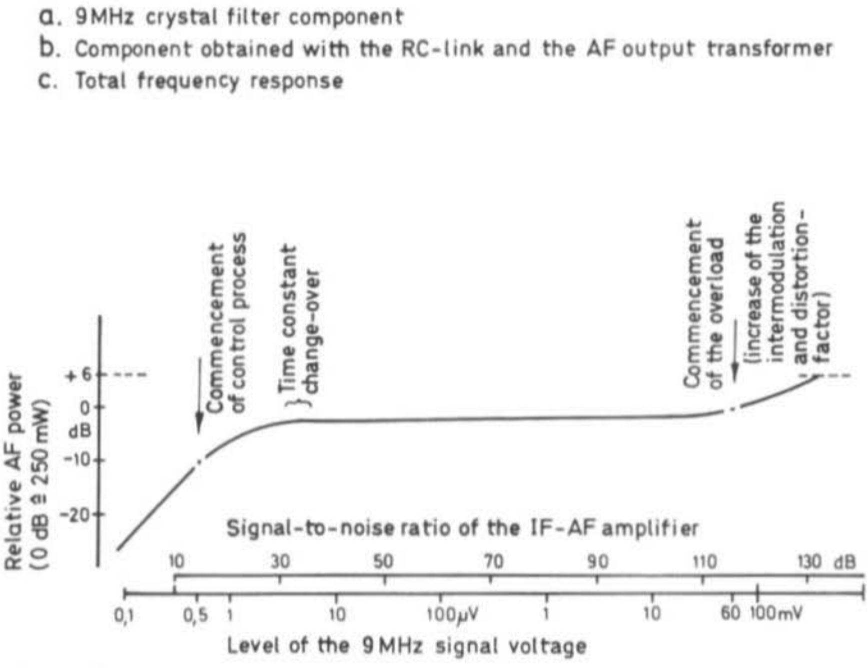


Fig. 6: Relative AF power as a function of the 9MHz input level of the IF-AF amplifier (static curve)

6. EQUIPPING THE PRINTED CIRCUIT BOARD DJ 9 ZR 005

The IF amplifier (Fig. 3) is built-up on a double-coated printed circuit board with the dimensions 105 mm x 50 mm (Fig. 4a and b). The printed circuit board is equipped with through contacts with a raster spacing of 2.5 mm and has silver-plated conductor lanes. Material: Epoxy glass fibre.

Miniature components must be used on the printed circuit board. This requires a certain amount of patience.

For means of convenience, the component location plan of the PC-board (Fig. 4c) is to be found on the centre pages of the magazine, where it appears in a larger scale.The title picture shows a photograph of the complete IF and AF amplifier (DJ 9 ZR 005).

6.1. SPECIAL COMPONENTS

Integrated circuits:
IC 1, IC 2, IC 3: CA 3028 A manufactured by RCA
IC 4, IC 5 : CA 3020 manufactured by RCA

Transistors:
T 1: BC 182 A, BC 184 A manufactured by Texas Instruments or
BC 108 A, 2 N 3904, 2 N 2926

T 2 = 2 N 706

Diodes:
D 1, D 2, D 3, D 4, D 5 = 1 N 914
D 6, D 7 = AA 143 manufactured by ITT, 1 N 277, manufactured by Texas-Instruments, or 0A 182 manufactured by AEG-Telefunken.
D 8 = BA 143 V (1 N 914), manufactured by Texas Instruments.

With the exception of the filter capacitor of 50 μF/15 V at connection point 11, all capacitors are:

12 V ceramic capacitors with a 2.5 mm raster spacing. The electrolytic capacitors are tantalum capacitors with 2.5 mm raster spacing.

Potentiometer P 1 = 500 Ω trimmer potentiometer manufactured by Amphenol (see Section 3.1.).

All resistors have a 1/8 to 1/10 W rating and should be suitable for 7.5 mm raster spacing.

7. ALIGNMENT OF THE PRINTED CIRCUIT BOARD DJ 9 ZR 005

Align inductance L 1 for maximum gain, further alignment is not required.

If the overall gain of the converter is greater than assumed, the overall gain of the IF portion can be reduced by adjusting the trimmer potentiometer P 1, which is connected to connection 4 of IC 1 via a 330 pF capacitor.

8. AVAILABLE COMPONENTS

The printed circuit board DJ 9 ZR 005, as well as various special components, is available individually or as a kit from the publishers or their national representatives (see advertising page).

LINEAR INTEGRATED CIRCUITS FOR AMATEUR APPLICATIONS

by D. E. Schmitzer, DJ 4 BG

1. GENERAL DETAILS

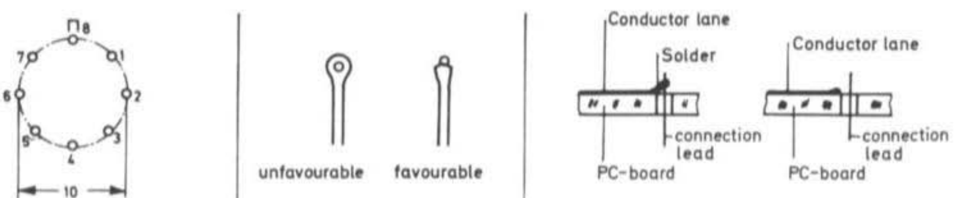
1.1. ADVANTAGES OF INTEGRATED CIRCUITS

Although the price of an integrated circuit is somewhat higher than the cost of purchasing the individual discrete components required to obtain the same function, if the time-saving, which is gained by avoiding the mounting, connection, testing and alignment of comparable discrete components, is considered, this will often more than compensate for the extra expense. The great advantages when using integrated circuits, for instance, in mobile or hand-held stations, are the very small space requirements. Since the circuit elements are imbedded in the silicon substrate, a good isolation is achieved in spite of the small dimensions. A number of the transistor systems are built up in a differential configuration. This results in a relatively great linear drive range and a low temperature dependence.

The article (1) described how the integrated circuit TAA 111 (Siemens) could be used as a tailored speech amplifier having a frequency response rising linearly with frequency in the voice pass band (approx. 6 dB/octave). Various applications using integrated circuits manufactured by RCA are to be given. The operating voltages, specifications and circuit details were partially extracted from RCA data sheets (2) and sometimes determined by experiments by the author. They do not necessarily represent the most favourable values.

1.2. MOUNTING INSTRUCTIONS

All of the described integrated circuits are enclosed in flat TO 5 casings (max. 8.5 mm dia., 4.6 mm high) having 8, 10 or 12 connections. The twelve connection leads are fed vertically through the base of the casing and from a circle of 4.8 mm to 6.2 mm diameter. When using such integrated circuits in conjunction with printed circuit boards, a diameter of 10 mm (Fig. 1) was found to be favourable so that some of the connection leads correspond to the 2.5 mm standard spacing. The holes on the PC-board have the conventional diameter of 1.3 mm. The PC-boards are designed so that the conductor lanes do not end in the usual round rings but end bluntly in front of the hole (see Fig. 1). This is a great advantage when removing the integrated circuits at a later date. This is carried out by unsoldering each connection lead individually and bending it back until the solder sets. After all connections are unsoldered, the integrated circuit can easily be removed.



Base connections of the integrated circuit as seen from below as well as the drill jig for the PC-board

Design of the conductor lanes

Unsoldering the connection leads of an integrated circuit

Fig. 1 : Mounting integrated circuits

1.3. VOLTAGE REQUIREMENTS

Many integrated circuits require two operating voltages having one common pole, e.g. 0 V, 6 V, 12 V or -6 V, 0 V, +6 V. It is possible to obtain the intermediate potential by series-connection of two voltage sources or by dividing the greater voltage using a zener diode or dropper resistor; good filtering is imperative. It is often advisable to connect the intermediate point to ground so that the base-leak resistors and the output resistors can be grounded. This configuration is mostly used by the author.

1.4. SPECIFICATIONS

Integrated circuits are sensitive to overload, especially voltage overload, in a similar manner to transistors. Details regarding the limit and recommended operating values are given in the data sheets. Typical circuits are given in the application notes. These information sheets are normally provided on purchasing the required integrated circuits.

2. APPLICATIONS USING THE INTEGRATED CIRCUIT CA 3028

2.1. CIRCUIT DIAGRAM

The circuit of this integrated circuit is given in Fig. 2. The external circuit represents a possible means of obtaining the required operating voltages.

Transistor system Q 3 operates as a constant-current source; the collector current is virtually independent of the collector voltage (impressed current) and distributes itself to the emitters of both Q 1 and Q 2. If the base voltage of system Q 2 is kept constant and that of system Q 1 increased, the emitter voltage will be influenced so that the emitter current of Q 2 is reduced by the same amount as the emitter current increase of Q 3. The collector current of Q 2 is correspondingly increased and that of Q 1 decreased.

If the voltage at the bases of Q 1 and Q 2 are simultaneously increased by the same amount, the collector currents will vary only slightly.

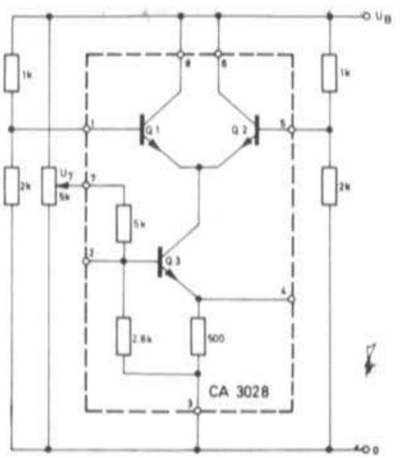


Fig. 2 : Integrated circuit CA 3028 with an external circuit to measure the static specifications

The collector currents at connections 6 and 8 of the integrated circuit vary only as a function of the voltage difference at the base connections 5 and 1, the sum voltage component is practically not amplified. The common mode rejection of the CA 3028 is not specified, but most certainly amounts to at least 60 dB.

The gain factor of the differential amplifier can be adjusted by varying the base voltage of system Q 3. This is because the forward transconductance Y_{21} (in mA/V) of transistors is dependent on the emitter current I_E; the following is a good approximation:

$$Y_{21} \approx 39 \times I_E \;/\; \text{volt}$$

System Q 1 receives negative feedback due to the interconnection with Q 2; the negative feedback factor amounts to approximately 2 at low drive values. The collector current I_{C3} of Q 3 is addition. y divided on to both systems. The active transconductance with low differential voltages is therefore:

$$Y_{21a} \approx 10 \times I_{C3} \;/\; \text{volt}$$

The integrated circuit can be used for a variety of applications, such as:

a) A variable gain amplifier with either an equal-phase or anti-phase output or as an amplifier with a push-pull output (Phase splitter), see Fig. 3.

b) As a variable gain amplifier in a cascode circuit (see Fig. 4).

c) As a differential amplifier to separate the push-push and push-pull component.

d) Measurement of AF and RF voltage differences, to compare voltage according to amount and phase in measuring bridges, reflectormeters, frequency and phase comparator circuits as well as an active element in filters.

e) As a mixer in either push-push or push-pull configuration or as product detector (Fig. 5 and 6).

f) As a push-push or push-pull oscillator

g) As a self-excited mixer (see Fig. 2)

Circuit examples are now to be given for some of these applications.

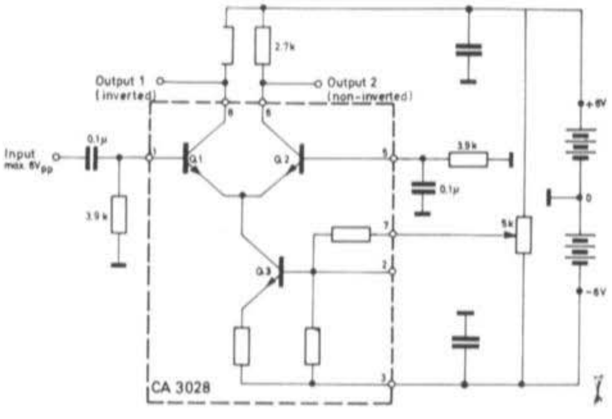


Fig. 3 : Integrated circuit CA 3028 as a variable gain phase splitter in the AF and RF range

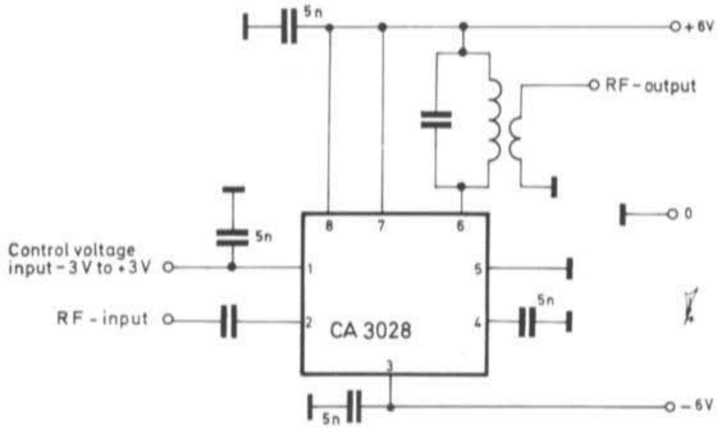


Fig. 4 : Integrated circuit CA 3028 as a variable gain cascode amplifier

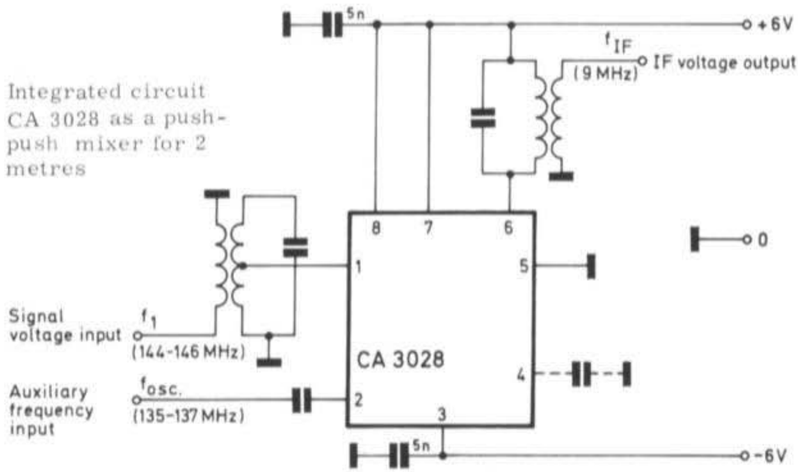


Fig. 5 : Integrated circuit CA 3028 as a push-push mixer for 2 metres

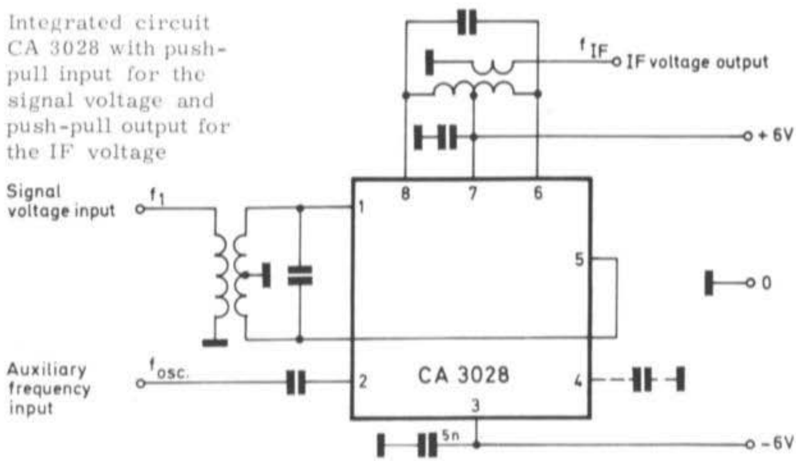


Fig. 6 : Integrated circuit CA 3028 with push-pull input for the signal voltage and push-pull output for the IF voltage

2.2. THE MOST IMPORTANT SPECIFICATIONS

The following typical values are obtained with an ambient temperature of 25° C in an external circuit as shown in Fig. 2 :

with $U_B = U_7 = 9$ V	I_6, I_8 = 2.5 mA ;
	I_1, I_5 = 29 μA
with $U_B = U_7 = 12$ V	I_6, I_8 = 3.5 mA;
	I_1, I_5 = 44 μA

Input impedance at connection 7 : approx. 7 kΩ

Power amplification

at U_B = 9 V, f = 100 MHz	20 dB (cascode)
	17 dB (differential)
f = 10.7 MHz	39 dB (cascode)
	32 dB (differential)

The signal voltage at point 1 or 5 must not exceed 6 V peak-to-peak.
The maximum permissible power dissipation P_{tot} = 300 mW.

2.3. VARIABLE GAIN RF AND AF AMPLIFIER

Figure 3 shows a circuit where the integrated circuit CA 3028 is used as an aperiodic amplifier. The base connection of system Q 2 is not required and may be directly connected to the bias voltage (0 V in the example). If an especially high temperature stability is required, the base may be connected to the bias voltage via a bypassed resistor, having the same value as the base leak resistor.

When used as an RF amplifier, the load resistor of the required output can be replaced by a resonant circuit. The load resistor of the unused output is directly connected to the operating voltage.

The control range amounts to approximately 60 dB at 10.7 MHz and is still more than 40 dB at 100 MHz. This range is obtained by varying the voltage at point 7 between -6 V and +3 V. The inherent stability of the integrated circuit is good and will not require further neutralization. Due to negative feedback at the emitter, a differential amplifier circuit is less affected by overload (cross modulation) in the IF range as individual transistors. However, the improvement is only slight.

2.4. CASCODE AMPLIFIER

Transistor system Q 3 possesses the same high transit frequency as systems Q 1 and Q 2. It may therefore be used as the input stage of a cascode circuit (see Fig. 4). The noise figure of such a circuit is somewhat less than that of the differential amplifier whereas the gain and control range are greater. However, the active component of the input impedance is reduced by factor 3. This must be taken into consideration at frequencies of over 20 MHz where the output impedance will be also negative. Oscillation, however, will only occur with output impedances of more than 8 kΩ at frequency of 145 MHz.

The unused transistor system Q 1 can be used to control the amplification. If the base voltage becomes more positive than -3 V, the emitter of Q 1 will take more and more of the collector AC voltage of Q 3. The operating point of Q 3 is virtually maintained which means that the input impedance is not altered. Such a low-reactive gain control is advisable, for instance, when the input impedance of a stage is simultaneously the terminating impedance of a filter. A gain variation could otherwise cause the filter to be detuned.

2.5. PUSH-PUSH AND PUSH-PULL MIXERS

The circuit of a push-push mixer for two metres is given in Fig. 5. A signal voltage is fed to connection 1. Connection 2 is connected via ground to the zero point of the operating voltage. An IF circuit (9 MHz) is located at the collector of system Q 2. The auxiliary frequency (135 - 137 MHz) is fed to the base of Q 3 where it causes a multiplicative mixing.

Fig. 6 shows a mixer circuit with a push-pull input for the signal voltage and a push-pull output for the IF voltage. The auxiliary frequency f_{osc} is suppressed by the push-pull configuration of the IF circuit. A similar circuit could therefore be used as balanced modulator in SSB equipment.

The negative feedback resistor in the emitter lead of system Q 3 should not be shorted in the RF sense. Of course, the required auxiliary voltage would fall from approximately 2 V to 50 mV due to the bypassing of connection 4 but a great number of unwanted conversion products would be generated at the same time. The transconductance without bypassing, i.e. with full negative feedback, amounts to approximately 1 mA/V with an auxiliary frequency voltage of 80 mV and 10 mA/V at 800 mV.

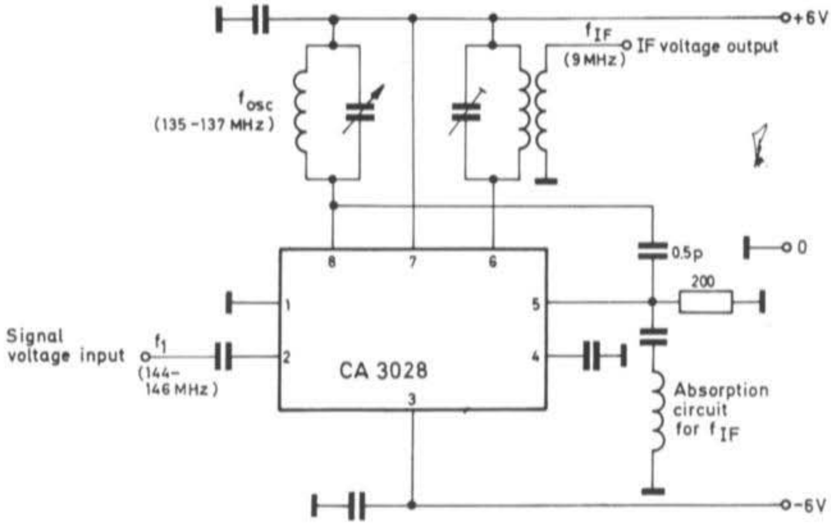


Fig. 7 : Using the integrated circuit CA 3028 as a self-excited mixer

2.6. SELF-EXCITED MIXER

Integrated circuits allow self-excited mixer circuits to be constructed whose oscillator frequency is virtually independent of the input signal value. Such a circuit is given in Fig. 7. Transistor system Q 1 operates in a common base circuit; the collector is coupled to the base of system Q 2. An absorption circuit, which is tuned to the intermediate frequency, is provided for neutralization. The multiplicative mixing occurs via the collector current of system Q 3.

To be continued in the next edition of VHF COMMUNICATIONS.

REFERENCES

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

(2) RCA Data Sheets : RF-Amplifiers CA 3028
RF-Amplifiers CA 3005, CA 3006
RCA Publications : ICAN 5337, ICAN 5022

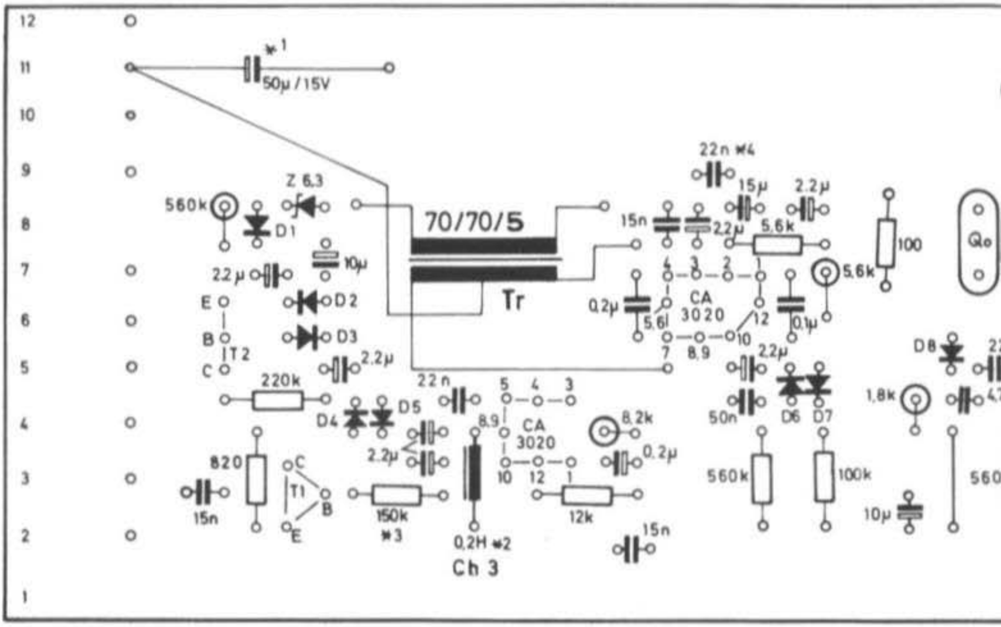


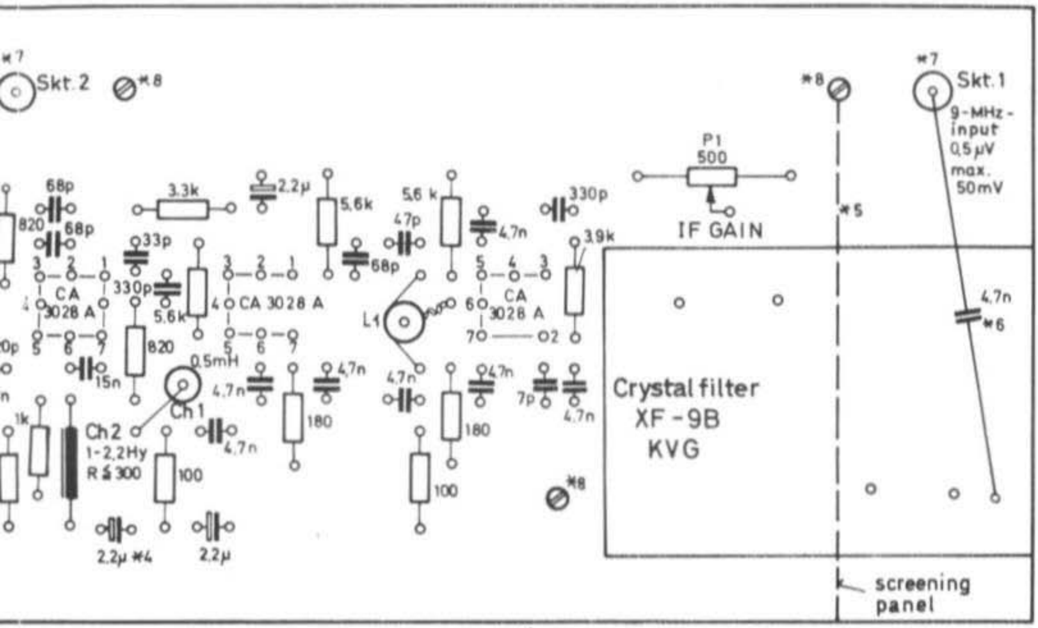
Fig.4c: Component location plan of the IF-AF amplifier

It should be pointed out that the operation of the described circuit is essentially dependent on the use of high quality components.

Inductance L 1	40 turns of 0.2 mm dia. (32 AWG) enamelled copper wire wound on a 4 mm coil former (standard coil former A) with SW core. Coil tap 12 turns from cold end.
Choke Ch 1	5 mH choke
Choke Ch 2	1 H choke; Z = 10 kΩ at 1 kHz; R = 300 Ω
Choke Ch 3	0.2 H choke; Z = 1 kΩ at 1 kHz; R = 100 Ω
Transformer	AF output transformer 70 Ω / 70 Ω / 5 Ω

The crystal filter type XF-9 B is available from the publishers or their national representatives.

The printed circuit board DJ 9 ZR 005 is provided with 12 connection points. It is possible for this board to be placed together with the 7 pole plug-in board DJ 9 ZR 006 (145 MHz board) into a 22 pole Amphenol connector. This allows both boards to be easily accomodated and exchanged. The order number of connector is 225-22221-101.



DJ 9 ZR 005

The twelve connections of the printed circuit board DJ 9 ZR 005 are as follows:

1 Ground
2 +9 V well filtered DC, switched off during "transmit"
3 S meter connection (see Fig. 3), fed via S meter to ground
4 Control voltage U_{C1} (0 V to +7 V) can be tapped off at this point
5 Control voltage U_{C2} (+1.2 V to +7.3 V) can be tapped off at this point
6 +9 V well filtered DC, not switched off during "transmit"
7 AF output, Z = 5 Ω
8 Ground
9 Connection of the 10 kΩ volume control (wiper)
10 Connection of the 10 kΩ volume control (hot end)
11 +9 V stabilized DC isolated from points 2 and 9 to avoid reaction on the AF preamplifier stages
12 Connection of switch S 1 for SSB-AM switching

The control voltages U_{C1} and U_{C2} can be tapped off for external use at connection points 4 and 5. See Section 7.

Notes regarding the component location plan Fig. 4 c:

+1 50 µF/15 V electrolytic capacitor will be necessary when a +9 V stabilized voltage is not available.
+2 Ch 3 may be mounted on the lower side of the PC-board for space reasons.
+3 If a BC 182 is chosen for T 1, increase this resistor to 270 Ω
+4 Drill necessary holes for this component.
+5 0.5 mm screening plate (brass) of 10 by 36 mm. Solder to the lower side of the PC-board
+6 4.7 nF capacitor accommodated on the lower side of the PC-board
+7 Coaxial sockets Skt 1 and 2 : BNC or similar
+8 Mounting holes for grounding screws.

DETERMINING THE IMPEDANCE OF QUARTER WAVE
GROUND PLANE ANTENNAS

by H.J. Dohlus, DJ 3 QC

INTRODUCTION

A description was given in Edition 2 of VHF COMMUNICATIONS (1) which showed what effect the dimensions of a rectangular counterpoise had on vertical rod antennas with lengths of $\lambda/4$, $\lambda/2$, $3\lambda/4$, $5\lambda/8$ and λ. The rod antennas were located in the centre of flat counterpoise plates having side-lengths of λ, $5\lambda/8$, $3\lambda/4$, $\lambda/2$, $\lambda/4$ and $\emptyset$. The 60 Ω feeder between slotted line and antenna was equipped with a quarter-wave coaxial sleath (balun) in order to avoid any sheath current.

The antenna configurations given in (1) are mainly of interest for mobile operation and for theoretical studies. The quarter-wave ground plane (and triple leg) antenna is, however, of more practical interest. Such antennas are equipped with a number of radials with the length l instead of a closed metal counterpoise. The radials are spread from the base of the rod antennas at an equal angle α from another and an angle β of 90° or more from the $\lambda/4$ rod (see Fig. 1). The counterpoise rods are referred to as radials even when β deviates from 90°. The number of radials is designated by n.

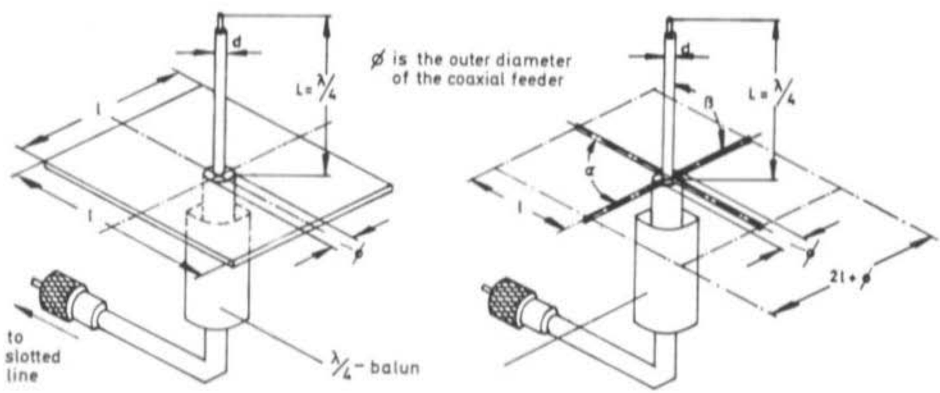


Fig. 1a: with counterpoise plate

n = ∞
α = 0
β = 90°
l = side length

Fig. 1b: with radial rods

n = 4
α = 90°
β = 90°
l = rod length

} One example of the nine measuring arrangements

Fig. 1 : Measuring arrangement to determine the feedpoint impedance of vertically polarized rod antennas

The following section explains how the impedance of $\lambda/4$ antennas with four, three and two radials is measured. The length 1 of the radials was $\lambda/2$, $3\lambda/8$, $5\lambda/16$, $\lambda/4$, $\lambda/8$, $\lambda/16$ and $\emptyset$. With four radials, the length 1 corresponds to approximately half the side-length of the counterpoise plate in (1) (see Fig. 1). The angle β between the radials and the antenna rod amounted to 90°, 120° and 135° during the various measurements. A quarterwave coaxial sleath was also used. The sign $\emptyset$ indicates that no radials were used.

The same measuring configuration was used during these measurements as was used in (1) i.e. a four-probe slotted line and an oscilloscope with direct Smith Chart representation. The radiation characteristics and gain were not measured.

1. RADIALS WITH AN ANGLE OF 90° TO THE $\lambda/4$ ROD

The metal counterpoise in (1) was located horizontally below the antenna rod and represents an infinite number of radials having a length approximately half that of the side-length. The angle α is zero, the angle β is 90°.

Measuring point	Radial length 1	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_X in Ω	Number of radials and angle α from another	Angle β to the vertical rod
1	$\lambda/2$	3.80 - j 1.90	228 - j 114		
2	$3\lambda/8$	0.62 + j 0.78	37 + j 47		
3	$5\lambda/16$	0.36 + j 0.12	21.5 + j 7	4 at α = 90°	90°
4	$\lambda/4$	0.27 - j 0.27	16 - j 16		
5	$\lambda/8$	0.28 - j 1.02	17 - j 61		
6	$\lambda/16$	0.30 - j 1.57	18 - j 94		
7	$\emptyset$	- j 3.30	- j 198		
1	$\lambda/2$	2.10 - j 2.90	126 - j 174		
2	$3\lambda/8$	0.88 + j 1.28	53 + j 77		
3	$5\lambda/16$	0.38 + j 0.20	23 + j 12	3 at α = 120°	90°
4	$\lambda/4$	0.27 - j 0.32	16 - j 19		
5	$\lambda/8$	0.28 - j 1.15	17 - j 69		
6	$\lambda/16$	0.30 - j 1.65	18 - j 99		
7	$\emptyset$	- j 3.30	- j 198		
1	$\lambda/2$	0.90 - j 2.78	54 - j 137		
2	$3\lambda/8$	4.40 + j 2.90	264 + j 174		
3	$5\lambda/16$	0.50 + j 0.54	30 + j 32.5	2 at α = 180°	90°
4	$\lambda/4$	0.27 - j 0.37	16 - j 22		
5	$\lambda/8$	0.28 - j 1.25	17 - j 75		
6	$\lambda/16$	0.30 - j 1.73	18 - j 104		
7	$\emptyset$	- j 3.30	- j 198		

Table I : Feedpoint impedance at an angle β of 90°.

Figure 2 and Table I show the result of the first measuring series with four, three and two radials of differing lengths having an angle β of 90° to the antenna rod in comparison to a rectangular metal counterpoise as shown in Fig. 1

or Fig. 3 in (1). Commencing with point 7 (counterpoise or radial length = ∅), the curves a, b, c and d have virtually the same run. This means that the feedpoint impedance of the $\lambda/4$ rod does not differ at low counterpoise or radial lengths as a function of the number of radials (four, three or two). The curve "a" leaves the common run at measuring point 4. Curve "b" runs together with "c" and "d" in excess of point 2. Curves "c" and "d" have a similar run.

The feedpoint impedance of the $\lambda/4$ rod with $\lambda/2$ radials is in the capacitive range with high impedance values. If the radials are slightly shortened in the direction of $1 = 3\lambda/8$, the impedance will become real and amount to approximately 330 Ω or 400 Ω according to the number of radials. At radial lengths between $1 = 5\lambda/8$ and $\lambda/4$ (between measuring points 3 and 4 of curves "b", "c" and "d" in Fig. 2), the feedpoint impedance is again real and amounts to approximately 20 Ω.

2. RADIALS WITH AN ANGLE OF 120° TO THE $\lambda/4$ ROD

Fig. 3 and Table II show the results of the second measuring series with four, three and two radials of differing lengths having an angle β of 120° to the $\lambda/4$ rod. This is compared to a rectangular counterpoise according to (1) having an angle $\beta = 90°$.

Measuring point	Radial length 1	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_X in Ω	Number of radials and angle α from another	Angle β to the vertical rod
1	$\lambda/2$	0.80 - j 2.20	48 - j 132		
2	$3\lambda/8$	5.50 + j 1.50	330 + j 90		
3	$5\lambda/16$	0.80 + j 0.55	48 + j 33	2 at $\alpha = 180°$	120°
4	$\lambda/4$	0.50 - j 0.39	30 - j 23.4		
5	$\lambda/8$	0.33 - j 1.33	20 - j 80		
6	$\lambda/16$	0.30 - j 1.79	18 - j 107.5		
7	∅	- j 3.30	- j 198		
1	$\lambda/2$	2.40 - j 2.40	144 - j 144		
2	$3\lambda/8$	1.40 + j 0.80	84 + j 48		
3	$5\lambda/16$	0.74 + j 0.18	44.5 + j 11	4 at $\alpha = 90°$	120°
4	$\lambda/4$	0.50 - j 0.25	30 - j 15		
5	$\lambda/8$	0.40 - j 1.03	24 - j 62		
6	$\lambda/16$	0.30 - j 1.62	18 - j 97		
7	∅	- j 3.30	- j 198		
1	$\lambda/2$	1.38 - j 2.85	83 - j 171		
2	$3\lambda/8$	1.80 + j 1.40	108 + j 84		
3	$5\lambda/16$	0.68 + j 0.24	42 + j 14.4	3 at $\alpha = 120°$	120°
4	$\lambda/4$	0.50 - j 0.30	30 - j 18		
5	$\lambda/8$	0.37 - j 1.20	22 - j 72		
6	$\lambda/16$	0.30 - j 1.71	18 - j 103		
7	∅	- j 3.30	- j 198		

Table II : Feedpoint impedance at an angle $\beta = 120°$.

β = 90°

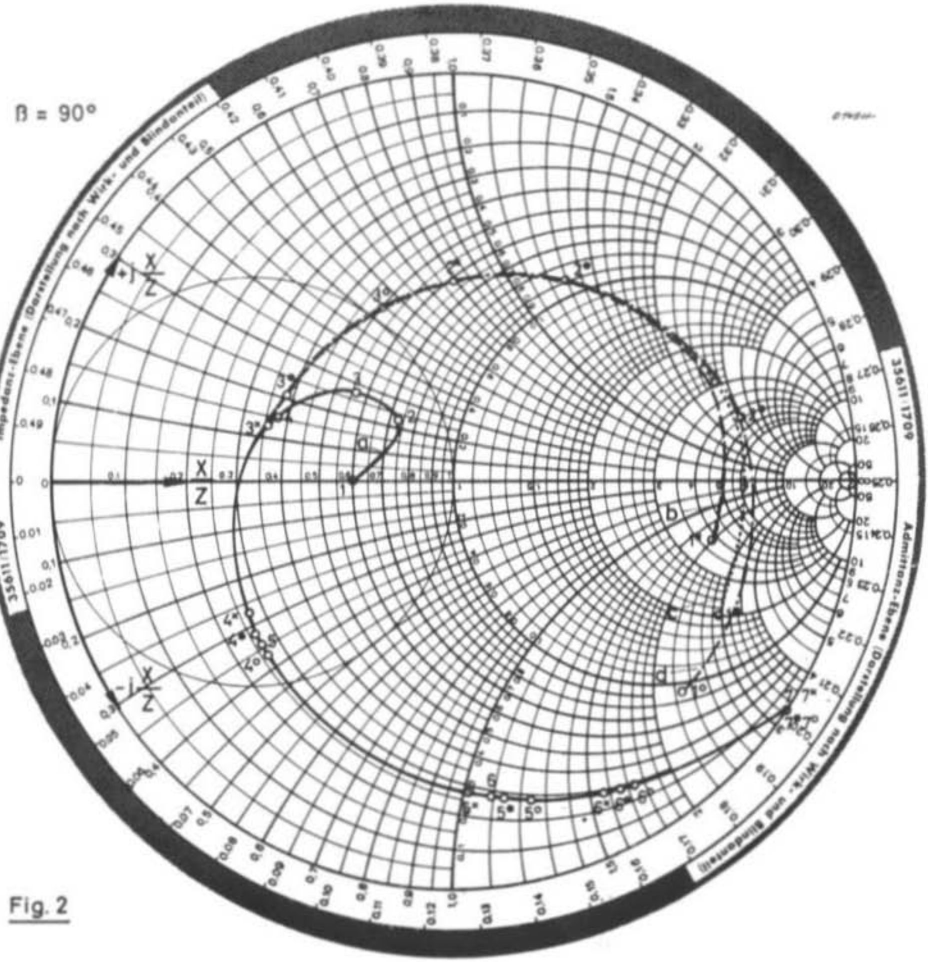


Fig. 2

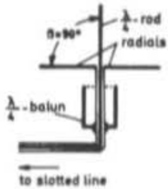


Fig. 2 :
Feedpoint impedance of vertically polarized λ/4 rod antennas having an angle β of 90° to the counterpoise or radial rods as a function of the radial length l.

——— a) rectangular metal plate (see Fig. 1 or 3 in (1)). n =
------- b) four radial rods, n = 4, ∝ = 90°, measuring points 1 to 7
.... c) three radial rods, n = 3, ∝ = 120°, measuring points 1 to 7
-.-.-.- d) two radial rods, n = 2, ∝ = 180°, measuring points 1 to 7

Measurement made with λ/4 coaxial balun.

Curves "b", "c" and "d", which have a common run until after point 4, deviate from the reference curve "a" (counterpoise plate β = 90°) very shortly after leaving the commencement point 7 (1 = ∅). The curves ("b", "c" and "d") cut the real axis between points 1 and 2 at values of 240 Ω and 380 Ω at radial lengths of between λ/2 and 3 λ/8.

At radial lengths between 5 λ/16 and λ/4, curves "b", "c" and "d" cross the real axis between points 3 and 4 at approximately 36 Ω, where they correspond to point 1 of curve "a".

3. RADIALS WITH AN ANGLE OF 135° TO THE λ/4 ROD

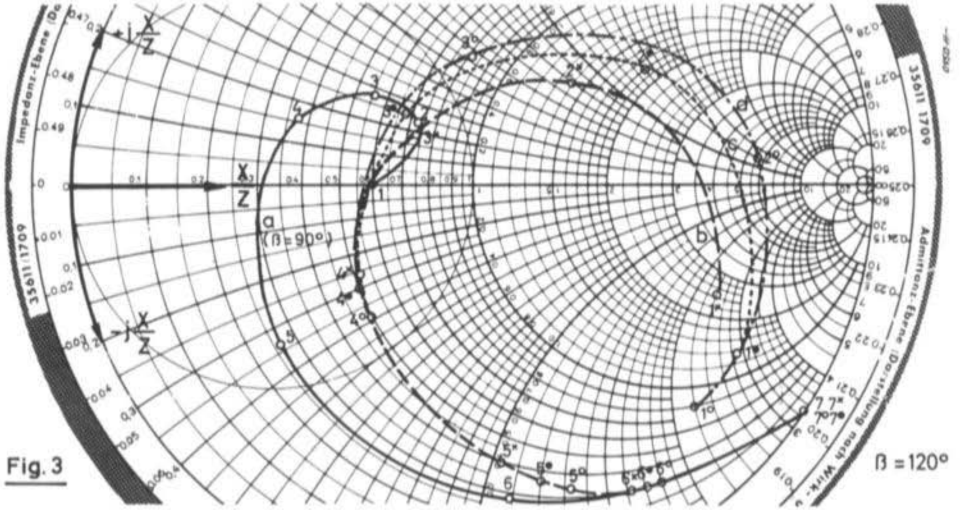
Fig. 4 and Table III contain the results of the third measuring sequence with four, three and two radials of various lengths having an angle β of 135° to the λ/4 rod. This is again compared to a rectangular counterpoise according to (1) having an angle β = 90°.

Measuring point	Radial length 1	Reference value X/Z in the Smith Chart	Value of the measured impedance Z_X in Ω	Number of radials and angle α from another	Angle β to the vertical rod
1	λ/2	2.00 - j 2.30	120 - j 138		
2	3 λ/8	1.62 + j 0.45	97 + j 27		
3	5 λ/16	0.90 + j 0.02	54 + j 1.2	4 at α = 90°	135°
4	λ/4	0.62 - j 0.35	37 - j 21		
5	λ/8	0.40 - j 1.07	24 - j 64		
6	λ/16	0.30 - j 1.62	18 - j 97		
7	∅	- j 3.30	- j 198		
1	λ/2	0.90 - j 2.90	54 - j 174		
2	3 λ/8	2.10 + j 1.00	126 + j 60		
3	5 λ/16	0.94 + j 0.05	56 + j 3	3 at α = 120°	135°
4	λ/4	0.62 - j 0.42	37 - j 25		
5	λ/8	0.40 - j 1.20	24 - j 72		
6	λ/16	0.30 - j 1.67	18 - j 100		
7	∅	- j 3.30	- j 198		
1	λ/2	0.45 - j 2.30	27 - j 138		
2	3 λ/8	5.50 + j 1.80	330 + j 108		
3	5 λ/16	1.05 + j 0.25	63 + j 15	2 at α = 180°	135°
4	λ/4	0.60 - j 0.50	36 - j 30		
5	λ/8	0.35 - j 1.37	21 - j 82		
6	λ/16	0.30 - j 1.74	18 - j 104		
7	∅	- j 3.30	- j 198		

Table III : Feedpoint impedance at an angle β = 135°.

Curves "b", "c" and "d" deviate from the reference curve "a" (1) very soon after leaving the common commencement point 7 and run in a tighter curve than in Fig. 2 or Fig. 3. The real axis is cut between points 1 and 2 (radial lengths of between λ/2 and 3 λ/8) at impedance values of 210 to 420 Ω.

Fig. 3



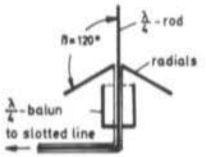
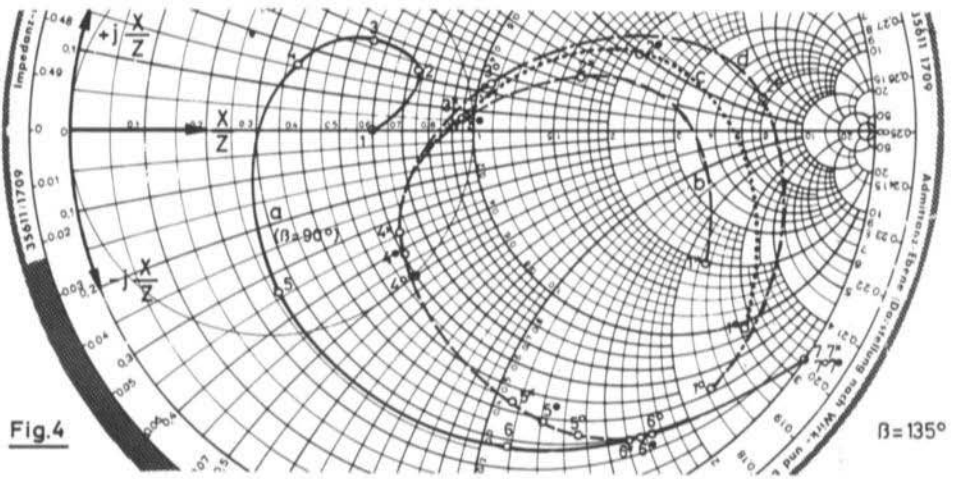


Fig. 3 :
Feedpoint impedance of vertically polarized λ/4 rod antennas having an angle β of 120° to the radials as a function of the radial length l. Counterpoise plate has β = 90°.

————	a) rectangular metal plate (see Fig. 1 or 3 in (1)). n = ∞ , β = 90°
- - - - - - -	b) four radial rods, n = 4, α = 90°, meas. points 1 to 7 β = 120°
· · · · · · · ·	c) three radial rods, n = 3, α = 120°, meas. points 1 to 7 β = 120°
- · - · - · -	d) two radial rods, n = 2, α = 180°, meas. points 1 to 7 β = 120°

with λ/4 coaxial balun.

Fig.4



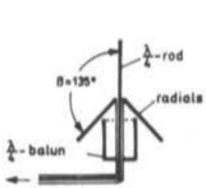


Fig. 4 :
Feedpoint impedance of vertically polarized λ/4 rod antennas having an angle β of 135° to the radials as a function of the radial length l. Counterpoise plate has β = 90°.

————	a) rectangular metal plate (see Fig. 1 or 3 in (1)). n = ∞ , β = 90°
- - - - - - -	b) four radial rods, n = 4, α = 90°, measuring points 1 to 7 , β = 135°
· · · · · · · ·	c) three radial rods, n = 3, α = 120°, measuring points 1 to 7 , β = 135°
- · - · - · -	d) two radial rods, n = 2, α = 180°, measuring points 1 to 7 , β = 135°

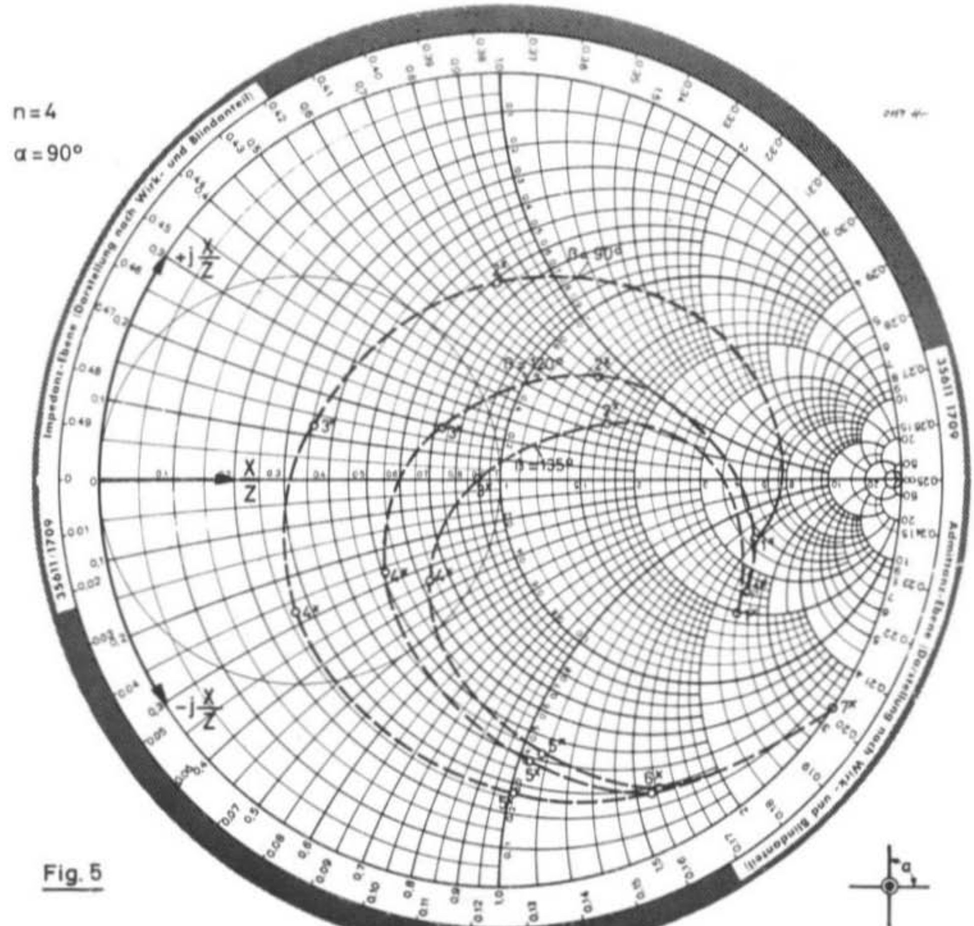


Fig. 5

Fig. 5 :
Feedpoint impedance of vertically polarized $\lambda/4$ rod antennas having n = 4 radials (α = 90°) as a function of the rod length l and the angle β. See Tables I, II and III as well as Fig. 2 to Fig. 4.

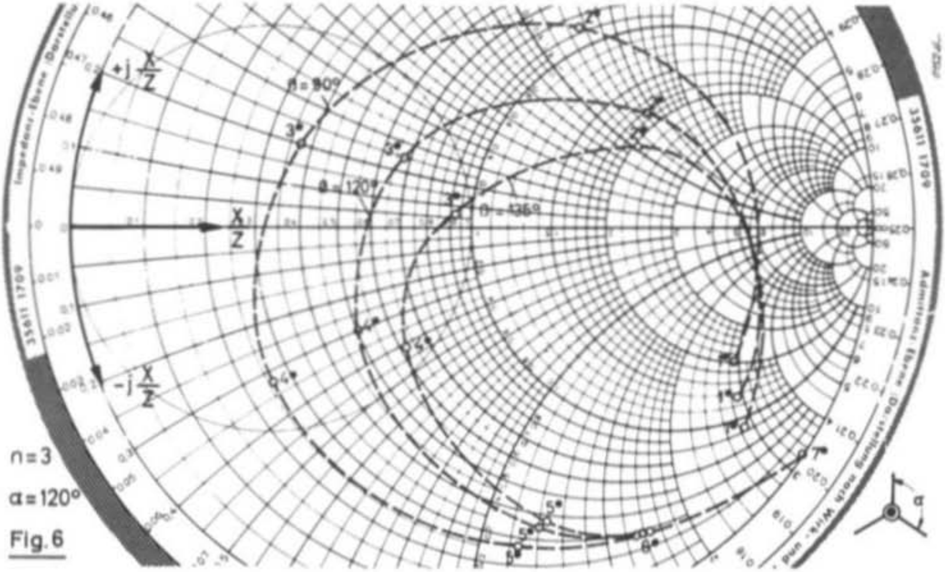


Fig. 6 :
Feedpoint impedance of vertically polarized λ/4 rod antennas having n = 3 radials (α = 120°) as a function of the rod length l and the angle β. See Tables I, II and III as well as Fig. 2 to Fig. 4.

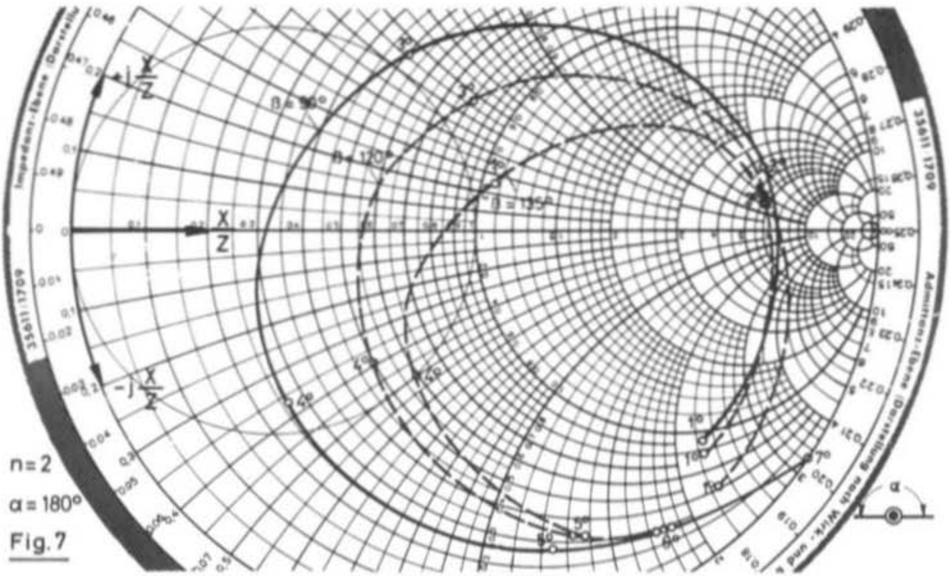


Fig. 7 :
Feedpoint impedance of vertically polarized λ/4 rod antennas having n = 2 radials (α = 180°) as a function of the angle β. See Tables I, II and III as well as Fig. 2 to Fig. 4.

Curves "b" and "c" cross the real axis at approximately 55 Ω (measuring point 3) with four or three radials of 5 $\lambda/16$. Curve "d" obtains this value between measuring points 3 and 4, i.e. between $l = 5\lambda/16$ and $\lambda/4$.

4. FEEDPOINT IMPEDANCE OF THE $\lambda/4$ ROD WITH n RADIALS AS A FUNCTION OF THE ANGLE β.

Figures 5, 6 and 7 show the dependence of a vertically polarized $\lambda/4$ rod antenna with a fixed number of radials n on the radial length l and the angle β. Further details are given in Tables I, II and III.

As can be seen in Figures 5, 6 and 7 the feedpoint impedance is shifted in the high impedance direction on increasing the angle β with radial lengths between 5 $\lambda/16$ and $\lambda/4$ (points 3 and 4). This is reversed at radial lengths of between $\lambda/2$ and 3 $\lambda/8$ (points 1 and 2). This curvature also increases together with the angle, independent of the number of radials n and their length l.

5. RESULTS OF THE MEASUREMENT

The feedpoint impedance of $\lambda/4$ ground plane antennas with a differing number of radials is somewhat dependent on the radial length l but far more so on the angle β . The number of radials n does not have very much effect.

By altering the angle β with respect to the $\lambda/4$ rod and correspondingly matching the length l it will be possible to obtain a real feedpoint impedance of 20 to 60 Ω. It is, on the other hand, also possible to obtain impedances in the order of 200 to 420 Ω. The vertical radiation characteristics were not examined. However, they seem to be at higher angles with a radial - to - rod angle β of 90° than at $\beta = 135°$. The maximum horizontal radiation is at intermediate angles α between the radials, however, the difference between maximum and minimum values is very small (no zero points).

The feedpoint impedance of $\lambda/4$ rod antennas with only two radials is effected far more by external influences than those with three or more radials - especially at high impedance values ($l > 5\lambda/16$). This antenna was only examined to complete the measuring series. Antennas with only two radials are not very favourable since the greater part of the radiated power is horizontally polarized, dependent on the length l and the angle β. This is not normally the case with three or more radials.

The rod length L can also be varied (1) to obtain a variety of feedpoint impedances. The vertical radiation angle can be very unfavourable if $L > \lambda/4$.

REFERENCES

(1) H. J. Dohlus: Determining the Impedance of Rod Antennas in the VHF Range. VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 98-109.

A CALIBRATED ATTENUATOR

by J. Wasmus, DJ 4 AU and G. Laufs, DL 6 HA

1. INTRODUCTION

The radio amateur is often faced with the problem of comparing voltage, current or power values. This requires sufficiently accurate measuring devices such as the versatile calibrated attenuator. An attenuator represents a passive four-pole placed between the test object (antenna, signal generator, converter etc.) and the indicating instrument (voltmeter, receiver etc.). Two easily assembled attenuators are described which will allow the amateur to carry out a great number of measurements in conjunction with his equipment.

2. ATTENUATOR I, 0 to 10 dB

This attenuator is variable in steps of 1 dB. It possesses ten different π sections switched in by a switch having two wafers each with eleven contacts. The mechanical assembly of the attenuator is given in Fig. 1. The switch is firstly dismantled and the distance pieces between the switch wafers are removed. The switch wafers are now assembled together with the circular screening plates as shown in the diagram. Both the wafers and the screening plates are held onto the tapped bolts by means of nuts. The diameter of the screening plates must be at least 2 to 3 cm greater than the switch wafers. The authors assembled the attenuator in a household can. The inner diameter of this can then determines the diameter of the screening plates. The lower screening may be made a little larger and slots cut along the edge at a spacing of 5 mm. This ensures that a good contact is made to the inner wall of the screening can. The centre screening plate possesses 11 holes which are arranged so that the resistor R 2 of the individual sections can be soldered vertically between the contacts of the switch wafers. After the resistors R 2 have been soldered into place, the resistors R 1 and R 3 are soldered between the switch contacts and the upper or lower screening plate. It is important that the resistor connection leads should be as short as possible.

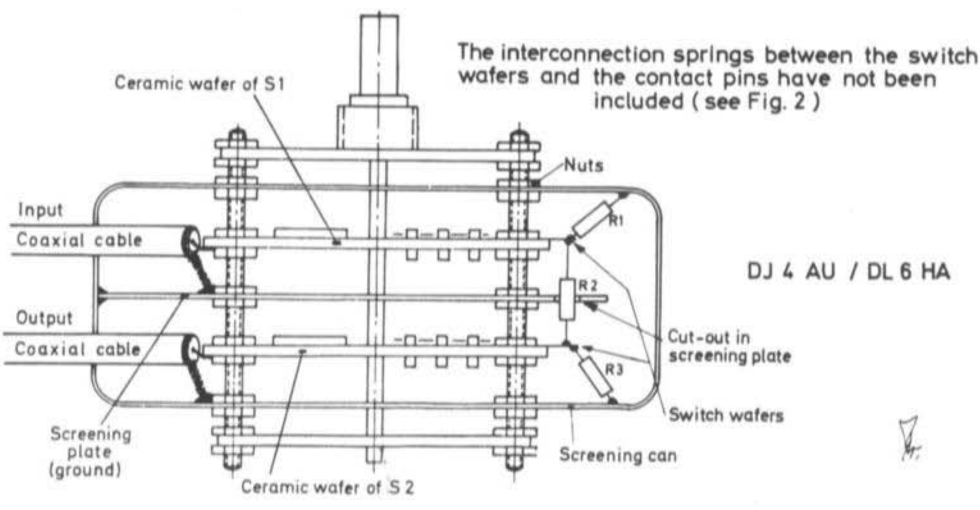


Fig.1: Mechanical build-up of switch S1/S2

The sliding contacts of both switch wafers are connected to the coaxial sockets by means of coaxial cable. The coaxial sockets should be spaced as far from another as possible and the screen of the coaxial cable brought right up to the switch contacts, where it is grounded to the nearest point of the screening plate. This ensures that the coupling attenuation between the input and output of the attenuator is as high as possible, so that the voltage to be measured is virtually only passed via the attenuator to the indicating instrument and not coupled in an unpredictable manner from the input to the output. The coupling attenuation of the described attenuator at a frequency of 145 MHz was 50 dB, which is sufficient for the selected measuring range.

The circuit and the resistance values of the attenuator sections for impedance values of 60 Ω and 52 Ω are given in Fig. 2. In order to maintain these resistance values even at high frequencies, it is necessary to use virtually non-inductive resistors. The required resistance values are given in Table 1. There are two methods of obtaining the necessary values. Firstly to use high-precision resistors with a tolerance of 1% or secondly to measure a number of 10% resistors with slightly different values until the required resistance is found.

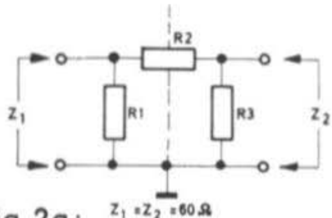


Fig. 2a: Fundamental diagram of an individual attenuator section

dB	R1,R3$_{(60)}$	R 2$_{(60)}$	R1,R3$_{(52)}$	R 2$_{(52)}$
1	1043 Ω	6.9 Ω	904 Ω	6.0 Ω
2	523	13.9	453	12.0
3	350	22	303	19
4	265	28.6	230	24.8
5	214	36.5	185	31.6
6	181	44.8	157	38.9
7	155	50	134	43.3
8	140	63.4	121	55
9	120	70	104	61
10	115	85	100	74
*) 20	73	297	63	257

*) Not built into the switch

Table 1: Resistance values of the attenuator sections

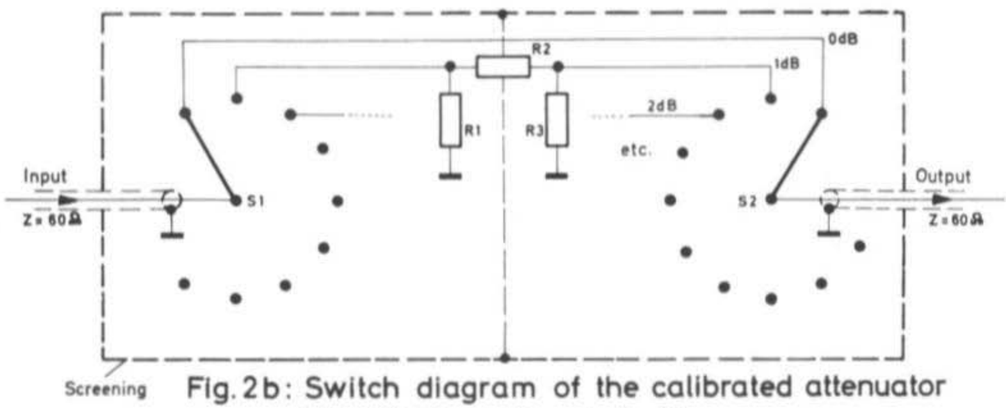


Fig. 2b: Switch diagram of the calibrated attenuator
(Switch build-up shown in Fig.1)

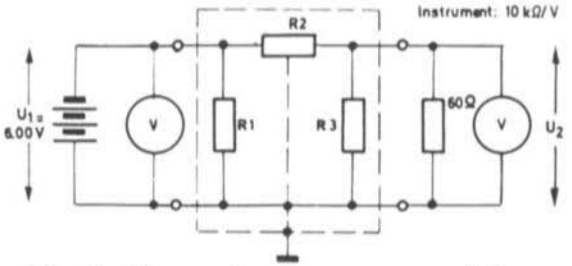


Fig. 3 : Measuring arrangement for checking the dB steps

Attenuation	Output voltage
0	6.00
1	5.35
2	4.75
3	4.20
4	3.78
5	3.36
6	3.00
7	2.68
8	2.38
9	2.10
10	1.88
dB	Volt

Table 2: Results of the DC measurement on attenuator I according to Fig. 3

Figure 3 shows a measuring arrangement allowing the attenuator to be checked with a DC voltage. The attenuator is connected to a DC source of 6 V and the output terminated with either a 60 Ω or 52 Ω resistor. Table 2 lists the voltage values which should be measured in the individual dB steps.

The impedance of the voltage source is not too important during this measurement but the termination must amount to the impedance value if the measured voltage is to coincide with the given values. The characteristic impedance is also formed at the input of the attenuator but this is only important when measuring in conjunction with a matched cable, for instance, with a feeder between antenna and receiver or if the generator is sensitive to variations of the load impedance, as is the case at the output of varactor multiplier stages.

3. ATTENUATOR II, 0 to 110 dB

A potentiometer type 4955 manufactured by the German firm of Preh is used as the attenuator pad in this attenuator. The mechanical assembly is shown in Fig. 4. The specifications of the RF potentiometer are given in Table 3.

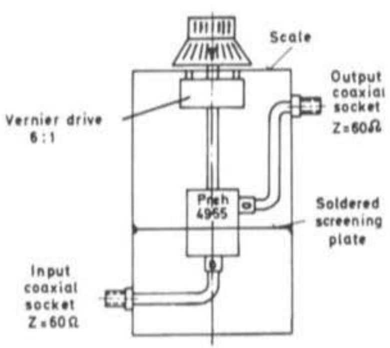


Fig. 4 : Attenuator II, 6 to 110 dB

Table 3:

Impedance	Z approx. 60 Ω
Total attenuation	110 dB ± 15 dB
Insertion loss	≦6 dB
Frequency limit	10 GHz
(noticeable frequency dependence above 300 MHz)	
Power rating	0.1 W
(0.1 W into 60 Ω = 2.45 V)	

Specifications of the screened RF potentiometer. Manufactured by Preh under the designation 4955.

However, it is often inadvisable at VHF to measure attenuation values up to 110 dB in one step. It is often very difficult even at HF to obtain a coupling attenuation of more than 110 dB between the input and output of the attenuator. At higher frequencies, it is not only this coupling attenuation that causes difficulties but especially the connected amateur radio equipment, which, with its sometimes insufficient screening and poor cable connections, does not allow coupling attenuation values of even 60 dB.

A measuring arrangement for the calibration of the variable attenuator is shown in Fig. 5. The variable output level of a signal generator is connected via attenuators I (0 to 10 dB) and II (0 to 110 dB) to a receiver equipped with a S meter. Attention must be paid that the input impedance of the receiver does not differ greatly from the characteristic impedance values at the frequency and voltage range in question. If it is not possible for the impedance to be determined, it is advisable to place an attenuator with a fixed attenuation of 10 dB (5 dB is even better) between attenuator II and the receiver.

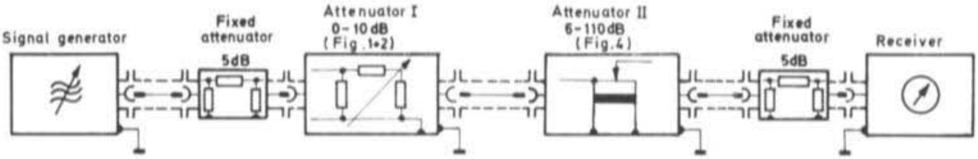


Fig.5: Measuring arrangement for the calibration of attenuator II with the aid of attenuator I

The input impedance of the Preh RF potentiometer deviates slightly from 60 Ω and can vary even more at low attenuation values. If the generator output impedance also deviates from 60 Ω, a second fixed attenuator of 5 dB should be used to keep the influence of such impedance variations on the calibration procedure at a minimum.

The calibration process is made as follows: Attenuator I is switched to 0 dB and attenuator II adjusted for the maximum possible attenuation. The output voltage of the power signal generator - such a generator is required since several volts are needed - is adjusted for an S meter deflection of approximately one third of the full scale value. The meter reading is now noted. Attenuator I is now switched to 10 dB which will cause a reduction of the S meter reading. The attenuation of attenuator II is reduced until the original meter reading is indicated. This point is then marked on the scale of attenuator II.

Attenuator I is now switched back to 0 dB and the output voltage of the signal generator is reduced until the original S meter reading is re-obtained. This calibration procedure is repeated again and again until the whole scale of attenuator II has been calibrated in 10 dB steps.

4. FIXED ATTENUATOR OF 10 dB

An attenuator was built-up as an accessory to attenuator I (0 to 10 dB). This attenuator has a fixed attenuation of 10 dB and is especially useful during antenna measurements. It extends the attenuation range of attenuator I from 1 to 10 dB up to 20 dB. The mechanical build-up is shown in the sectional diagram Fig. 6. The use of a coaxial plug PL 259 and a matching socket SO 239, from which the flange has been removed, allows this attenuator to be easily connected to any coaxial cable. The resistance values of R 1, R 2 and R 3 are given in Table 1.

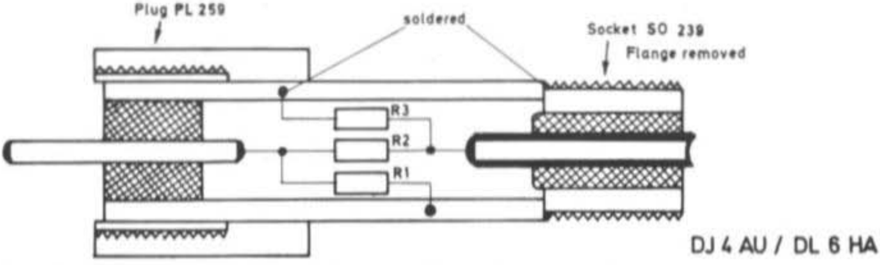


Fig. 6 : Sectional view of the 10 dB fixed attenuator

MODULATION INDEX FOR NBFM TRANSMISSIONS

At the Region I Conference of the IARU in Brussels, May 1969, agreement was reached regarding the modulation index for amateur radio transmissions in the narrow band frequency modulation (NBFM) mode. A modulation index of 1 and a modulation frequency of 3 kHz was laid down. This means that the maximum deviation may also amount to 3 kHz which results in a transmission (or filter) bandwidth of 12 kHz.

This decision was made early enough to ensure that amateur radio FM transmissions do not use differing modulation indices - at least in Region I of the IARU. In addition to this, a maximum frequency deviation of 3 kHz usually allows FM demodulation with AM receivers (skirt demodulation) until VHF receivers equipped with FM discriminators are in general use. VHF COMMUNICATIONS will be describing two modern FM demodulators (a digital discriminator and quartz crystal discriminator) in one of the next editions as well as an article discussing the advantages and disadvantages of the various modulation modes.

A SIMPLE ELECTRONIC FUSE

by R. Lentz, DL 3 WR

1. INTRODUCTION

Those readers who have experimented with power transistors will know the "sudden death" that can occur with these (usually the most expensive) transistors if they break into oscillation due to a tuning or loading error. The power output and modulator stages of transistor transmitters are mostly endangered because they are usually built-up without current-limiting emitter resistors - especially at low operating voltages - in order to achieve a high efficiency. Conventional fuses do not offer any protection since they are too inert. This means that the semiconductor will be destroyed before the fuse is able to break the current flow.

A simple electronic fuse (1) is to be described which guarantees a rapid cut-out within approximately 100 μs.

The characteristics of this electronic fuse are three-fold:

a) It does not completely break the current flow but allows approximately 5% of the selected cut-out current value to flow. This is, however, not very important for amateur applications.

b) It is connected into the circuit as a two-pole in the same manner as a conventional fuse and does not require an additional power supply.

c) It is possible to build up the fuse on a small printed circuit board.

2. CIRCUIT DESCRIPTION

As can be seen in Fig. 1, the power transistor T 1 is in the main current circuit. This transistor is maintained in its conducting state under normal conditions by transistor T 2. The (low) voltage drop across T 1 and resistor R 1 also appears across the series-connected resistors R 4 and P 1. A portion of this voltage is tapped off. Transistor T 3 will remain blocked as long as this voltage is lower than the sum of the diffusion potentials of diode D 1 and the base-emitter path of transistor T 3. In this state, transistors T 2 and T 1, which represent a Darlington circuit, will conduct.

The voltage drop across R 1 and T 1, and thus across R 4 and P 1, increases on advancing the current flow. The tapped off voltage will cause T 3 to conduct as soon as the threshold is reached which in turn causes T 2 and T 1 to block. This means that the whole operating voltage is virtually dropped across the electronic fuse. This represents the maximum voltage at P 1 and thus ensures that the cut-out condition is maintained until either the operating voltage or the (over) load are momentarily disconnected.

In the cut-out state, a small residual current will flow through the electronic fuse and thus the protected circuit. The value of this residual current is determined by resistor R 3. If transistor types having a current amplification of $B \geq 100$ are used for transistors T 1 and T 2, it will be possible to increase the value of R 3 to 1.2 kΩ/1 W (or even more) which will reduce the residual current (see Section 3.).

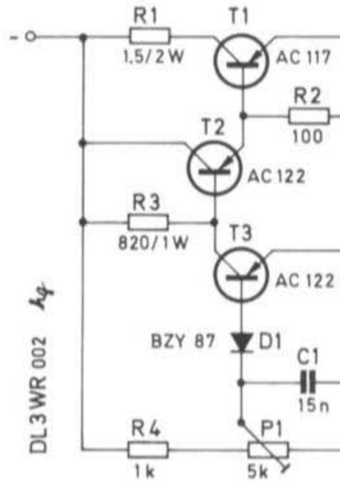


Fig. 1: Circuit of the electronic fuse for a maximum of 1 A at 24 V

Capacitor C 1 ensures that steep current surges do not control transistor T 3 via the capacity of diode D 1 which would cause the electronic fuse to be actuated too quickly.

The transistors are used merely as switches, which is the reason why their dissipation is low. The transistor selection is therefore mainly dependent on the expected current values. With continuous currents of more than 0.5 A through the electronic fuse, the dissipation ($I_{fuse} \times U_{CE}$) of the fully conducting transistor T 1 will reach values that require some form of heat sink. This is also the case with resistor R 1 : a rating of 0.3 W will be sufficient for continuous currents of up to 0.5 A; a 2 W rating will be required for currents up to 1 A.

Since the full operating voltage is connected across the fuse when actuated, the selected transistors must be able to handle this voltage. The following table allows a suitable semiconductor complement to be found for operating voltages U_{op} of up to 24 V or 36 V.

Silicon PNP transistors are not suitable due to their higher diffusion voltage.

T 1	T 2, T 3	D 1	U_{op}
AC 117	AC 122	BZY 87	
AC 128	AC 125, AC 126	BAY 86 to BAY 88	
OC 318	OC 304	1 N 816, 1 N 912 M	
AC 153 K, AC 154	AC 151	1 N 2938, 1 N 3896	up to 24 V
2 N 672	2 N 187, 2 N 188	1 N 4362, 1 N 4828	
2 N 2001, 2 N 4105	2 N 2428, 2 N 2953	(or any silicon diode	
AC 124	AC 122/30	having an U_F= 0.6 -	
2 N 674	2 N 217, 2 N 280	0.75 V at I_F = 5 mA)	up to 36 V
2 N 2000	2 N 1274, 2 N 2447		

3. SPECIFICATIONS

The described version was found to have the following characteristics:

Max. switchable voltage U_{op}:	24 V or 36 V (see transistor table)
Max. continuous current:	1 A
Voltage drop across the fuse:	1 V at 0.5 A or 1.9 V at 1 A
Cut-out threshold adjustable between:	0.1 A and 1 A.
Residual current after cut-out	

R 3 = 820 Ω :	17 mA at U_{op} = 12 V
	34 mA at U_{op} = 24 V
R 3 = 1.2 kΩ:	12 mA at U_{op} = 12 V
	24 mA at U_{op} = 24 V

4. MECHANICAL ASSEMBLY

The fuse can be built up on a printed circuit board having the dimensions 50 mm by 35 mm. As has already been mentioned, it will be necessary to mount transistor T 1 on a heat sink if continuous currents of more than 0.5 A are to flow through the electronic fuse. Three solder tags are then connected to the PC-board instead of the transistor to which the connection leads are soldered. The same is valid for potentiometer P 1 if the current threshold is to be often varied which would require P 1 to be located on the front panel.

A photograph of the electronic fuse is given in Fig. 2. The printed circuit board is so dimensioned that it does not only accept miniature resistors but also somewhat larger types (from the junk box). A diagram of the printed circuit board is given in Fig. 3 and the associated component location plan in Fig. 4.

The most space is taken up by resistors R 1 and R 3 as well as potentiometer P 1.

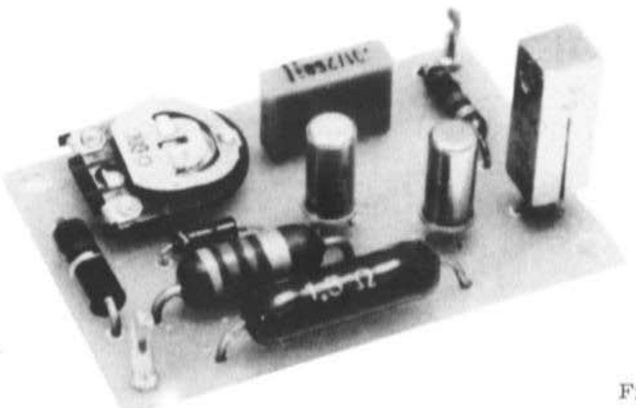


Fig. 2 : Photograph of the electronic fuse

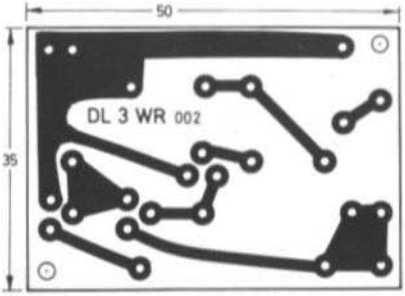


Fig. 3 : Printed circuit board of
the electronic fuse

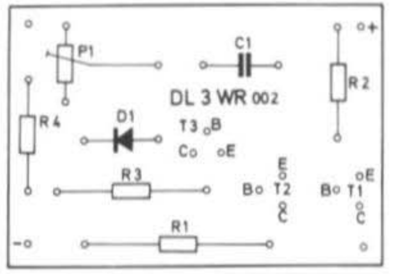


Fig. 4 : Component location plan
associated to Fig. 3

5. NOTES

Since the electronic fuse causes a noticeable voltage drop, it is advisable to only use the fuse during alignment procedures or test periods. During mains (line) operation, however, it has been found very favourable to place the electronic fuse between the rectifier and pass transistor of a stabilized power supply. The impedance of the fuse does not have any adverse effect due to the subsequent stabilization. The authors prototype used such a system with the transistor complement: AC 124, AC 122 (2), BZY 87 connected in such a configuration. This fuse was found to be capable of handling an operating voltage rising to 30 V after cut-out. Transistor T 1 is mounted on the chassis. Resistor R 1 had a value of 1.8 Ω due to the fact that no more suitable resistor could be found. The circuit will not operate if R 1 = 1.2 Ω.

6. AVAILABLE PARTS

The printed circuit board DL 3 WR 002 of the electronic fuse is available from the publishers or their national representatives (see advertising page).

7. REFERENCES

(1) G. Günzel : Electronische Sicherung
Internationale Elektronische Rundschau 1966, Edition 11, pages 628-629.

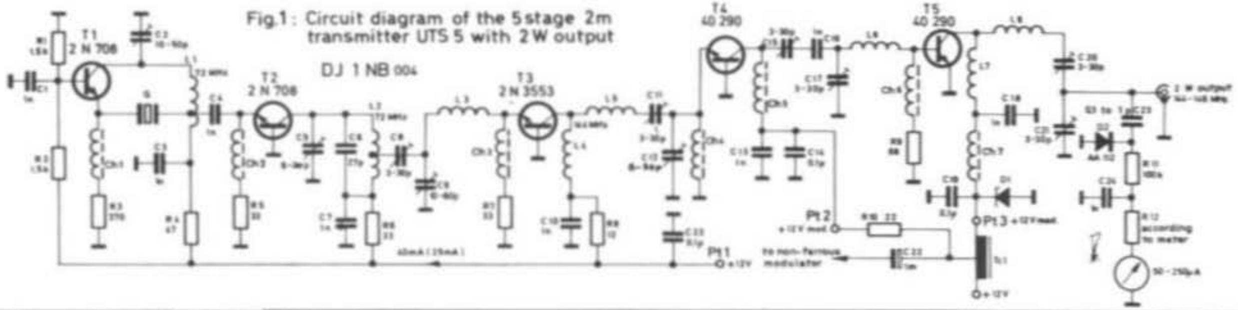


Fig.1: Circuit diagram of the 5 stage 2m transmitter UTS 5 with 2W output

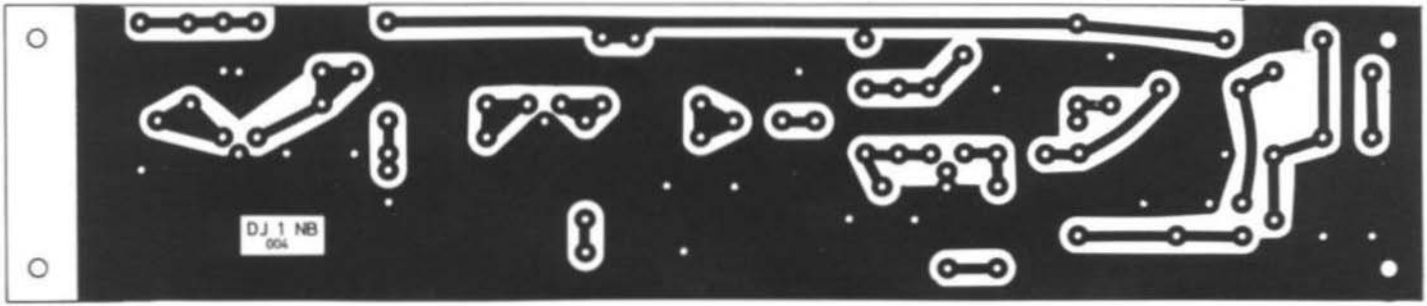


Fig.2: Printed circuit board DJ1NB004(centre)

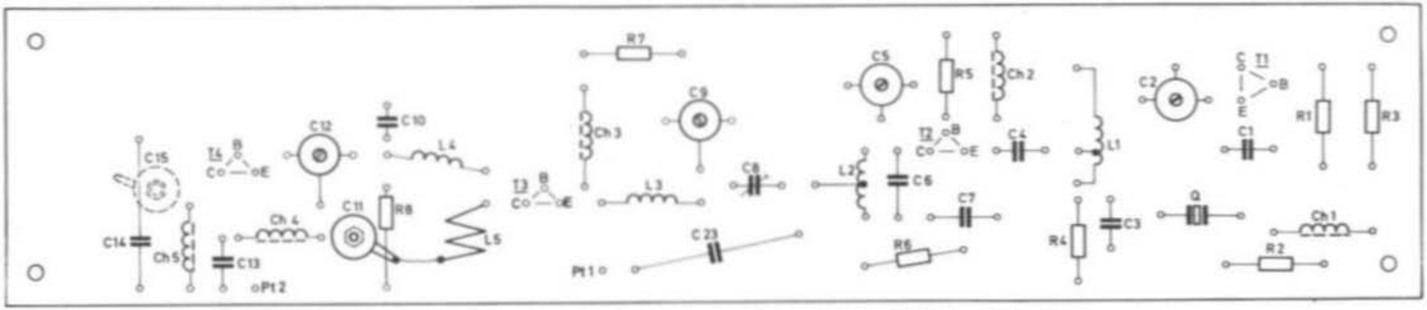


Fig.3: Component location plan DJ1NB004(below)

THE 2 METRE TRANSMITTER UTS 5 WITH 2 WATTS MEAN OUTPUT AT AN OPERATING VOLTAGE OF 12 V

by E. Flügel, DJ 1 NB

1. INTRODUCTION

Most of present transistor transmitters designed for output power levels of more than 0.5 W to 1 W at 144 MHz usually require an operating voltage of at least 18 V. However, an operating voltage of 12 V would be more favourable since not only battery costs would be lower, but it would also be possible to use a 12 V car battery as power source - expecially during mobile operation - without having to use a DC-DC converter. Modern overlay transistors especially developed for amplitude modulated operation from an operating voltage of 13.5 V allowed the construction of the described two metre transmitter. The transistor type 40 290 (RCA) is used in the driver and power amplifier stages. This transistor provides a mean carrier output of 2 W to 2.25 W at an operating voltage of 13.5 V. With a modulation depth of 100%, this will correspond to a peak envelope power (PEP) of 8 W to 9 W.

The five-stage transmitter has enough power reserve in each of the individual stages to compensate for transistor fluctuations and differing crystal characteristics so that it is not necessary to retune after changing frequencies within the band. This guarantees a constant, high quality modulation, even when switching from the extreme band limits, without loss of power.

The overall efficiency of the transmitter is 30% to 40% (according to the individual transistors). This value also includes the quiescent current of a 7 stage modulator. The first four stages of the transmitter are mounted on a printed circuit board; the power amplifier stage is mounted on a brass plate.

2. CIRCUIT DESCRIPTION

2.1. FIVE STAGE TRANSMITTER

The circuit diagram of the five stage transmitter is given in Fig. 1. It will be noticed that the 72 MHz crystal oscillator with transistor T 1 operates in an overtone circuit. The emitter of the buffer amplifier (transistor T 2) is driven simultaneously from the very low impedance feedback link. Both stages operate in a common-base configuration with a low emitter impedance. Due to the emitter coupling, defined drive characteristics result for the buffer stage which means that a relatively low reaction and minimum coupling loss result. The low emitter impedance of the oscillator stage, which makes neutralization of the crystal holder capacity superfluous, is only achieved at high collector currents when using low-gain transistors (e.g. 2 N 708). It also causes a certain damping of the collector circuit which means that trimmer capacitor C 2 need not be realigned when exchanging crystals within the two metre band.

The amplified 72 MHz signal is available at the collector circuit of the second stage. The signal is matched to the low impedance of the subsequent frequency doubler stage by a network comprising C 8, C 9 and L 3. The frequency doubler stage (transistor T 3) also operates in a common-base circuit and is equipped with the well-known overlay transistor 2 N 3553. This transistor provides sufficient drive power for the subsequent driver stage. The coupling from the collector circuit of the frequency doubler stage to the driver stage is made in a manner recommended by RCA (1), (2).

These stages have been especially designed for these transistor types. If different types are used, it will be necessary for the circuit to be redimensioned - especially for transistors T 1 and T 2.

The driver stage is equipped with the overlay transistor 40 290 or the planar-epitaxial transistor BFY 44 (Philips). This stage amplifies the 145 MHz signal to approximately 500 mW. Since this stage is also collector modulated, it should be equipped with a transistor type having good modulation characteristics at an operating voltage of 12 V. The driverstage also operates in a common-base configuration.

The matching of the driver to the input impedance of the output stage is made by the transformation link comprising capacitors C 15, C 17 and the inductance L 6. The capacitor C 16 is merely provided as a safety measure, so that the full operating voltage cannot reach the base of the output transistor (where the transistor would be destroyed) if a short circuit should occur in the trimmer capacitor C 15.

The output transistor T 5 (40 290) operates in a common-emitter configuration and possesses a similar output circuit to the frequency doubler stage (T 3). As is also the case with the previous stages, a pass inductance (L 7) in the matching circuit suppresses unwanted harmonics and spurious signals. Since this circuit configuration also reduces the current amplitude of the harmonic contents, the collector power dissipation will be reduced and, especially during class C operation, the efficiency increased (1).

The three supply voltages at connection points Pt 1, Pt 2 and Pt 3 are bypassed by large, low-inductive capacitors so that no coupling effects can occur via the power supply connections (C 23, C 14, C 19).

The crystals are switched by a multiposition switch with relatively short connection leads. The crystals are mounted directly around the switch for the same reason.

2.2. MODULATION

The modulation transformer is built-up in the form of an economy transformer so that thickish wire could be used on a relatively small core. This is to ensure that the lowest possible voltage drop occurs across the transformer. The driver is modulated with approximately 40% of the AF voltage. The electrolytic capacitor C 22actually belongs to the modulator.

Any ferrous-free AF amplifier can be used as modulator that provides an AF output power of 2 W into 5 Ω. This means that the complementary transistor pair AC 117/AC 175 can be used (4).

The 33 V zener diode D 1 shorts excessive voltage peaks from the modulator transformer to ground. The zener voltage, however, allows 100% modulation up to an operating voltage of 16 V.

2.3. INSTRUMENTATION

A low impedance moving coil meter with a full scale deflection of 0.6 A to 1 A (or a suitably shunted μA meter) can be used to monitor operation of the transmitter. The meter can either indicate the total current or the collector current of the output transistor. A 100 μA meter together with a dropper resistor and diode can be connected via a capacitor of approximately 0.5 pF to the transmitter output to monitor the RF output signal. It should, however, be noted that all leads should be kept as short as possible so that no RF injection occurs, since a phase-shifted injection could falsify the reading so that the RF voltage maximum at the antenna connector does not correspond to maximum meter reading.

2.4. POWER SUPPLY

The power supply during portable operation is formed by nine 1.5 V (U 2) batteries, which are series connected to provide an operating voltage of 13.5 V. These batteries have been found to be sufficient to cover an active contest without exchanging batteries. Flat batteries cannot be used since they are not able to handle the high current requirements of the transmitter.

For fixed operation, the transmitter can be operated from a stabilized power supply in conjunction with a simple electronic fuse (4). Such a fuse is especially recommended during alignment and preliminary operation of the transmitter to avoid destruction of the power transistor due, for instance, to self-oscillation. Conventional fuses are too inert for this purpose.

3. MECHANICAL ASSEMBLY

Most of the transistor transmitter is accommodated on the epoxy printed circuit board DJ 1 NB 004. This PC-board possesses the dimensions 215 mm x 45 mm. Fig. 2 shows the conductor side and Fig. 3 the component side of this board. Space for one quartz crystal is provided. If more crystals are to be used, these are connected to the change-over switch and fed via two (short) leads to the PC-board. Transistors T 3 and T 4 are depressed onto the PC-board when soldering so that the connection leads are as short as possible. Transistor T 4 must be provided with a cooling fin.

Trimmer C 15 and the subsequent components given in the circuit diagram are mounted either on the conductor side or on the brass plate B of the output transistor. Figures 4a and 4b show the build-up of the output stage. The two brass plates A and B are made according to Figures 5 and 6. Whereas plate A only completes the screening of the output circuit and is used to mount the transmitter in a cabinet, plate B is provided for mounting the output stage components. Both brass plates are screwed to the end of the PC-board as shown in Fig. 4a. After this, part B is soldered to the ground area of the PC-board. Trimmer capacitors C 17 and C 21 are now soldered into the holes given in Fig. 6 and the ceramic support is screwed into place. Trimmer capacitor C 15 is inserted from the conductor side of the PC-board and soldered into place. This is followed by soldering C 16, L 6 and the bypass capacitor C 18. The author used a mica capacitor for C 18 which could be soldered to the chassis. However ceramic capacitors with ribbon connections could also be used.

It is now possible to solder trimmer capacitor C 20 to the ceramic support and the outer connection of C 21. This is followed by soldering L 7 to bypass capacitor C 18 and L 8 to the ceramic support. The collector circuit connections are joined together and soldered in the last process to the collector lead of the output transistor T 5. The collector connection must be previously shortened to 8 mm, covered with insulating tubing and placed through the cut-out shown in Fig. 6.

The emitter lead of transistor T 5 is shortened to a length of 6 mm and is soldered to the outer surface of brass plate B. The base lead is connected to the bent-up end of inductance L 6. Choke Ch 6 is also connected to this point and the other end connected to resistor R 9, which is in turn connected to brass plate B. R 9 has a rating of 0.5 W so that the connection leads are sufficient for self-support.

A large TO 5 cooling fin (30 mm outer diameter) will be sufficient to cool the output transistor for short transmission periods at 12 V. For longer transmissions and higher operating voltages, it will be necessary to use a heatsink as given in Fig. 7. It is made from aluminium plate or similar material. The dimensions are approximate values; however, the heatsink dimensions should not be reduced too far. It is important that the transistor (before soldering) is pressed into the heatsink so that a good thermal contact is guaranteed. After wiring, the heatsink is fixed to brass plate B using a trolitul strip and a two-component adhesive.

The soldering of choke Ch 7, capacitor C 19 and zener diode D 1 represents the end of the assembly. The ground sides of the above mentioned components are soldered to brass plate A. The modulation transformer and the resistor R 10 are accommodated and connected externally. After soldering the coaxial cable or output connector and the meter, the transmitter will be ready for alignment.

3.1. COMPONENT LIST

T 1, T 2 : 2 N 708 (BSY 19, AEG-Telefunken)
T 3 : 2 N 3553, RCA
T 4, T 5 : 40 290, RCA

D 1 : BZY 92/33 (33 V/ 1 W zenerdiode)
D 2 : AA 112, OA 90 etc. (Germanium point-contact diode)

R 1 : 1.5 kΩ	R 4 : 47 Ω	R 9 : 68 Ω
R 2 : 1.5 kΩ	R 5, R 6, R 7 : 33 Ω	R 10 : 22 Ω
R 3 : 270 Ω	R 8 : 12 Ω	R 11 : 100 kΩ

R 12 : according to meter. Reading to scale centre at, i.e. 12 V operating voltage

C 1, C 3, C 4, C 7, C 10, C 13, C 16, C 24 : 1 nF ceramic disc trimmers
C 2, C 9 : 10 - 60 pF ceramic disc trimmer (10 S-Triko 06/10 - 60 pF)
C 5, C 12 : 6 - 30 pF ceramic disc trimmer (7 S-Triko 02/6 - 30 pF)
C 6 : 27 pF tubular ceramic capacitor
C 8, C 11, C 15, C 17, C 20, C 21 : 3 - 30 pF air spaced trimmer C 005 CA/ 30 E or WN 40 163 Philips

C 14, C 19, C 23 : approx. 0.025 - 0.1 μ F plastic-foil capacitors
e.g. 0.1 μ F/50 V (type MKL, Siemens)

C 18 : 1 nF mica capacitor or ceramic capacitor with ribbon connections

C 22 : 1 mF (1000 μ F), 15/18 V electrolytic capacitor

C 23 : 0.3 to 1 pF

3.2. COIL DATA

L 1 6 turns of 1 mm dia (18 AWG) silver-plated copper wire wound on a 8 mm former, self-supporting. Coil tap two turns from ground end. Coil length 16 mm.

L 2 4 turns with centre tap. Coil length 10 mm, otherwise as L 1.

L 3 5 turns. Coil length 12 mm, otherwise as L 1.

L 4 3 turns wound on 5 mm former. Coil length 15 mm, otherwise as L 1

L 5 2 turns. Coil length 6 mm, otherwise as L 1

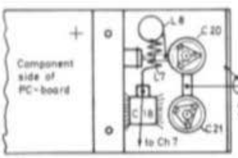
L 6 2 turns wound on 7 mm former. Coil length 5 mm, otherwise as L1

L 7 3 turns wound on 5 mm former. Coil length 15 mm, otherwise as L 1

L 8 3.75 turns. Coil length 13 mm, otherwise as L 1

Tr 1 E I 48 core with air gap (0.3 mm insulating material between "E" and "I" portion) continuous winding. 400 turns of 0.4 mm dia (26 AWG) enamelled copper wire. Tapping point 180 turns.

Ch 1 to Ch 7 ferrox cube wideband chokes VK 200 (Philips)



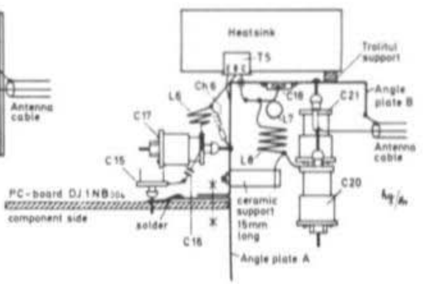
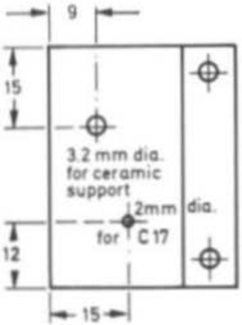


Fig. 4 Output stage, collector circuit

4. ALIGNMENT

Reduce the capacity of C 8 to minimum and tightly couple a 6 V/0.6 W lamp with a 3 turn link to inductance L 2. An operating voltage of 12 V is fed via a mA meter to connection Pt 1. A two metre receiver with shorted (terminated is better) input is tuned to the transmit frequency. If the oscillator does not already oscillate, a current of approximately 25 mA should be indicated.

The crystal oscillator will commence oscillation on adjusting C 2. By alternately tuning capacitor C 5 and altering the coupling of the 6 V lamp, the point of maximum brightness should be found (slight glow). The oscillator is then tuned from maximum point towards the flatter slope of the collector current. The signal should now be stable and no spurious signals present (tune receiver). The current should read approx. 40 mA in this condition.



Material: Brass plate 0.5 mm thick
All dimensions in mm

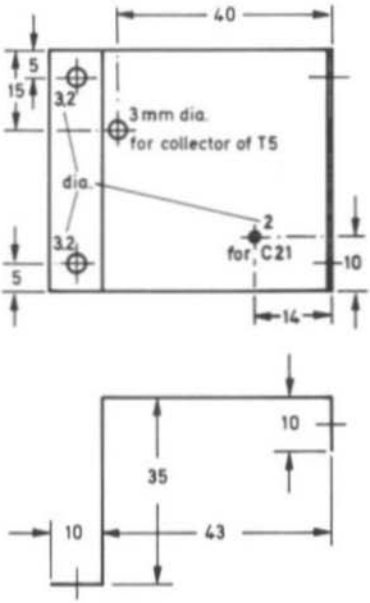
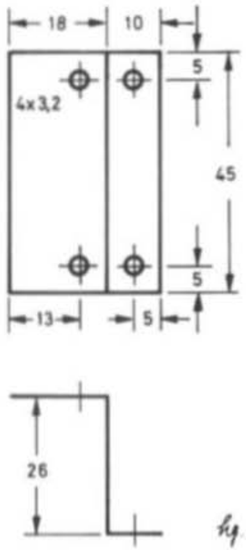
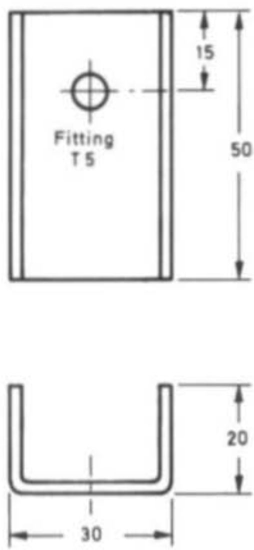


Fig.6 Angle plate B



Material: Brass plate 0,5mm thick
All dimensions in mm

Fig.5 Angle plate A



All dimensions in mm
Material: Aluminium, 2 mm thick

Fig. 7 Heatsink for T5

A lamp is now tightly coupled with only two turns to inductance L 5 for alignment of the frequency doubler stage. The capacitance of trimmer capacitor C 8 is now increased in steps, aligning firstly C 9 and then C 5 for maximum brightness, until the most favourable position is found. If no clear maximum is obtainable with C 5, this will indicate that the capacitance of C 8 is too great. The tuning of capacitor C 9 is not critical and it may be replaced by a fixed capacitor after the final alignment (with modulation) has been completed. Capacitors C 11 and C 12 are also aligned for maximum brightness.

After this, connection point 2 is connected via a mA meter and the electronic fuse to the operating voltage. It is advisable to set the electronic fuse to a threshold of approximately 200 mA. Choke Ch 6 and the base connection of transistor T 5 are removed during alignment of the driver stage and the lamp is connected between the free end of L 6 and ground. After rough alignment of the trimmer capacitors C 11, C 12, C 15 and C 17 it is necessary to operate the output stage because the final alignment can only be made with the whole transmitter operative.

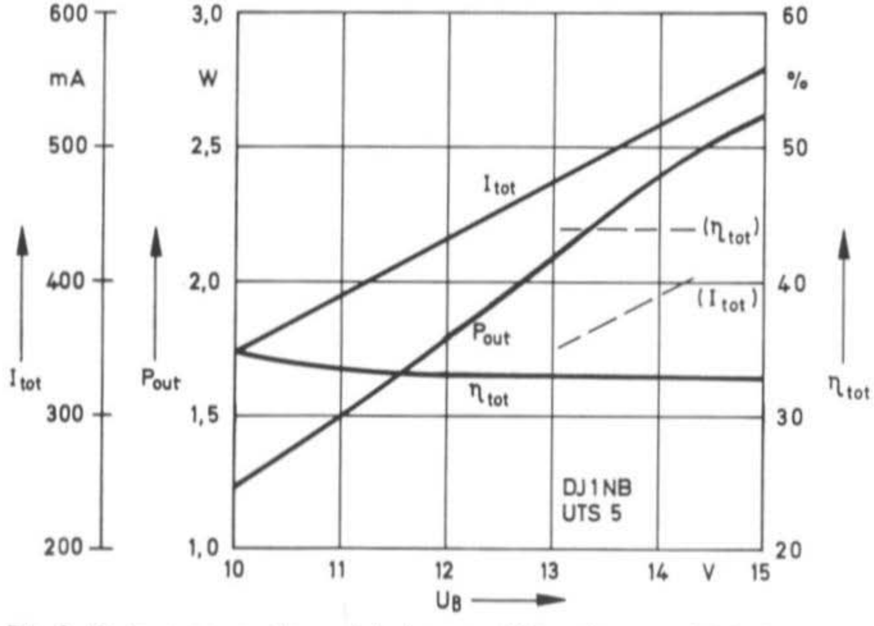


Fig. 8 Output power P_{out}, total current flow I_{tot} and total efficiency η_{tot} of the transmitter UTS 5

A 6 V/1.8 W lamp is now soldered directly to the transmitter output. It is advisable that each of the three operating voltage connections should be provided with a mA meter. The total current should be fed via the electronic fuse; the threshold current should be 500 mA.

A certain degree of patience is required for the following peak-power alignment because several components must be adjusted in order to find the most favourable combination of their values. It will be found that the alignment of trimmers C 8, C 11, C 15 and C 17 is critical, whereas especially the output trimmers C 20 and C 21 can be varied greatly without hearing noise points or spurious signals in the receiver.

If the transmitter is operating correctly, alignment of trimmers C 2, C 5 and C 9 will no longer indicate any clear maximum adjustment since a saturation effect occurs. This is provided as drive reserve for operation at low temperatures, for instance, during mountain field days and to obtain a certain wide-band capability. The final alignment of the first three stages should therefore be carried out separately and not subsequently altered.

Finally the transmitter is fed with modulation. The modulation depth is kept low at first. Trimmers C 11, (C 12 may have been replaced with a fixed capacitor) C 15, C 17, C 20 and C 21 are carefully aligned for maximum brightness. The modulation is increased in steps and the trimmers carefully realigned for maximum. The modulation should be switched off several times during the alignment to determine that the brightness of the lamp increases during modulation.

It is important that the signal is monitored during this process. A sinusoidal (sine wave) modulation is advisable since this waveform allows distortion products to be heard more easily. The most favourable means of adjusting the modulation is with the aid of a DC oscilloscope, which allows the modulation depth and distortions to be simultaneously observed.

Since the lamp used for the alignment does not represent a real 60 Ω (or 50 Ω) impedance, it is necessary to repeat the modulation alignment using an antenna (dummy load). If a DC oscilloscope with absorption circuit is not available, a reliable test can be made as follows: Tune a selective receiver to a frequency approximately 10 to 25 kHz from the transmit frequency and check that no sidebands are audible when aligning the sine-wave modulated transmitter for maximum output. Trimmers C 20 and C 21 have the greatest influence; however, C 15 and C 17 may also require realignment. Although the alignment requires a certain patience and feeling, the transmitter has been found to be very stable after completion of the alignment. As can be seen from the following measured values, the transmitter is able to cope with large voltage variations; it is sufficiently wideband and is not endangered if one forgets to connect the antenna.

5. MEASURED VALUES

The described transmitter was measured in an industrial laboratory by R. Lentz, DL 3 WR. The determined values are given in Fig. 8. The transmitter had been previously aligned for 100% modulation. The linear increase of the output power with the operating voltage shows that the higher power levels are achieved without feedback effects. The capacity of the dry cells can be well utilized, since the transmitter can (if correctly aligned) be modulated correctly at an operating voltage of down to 10 V. This is also where the high power level of the first three stages is advantageous. An operating voltage of 15 V should not be exceeded since this would endanger the driver and output transistors.

The characteristics of another transistor of the 40 290 series is drawn as a dashed line in Fig. 8 - it was destroyed by a voltage peak before installing diode D 1. This transistor produced the same output power at a lower current drain, which led to an overall efficiency of 44%. This example clearly shows the fluctuations between individual transistors of the 40 290 series. The transmitter having the measured values shown in Fig. 8 has proved itself in more than two years of operation and three mountain field days.

6. AVAILABLE PARTS

The printed circuit board DJ 1 NB 004, trimmer capacitors and other components as well as a complete kit of parts are available from the publishers or their national representatives (see advertising page).

7. REFERENCES

(1) RF POWER TRANSISTORS;
RCA Technical Presentation 4/68 ST-3701

(2) F. Münzel: Erfahrungen mit "Overlay"-Transistoren,
Das DL-QTC 39 (1968), Edition 5, pages 280-285

(3) Data Sheet of the overlay transistor types 40 290, 40 291, 40 292

(4) R. Lentz: A Simple Electronic Fuse
VHF COMMUNICATIONS 1 (1969), Edition 3, pages 174-178

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)		
Crystal	46.0000 MHz (HC-18/U) (DJ 7 ZV / DJ 9 ZR)) set	DM	49.--
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)		

DL 6 HA KITS TWO METRE SSB TRANSCEIVER

DL 6 HA 001/28	2 m MOSFET Converter (IF = 28-30 MHz)	Ed. 1/70	
PC-board	DL 6 HA 001 (with printed plan)	DM	6.--
Minikit	DL 6 HA 001 (coil formers and trimmer set)	DM	5.20
Semiconductors	DL 6 HA 001 (5 transistors)	DM	20.50
Crystal	38.66667 MHz (HC-6/U)	DM	13.70
Kit	DL 6 HA 001 with above listed components	DM	45.40

DL 6 HA 001/14	2 m MOSFET converter (IF = 14-15 MHz)	Ed. 1/70	
PC-board	DL 6 HA 001 (with printed plan)	DM	6.--
PC-board	DL 3 YK 001 (with printed plan) (Ed. 4/70)	DM	4.--
Minikit	DL 6 HA 001/14 (coil formers and trimmer set, 2 crystal holders)	DM	11.30
Crystal	65.00000 MHz (HC-6/U)	DM	16.50
Semiconductors	DL 6 HA 001/14 (7 transistors)	DM	24.50
Crystal	65.50000 MHz (HC-6/U)	DM	16.50
Kit	DL 6 HA 001/14 with above listed components	DM	78.80

DL 6 HA 002	9 MHz SSB Transceiver	Ed. 2/70	
PC-board	DL 6 HA 002 (with printed plan)	DM	16.--
Minikit	DL 6 HA 002 (coil formers and trimmer set)	DM	6.--
Semiconductors	DL 6 HA 002 (12 transistors, 7 diodes)	DM	46.80
Kit	DL 6 HA 002 with above listed components	DM	68.80
Crystal filter	XF-9A with both sideband crystals	DM	106.--
Crystal filter	XF-9B with both sideband crystals	DM	137.--

DL 6 HA 003	9 MHz Carrier Oscillator	Ed. 2/70	
PC-board	DL 6 HA 003 (with printed plan)	DM	3.--
Minikit	DL 6 HA 003 (coil formers and trimmer set, 2 transistors, 2 diodes)	DM	7.50
Kit	DL 6 HA 003 with above listed parts	DM	10.50

DL 6 HA 004	9/14 MHz Transmit-Receive Converter	Ed. 3/70
PC-board	DL 6 HA 004 (with printed plan)	DM 16.--
Minikit	DL 6 HA 004 (coil formers, trimmer and choke set)	DM 8.70
Semiconductors	DL 6 HA 004 (8 transistors, 5 diodes)	DM 49.10
Kit	DL 6 HA 004 with above listed components	DM 73.80

DL 6 HA 005	14/145 MHz Transmit Converter	Ed. 3/70
PC-board	DL 6 HA 005 (with printed plan)	DM 8.--
Minikit	DL 6 HA 005 (coil formers, trimmer and choke set)	DM 4.80
Semiconductors	DL 6 HA 005 (4 transistors)	DM 18.--
Kit	DL 6 HA 005 with above listed components	DM 30.80

DL 6 HA 006	V F O (5-6 MHz)	Ed. 3/70
PC-board	DL 6 HA 006 (with printed plan)	DM 8.--
Minikit	DL 6 HA 006 (2 transistors, 1 ceramic coilformer) ..	DM 6.50
Variable capacitor	100 pF (48.5 x 48.5 x 49.5 mm, two bearings)	DM 13.50 +
Kit	DL 6 HA 006 with above listed components	DM 28.--

DL 6 HA 007	Low-pass Filter	Ed. 3/70
PC-board	DL 6 HA 007 (with printed plan)	DM 3.--
Minikit	DL 6 HA 007 (coil former set, 1 transistor)	DM 3.20
Kit	DL 6 HA 007 with above listed components	DM 6.20

DL 6 HA 008	Stabilized Power Supply	Ed. 4/70
PC-board	DL 6 HA 008 (with printed plan)	DM 8.--
Minikit	DL 6 HA 008 (choke set, 5 transistors, 6 diodes) ...	DM 39.70
Kit	DL 6 HA 008 with above listed components	DM 47.70

DL 6 HA 009	Transformerless AF Amplifier (approx. 2.5 W)	Ed. 4/70
PC-board	DL 6 HA 009 (with printed plan)	DM 8.--
Semiconductors	DL 6 HA 009 (4 transistors, 1 diode)	DM 14.50
Kit	DL 6 HA 009 with above listed components	DM 22.50

All Kits	DL 6 HA 001/14 to 009 (without crystal filter)......	DM 350.--
All PC-boards	DL 6 HA 001 to 009 and DL 3 YK 001	DM 70.--

DJ 9 ZR KITS — TWO METRE SSB TRANSCEIVER

DJ 9 ZR 001	5 W SSB Transmitter for 145 MHz	Ed. 2/69
PC-board	DJ 9 ZR 001 (with printed plan)	DM 15.--
Minikit	DJ 9 ZR 001 (coil formers, trimmer and choke set)..	DM 7.50
Semiconductors	DJ 9 ZR 001 (9 transistors, 3 diodes)	DM 92.30
Kit	DJ 9 ZR 001 with above listed components	DM 114.80
Potted core kit	DJ 9 ZR 001 (for DC-DC converter)	DM 15.--
Crystal filter	XF-9A with both sideband crystals	DM 106.--
Crystal filter	XF-9B with both sideband crystals	DM 137.--

DJ 9 ZR 002	V X O (136.446 MHz $\pm$ 20 kHz)	Ed. 2/69
PC-board	DJ 9 ZR 002 (with printed plan)	DM 3.--
Minikit	DJ 9 ZR 002 (coil former set, 3 transistors, 2 diodes)	DM 13.40
Crystal	45,478 MHz (HC-18/U)	DM 24.50
Kit	DJ 9 ZR 002 with above listed components	DM 40.90

DJ 9 ZR 005	IF-AF Portion (9 MHz)	Ed. 3/69
PC-board	DJ 9 ZR 005 (double coated with through contacts) ...	DM 42.--
Minikit	DJ 9 ZR 005 (coil formers, transformer and choke set)	DM 43.20
Semiconductors	DJ 9 ZR 005 (5 ICs, 2 transistors, 8 diodes)	DM 74.10
Kit	DJ 9 ZR 005 with above listed components	DM 159.30
Crystal filter	XF-9B with both sideband crystals	DM 137.--
Precision spindle	potentiometer 500 Ohm	DM 9.--

DJ 9 ZR 006	VHF Portion (145/9 MHz)	Ed. 3/69
PC-board	DJ 9 ZR 006 (teflon (PTFE) with printed inductances).	DM 43.50
Minikit	DJ 9 ZR 006 (coil formers and trimmer set)	DM 10.80
Semiconductors	DJ 9 ZR 006 (3 transistors)	DM 31.50
Kit	DJ 9 ZR 006 with above listed components	DM 85.80

DJ 9 ZR 008	Electronically Stabilized Power Supply	Ed. 3/70
PC-board	DJ 9 ZR 008 (with printed plan)	DM 14.--
Semiconductors	DJ 9 ZR 008 (5 transistors, 14 diodes, heatsink)....	DM 58.30
Potted core kit	DJ 9 ZR 008 (for DC-DC converter).............	DM 12.50
Kit	DJ 9 ZR 008 with above listed components	DM 84.80

CONVERTERS and TRANSVERTERS (See also DL 6 HA and DJ 9 ZR kits)

DL 6 SW 004/145	Two Metre FET Converter	Ed. 1/69
PC-board	DL 6 SW 004 (with printed plan)	DM 6.--
Minikit	DL 6 SW 004 (coil formers and trimmer set)	DM 5.20
Semiconductors	DL 6 SW 004 (5 transistors)	DM 19.--
Crystal	38.66667 (HC-6/U)	DM 13.70
Kit	DL 6 SW 004 with above listed components	DM 43.90

DL 6 SW 004/70	FET Converter for 70 MHz	Ed. 2/69
PC-board	DL 6 SW 004 (with printed plan)	DM 6.--
Minikit	DL 6 SW 004 (coil formers and trimmer set)	DM 5.20
Semiconductors	DL 6 SW 004 (5 transistors)	DM 19.--
Crystal	42.00000 MHz (HC-6/U)	DM 15.60
Kit	DL 6 SW 004/70 with above listed components......	DM 45.80

DJ 6 ZZ 001	28 MHz - 145 MHz Transverter with FET Mixers	Ed. 4/69
PC-board	DJ 6 ZZ 001 (with printed plan)	DM 15.--
Minikit	DJ 6 ZZ 001 (coil formers and trimmer set)	DM 8.40
Semiconductors	DJ 6 ZZ 001 (12 transistors)	DM 39.--
Crystal	38.66667 (HC-6/U)	DM 13.70
Kit	DJ 6 ZZ 001 with above listed components	DM 76.10

DL 9 GU 001	70 cm Converter (IF = 145 MHz)	Ed. 2/69
PC-board	DL 9 GU 001 (with printed plan)	DM 6.--
Minikit	DL 9 GU 001 (coil formers and trimmer set, 2 disc and 3 tubular trimmers)	DM 8.10
Semiconductors	DL 9 GU (5 transistors)	DM 35.--
Crystal	96.000 MHz (HC-6/U)	DM 21.50
Casing	DL 9 GU 001 silver-plated with inner conductors and trimmers (delivery max. 6-8 weeks)	DM 65.--
Kit	DL 9 GU 001 with above listed components (but without tubular trimmers)	DM 129.60

DC 6 HY 001	Receive Converter 432/144 MHz)	Ed. 4/70
PC-board	DC 6 HY 001 (with printed plan)	DM 10.--
Minikit	DC 6 HY 001 (coil formers and trimmer set)	DM 13.20
Semiconductors	DC 6 HY 001 (5 transistors)	DM 20.30
Crystal	95.833 MHz (HC-18/U)	DM 28.--
Kit	DC 6 HY 001 with above listed components	DM 71.50
Ready-to-operate	receive converter DC 6 HY 001 (slight delay)	DM 179.60

DC 6 HY 002	Transmit Converter 144/432 MHz	Ed. 4/70
PC-board	DC 6 HY 002 (with printed plan)	DM 10.--
Minikit	DC 6 HY 002 (trimmer set and ferrite beads)	DM 21.50
Semiconductors	DC 6 HY 002 (3 transistors)	DM 18.50
Kit	DC 6 HY 002 with above listed components	DM 50.--
Ready-to-operate	transmit converter DC 6 HY 002 (slight delay)......	DM 149.50

DC 6 HY	432 MHz Linear Amplifier	Ed. 4/70
Minikit	DC 6 HY/Lin (trimmer set and tube socket)	DM 10.90
Semiconductors	DC 6 HY/Lin (1 transistor, 1 diode)	DM 4.20
Tube	EC 8020	DM 27.--
Kit	DC 6 HY/Lin with above listed components	DM 42.10
Ready-to-operate	linear amplifier DC 6 HY (slight delay)	DM 142.40

TRANSMITTERS, VFOs and OSCILLATORS
(see also DL 6 HA- and DJ 9 ZR Kits as well as DJ 7 ZV, DC 6 HY

DJ 1 NB 004	2 W AM Transmitter for 145 MHz	Ed. 3/69
PC-board	DJ 1 NB 004 (with printed plan)	DM 10.--
Minikit	DJ 1 NB 004 (trimmer and choke set)	DM 12.60
Semiconductors	DJ 1 NB 004 (5 transistors, 2 diodes)	DM 56.50
Kit	DJ 1 NB 004 with above listed components	DM 79.10
Crystal	72,.... MHz (HC-6/U) on request, please state required frequency (delivery approx. 6 weeks)	DM 21.50
Crystals available ex. stock	72.0500 MHz / 72.0750 MHz / 72.1000 MHz / 72.1250 MHz 72.1500 MHz / 72.2000 MHz / 72.2500 MHz / 72.3000 MHz 72.3500 MHz / 72.4000 MHz / 72.4500 MHz / 72.5000 MHz when ordering please indicate frequency	DM 21.50

Modulation transformer kit: see DL 3 WR 003 transmitter

DL 3 WR 003	2 W Transmitter for 145 MHz for VFO Operation	Ed. 2/70
PC-board	DL 3 WR 003 (with printed plan)	DM 17.50
Minikit	DL 3 WR 003 (trimmer and choke set)	DM 14.40
Semiconductors	DL 3 WR 003 (11 transistors, 5 diodes)	DM 93.10
Modulation transformer kit	DL 3 WR 003 L for PC-board mounting	DM 7.50
Kit	DL 3 WR 003 with above listed components	DM 132.50
Crystals	72,.... MHz see DJ 1 NB 004	DM 21.50

DL 3 WR 007	24 MHz Synthesis VFO with FM Attachment	Ed. 3+4/70
PC-board	DL 3 WR 007 (with printed plan)	DM 5.--
Minikit	DL 3 WR 007 (4 trimmers and 1 variable ceramic coil 1.8 - 2.5 µH)	DM 25.--
Semiconductors	DL 3 WR 007 (5 transistors, 2 diodes)	DM 30.50
Variable capacitor	100 pF (48,5 x 48,5 x 49,5 mm) two bearings	DM 13.50
Potted core kits	DL 3 WR 007 (4 potted cores)	DM 16.40
Crystal	27.800 MHz (HC-18/U)	DM 25.--
Kit	DL 3 WR 007 with above listed components	DM 115.40

DJ 8 PG 001	72 MHz VFO for FM Transmitters	Ed. 4/70
PC-board	DJ 8 PG 001	DM 2.50
Minikit	DJ 8 PG 001 (coil formers, trimmer and choke set)..	DM 3.85
Semiconductors	DJ 8 PG 001 (2 transistors, 1 diode)	DM 13.--
Kit	DJ 8 PG 001 with above listed components	DM 19.35

DL 3 YK 001	Oscillator Board for DL 6 HA 001	Ed. 4/70
PC-board	DL 3 YK 001 (with printed plan)	DM 4.--
Minikit	DL 3 YK 001 (coil former set, 2 crystal holders) ...	DM 6.10
Semiconductors	DL 3 YK 001 (2 transistors)	DM 4.--
Crystal	65.000 MHz (HC-6/U)	DM 16.50
Crystal	65.500 MHz (HC-6/U)	DM 16.50
Kit	DL 3 YK 001 with above listed components	DM 47.10

X Kit included in DL 6 HA 001/14. Serves to modify the 28-30 MHz IF to 14-15 MHz.

MEASURING, AUXILIARY EQUIPMENT and ACCESSORIES

DL 3 WR 004	RF-Voltage Indicator	Ed. 2/70
PC-board	DL 3 WR 004 (with printed plan)	DM 2.50
Kit	DL 3 WR 004 (PC-board, trimmer and diode)	DM 4.25
DL 9 FX 001	Simple Band Scanner	Ed. 4/70
PC-board	DL 9 FX 001 (with printed plan)	DM 4.--
Minikit	DL 9 FX 001 (coil formers and trimmer set, 5 transistors, 2 diodes)	DM 18.80
Kit	DL 9 FX 001 with above listed components	DM 22.80
DC 6 HY 003	Calibration Spectrum Generatur	Ed. 4/70
PC-board	DC 6 HY 003 (with printed plan)	DM 4.--
Minikit	DC 6 HY 003 (1 trimmer, 5 transistors)	DM 16.50
(Standard frequency	crystal 1.000 MHz (HC-6/U)	DM 20.50
(Kit	DC 6 HY 003/1 MHz with above listed components ...	DM 41.20
(Standard frequency	crystal 100 kHz (HC-13/U)	DM 28.--
(Kit	DC 6HY 003/100 kHz as above	DM 48.70
(Standard frequency	crystal 5 MHz (HC-6/U)	DM 25.--
(Kit	DL 6 HY 003/ 5 MHz as above	DM 45.70
DJ 4 BG 002	Digital Discriminator	Ed. 2/70
PC-board	DJ 4 BG 002 (with printed plan)	DM 5.--
Semiconductors	DJ 4 BG 002 (1 IC, 4 transistors, 2 diodes)	DM 17.50
Kit	DJ 4 BG 002 (with above listed components)	DM 22.50
DJ 4 BG 006	Speech Processor	Ed. 1/71
PC-board	DJ 4 BG 006 (with printed plan)	DM 7.--
Semiconductors	DJ 4 BG 006 (1 IC, 2 transistors)	DM 17.80
Kit	DJ 4 BG 006 with above listed components	DM 24.80
Connectors	Siemens (13-pole) for PC-board DJ 4 BG 006, set ...	DM 7.40
DL 3 WR 002	Electronic Fuse 24 V	Ed. 3/69
PC-board	DL 3 WR 002 (with printed plan)	DM 1.50
Semiconductors	DL 3 WR 002 (3 transistors, 1 diode)	DM 11.20
Kit	DL 3 WR 002 with above listed components	DM 12.70

TERMS OF DELIVERY AND BANK ACCOUNTS

All prices are given in West German Marks. The prices do not contain post and packing for which an extra charge will be made: .... DM 2.--

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

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

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

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

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

VERLAG UKW-BERICHTE, Hans J. Dohlus, DJ 3 QC, D-8520 ERLANGEN, Gleiwitzer Str. 45
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