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VHF COMMUNICATIONS

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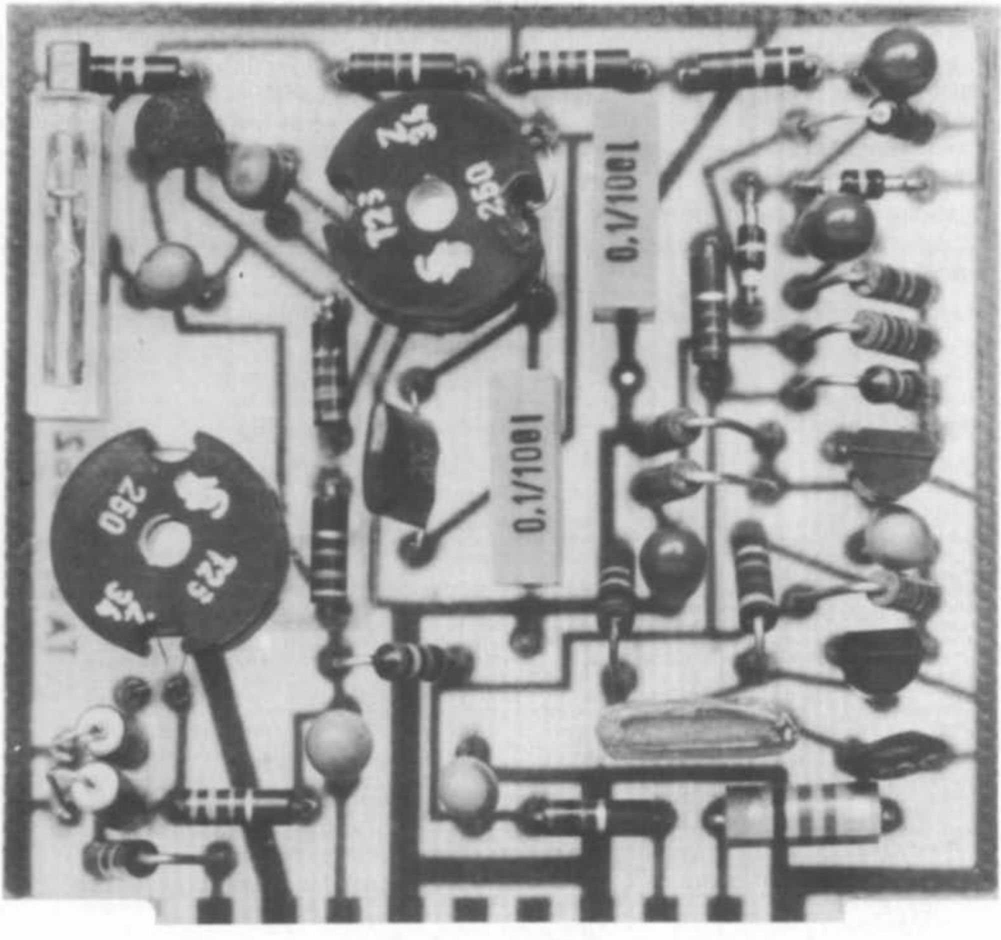
ESPECIALLY COVERING VHF, UHF AND MICROWAVES

VOLUME NO. 3 EDITION 4 NOVEMBER 1971 DM 4.00



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Unfortunately we have been forced to increase the subscription price of VHF COMMUNICATIONS to DM 14.00 due to considerable increases of postal rates and printing costs. This is the first increase that we have made from the original DM 12.00.

Volume 4 (1972) will be full of interesting articles such as: An SSB transceiver for 2 m with FM-attachment. 200 kHz Droitwich receiver for phase-locking 1 MHz crystal oscillators. A 23 cm synthesis transmitter for SSB, AM and FM. A 70 cm PA with the 2 C 39. A 9 MHz IF module with three separate IF strips for AM, SSB/CW and FM. An 18 W transistorized stripline PA for 70 cm. Transistorized linear amplifier for 2 m, and much more.

A DIGITAL CALIBRATION-SPECTRUM GENERATOR

by D. E. Schmitzer, DJ 4 BG

1. INTRODUCTION

Calibration spectrum generators have proved themselves to be a valuable measuring aid. The large number of applications and the reduction of the semiconductor - especially integrated circuit - prices allow extremely versatile equipment to be constructed. For instance, the calibration spectrum generator that is to be described allows a calibration spectrum to be generated whose calibration lines can be switched for a spacing of 1 MHz, 500 kHz, 100 kHz, 50 kHz, 10 kHz, 5 kHz and 1 kHz. In addition to this, a second harmonic generator can be selected which oscillates at a fundamental frequency of 1.001 MHz. Since the frequency spacing of 1 kHz with respect to the main oscillator is also multiplied, a frequency determination is possible by counting the multiplication factor. This method will be explained further in part II.

In addition to this, the frequency of the main oscillator can be pulled slightly using a varactor diode. This allows the calibration spectrum generator to be brought into zero-beat with a standard frequency transmitter without having to open the unit. In part III of this series of articles, a fixed-frequency receiver is to be described for 200 kHz which synchronizes the calibration spectrum generator automatically with the aid of the varactor diode so that the frequency accuracy is the same as that of the Droitwich, England transmitter. The frequency accuracy and stability represented by this is in the order of the time bases of very expensive frequency counters and is especially of interest for the very high frequency bands as well as for EME and meteor-scatter enthusiasts. Figure 1 shows the block diagram of the complete unit.

The described characteristics of the equipment can, according to the requirements and financial expenditure, be obtained step by step. According to the modular principle, a number of modules can be constructed without requiring later modification. Since a construction using discrete components would be too extensive, integrated circuits are used. This results in a simple and clear construction on printed boards which can be included in modular cases as described in (1).

2. CIRCUIT DETAILS OF THE CALIBRATION SPECTRUM GENERATOR

The circuits are built up of integrated digital circuits in the TTL-technology. The operation and application of the basic digital circuits is explained in detail in (2) so that it is not necessary for it to be explained here. For clarity, the description will be made separately for each printed circuit board. The power supply and the 1.001 MHz oscillator will be described in Part II. The description of the 200 kHz Droitwich-receiver and its auxiliary circuit will be described later in Part III.

The following description, Part I, contains the actual calibration spectrum generator.

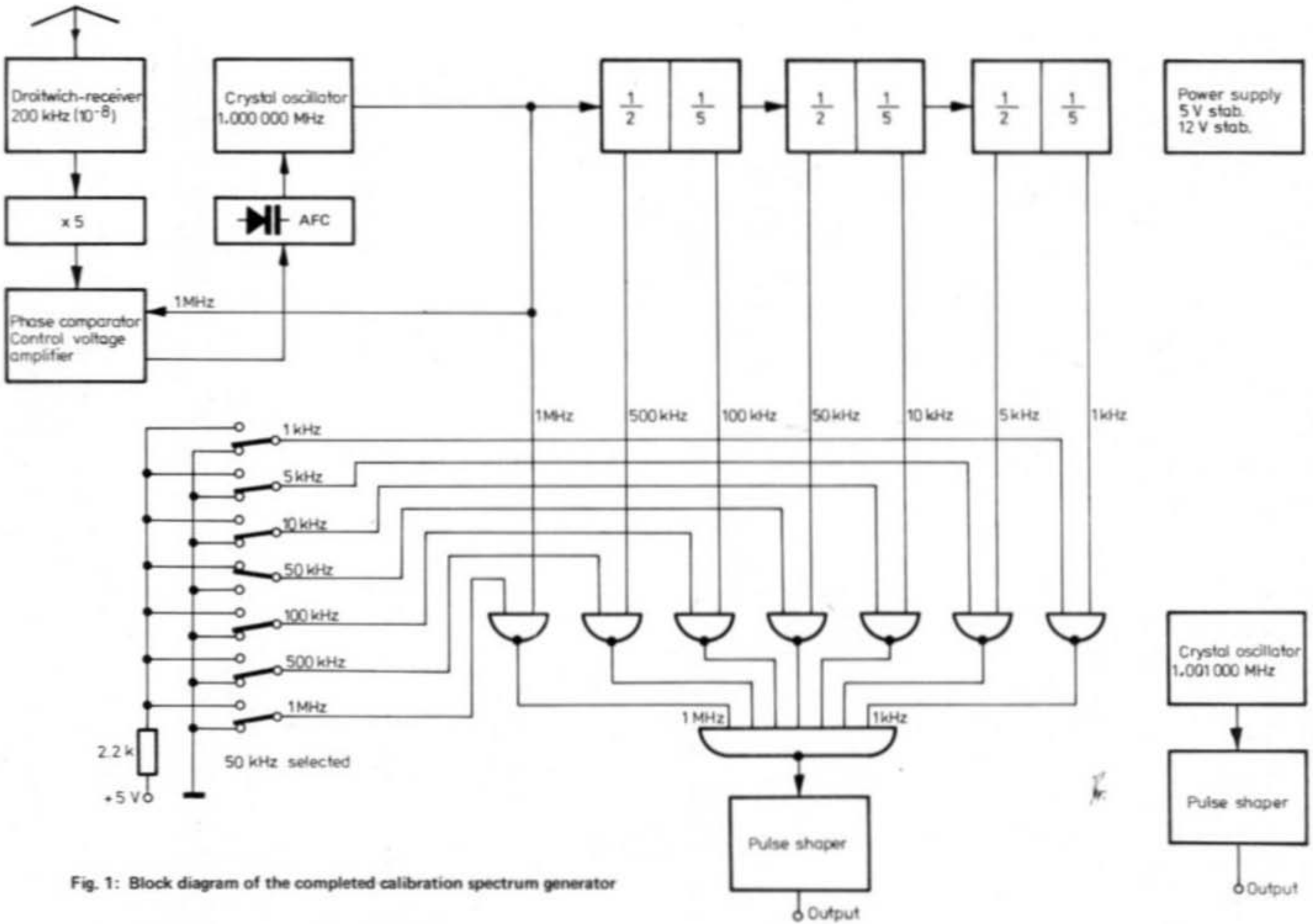


Fig. 1: Block diagram of the completed calibration spectrum generator

2.1. THE OSCILLATOR (First step)

A circuit consisting of two inverters (gates with parallel-connected inputs) is used as oscillator as was described in (3). In this circuit, a feedback link ensures that a DC operating point is adjusted which guarantees a secure commencement of oscillation. In our case, two gates of the integrated circuit SN 7400 N are used and one of the two remaining gates is used as buffer. The last gate will be required on extending the calibration spectrum generator.

The basic circuit of the oscillator is given in Figure 2; the operation roughly corresponds to that of a multivibrator by which a feedback link is provided via the crystal. The same crystal is used as was given in (4) where it was used at parallel resonance with a loading of approximately 30 pF; whereas it is used here at the somewhat lower series resonance which allows it to be aligned to the exact frequency using a series capacitor C_s of approximately 35 pF.

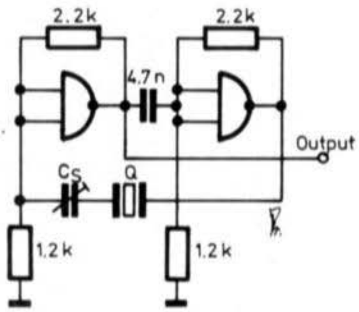


Fig. 2: Crystal oscillator from two NAND - gates ($^1/_2$ SN 7400)

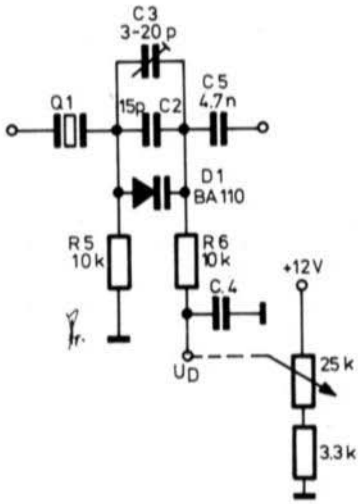


Fig. 4: Circuit extract, fine tuning

After completion, this oscillator is fully ready for operation on the printed circuit board. The output signal is a square wave with a duty cycle of virtually 1 : 1 which is available at connection point Pt 2. The complete oscillator circuit is contained in the circuit diagram (Fig. 3) where it is designated step 1.

It is possible for the frequency to be exactly aligned not only with the series capacitor C_s on the board but also with the aid of a varactor diode D 1 using an adjustable voltage. This allows the calibration process to be carried out without opening the case since the tuning voltage can be varied using a potentiometer on the front panel, see Figure 4.

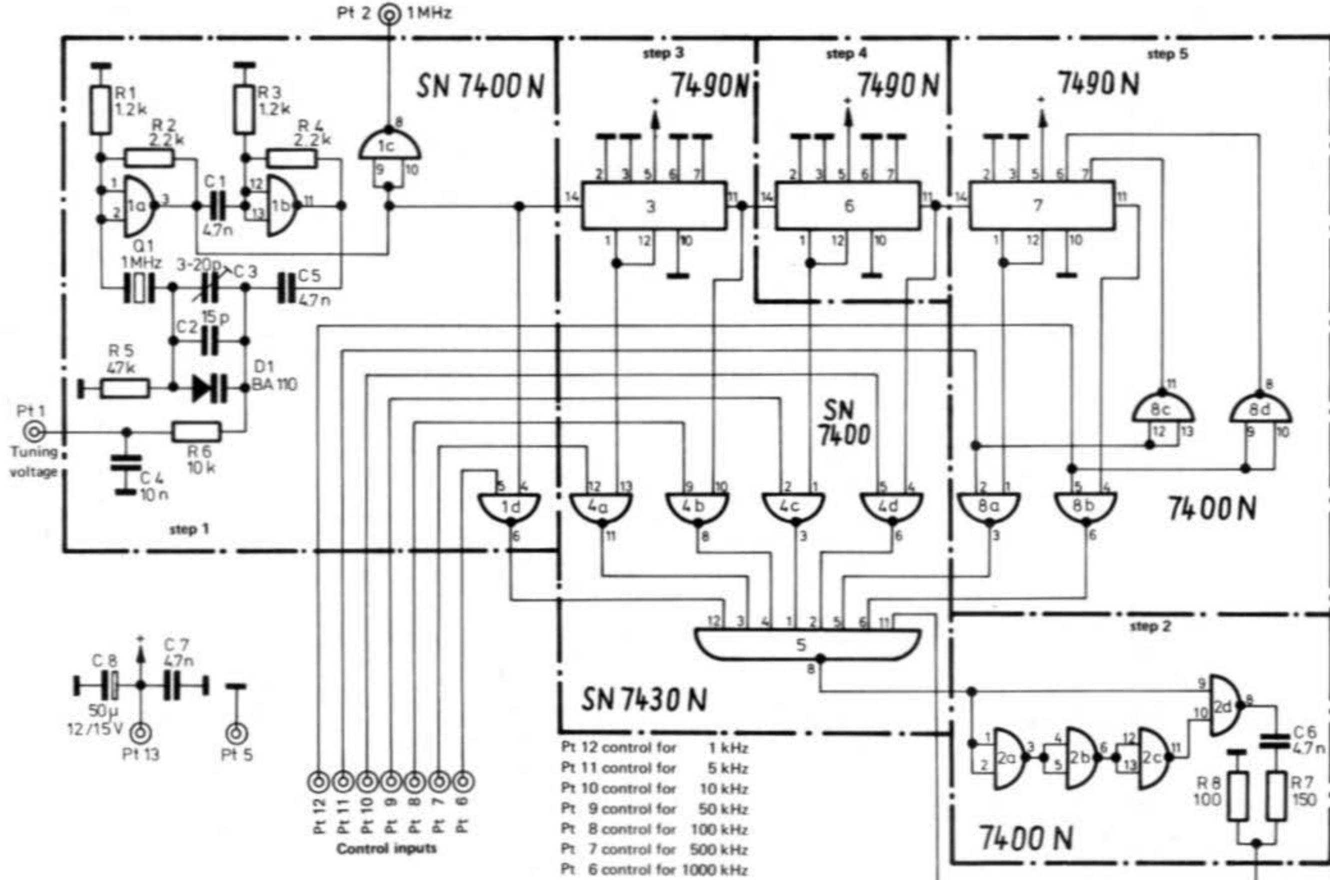


Fig. 3: Circuit of the calibration spectrum generator

2.2. PULSE-SHAPER CIRCUIT (Second step)

The output signal of the oscillator possesses a duty cycle of virtually 1 : 1 which is also true of the frequency-divided signals of the subsequent stages because they originate directly from flip-flops. Only uneven harmonics (e.g. 1f; 3f; 5f; etc.) are contained in a square-wave voltage having a duty cycle of exactly 1 : 1; whereas even harmonics (e.g. 2f; 4f; 6f; etc.) are not present or of a very low level due to a certain unbalance of the circuit. In addition to this, the lower harmonics appear with a far higher energy with such a squarewave signal than the higher order harmonics. This can lead to difficulties in the required application. However, if the squarewave signal is differentiated in a CR-link, all uneven harmonics will be available with approximately the same amplitude according to the short rise time of the needle pulse generated in this manner. If the pulses of negative polarity are removed by a diode after this differentiation and if the positive pulses are passed (or vice versa), the resulting voltage will contain both odd and even harmonics of the fundamental signal of approximately the same energy.

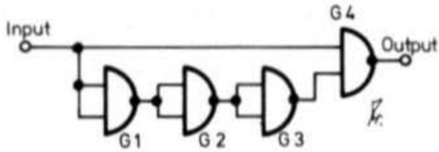


Fig. 5: Pulse shaper from four NAND - gates (1 x SN 7400 N)

Such pulses can be obtained with a circuit comprising four gates as shown in Figure 5. It utilises the inavoidable delay of the signal travelling trough the gates. If the input of this circuit receives a zero-signal, a 1-signal will appear at the output of gate 4, independent of the fact that an 1-signal is present at the second input of gate 4 via the second branch (G 1 to G 3). If the input of the circuit now receives a 1-signal, a 1-signal will exist for a short time at both inputs of gate 4 and the output of G 4 will jump to zero. After the input pulse has also passed via branch G 1 to G 3, zero will be present at the second input of G 4 so that the output jumps back to 1. The 1-0-1-pulse achieved in this manner at the output of G 4 has a pulse length of approximately 30 ns when using the quadruple gate SN 7400 N; the rise time is approximately 8 ns. The extremely steep slope of the pulse means that harmonics of sufficient energy will be available far into the VHF range.

If the crystal oscillator is only to be used with the pulse-shaper stage, the output of the oscillator should be connected to the input of the pulse-shaper using a wire bridge on the printed circuit board. The output signal is then present at connection point Pt 3 at an amplitude of approximately 0.7 V peak-to-peak at an impedance of 60 Ω.

2.3. FREQUENCY DIVIDER FOR 500 kHz and 100 kHz (Third step)

The integrated circuit SN 7490 N comprises frequency dividers of 2 : 1 and 5 : 1. When driven with the 1 MHz signal of the oscillator, output signals of 500 kHz and 100 kHz are obtained without further measures. (Of course, one could connect the 5 : 1 frequency divider first, followed by the 2 : 1 divider and thus obtain 200 kHz and 100 kHz). Three different frequencies are now available which means a switching system is required so that only the required signal is present at the output. This is achieved using NAND-gates. The 1 MHz signal is passed via the remaining gate 1d of IC 1 not required in step 1 (oscillator). The 500 kHz and 100 kHz signals are connected via two gates (4a and 4b) of a further SN 7400 N (IC 4). The two remaining gates remain for the third step. If the second input of the connected gates are now connected to ground, no signal will be present at the output (see the function of the gate as described in (2). If the grounded connection is disconnected by one of the gates or better still, the input connected to + 5 V via a protective resistor, the required signal will be available at the output.

The integrated circuit IC 5 of type SN 7430 N is used for combining the three outputs. This IC comprises a gate with eight inputs and serves as a NOR-circuit. The other inputs are not required at present. On the printed circuit board, the output of the SN 7430 N (IC 5) is connected to the input of the pulse-shaper stage. When adding step 3, it is therefore only necessary for the wire bridge mentioned in Section 2.2. to be removed from the printed circuit board. After equipping step 3, the required output frequency will be available at the output Pt 3.

2.4. FREQUENCY DIVIDER FOR 50 kHz AND 10 kHz (Fourth step)

A further SN 7490 N (IC 6) is used in step 3 as frequency divider. This is fed from the 100 kHz output and supplies 50 kHz and 10 kHz signals. These signals are then fed via the two remaining gates 4c and 4d remaining in step 3 and fed to the output via two remaining inputs of IC 5.

2.5. FREQUENCY DIVIDER FOR 5 kHz AND 1 kHz (Fifth step)

This module may not be required because the frequency stability or scale accuracy of a receiver or transmitter under test may make a calibration to an accuracy of 5 kHz or 1 kHz rather superfluous. As in the previous steps, a SN 7490 N (IC 7) is also used whereas the two gates (8a and 8b) of a further SN 7400 N (IC 8) are used as switch. The outputs are also combined in gate IC 5 which operates as a NOR-circuit.

Since the decoupling of the gates used as switches is not too high, all spectral lines are more or less modulated with the 1 kHz signal after equipping with the last frequency divider (IC 7). This can lead to errors. The two last remaining free gates 8c and 8d in IC 8 are therefore used for controlling IC 7 so that it only divides when the 5 kHz or 1 kHz signal is required. In all other cases, it remains blocked via the "reset" inputs.

If one selects for instance the "1 kHz" switch in addition to a higher frequency of for example 1 MHz, the 1 MHz signal will be modulated with 1 kHz. This, however, does not occur in a sinusoidal manner but as a squarewave so that a 1 kHz spectrum is generated above and below each 1 MHz line. This can be utilized when at very high frequencies the low-frequency lines are too weak, whereas the high frequency lines are still sufficiently strong. By modulating a high frequency spectral line, it is therefore possible for low-frequency lines to be heard.

3. MODIFICATIONS OF THE BASIC CONCEPT

If the spectrum generator is only to be extended for 10 kHz calibration markers, it is possible to utilize the switching possibility mentioned in Section 2.5. for the 50 kHz/10 kHz divider. In this case, step 4 (IC 6) is deleted and a bridge is made on the printed circuit board so that the input of the IC 6 (connection 14) is connected to the output (connection 11). IC 7 is now directly fed with the 100 kHz signal and will supply 50 kHz and 10 kHz. The control inputs 9 and 10 are grounded. The control for 50 kHz is now connected to 11, and 12 is responsible for the control for 10 kHz. As was explained in Section 2.5., IC 7 is only switched on when the 50 kHz or 10 kHz calibration markers are required. In other cases it is blocked so that no errors can be caused due to spurious 10 kHz or 50 kHz markers during operation of the 1 MHz, 500 kHz and 100 kHz markers.

4. OPERATIONAL NOTES

The integrated circuits used require a very accurate operating voltage of 5 V ± 0.25 V for correct operation. This voltage may not be exceeded considerably (upper limit for short periods 7 V) since the integrated structure will be destroyed. This is also valid for short interference pulses. The whole circuit must therefore be operated from a sufficiently stable and well filtered voltage source. The simple stabilizer given in Figure 6 is suitable for this. The zener diode should be found experimentally so that the voltage at the output of the stabilizer is well within the tolerance limits of 4.75 V to 5.25 V. The adjustment of the voltage must be made under load (approx. 150 mA) for which a resistor of 33 Ω/1 W is suitable. Only after the voltage is within the tolerance limit, is it possible for the load resistor to be removed and the voltage source connected to the spectrum generator. As mentioned in the introduction, a special voltage stabilizer will be described in part II of this series.

If a lower voltage is fed to the integrated circuits, they will not be damaged, but correct operation cannot be guaranteed.

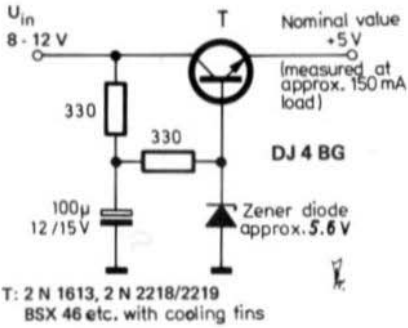


Fig. 6: Simple stabilizer

5. CONSTRUCTION

A double-coated printed circuit board must be used for construction of the calibration spectrum generator otherwise the necessary connections cannot be made. If a single-coated PC-board were to be used, it would be necessary for a large number of wire bridges to be provided. This, however, would result in error sources and would make the construction more difficult.

The construction of the very extensive circuit is simplified greatly by using the relatively expensive double-coated board with through-contacts. The printed circuit board possesses the dimensions 65 mm x 130 mm and is therefore suitable for accommodation in a modular case such as the TEKO 4 A. The connections are made via a 13 pole connector. This means that a completely screened plug-in module is obtained. Further details of this system were described in (1). If the module is not to be plugged, solder tags can be inserted instead of the connector strip. The printed circuit board is designated DJ 4 BG 004; Fig. 7 shows the conductor side. The conductor lanes on the component side as well as the component location plan are given in Figure 8.

5.1. NOTES REGARDING THE CONSTRUCTION

When observing the usual precautions with semiconductors, no difficulties should occur. The construction is made favourably by completing one stage after the other and checking the function after completion of each stage. Commencing with the oscillator, the construction is made as follows:

After soldering IC 1 onto the PC-board in the correct position indicated by the notch on the upper surface, the nominal operating voltage is connected. A power supply is suitable whose output voltage can be varied from 0 V. If this is not possible, a potentiometer should be connected in the operating voltage lead (approx. 1 kΩ) and the resistance value reduced in steps. The current flowing through the circuit is now measured so that the full short-circuit current cannot flow through the circuit if the IC has been incorrectly connected. On obtaining the nominal voltage (potentiometer to 0 Ω), a current of approximately 12 mA should be measured (deviations up to a maximum of 22 mA are permissible). With higher current values, it should be checked to see whether the integrated circuit has been incorrectly inserted or whether any short-circuits have been made on soldering the closely spaced legs of the integrated circuits.

After this, resistors R 1 to R 4, capacitors C 1, C 2 and C 3 as well as the crystal holder are soldered into place; however, the crystal is not inserted. A current of approximately 35 mA will then flow which will be reduced to approximately 25 mA after inserting the crystal. The fundamental signal of the crystal oscillator can now be monitored on a medium wave broadcast receiver, or the harmonics on a shortwave receiver. The frequency should be slightly variable with the trimmer capacitor C 3 without causing the oscillation to cease. The receiver can be connected to Pt 2 via a capacitor of a few picofarad. It is recommended that a ground connection be provided so that no transmitters can be induced into the receiver (caution should be made with AC-DC-receivers). In step 2, e.g. in the pulse-shaper, the output signal will appear at Pt 3. Since a source impedance of approximately 60 Ω is present at this position, it may be more favourable for receivers with coaxial inputs.

The construction is extended in a similar manner stage for stage in the order of the integrated circuit numbers. After installation of IC 1 to IC 5, the switching behaviour of the first switching stages is checked by selecting, for instance, 500 kHz (the other inputs must be grounded) and observing an harmonic. If the input of the gate associated with 500 kHz is grounded, the signal strength should be reduced noticeably. A wideband oscilloscope is very useful for this test (probe to Pt 2 or 3; horizontal deflection according to the frequency in question; sensitivity 1 V/cm). However, if care is taken, the calibration spectrum generator can be checked stage for stage by listening to the harmonics.

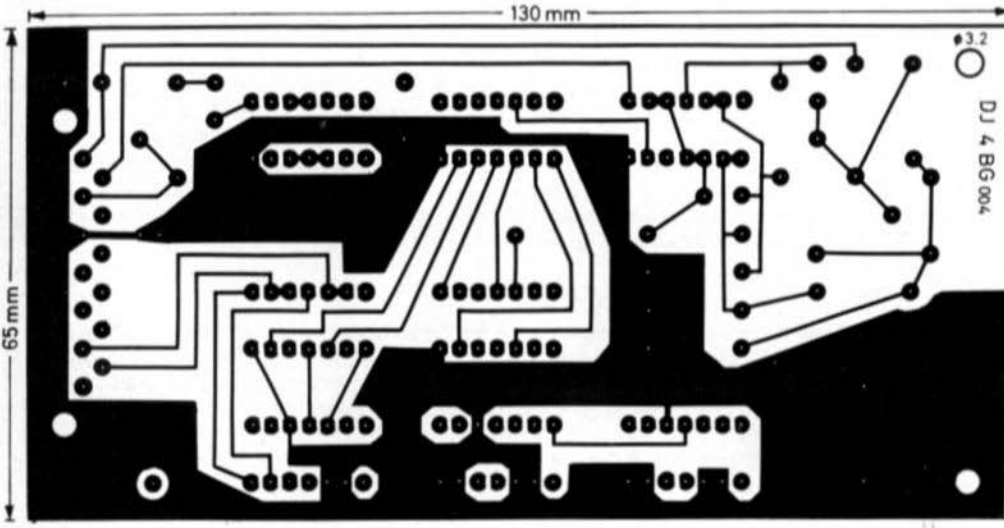


Fig. 7: PC - board DJ 4 BG, double-coated, with through-contacts

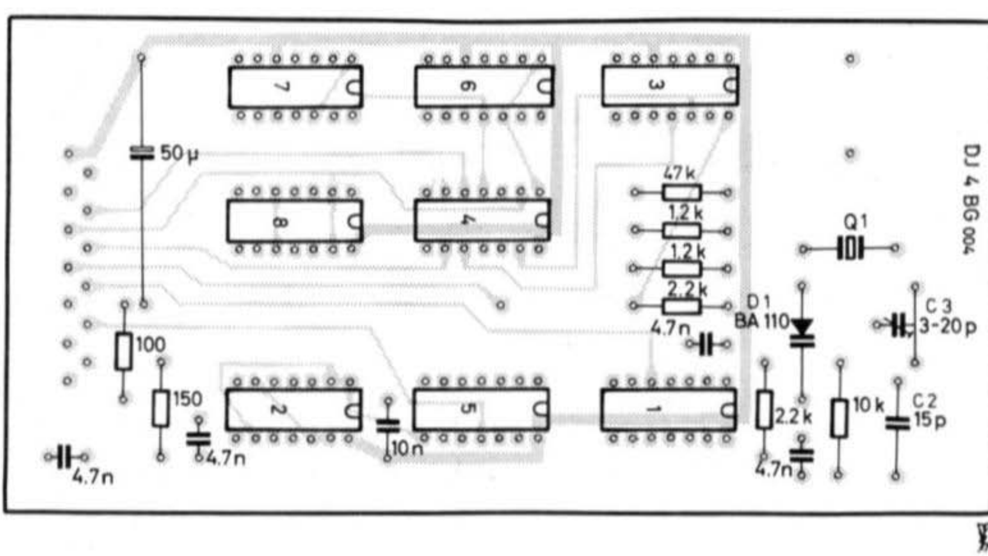


Fig. 8: Component side of DJ 4 BG 004

6. CALIBRATION

As long as the possibility does not exist for the alignment to be made exactly to 1 MHz using a frequency counter, it is usually made by beating one of the harmonics with a standard frequency transmitter. If a shortwave receiver is not available with which the 5 MHz or 10 MHz harmonic can be compared with WWV or other standard frequency transmitters, it is possible for the harmonic of the 100 kHz output of step 3 to be compared with the long-wave transmitter Droitwich on 200.000 kHz. The frequency error of this transmitter is less than 10^{-8} at the receive location. It is only necessary for a broadcast receiver covering the long-wave band that possesses a "magic eye" or similar tuning indicator to be used. On switching on the calibration spectrum generator, a beat frequency is usually heard. Trimmer capacitor C 3 is now aligned as accurately as possible to zero beat. However, the alignment should be made after the crystal oscillator has been in operation for some time so that transient effects will not occur. If the calibration spectrum generator is aligned so that a beat frequency of less than 1 Hz results when comparing it to the 200 kHz transmitter Droitwich, (e.g. the magic eye requires longer than 1 s in between the maximum and minimum signal strength indicated), the frequency of the calibration spectrum generator will deviate by less than 5×10^{-6} from the standard frequency. A deviation of 1 Hz from the standard frequency transmitter WWV on 10 MHz would mean a frequency accuracy of as great a 1×10^{-7}. However, this value cannot be maintained over a long period since the crystal is not enclosed in an oven; this means that a temperature variation of 1/10 °C will cause an equally great frequency variation.

This shows that it is more accurate to compare the output of the calibration spectrum generator to WWV than to the long-wave transmitter. When regular calibration is made to WWV, one will always be able to maintain the frequency deviation lower than 10^{-6}. This should be sufficient for virtually all amateur applications. Higher accuracies are possible when the crystal oscillator is synchronized to a standard frequency transmitter using an additional attachment. Such an attachment is to be described in Part III of this series.

7. THE INTEGRATED CIRCUITS AND EQUIVALENT TYPES

Only three different types of integrated circuits are used in the described calibration spectrum generator. These are the SN 7400 N which comprises four gates each having two inputs (see Fig. 9a for the socket connections), SN 7430 N: a gate with 8 inputs (see Fig. 9b), as well as the decade counter SN 7490 N. The latter integrated circuit consists of a chain of 4 flip-flops that can be used as a 2 : 1 and a 5 : 1 frequency divider (Fig. 9c).

If the original Texas Instrument's types are not available, the following table lists equivalent types of other companies:

Texas Instr.	AEG-TFK	National Semicond.	Siemens	Sprague	Transitron
SN 7400 N	TG 7400 N	DM 8000 N	FLH 101	SN 7400 N	TG 7400 E
SN 7430 N	TG 7430 N	DM 8030 N	FLH 131	SN 7430 N	TG 7430 E
SN 7490 N	TG 7490 N	DM 8530 N	FLJ 161	SN 7490 N	TG 7490 E

National Semiconductor also uses the same SN-type numbers as Texas Instruments in addition to the DM designations.

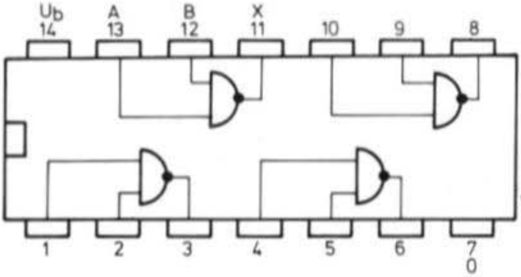


Fig. 9 a: **Socket connections of SN 7400 N**

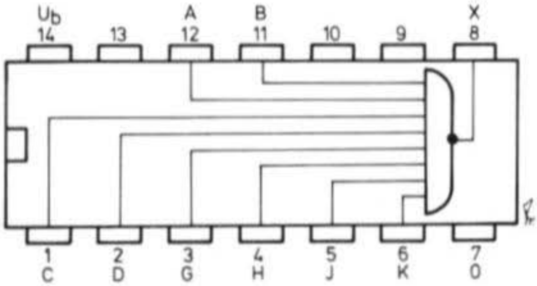


Fig. 9 b: **Socket connections of SN 7430 N**

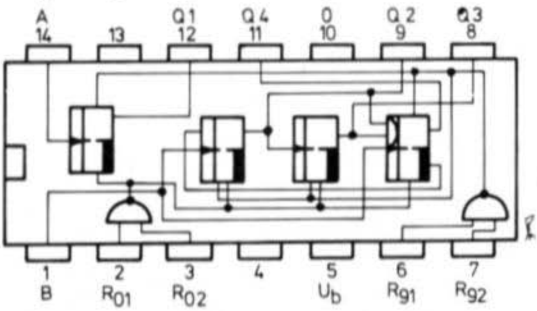


Fig. 9 c: **Socket connections of SN 7490 N**

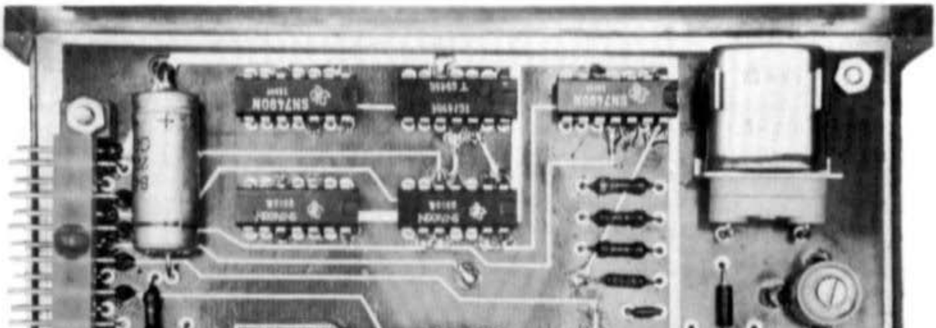


Fig. 10: Photograph of a prototype calibration spectrum generator DJ 4 BG 004
(no through contacts)

9. AVAILABLE PARTS

The printed circuit board DJ 4 BG 004, various individual components and a kit are available from the publishers or their national representatives; please see advertising page.

10. REFERENCES

(1) D. E. Schmitzer: Plug-in Modular Equipment
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 107-109

(2) D. E. Schmitzer: Basic Digital Circuits
VHF COMMUNICATIONS 3 (1971), Edition 3, Pages 150-155

(3) Texas Instruments: Series 54 Crystal Controlled Oscillator
The Network News, No 102 (22. Dec. 1966)

(4) H. Götting and D. E. Schmitzer: A Transistorized Calibration Spectrum Generator for Two Metres
VHF COMMUNICATIONS (2) 1970, Edition 1, Pages 41-44

NOTES AND IMPROVEMENTS TO THE DC 9 MD MINI WALKY-TALKY

Several improvements have come to the notice of the publishers which have been evaluated.

1. PRINTED CIRCUIT BOARD

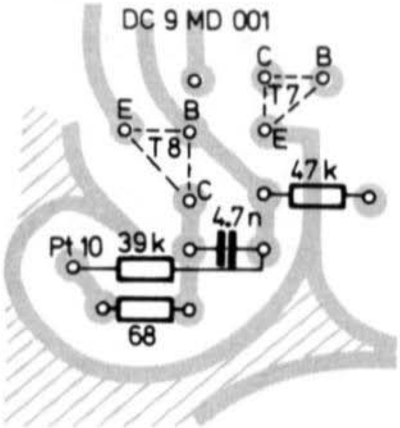
The manufacturer of the printed circuit board DC 9 MD 001 unfortunately use the old, preliminary drawing for printing the board. A number of these printed circuit boards were delivered before the fault was noticed. These printed circuit boards have two errors:

a) The emitter of transistor T 50 is not grounded; a small solder bridge is necessary.

b) The resistors R 30, R 31, R 32 and capacitor C 43 in the vicinity of transistor T 8 must be soldered as shown in the diagram to the conductor lanes which are somewhat different than the printed plan. The diagrams of the printed circuit board given in Edition 2/1971 are correct.

2. SEMICONDUCTORS

In some cases, it has been noticed that the receive oscillator (T 6) does not oscillate strong enough or not at all when using the given transistor BC 108 (low conversion gain and thus reduced sensitivity of the receiver). The reason for this is the low collector current of this stage which causes too low a transit frequency. Transistor types BF 224, 2 N 918 or other RF transistors are preferable. It has been found that diode 1 N 4148 is more suitable for frequency modulation of the transmitter than the older type 1 N 914.



3. COIL DATA

L 5 : 28 turns of 0.2 mm dia. (32 AWG) and not 0.4 mm dia. (26 AWG) enamelled copper wire.

L 51: 3 turns of 0.8 mm dia. (20 AWG) instead of 1 mm dia. (18 AWG) silver-plated copper wire.

4. CAPACITORS

The transmitter can be aligned more easily when a lower capacitance type is used for C 58 (e.g. 3.5 - 13 pF).

It is preferable for capacitor C 14 to be 15 pF instead of 20 pF so that the first IF transformer IF 1 is not loaded too greatly. Otherwise, it is sometimes difficult for the four intermediate circuits ZF 1 to ZF 4 to be aligned to a common frequency.

5. INTERCONNECTIONS

Some screened cables seem to be unsuitable for connecting the transmitter output and receiver input. Too thin a microphone cable should not be used.

6. MICROPHONE

The original microphone type MM 21 (Sennheiser) is no longer manufactured. Type B.4-22 (Beyer) is designed for noise compensation when installed in a headset. After mounting into the casing of the walky-talky, it was found that the response was rather muffled. This effect can be avoided by covering the rear of the microphone with a soft adhesive material.

7. SENSITIVITY

Too low a sensitivity can, as previously mentioned, be due to the fact that the local oscillator of the receiver is not providing enough voltage. Another means of increasing the sensitivity is by reducing the value of the emitter resistor R 1 of transistor T 1 to 680 Ω - 820 Ω.

STRIPLINES FOR VHF AND UHF

by K. Hupfer, DJ 1 EE

The following article describes how stripline circuits can be designed easily and with sufficient accuracy. Basic principles are to be given for the dimensioning of striplines of the required impedance, inductivity or capacitance.

1. GENERAL

The electronics industry have been using the well-known printed circuit boards for some time now. Of course, at low frequencies, the whole aim of this technique is to provide interconnections in the most economical and reliable manner. This has resulted in great space savings since the wiring is limited to a single plane under the components. Furthermore, the stability of the wiring has been increased considerably. One only has to consider an oscillator and the frequency variations that would be caused by slightly shifting the wiring.

However, the following considerations are to be limited especially to the electrical behaviour and application of stripline circuits in the frequency range of 100 MHz to 4 GHz and to show how lines and matching networks can be manufactured with the correct impedance.

2. IMPEDANCE

The impedance Z of a radio frequency line (balanced or coaxial cable) is defined by its inductivity and capacitance per unit of length. If the line losses are n eglected, the impedance Z can be calculated according to the following equation:

$$Z_{(\Omega)} = \sqrt{\frac{L'_{(H)}}{C'_{(F)}}}$$

Coaxial cables are preferred for the transmission of RF-energy due to the good screening and unbalance to ground and because most amplifiers are built up in a non-symmetrical configuration. Figure 1 shows the equivalent circuit diagram of a coaxial cable.

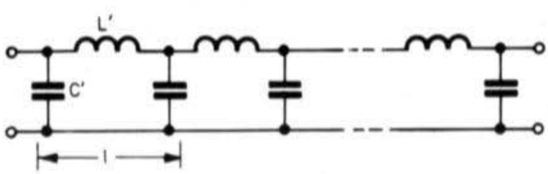


Fig. 1: Equivalent circuit diagram of a coaxial cable

Figures 2a to 2e illustrate several mechanical forms of RF-lines. The transition from coaxial cable to stripline can be followed step-by-step from a to e. By adding a material with a high relative dielectric constant ε_r and shifting the inner conductor, the line shown in Figure 2c is modified to a configuration as given in Figure 2d. One can imagine that virtually the whole RF-energy is confined within the space limited by the two conducting surfaces A and B. If the relationship of H/h is great enough, walls A , C and D can be neglected, which will result in a stripline as shown in Figure 2e having approximately the same RF-characteristics as a coaxial cable. The impedance Z is dependent on the dimensions w, h, d and B as well as on the relative dielectric constant ε_r.

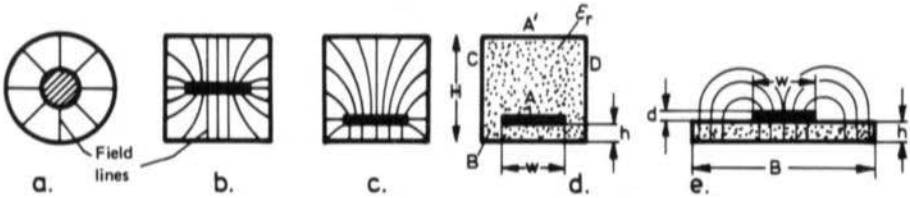


Fig. 2: Transition from coaxial cable to stripline

The calculation of the resulting impedance Z is complicated in comparison to the simple build up of the line in Figure 2e. The practical designer therefore uses a graphic representation from which the dimensions required to obtain a given impedance can be calculated easily. An error of 5% is still acceptable. Figure 3 gives a family of curves from which it is possible to read off the impedance Z of the stripline as a function of the dimensions w, h and the relative dielectric constant ε_r.

The base area of the stripline should have a minimum width of B $\geqq$ 3w for an undisturted transmission of the RF-energy. However, this is only valid for low impedances of up to 50 Ω. At impedances of 70 Ω to 120 Ω, the width should be B $\geqq$ 10w. If this rule is followed, no unwanted coupling will be made to other circuits.

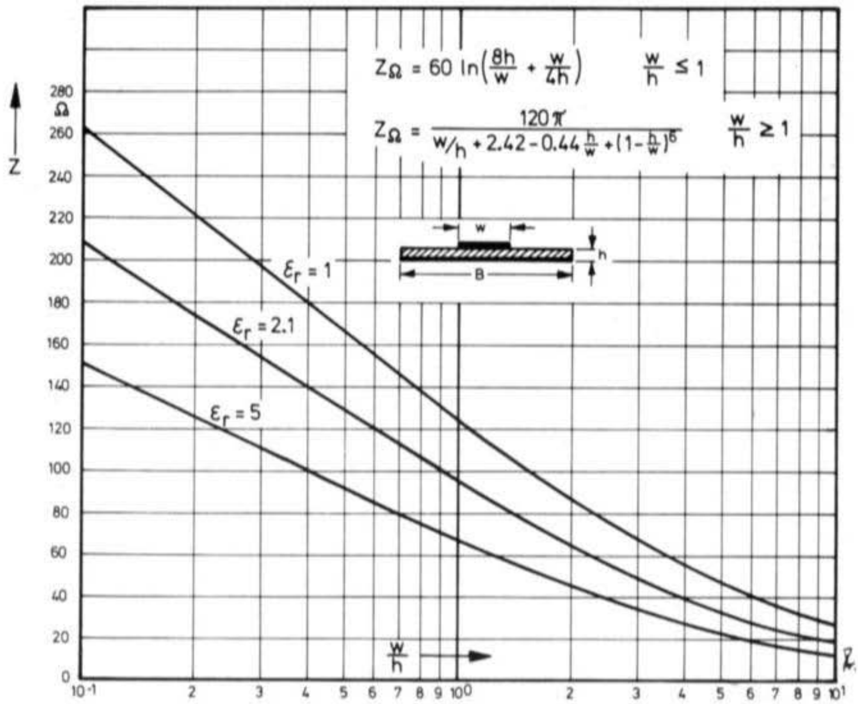


Fig. 3: Family of curves for the impedance Z of a stripline as a function of w, h and ε_r

It is advisable to keep the thickness h as small as possible. The lower dispersion of the field lines results in less coupling to other circuits. In addition to this, the surface currents on the outsides of both conductors (stripline and ground) will be reduced and with it the loss.

The relationship d/h is also of importance for the exact determination of the impedance Z. With the materials used in our example with $h = 2$ mm and $d \sim 35\,\mu$, $d/h \ll 0.5$ and need no longer be considered. This means that the curves given in Figure 3 are sufficient for determination of the impedance Z of a stripline.

The calculation of the impedance Z of a stripline is now to be described with the aid of an example: A 50 Ω line is to be manufactured on a double-coated PC-board having the thickness $h = 1.6$ mm.

What are the widths of the stripline and the base area? A value of $\varepsilon_r = 5$ can be assumed for epoxy PC-board material.

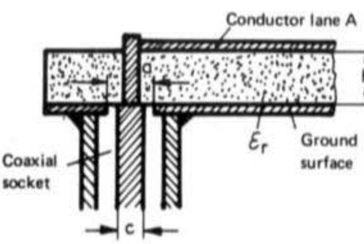
The required values are found in the graph by following the 50 Ω line on the vertical scale of Figure 3 until the intersection with the $\varepsilon_r = 5$ curve. The relationship of $w/h = 1.80$ can then be read off on the horizontal scale.

The result is: $w = h \times 1.8 = 1.6 \times 1.8 = 2.87$ mm

This means that a 2.87 mm wide stripline must be etched onto the printed circuit board. The base area width should be approximately $3 \times 2.87 \approx 9$ mm. This means that other lines and circuits should be spaced at least 4.5 mm from the stripline in our example.

2.1. TRANSITION: STRIPLINE TO COAXIAL LINE

A 50 Ω coaxial connector (e.g. BNC) is to be connected to a 50 Ω stripline at low impedance-discontinuity. If the transition is made as shown in Figure 4, a standing wave ratio of less than 1.2 will be obtained for frequencies up to 5 GHz.



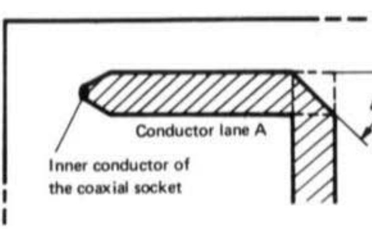


Fig. 4: Low-discontinuity transition between stripline and coaxial connector

The hole "a" on the ground side of the board in Figure 4 may not be as great as the inner diameter of the outer conductor of the coaxial connector. A good approximation is $a/c \approx 1.5$.

2.2. PLANNING THE CONDUCTOR LANE

As mentioned in Section 2, the dimensions of the base area are given as B $\geq$ 10w. No other lines may be located within this area.

As can be seen in Figure 4a, the corners are deleted when the conductor lane is bent by 90° on the PC-board. This is because strong electrical fields would otherwise be present which would have the same effect as increasing the line capacitance C' (Fig. 1). By cutting the corners diagonally, a certain compensation occurs which allows a SWR $\leq$ 1.05 : 1 to be obtained.

3. RESONANT LENGTH OF THE LINE

It is well known in transmission line theory that the propagation speed at $\varepsilon_r > 1$ no longer corresponds to c but is to the value of $1/\sqrt{\varepsilon_r}$ less. This means that the mechanical length of a line is by factor $1/\sqrt{\varepsilon_r}$ less than the wavelength λ.

Example: A $\lambda/2$ loop is used as balun. If air was used as dielectric, the following would be valid at 1.3 GHz:

$$\lambda = \frac{c}{f} = \frac{3000\,000}{1300} \left(\frac{\text{km}}{\text{s x GHz}}\right) = 0.23 \text{ m or } 23 \text{ cm}$$

and $\lambda/2$ = 11.5 cm

However, if RF-cable is used with ε_r = 2.3 the mechanical length of the $\lambda/2$ loop is only:

$$\lambda/2 \times \frac{1}{\sqrt{\varepsilon_r}} = \frac{23}{2\sqrt{2.3}} = 7.9 \text{ cm}$$

This calculation of the resonant wavelength is only valid for the triplate type of striplines (Fig. 5). In this case, the whole volume of the RF-line is filled with the dielectric which means that the same conditions exist as for coaxial cable with plastic dielectric.

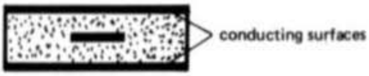


Fig. 5: Triplate stripline

This is not valid for the striplines shown in Figures 3 and 4. As can be seen in Figure 2e, not all field lines are completely within the dielectric. The magnitude ε'_r used for the calculation of the resonant length of the line is only at very small spacings corresponding to ε_r. In all other cases, $\varepsilon'_r < \varepsilon_r$. This results in a resonant line length $\lambda\varepsilon'_r$ that is between $\lambda\varepsilon_r = 1$ and $\lambda\varepsilon_r$.

Two curves are given in Figure 6 for practical purposes from which it is possible to determine the resonant line length l as a function of the construction w/h and the dielectric constant ε_r.

The use of this diagram is to be explained with the aid of an example:
Required is a $\lambda/4$ transformer for matching a transistor to a source impedance of Z = 50 Ω.
R_{BE} (assumed to be real) = 150 Ω; Z_{source} = 50 Ω; f = 432 MHz.
The impedance of the $\lambda/4$ line is calculated as follows:

$$Z = \sqrt{R_{BE} \times Z_{source}} = \sqrt{50 \times 150} = \sqrt{7500} = 86.6\ \Omega$$

If the stripline is to built up on a double-coated PC-board material of $\varepsilon_r = 5.0$, a ratio of $w/h \approx 4.8$ will result.

If the dielectric is 1.6 mm thick, the stripline will have a width of $w = 4.8 \times 1.6 = 7.7$ mm.

The curves given in Figure 6 allow the required length 1 to be determined:

$$\frac{\lambda_{SL}}{\lambda_o} = f(\varepsilon_r, h, w)$$

At $\varepsilon_r = 5$ and $w/h = 4.8$, a $0.55\ \lambda_o$ is obtained for λ_{SL}/λ_o. The resonant wavelength with air dielectric ($\varepsilon_r = 1$) amounts to 70 cm.

The length of the $\lambda/4$ stripline is:

$$1 = \frac{\lambda_{SL}}{4} = \frac{0.55 \times \lambda_o}{4} = \frac{0.55 \times 70}{4} = 9.6 \text{ cm}$$

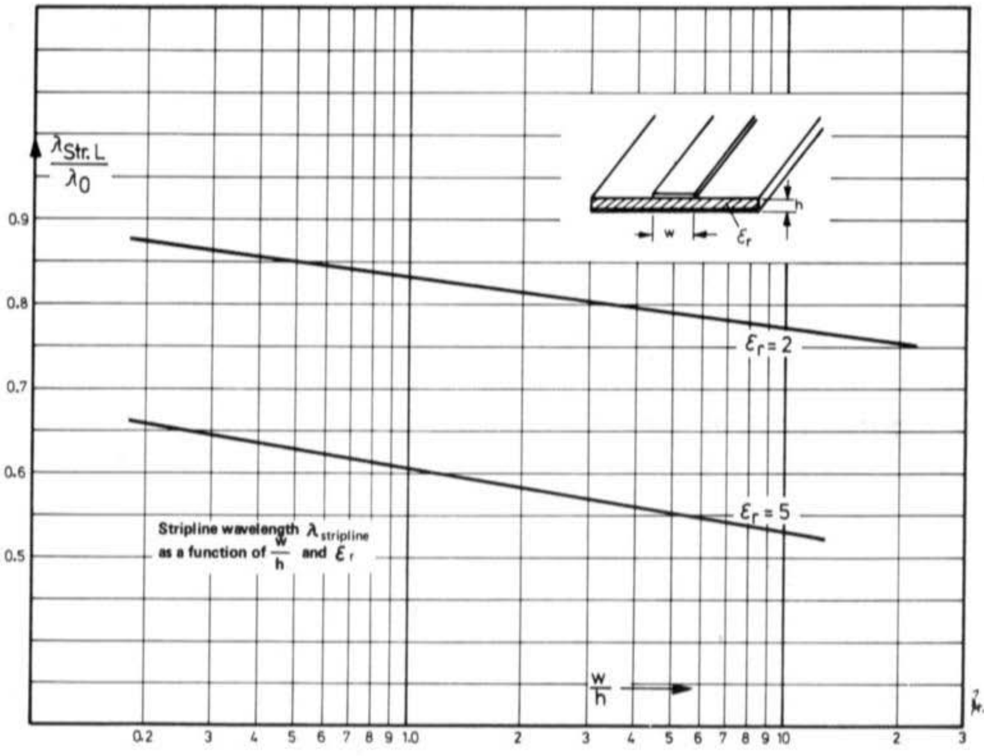


Fig. 6: Stripline wavelength as a function of the magnitudes of w/h and the dielectric constant ε_r

4. REACTIVE IMPEDANCE OF A STRIPLINE
4.1. INDUCTIVITY AND CAPACITANCE

As can be seen in Figure 1, an RF-line can be assumed to be an LC-circuit. A piece of line having a length of $l \leqq \lambda/8$ is equivalent to a π or T-circuit as shown in Figure 7. The inductivity L and the capacitance C can be calculated as follows:

$$L_{\pi} = \frac{Z}{2\pi f} \text{ sine } \frac{2\pi f l}{v}$$

whereby v is the velocity of electromagnetic waves.

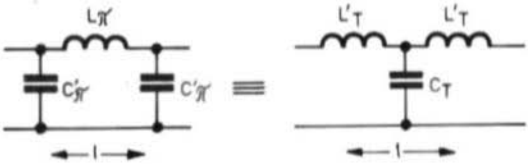


Fig. 7: Equivalent circuits

Fig. 7 a: π - circuit Fig. 7 b: T - circuit

As already mentioned, only lengths of $l \leqq \lambda/8$ are to be considered. A simple equation can be given for the circuit:

$$L = L_{\pi} = \frac{Z_o \times l}{v} \quad \left(v = \frac{c}{\sqrt{\varepsilon_r}} \right)$$

whereby c is the speed of light (3×10^{10} cm/s)

In the case of the T-circuit, the following is valid for the inductivity L'_T :
$L'_T = \frac{Z_o \times l}{2 v}$.

From the two equations and Figure 7 it can be seen that: $L_{\pi} = 2 L'_T$.

The capacitance components C'_{π} and C_T can also be calculated for short lengths of line:

$$C'_{\pi} = \frac{l}{Z \times 2v} \qquad C_T = \frac{l}{Z \times v}$$

This means that $C_T = 2C'_{\pi}$.

For RF lines of $Z > 60\ \Omega$, the value C'_{π} can be neglected in the equivalent circuit given in Figure 7 providing that the source and terminating impedances are not too great. It may be necessary for it to be included in the calculation in some cases, which is usually possible.

As an example, the L_{π} and C'_{π} values occuring with a line of $l = \lambda/16$ at 1296 MHz and an impedance of $Z = 100\ \Omega$ ($\varepsilon_r = 1$) are to be calculated. At a wavelength of 23 cm (1296 MHz), $\lambda/16$ corresponds to approximately 14.5 mm.

$$L_{\pi} = \frac{Z \times l}{v} = \frac{100\ (\Omega) \times 145\ (\text{mm})}{3 \times 10^{11}} = 5 \text{ nH}$$

$$C'_{\pi} = \frac{l}{Z \times 2v} = \frac{145\ (\text{mm})}{100 \times 2 \times 3 \times 10^{11}} = 0.23 \text{ pF}$$

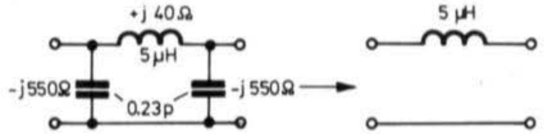


Fig. 8: L_π and C_π values for a stripline at 23 cm

Fig. 8 a: Exact equivalent circuit Fig. 8 b: Simplified circuit

The circuit as shown in Figure 8 results at 1.3 GHz for the 14.5 mm of 100 Ω line. The inductive and capacitive impedances are given for comparison. Since the capacitive impedance is approximately ten times greater than the inductive impedance it can be neglected with conventional transistor circuits. The result is the simplified equivalent circuit given in Figure 8b.

An impedance of approximately 120 Ω is advisable for construction of inductances in the stripline technology. The value of C'_π will then be negligible. The length 1 in mm for any required inductance value is given by:

$$l = \frac{L \times v}{Z_o}$$

whereby: Z_o is the impedance in Ω with $\varepsilon_r = 1$; $v = 3 \times 10^{10}$ cm/s. The inductance L is given in nH.

This equation is given in the form of a graph for practical use (see Fig. 9). The impedance Z_o is already contained in the ratio w/h so that it is possible for the dimensioning to be made immediately.

The previous example is now to be checked according to this curve: The assumed impedance of $Z_o = 100\ \Omega$ requires a w/h ratio of 1.65. The vertical line corresponding to this ratio is followed until the curve is met. The inductance value is read off on the vertical scale; in the case in question it is 3.4 nH/cm. At a length of $l = 14.5$ mm, the inductance L is:

$$L = 3.4\ \frac{\text{nH}}{\text{cm}} \times 1.45\ \text{cm} = 4.95\ \text{nH} \approx 5\ \text{nH}.$$

It can be seen from the previous details that ε_r is of no importance to the inductance of such short lines.

A practical example is now to be given where a transistor is to be matched a circuit: A power transistor for an output power of 6 W at 432 MHz exhibits a complex base-emitter resistance R_{BE} of approx. $2\ \Omega + j3\ \Omega$ (Fig. 10).

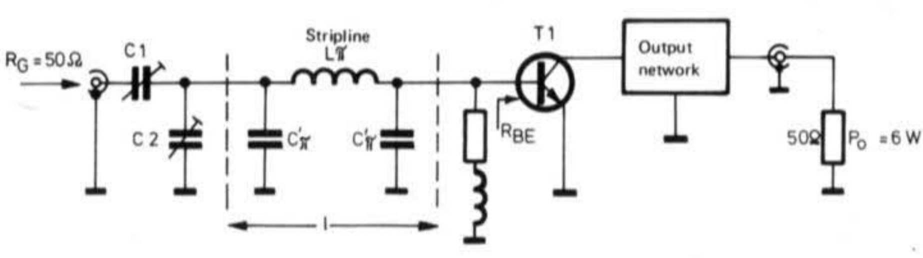


Fig. 10: Matching a transistor circuit via a stripline to a source impedance of 50 Ω

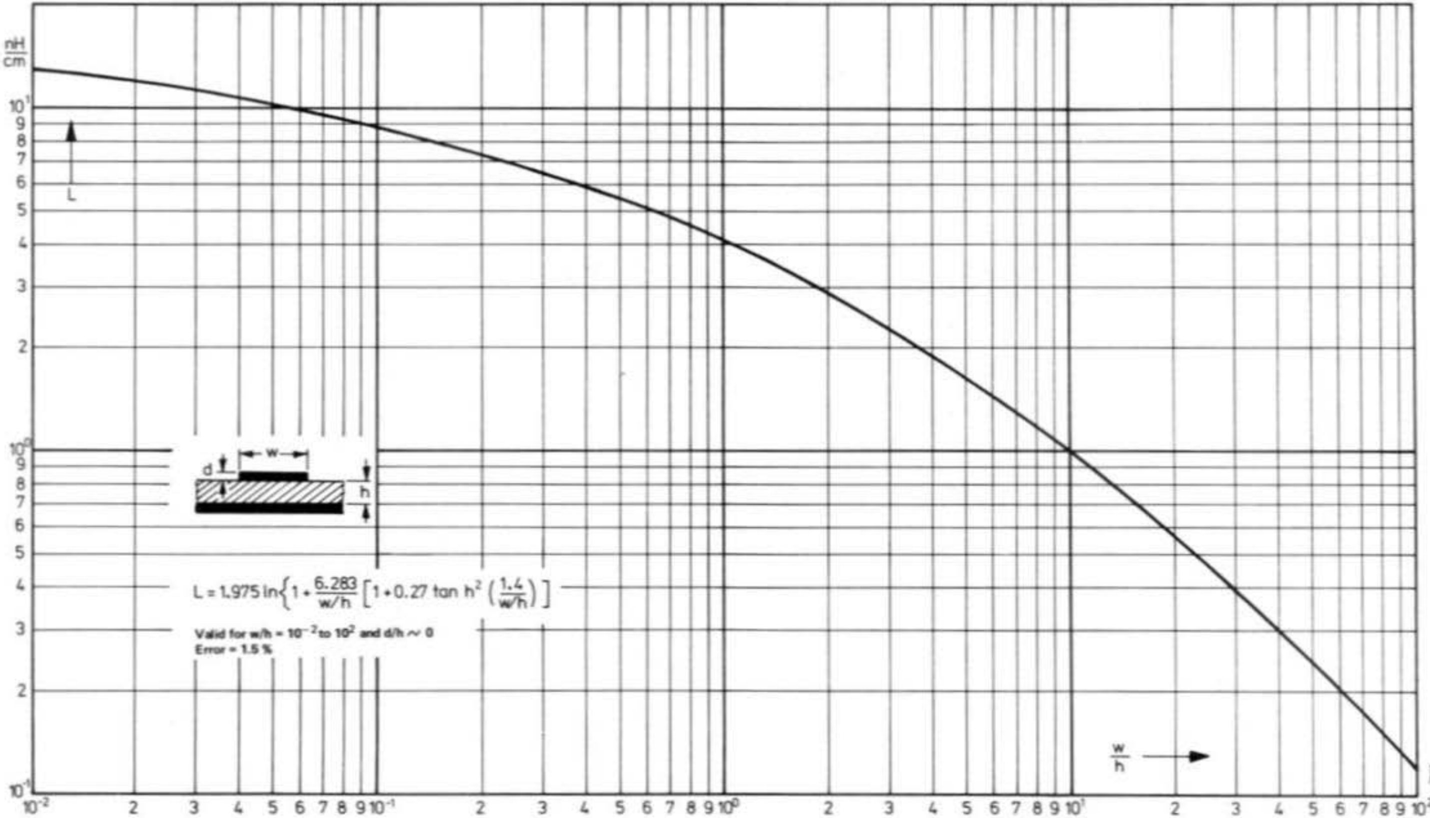


Fig. 9: Determination of the inductance L from the relationship w/h

The low, complex base-emitter impedance of transistor T 1 is to be transformed to the source impedance Z_s of 50 Ω with the aid of circuit elements L_π, C 2 and C 1. The required inductance is to be provided with a stripline of the impedance Z and the length 1.

The values of C 1, C 2 and L_π are determined as follows:
The impedance - admittance diagram shown in Figure 11 can be used. The basic values R_{BE} = 2 + j3 Ω and Z_s = 50 Ω are standardized at 10 Ω in order to work with the diagram. R_{BE} is then: 0.2 + j 0.3 Ω and Z_s is 5 Ω.

It is now necessary to transform 0.2 + j 0.3 Ω to 5 Ω. This is shown in Fig. 11.

X_L is determined as : $X_{L\pi} = 8.0\ \Omega$

$$L = \frac{X_{L\pi}}{\omega} = \frac{8\ (\Omega)}{2\pi \times 432\ (\text{MHz})} = 3\ \text{nH}$$

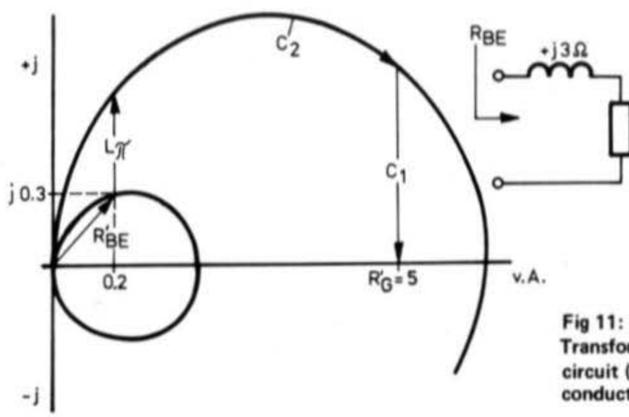


Fig 11:
Transformation process of the matching circuit (standadized) in the form of a conductance diagram

An inductance of L = 3 nH is obtained, according to the previously mentioned representation in Figure 9, at a w/h ratio of 2 and a length 1 = 10 mm. The mechanical dimensions of L_π are thus known.

If the calculation of the circuit elements for the matching network is continued, the following values are obtained for the trimmer capacitors: C 2 = 10 to 20 pF and C 1 = 5 to 20 pF. It will be seen that C'_π in the equivalent diagram of the stripline shown in Figure 7 has no influence.

4.2. CAPACITANCE

As was already mentioned in Section 4.1., a stripline can also be used to form capacitance. According to the equivalent diagram the following is valid:

$$C_T = \frac{1}{Z \times v} = \frac{1 \times \varepsilon_{eff.}}{Z_o \times c}$$

If Z is very low (wide stripline and high ε_r), C_T will be great. Z-values of approximately 10 Ω are normally used for forming capacitance in the stripline technology. The unwanted line inductance L'_T in Figure 7b will then be negligible.

As was previously mentioned, not all lines of flux run within the epoxy-glass fibre material. This means that a mean dielectric constant $\varepsilon_{r\ eff}$ must be employed that is smaller than ε_r. Only in this case will it be possible for a calculation to be made. However, $\varepsilon_{r\ eff}$ is dependent on the ratio w/h. It is therefore advisable for the capacitance of a stripline to be determined graphically for a given ε_r as was also the case on determining the inductance of the stripline. The capacitance per unit of length for $\varepsilon_r = 5$ is given in Figure 12. The capacitance of a stripline of $Z = 20\,\Omega$ ($w/h = 7.2$) and 10 cm in length is to be determined as an example. The stripline is accommodated on a double-coated epoxy board with an $\varepsilon_r = 5$. With the aid of the curve given in Fig. 12, a capacitance of 380 pF/m is obtained for a w/h value of 7.2 on the horizontal axis. A stripline length of 10 cm therefore provides a "printed capacitance" of 38 pF. This last example shows that it is only possible for relatively small capacitance values to be obtained when using normal thickness of PC-board material. Greater capacitance values can only be obtained using thinner PC-boards or larger areas in the case of normal boards. Anyway, it is usually necessary for variable capacitors to be used in conjunction with printed inductances in order to align the circuits to resonance at VHF and UHF. In most cases, printed capacitances are not required or are not absolutely necessary, and striplines need only be dimensioned for the required impedance and inductance.

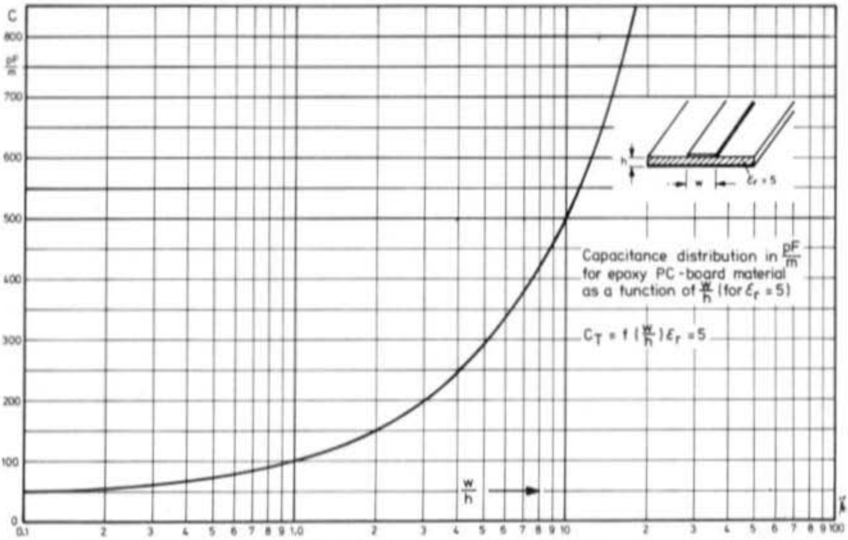


Fig. 12: Capacitance of stripline as a function of the relationship w/h

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Microwave Filters, Impedance-Matching Networks and Coupling Structures

SIMPLE 70 cm TRANSVERTER FOR PORTABLE EQUIPMENT

by J. Reithofer, DL 6 MH

Two meter portable transceivers can be extended for operation on the 70 cm band with the aid of this transverter. The transmit signal is tripled with the aid of a varactor diode which is able to handle CW, AM and FM signals (reduce the frequency deviation of FM). The same varactor diode is provided with a 288 MHz crystal-controlled signal in the receive mode so that it operates as a mixer diode and converts the 70 cm input frequencies to an intermediate frequency band of 144 to 146 MHz. All operating modes can be received on the receiver for which it is equipped (even SSB).

A similar transverter has already been described in (1). The described transverter is built up on a printed circuit board so that the construction is greatly simplified. Coaxial line circuits are somewhat difficult to manufacture and have been avoided in this case by using printed stripline circuits. When using the economic tuning diodes of UHF television tuners, a peak power of 3 W can be handled. With amplitude modulation, the unmodulated carrier level should amount to a quarter of this.

Approximately 30 to 40% of the input power is available on the 70 cm band. The sensitivity of the receiver corresponds to this power level. From good locations, communication has been made using a 4-element antenna to stations located more than 150 km away. The described transverter is extremely suitable for portable operation and the operator will be surprised how many stations can be worked with it.

Of course, the transverter is also suitable for local contacts. The main idea of this article is to assist VHF amateurs to become active on the 70 cm band.

Figure 1 shows the author's prototype with the cover removed.

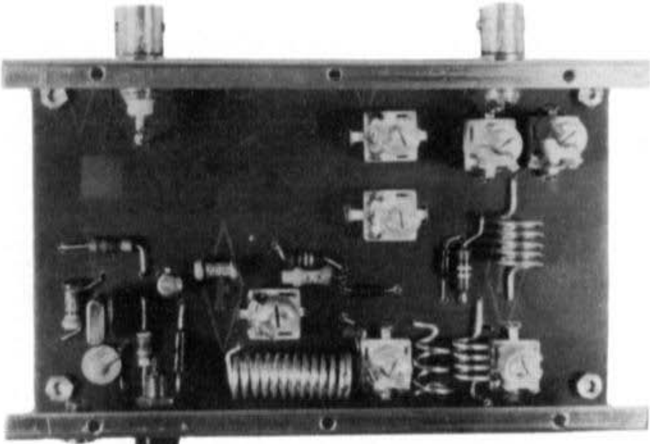


Fig. 1: A photograph of the 70 cm stripline transverter

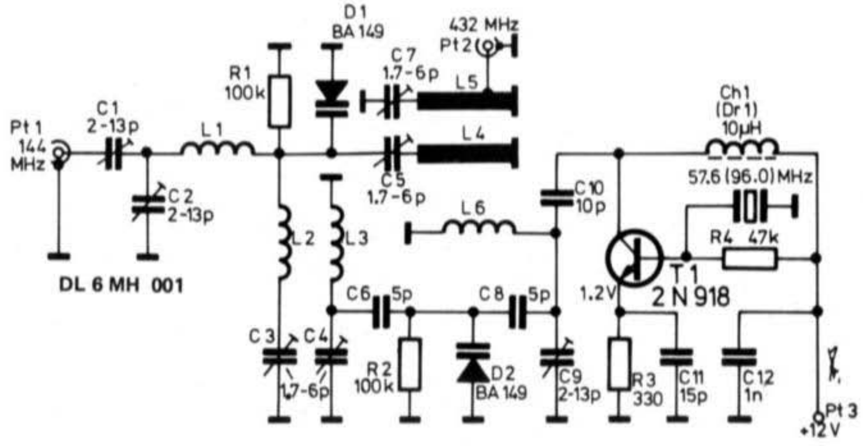


Fig. 2: Circuit diagram of the 70 cm transverter

1. CIRCUIT DETAILS

The circuit diagram of the 70 cm transverter is given in Figure 2. The varactor tripler will be seen upper left with a matching network tuned to 144 MHz (C 1, C 2 and L 1), a 432 MHz stripline bandpass filter (C 5/L 4 and C 7/L 5) at the output, and the idler circuit (L 2/C 3). The idler circuit is tuned to the first harmonic of 144 MHz, e. g. 288 MHz, where it represents a series resonant circuit shorting this frequency to ground. This virtually suppresses this unwanted frequency. In addition to this, the high short circuit current at 288 MHz improves the efficiency of the tripling process.

In the receive mode, varactor diode D 1 receives a local oscillator frequency of 288 MHz via the idler circuit. This means that the 432 MHz input signal is mixed with the local oscillator signal of 288 MHz so that an intermediate frequency of 144 MHz results. The auxiliary frequency is taken from a transistor oscillator which is controlled from a 57.6 MHz crystal. The resonant circuit at the collector (L 6/C 9) is connected to a varactor tripler via the small capacitance of C 8. This tripler operates in a similar manner to the tripler in the transmit branch but is, for simplicity, not provided with an idler circuit. If a 57.6 MHz crystal is used, the oscillator frequency will be multiplied five times. On the other hand, if a 96.0 MHz crystal is used it will only be necessary for the frequency to be tripled. The parallel circuit comprising inductance L 3 and trimmer capacitor C 4 is inductively coupled to the idler circuit. This parallel circuit is tuned to 288 MHz immaterial whether a 57.6 MHz or a 96.0 MHz crystal is used. It is only necessary for the resonant circuit of the oscillator (L 6/C 9) to be modified to suit the crystal frequency. The transistor oscillator only requires 8 to 12 mA which means that it can continue to oscillate in the transmit mode. Since the 70 cm antenna is connected to the 432 MHz output/input, it is only necessary for the 2 m station to be switched between transmit and receive.

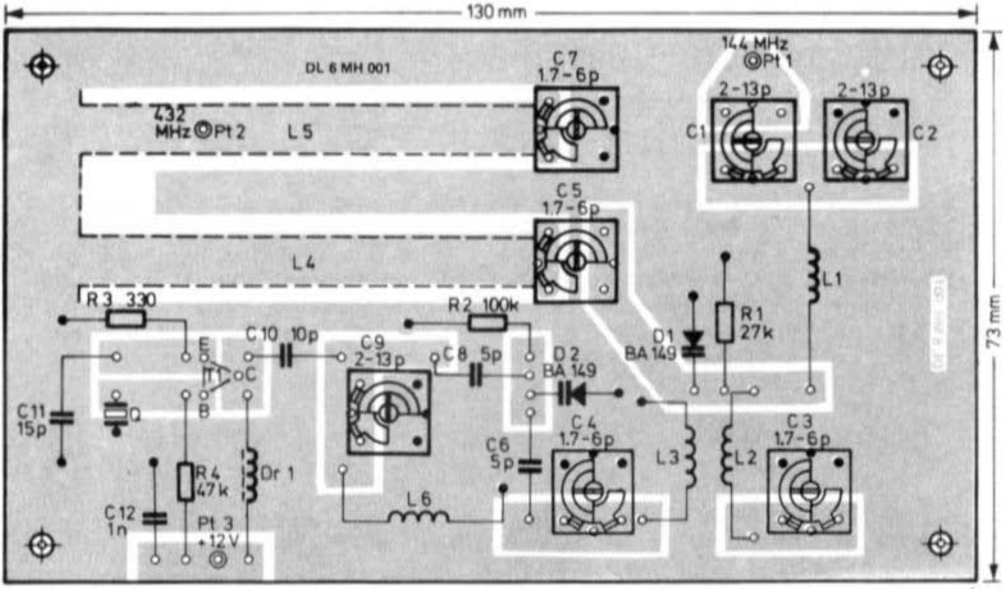


Fig. 3: Printed circuit board DL 6 MH 001 showing the component locations

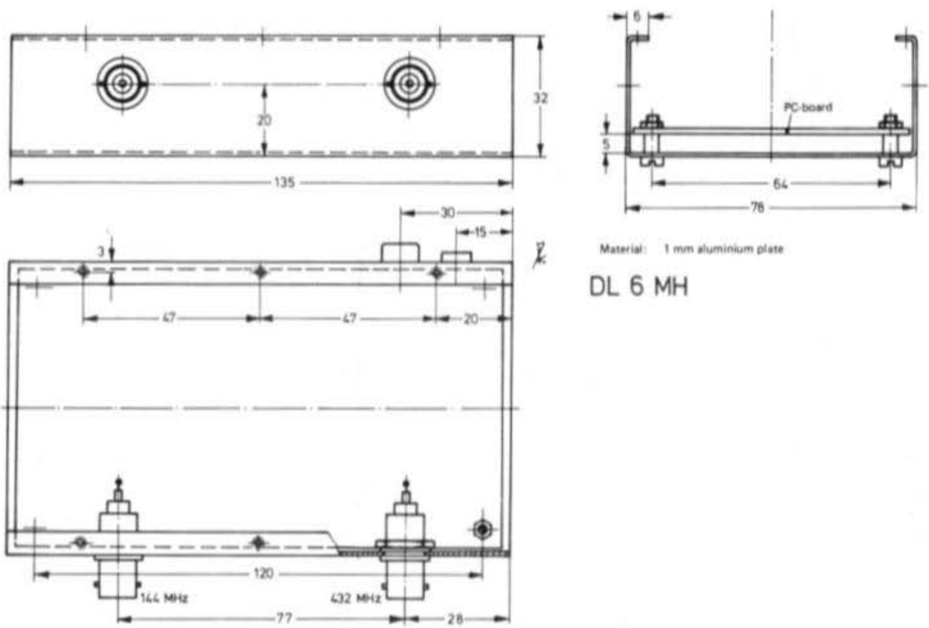


Fig. 4: Metal casing for the 70 cm transverter

2. CONSTRUCTION

The whole transverter circuit is built up on a single-coated printed circuit board having the dimensions 130 mm x 75 mm. Figure 3 shows this printed circuit board and the location of the components. The PC-board has been designated DL 6 MH 001. The completed printed circuit board is accommodated in a simple aluminum case whose dimensions are given in Figure 4.

An U-shaped cover, which is not shown, can be used to enclose the unit. A note to the ground connections is deemed necessary: Although the ground surface of the printed circuit board is grounded via the four mounting screws to the metal casing, it is necessary for an additional, low-inductive ground connection to be made to the two coaxial sockets. This is done by placing a piece of silver-plated copper wire around each coaxial socket and fixing it under the appropriate nut. Both ends of the wire are connected in the shortest possible manner to the ground surface of the printed circuit board, and without bend.

2.1. SPECIAL COMPONENTS

T 1: 2 N 918, 40 405, 40 519 (RCA)
D 1, D 2: BA 149 or BAY 70 (AEG-Tfk), BA 110 (ITT-Intermetall), 1 N 5462 A
Crystal: 57.600 MHz or 96.000 MHz in HC-6/U or HC-25/U holder

Inductances L 1, L 2, L 3, L 6 are made from 1.0 mm to 1.3 mm diameter (16-18 AWG) silver-plated copper wire. Turn spacing = wire diameter. Self supporting.

L 1: 5 turns on a 8 mm former
L 2: 4 turns on a 6 mm former
L 3: 3 turns on a 6 mm former, coil length extended to 17 mm
L 6: 12 turns on a 8 mm former (9 turns with a 96 MHz crystal)
L 4 and L 5: are printed striplines

Ch 1: 5-10 μH ferrite choke (Delevan) or ferrox beads with approximately 5 turns of approx, 0.2 mm (32 AWG) enamelled copper wire pulled through in the longitudinal direction.

C 1, C 2 and C 9: 2-13 pF air-spaced trimmer
C 3, C 4, C 5 and C 7: 1.7 - 6 pF air-spaced trimmer

Two BNC coaxial sockets for single hole mounting (UG 1094/U).

3. ALIGNMENT

The alignment is commenced with a preliminary alignment of the transmit branch. A good terminating resistor or 70 cm antenna, as well as RF output meter are required. The reflectometer described in (2) is very useful. A 70 cm bandpass filter, e.g. (3) connected between the transverter and the reflectometer simplifies the alignment considerably. After this measuring arrangement has been set up, it is only necessary for trimmers C 1, C 2, C 3, C 5 and C 7 to be aligned for maximum reading. If a bandpass filter is not available, it will be necessary for a dip-meter tuned to 288 MHz or an absorption wavemeter to be used to check whether it is this frequency that is present at the output. If this is the case, it is necessary for the transverter to be aligned so that the output power is available at a different position of trimmers C 3, C 5 and C 7.

A weak 70 cm signal (harmonic of a 2 m transmitter or a crystal oscillator fed to the input via an attenuator and, if possible, a bandpass filter) is used for the alignment of the receive branch. It is important that the signal only reaches the transverter via the antenna input socket and not directly. The 57.6 MHz oscillator is aligned with the aid of the dip-meter. If the 96.0 MHz crystal is used, the signal can be checked on a VHF-FM broadcast receiver. The oscillator must commence oscillation immediately on connecting the operating voltage. If this is not the case, the resonant circuit should be tuned with the aid of trimmer capacitor C 9 on the flat resonant slope. After this, the 288 MHz circuit should be tuned with trimmer C 4 for the best reception of a 70 cm signal which is continuously reduced in strength. The idler circuit with C 3 as well as the 432 MHz and 144 MHz circuits require a fine alignment. Finally alternate tuning in the transmit and receive modes should be made to find the best compromise. It may be advisable to finally align for best reception since the small loss of transmit power is not noticeable in practice.

4. REFERENCES

(1) L. Wagner: 144 MHz/432 MHz Transverter for Low Power and Field Day
VHF COMMUNICATIONS 1 (1969), Edition 1, Pages 32-35

(2) R. Griek: Simple Stripline Reflectometers for 144 and 432 MHz
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 89-92

(3) J. Reithofer: A Stripline Bandpass Filter for 70 cm
In this edition of VHF COMMUNICATIONS

INEXPENSIVE "VARACTOR DIODES" FOR FREQUENCY MULTIPLIERS

It was mentioned at the 1971 annual UHF-SHF Amateur Meeting in Munich, West Germany that several amateurs were successfully using overlay power transistors instead of varactor diodes in frequency multipliers. Since only the collector base diode is used, it is even possible to use transistors whose base-emitter diode has been destroyed and are therefore useless for their normal applications. Up till now, it is mainly transistor types 2 N 3375 and 2 N 3632 that have been used in frequency triplers from 144 MHz to 432 MHz. In the case of the 2 N 3632 input power levels of up to 10 W are possible.

Fundamentally speaking, all overlay transistor types should be suitable for replacing varactor diodes with input power levels in the order of the normal output power, and the multiplied output frequency as high as the usual operating frequency. The emitter remains unconnected; either the base or collector is grounded low-inductively using a metal strip.

Using this principle, it is possible for stripline triplers such as (1) and (2) to be constructed for higher power levels even if the original diodes, e.g. 1 N 5149 or MV 1806 (Motorola), BAX 11 (AEG-Telefunken), MA 4952 or MA-4953 (Microwave Associates), or BXY 14 F (Siemens) are too expensive.

REFERENCES

(1) H. J. Franke: A Ten Watt Transmitter for 70 cm
VHF COMMUNICATIONS 1 (1969), Edition 4, Pages 243-248

(2) H. J. Franke: 70 cm-23 cm Stripline Varactor Tripler
VHF COMMUNICATIONS 2 (1970), Edition 3, Pages 160-165

STRIPLINE BANDPASS FILTER FOR 70 cm

by J. Reithofer, DL 6 MH

It is very advisable for a 70 cm bandpass filter to be placed after varactor triplers and previous to the driver or power output stage in SSB transmitters in order to suppress subharmonic, harmonic and spurious frequencies. The two-circuit bandpass filter described here is extremely easy to construct since printed stripline circuits on an epoxy printed circuit board are used. As can be seen in Figure 1, it is only necessary for two trimmers to be soldered onto the PC-board and for two coaxial sockets for input and output to be connected. The filter is accommodated in a small casing made from aluminum plate or PC-board material.

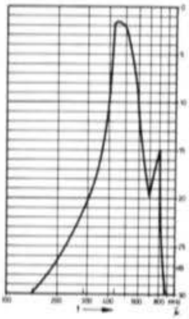


Fig. 2: Passband curve of the 2-circuit bandpass filter

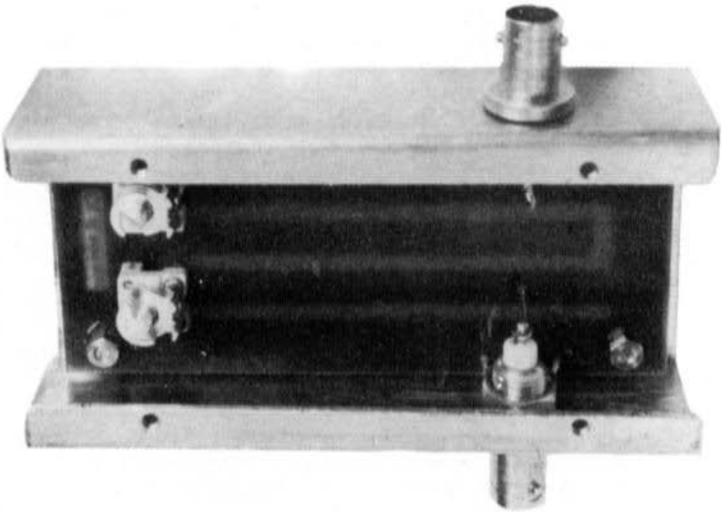


Fig. 1: Photograph of the stripline filter for 70 cm

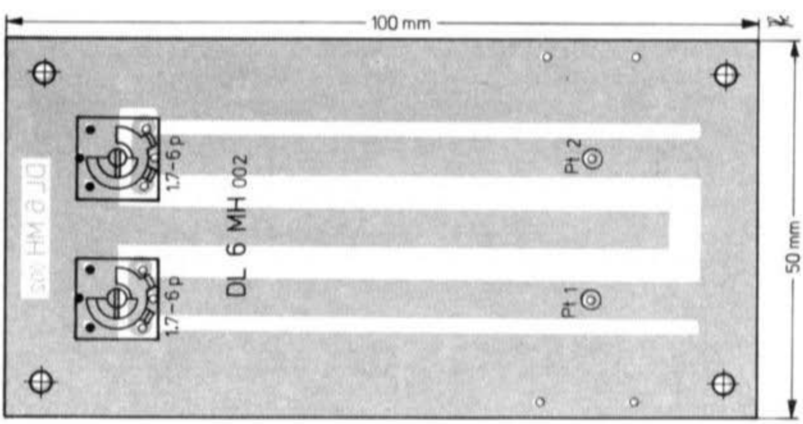


Fig. 3: Printed circuit board DL 6 MH 002 of the 432 MHz filter

The effectiveness of the filter can be seen by studying the passband curve in Figure 2. Two-meter signals are suppressed by more than 30 dB and their first harmonics at 288 MHz by approximately 20 dB. The first harmonic of the 70 cm band at 864 MHz will be suppressed by approximately 30 dB. The insertion loss amounts to approximately 2 dB which can easily be compensated for by the linear amplifier stages of a SSB transmitter.

The printed circuit board is designated DL 6 MH 002 and has the dimensions 100 mm long by 50 mm wide. The printed circuit board and component location are shown in Figure 3. The trimmers have a capacitance of 1.7 - 6 pF (or less). Air-spaced trimmers are used from which it may be necessary to remove one stator plate.

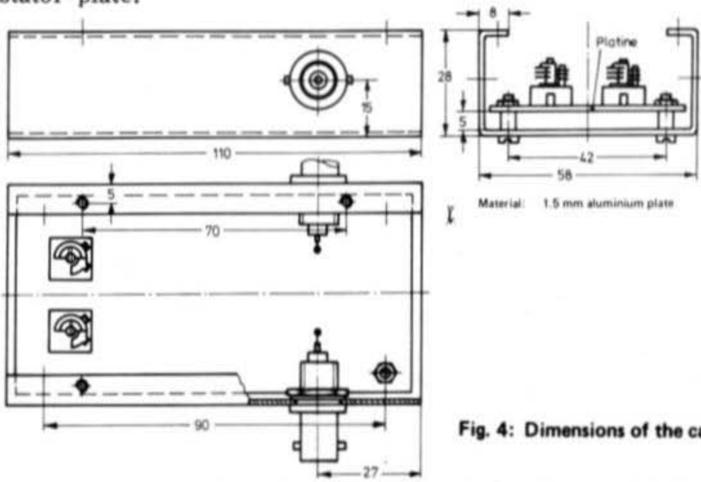


Fig. 4: Dimensions of the casing

After soldering the trimmers into place, the printed circuit board is built into a casing in which the coaxial sockets have already been mounted. Figure 4 shows the required dimensions. The printed circuit board should be spaced 5 mm from the floor of the case. Metal spacing bushings or nuts are used for this. Although the ground surface of the printed circuit board is grounded via the mounting screws, it is necessary for additional, low-inductive ground connections to be made between the coaxial sockets and the connection points marked on the printed circuit board. This is made by screwing a piece of silver-plated copper wire securely to the coaxial socket and soldering the ends to the ground surface of the PC-board. BNC sockets for single-hole mounting (UG 1094/U) are suitable. The filter is enclosed by adding a U-shaped cover.

For alignment of the filter, it is necessary for the output power of a 70 cm transmitter to be indicated between the filter and a terminating resistor (antenna). The reflectometer described in (1) can be used for this. The trimmers of the bandpass filter are aligned for maximum RF power. Finally, it should be pointed out that any impedance variation differing from those existing during the alignment will more or less load the filter circuit, or will detune it if the termination is not real. This can cause the insertion loss to increase considerably. The filter should be aligned in the circuit in which it is to be used.

REFERENCES

(1) R. Griek: Simple Stripline Reflectometers for 144 and 432 MHz
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 89-92

AN AUDIO-FREQUENCY RTTY-CONVERTER

by Dr. A. Gschwindt, HA 8 WH

This RTTY converter was mainly designed by the author for reception of the AFSK (audió frequency-shift keying) telemetry of OSCAR 6. However, the small transistorized unit is equally suitable for reception of amateur RTTY signals. It is hoped that OSCAR 6 will be launched during 1972 but there is still no confirmation of this.

1. PRINCIPLE OF OPERATION

The RTTY converter consists of two independent modules:

a) Digital FM discriminator

b) AF-filter

The digital discriminator (HA 8 WH 001) allows to receive and resolve any frequency shift between 100 Hz and 1.3 kHz without a bandpass filter.
The separate audio filter (HA 8 WH 002) is provided to suppress the interfering noise power outside of the useful pass band. The two filters are designed to allow a pass band at 2125 Hz and 2975 Hz, matching it to the most common frequency shift of 850 Hz.

The block diagram of module a) is given in Figure 1. The AFSK-signal to be decoded will be within one of the two audio frequency bands 900-2200 Hz and 1700-3000 Hz. Two circuit configurations have been designed to match these two AF-bands.

The audio output signal from the receiver is limited, amplified and limited again in order to obtain a square-wave signal of constant amplitude. This signal is now fed to the input of the digital FM discriminator which consists of two main circuits: a) differentiating network and limiter, b) integrating circuit.

The output voltage of the FM discriminator is linearly dependent on the frequency of the input signal. Two different types of discriminator can be used in this RTTY-converter. The one given in the complete circuit diagram operates in the audio frequency range of 900-2200 Hz. The circuits are to be described in more detail when covering each individual module.

The discriminator is followed by a post-detection filter to eliminate any residual jitter originating from the input signal and to provide an optimum bandwidth for the RTTY-pulses. The output signal to-noise ratio is improved in this manner. After filtering, the signal is fed to the limiter after which the limited signal is amplified in DC-coupled stages. The output stages of the amplifier operate as electronic switches for driving the relay of the RTTY machine.

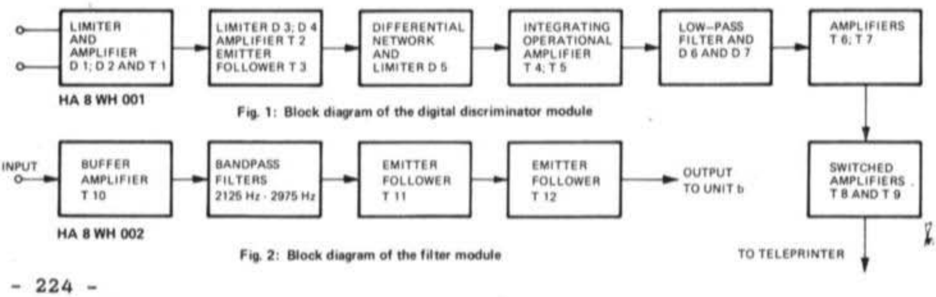


Fig. 1: Block diagram of the digital discriminator module

Fig. 2: Block diagram of the filter module

A switch is provided for switching the polarity of the DC output signal. It is therefore not necessary to tune the BFO of the receiver to the other side of the received signal.

The block diagram of the audio frequency filter (HA 8 WH 002) is given in Figure 2. The audio signal from the receiver is fed to a buffer which drives the bandpass filters tuned to 2125 Hz and 2975 Hz respectively. The filter is followed by two emitter-followers which feed the signal to the FM-discriminator.

2. CIRCUIT DETAILS

2.1. DIGITAL FM-DISCRIMINATOR

The circuit diagram of this module is given in Figure 3. The minimum input voltage for correct operation is 1 V_{rms}. The source impedance should be in the order of 600 Ω. Since fading often occurs on the received signal, it is very useful to increase the level of the input signal even up to 8 V-10 V_{rms}.

Diodes D 1 and D 2 are limiters to maintain the input signal at 1.4 V_{pp}. Transistor T 1 is a class A amplifier which is provided with a second limiter at the output (D 3, D 4).

The limited signal has satisfactory rise-time/fall-time characteristics but the amplitude is very low. Transistor T 2 now amplifies the square-wave signal and passes it to the emitter-follower T 3 which in turn feeds the differentiating RC network. The actual demodulation process is described in detail in (1).

It was mentioned previously that the discriminator is operative in two frequency bands. With the components given in Figure 3, the useful frequency band will be 900 Hz-2200 kHz. Some voltages and waveforms are given in Figure 4.

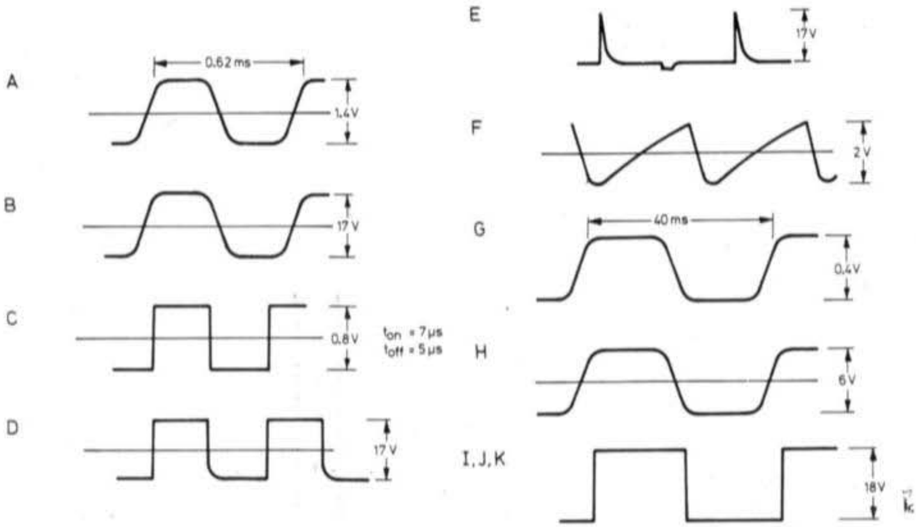


Fig. 4: Signal waveforms at various points of the discriminator module. With curves A to F, the input signal is sinusodial at 1.6 kHz; curves G to K the AFSK-signal is modulated with a 25 Hz squarewave signal, 850 Hz shift.

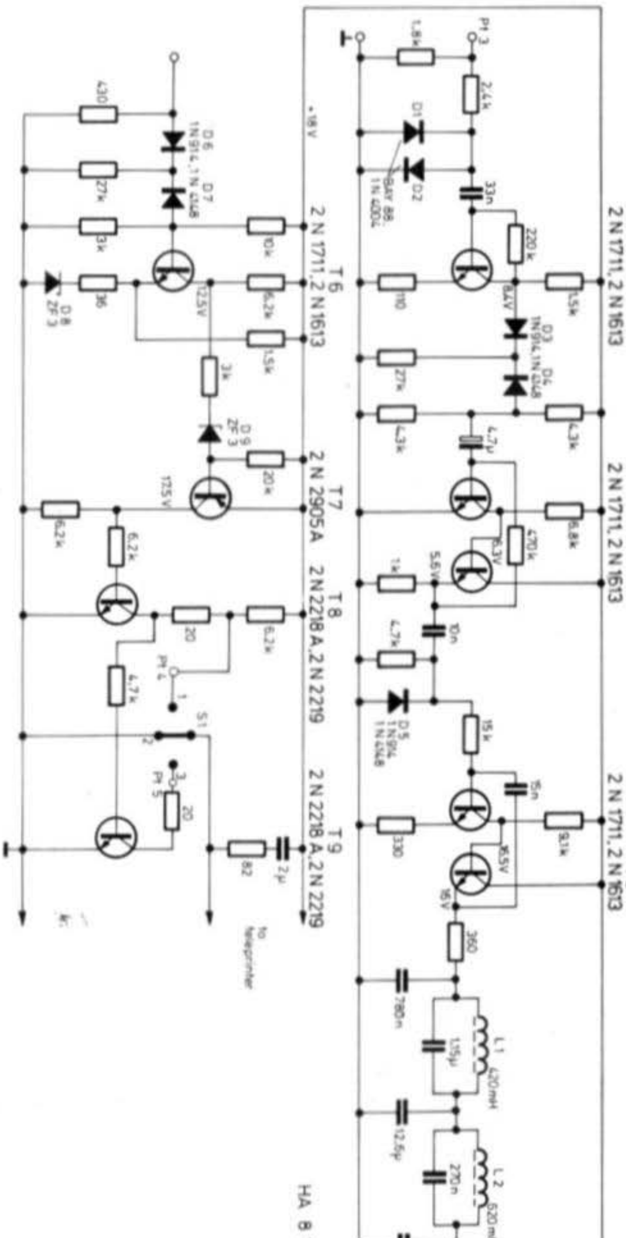


Fig. 3: Circuit diagram of discriminator module (DC voltages measured without input signal)

After differentiating and removing the negative pulses, the signal is fed to the integrating circuit comprising transistors T 4 and T 5. The time constant of this integrating amplifier is dependent on the speed of the transmitted information. In the 50 Baud mode, the highest square-wave frequency originating from an RTTY machine will be 25 Hz.

It would be advantageous to use a long integration time. However, if the integration time is too long the information will be lost. The dimensioning of the circuit has been chosen to be suitable for the mentioned parameters.

Whilst designing the converter, one attempted to obtain the same sensitivity for both the 900-2200 Hz and the 1.7-3 kHz discriminators. This is why the components of the differentiating and integrating circuits vary in conjunction with each discriminator. Figure 5 shows both configurations together with their transfer characteristics.

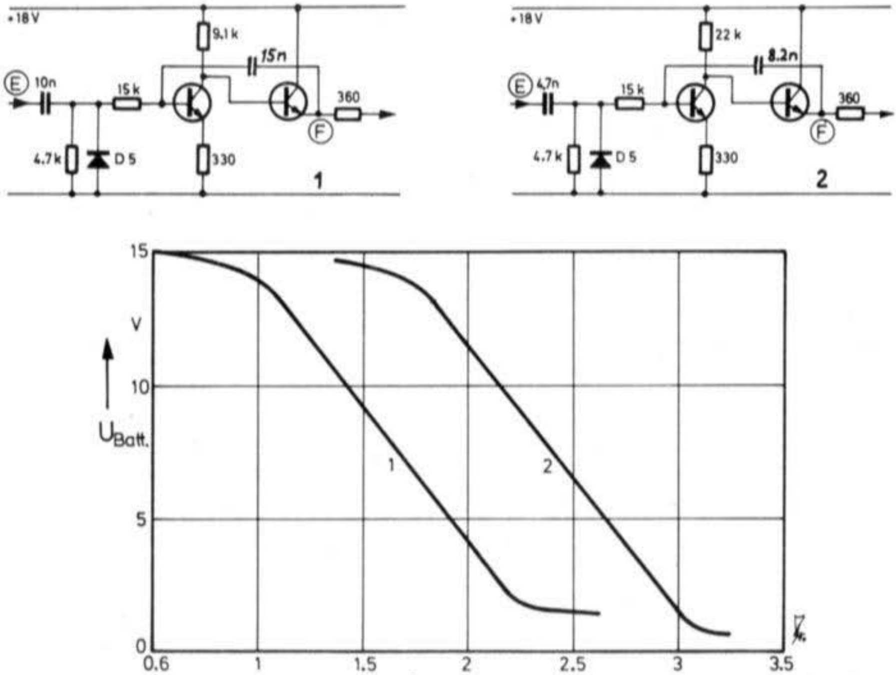


Fig. 5: Two discriminator circuits together with their transfer characteristic

The integrating stages T 4 and T 5 are followed by a low-pass filter which should possess the following characteristics: sharp cut-off and minimum waveform distortion. The bandwidth has been chosen which is suitable for a transmission speed of 50 Baud (25 Hz square-wave). The filter passes the third harmonic of the 25 Hz-signal at 75 Hz which will be 3 dB down. The filter is an L-type having minimum distortion of the waveform but no sharp cut-off characteristic. The circuit and measured transfer characteristics of this filter are given in Figure 6.

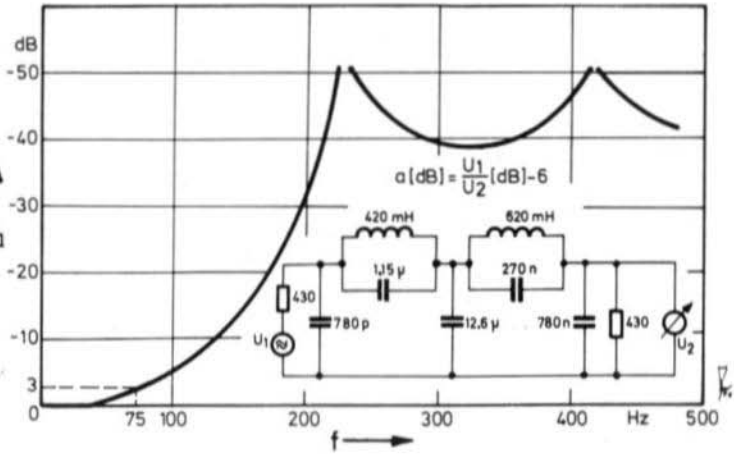


Fig. 6: Low-pass (post detection) filter with transfer characteristic

The filter drives the series limiter comprising diodes D 6 and D 7. This limiter is designed to determine the mean value of the discriminator and to keep the input level for transistor T 6 constant. The pulses are then amplified in the class A amplifier comprising T 6; the operating point is shifted in the emitter circuit to match the DC levels of the limiter at the base of transistor T 6. The amplified pulses switch the collector current of transistor T 7. The relay of the RTTY-machine is directly switched by the electronic switch comprising transistors T 8 and T 9. The given circuit provides a DC-voltage of 16 V which is sufficient for the RTTY-machine used by us. If higher DC-voltages are required, the collector voltages of T 8 and T 9 can be increased up to the maximum permissible level.

The RTTY-machine represents an inductive load for the electronic switches. The transistors could be destroyed on switching off due to the induced voltages. In order to avoid this, a series RC-transient suppressor is connected to the output of the electronic switches.

Switch "S" allows the polarity of the pulses to be inverted or for the RTTY-machine to be muted with a constant stop-signal.
A stable, well filtered power supply should be used.

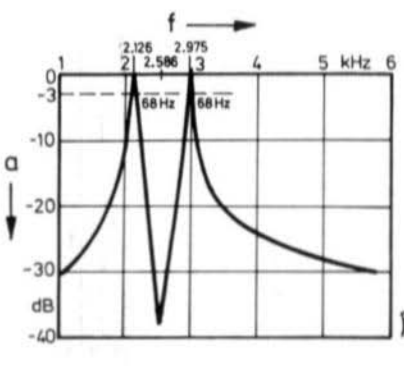


Fig. 7: Transfer characteristic of the filter module

2.2. AUDIO FREQUENCY FILTER

The circuit diagram of this module is given in Figure 8. The telemetry signals from OSCAR 6 will be transmitted to ground in the A3-mode. The mark and space signals will have a frequency of 2125 Hz and 2975 Hz respectively. It is important to use as low a bandwidth as possible for reception in order to improve the signal-to-noise ratio. The bandwidth is dependent on the speed (Baud). of the RTTY-signal. A bandwidth of 70-75 Hz around the mark and space signals will be suitable for 50 Baud.

The described filter has two pass band frequencies at 2125 Hz and 2975 Hz. The same bandwidth and amplitude is required for both signals.

These conditions are fulfilled in our circuit using two parallel-tuned LC-circuits with the same capacity (Fig. 8). This means that their resistance is equal at resonance. Of course, their Q-factors vary: The Q-factor of the LC-circuit tuned to 2975 Hz is 42 (bandwidth 70 Hz). In practice, a higher Q-factor can be realized than is required in this application. The filters in question are loaded with a variable resistor in order to maintain the required Q-factor.

The frequency response of the two filters is shown in Figure 7.

Transistor T 10 drives the two resonant circuits at a constant current and high source impedance. The input level is controlled by potentiometer P 1. The input signal must be maintained at a level that will not overload transistor T 10. Required is an input voltage of approximately 0.2 V_{rms}. The high input impedance of transistor T 11 ensures that T 10 is not loaded. The maximum output voltage is 15 V_{pp}.

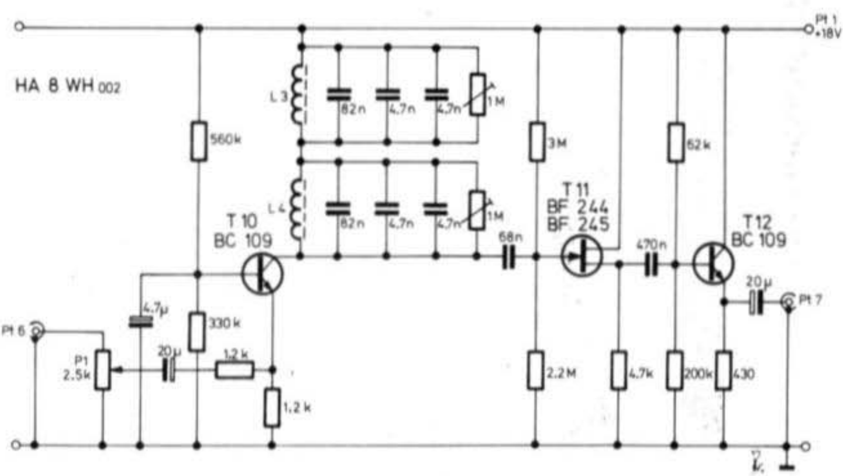


Fig. 8: Circuit diagram of the filter module

3. CONSTRUCTION

The RTTY-converter is built up on two separate PC-boards. HA 8 WH 001, which is shown in Figure 9 accommodates the digital FM-discriminator. The dimensions of PC-board HA 8 WH 001 are 82.5 mm x 182.5 mm. The audio frequency filter is accommodated on PC-board HA 8 WH 002 with the dimensions 82.5 mm x 102.5 mm. This board is given in Figure 10.

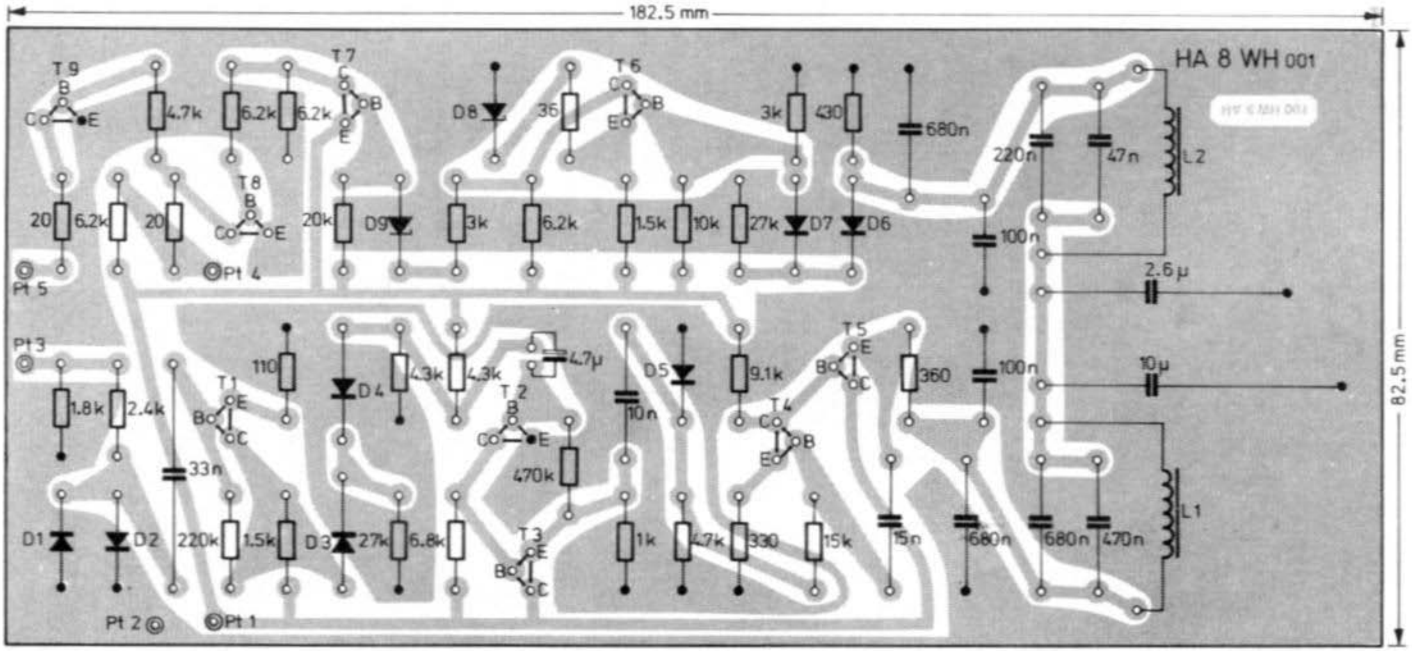


Fig. 9: Printed circuit board HA 8 WH 001 (discriminator)

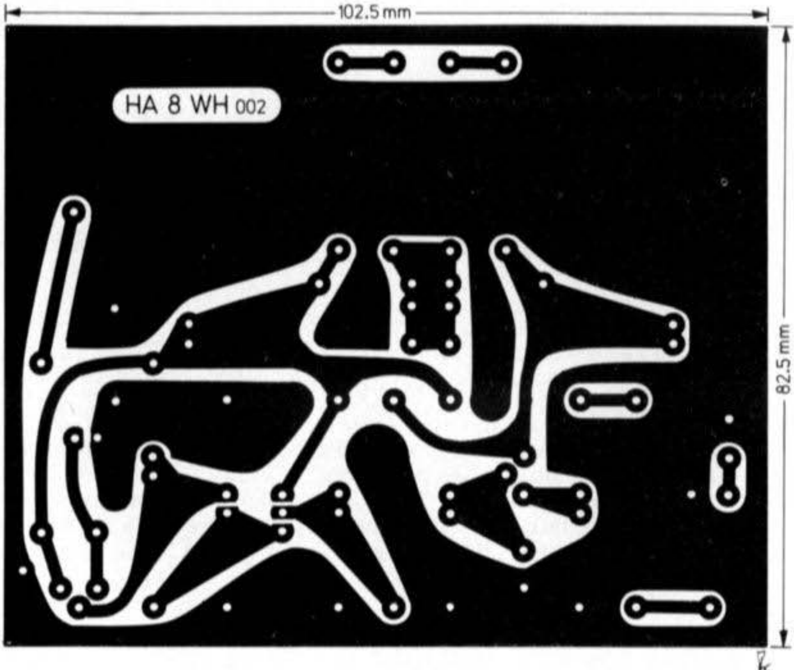


Fig. 10: Printed circuit board HA 8 WH 002 (filter)

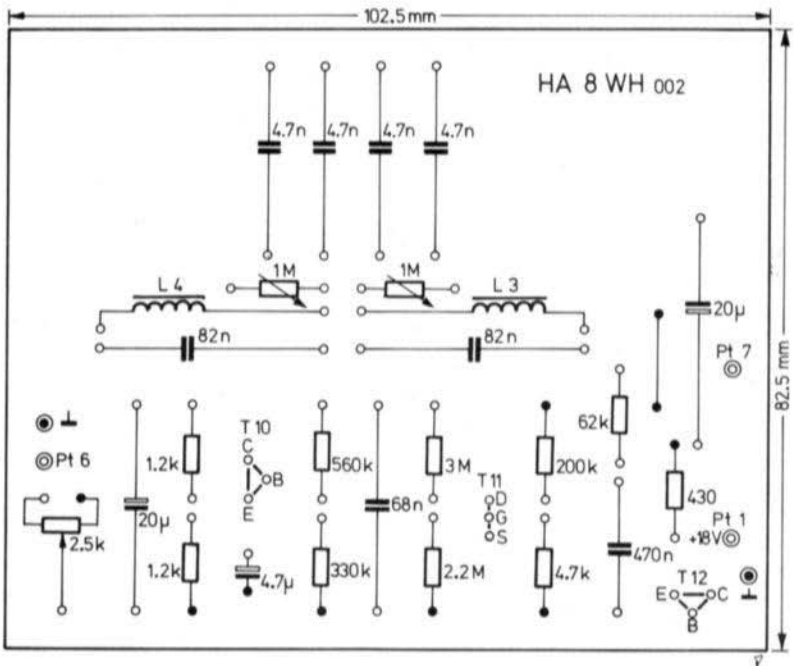


Fig. 11: Component location plan for HA 8 WH 002

3.1. COMPONENT DETAILS

T 1, T 2, T 3, T 4, T 5, T 6: BFY 46, 2 N 1711, 2 N 1613
T 7: BC 160, 2 N 2905 A
T 8, T 9: 2 N 2218 A, 2 N 2219
T 10, T 12: BC 109
T 11: BF 244, BF 245

D 1, D 2: BAY 45, BAY 88, 1 N 4004
D 3, D 4, D 5, D 6, D 7: OA 1160, 1 N 914, 1 N 4148

L 1: 420 mH
L 2: 620 mH
L 3: 65 mH
L 4: 32 mH

4. ALIGNMENT

Connect the operating voltage to the discriminator module and check the DC-voltages given in the circuit diagram (Fig. 3). If these are correct, an audio generator should be connected to the input of the discriminator and tuned over the frequency range of interest. Plot the output voltage as a function of the input frequency. The nominal curve is given in Figure 5. The magnet of the teleprinter will be actuated when the centre frequency is passed.

If an oscilloscope is available, the curves given in Figure 4 can be checked at the test points A to K.

The discriminator module is now ready for operation.

The most critical alignment for the low-pass filter is the accurate tuning and loading of L 3 and L 4. The fixed capacitance across these potted cores should be identical; the frequency alignment is made by varying the cores.

5. REFERENCES

(1) D. E. Schmitzer: A Digital Discriminator Acessory for FM Demodulation
VHF COMMUNICATIONS 2 (1970), Edition 2, Pages 105-110

A SIMPLE MODULATOR FOR FM-TRANSMITTERS

by K. P. Timmann, DJ 9 ZR

An AF-amplifier module is to be described having similar characteristics to the FM-modulators of commercial mobile FM-systems. The modulator requires a minimum of components and is easy to construct. A circuit diagram of the module is given in Figure 1. The LC combination (Ch 1/C 1) stops any RF-voltage from reaching the input of the two-stage preamplifier; this stage is followed by a balanced amplitude limiter comprising two silicon diodes which are provided with a stabilized bias voltage. A passive low-pass filter comprising potted inductance L 1 and capacitors C 7, C 8 and C 9 suppresses all frequencies in excess of 3 kHz. The subsequent amplifier stage amplifies the limited and filtered audio signal to a level suitable for modulation of varactor diodes.

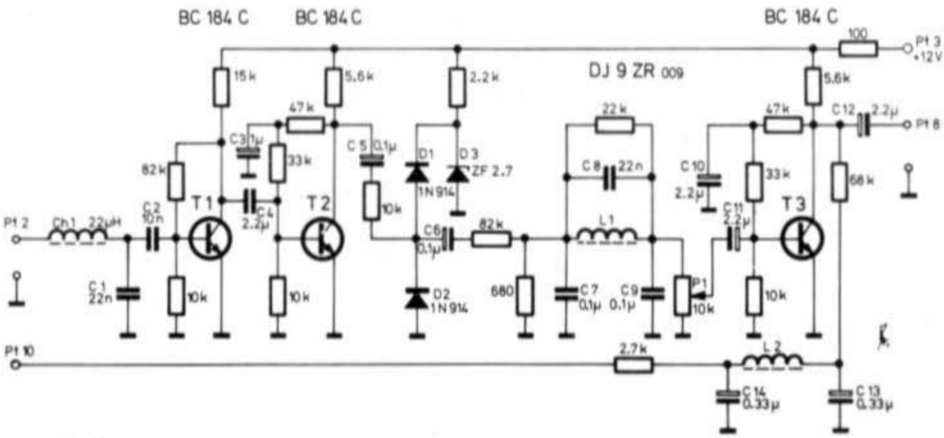


Fig. 1

A special feature of this circuit is the feedback loop via an AF-resonant circuit comprising potted inductance L 2: if connection point Pt 10 is connected to the input Pt 2, the feedback loop will be closed and the circuit will produce a 2.5 kHz oscillation. This calling or test tone will only be generated if the input (Pt 2) is not connected to a low-impedance source. If a microphone having an impedance of less than 200 Ω is connected to the input, the feedback voltage will be reduced to such a level that oscillation will cease and the modulator will operate normally. This means that it is only necessary for the microphone to be disconnected with the aid of a pushbutton contact when the calling or test tone is required.

Figure 2 illustrates the frequency response of the modulator module at low voltage levels. It represents a typical frequency response necessary for FM voice transmissions including the required pre-emphasis. The gain drops off by 6 dB/octave at frequencies lower than 3 kHz and at 25 dB/octave toward higher frequencies. With an infinite drive (clipping), the maximum output level is obtained at 800 Hz; it is 1.2 dB (10%) higher than the undistorted maximum signal at 3 kHz.

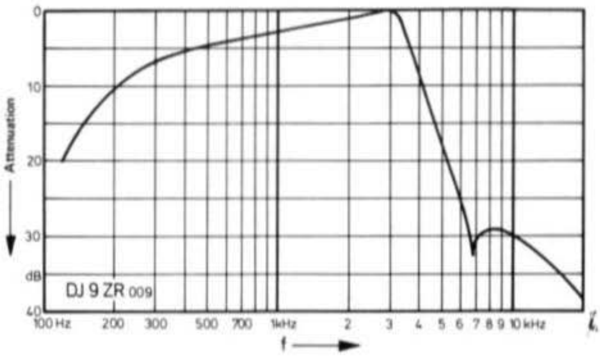
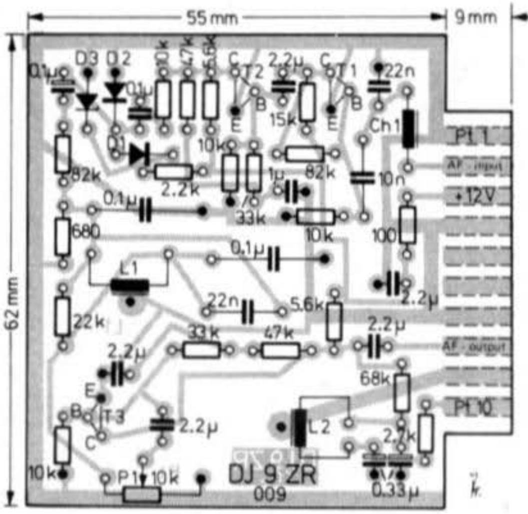
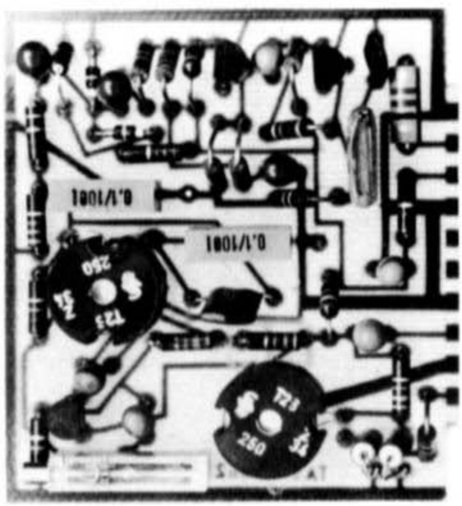


Fig. 2

If the module is to be used for phase modulation or as a clipper for an SSB transmitter, a low-pass link comprising a series resistor of 1 kΩ and a parallel capacitor of 0.68 μF connected to the output of the low-pass filter instead of P 1 will ensure that the linear frequency response is restored. The modulator is accommodated on the printed circuit board DJ 9 ZR 009 with the dimensions 55 mm x 62 mm. The board is designed as a plug-in card for a 20 pole Amphenol connector of which only one of the two rows of contacts is used. If another form of connection is required, the connections can be soldered to the appropriate point on the contact strip. The photograph given in Figure 4 shows the components and arrangement of the PC-board.





L 1: 300 turns of 0.1 mm dia. (38 AWG) enamelled copper wire
L 2: 270 turns as L 1 both wound on Siferrit-potted cores (14 x 8 mm)
Material: T 26, AL = 250; without holder, directly glued to the PC-board.
Ch 1: 22 μH (or similar value), miniature ferrite choke (Delevan 1025-52)
T 1 to T 3: BC 109 C, BC 184 C, 2 N 3707 or similar

D 1, D 2: 1 N 914 or similar D 3: 2.7 V zener diode

All capacitors of more than 0.1 μF are miniature tantal electrolytics. Capacitors C 2, C 7, C 8 and C 9 are plastic-foil types. All resistors are 7.5 mm.

A TRANSISTORIZED POWER AMPLIFIER FOR 2 METRES USING THE 2 N 3632

- Concluding second Part -

by H. J. Brandt, DJ 1 ZB

5. A PRACTICAL AMPLIFIER
5.1. A TWO-STAGE POWER AMPLIFIER

Figure 7 illustrates a 2-stage power amplifier for 2 metres that has been dimensioned according to the information given in Part I of this article. At an operating voltage of 28 V, this power amplifier will provide an output power of 15 W. The required input power level is 450 mW.

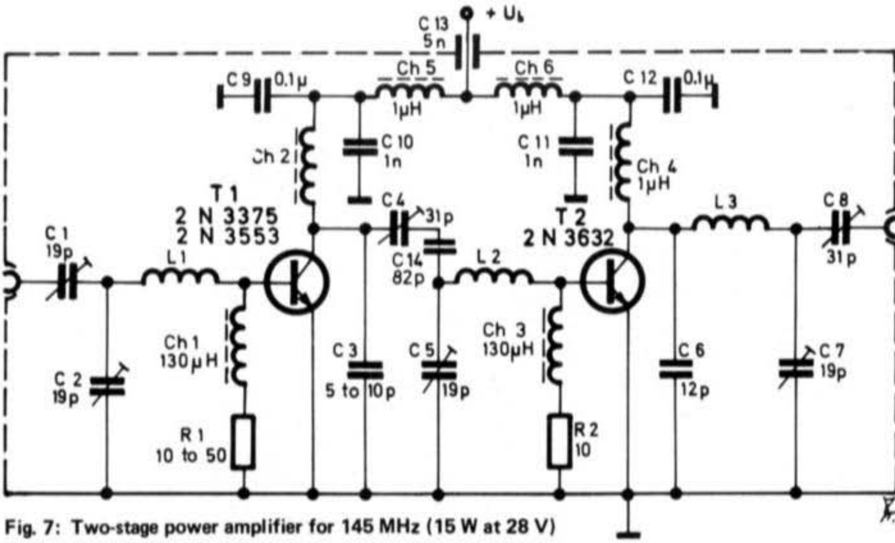


Fig. 7: Two-stage power amplifier for 145 MHz (15 W at 28 V)

The power amplifier is equipped with a transistor 2 N 3632 in the output stage. Several different types are suitable for the driver transistor, however, the most suitable with regard to power, simple mounting and cooling is probably a 2 N 3375 transistor. Sufficient drive would also be provided by a transistor type 2 N 3553. Type 40 290 can also be used if AM is to be operated at an operating voltage of 12 V. If the amplifier is only to be used with amplitude modulation, TO-5 cooling fins (Wakefield Engineering Type NF 207 or 209, or similar) will be sufficient for cooling. With continuous CW or FM operation, a better cooling will be required, especially when higher ambient temperatures are to be expected. Beryllium or aluminum-oxide washers are available for TO-5 cases with which the heat can be transferred from the casing to the chassis. Of course, there are a number of different types of heat sinks that can be used. The author used an anodised TO-5 heat sink type 1101 A for screw-mounting manufactured by Jermyn. The use of this heat sink causes an additional capacitance of nearly 20 pF between collector and ground, which the described circuit is just able to handle. Of course, the additional capacitance C/3 will then be deleted. In addition to this, it is advisable for inductance L 2 to be only wound with 2 turns.

Both transistors are fed with a low bias voltage via the resistors in the base current circuit (R 1 or R 2). The value is selected so that a reduction of the gain can just be noticed. The collector currents are filtered twice and are bypassed with a capacitor for VHF and for low frequencies (C 9 - C 12). Approximately 1.5 μF represent the upper limit for the total bypass capacitance if this circuit is to be modulated since the modulator works into an impedance of only 30 Ω.

5.2. CONSTRUCTION

With power amplifier stages of this power category, the required cooling virtually determines the construction. The author's prototype was dimensioned for continuous operation and is built up on a heat sink type KS 97-150-A of Austerlitz, Nuremberg. A small aluminum casing (Teko model 4 A), in which the actual circuit is accommodated, is mounted on this heat sink. The case is provided to ensure that the circuit is adequately screened. Figure 8 shows a photograph of the author's prototype.

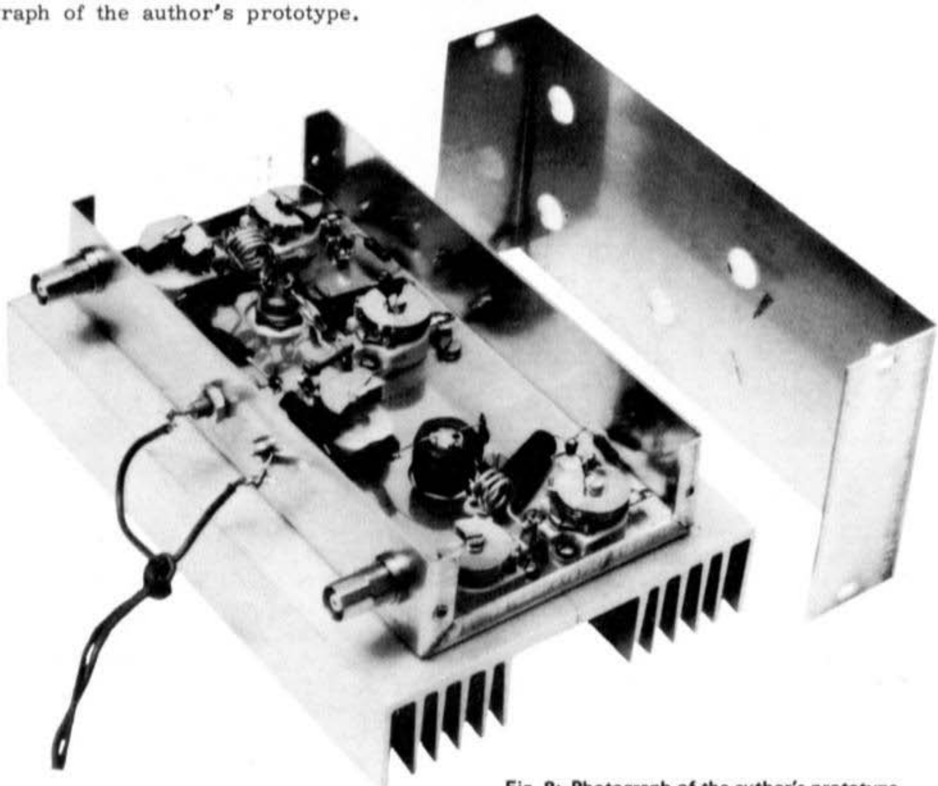


Fig. 8: Photograph of the author's prototype

Heat-conducting paste (e.g. type 120 Thermal Joint Compound of Wakefield Engineering) is to be used on mounting the transistors in order to ensure a good transfer of heat to the heat sink. Conventional wiring has been used for this amplifier since a large amount of room is available for the relatively few components due to the large heat sink and casing. In addition to this, the large ground surface is important for correct operation of the circuit at such high power levels.

If the amplifier is only to be used for amplitude modulation, the heat to be dissipated will be considerably lower and it would be possible to reduce the size of the heat sink. However, this is not to be recommended since it could have an adverse effect on the life-expectancy of the transistors. The temperature of the relatively small transistor systems will follow the power variations caused by the modulation up to frequencies of approximately 20 kHz. This continuous heating and cooling places a certain amount of stress on the system and contacts which increases with the degree of the temperature fluctuations. For this reason, an adequate cooling of the expensive power amplifier transistors is recommended even in the AM-mode.

5.2.1. SPECIAL COMPONENTS FOR THE CIRCUIT GIVEN IN FIG. 7

T 1: 2 N 3375 or 2 N 3553 (RCA); T 2: 2 N 3632 (RCA)

All inductances are wound from 1 mm diameter (18 AWG) silver-plated copper wire wound on a 6 mm diameter former, self-supporting, turn spacing 1 mm.

L 1: 4 turns; L 2: 3 turns; L 3: 4 turns.

Ch 1, Ch 3: approx. 130 μH, ferrite wideband chokes (Siemens type B 82501-A-C 27, Delevan type 1025-70)

Ch 2, Ch 4 - Ch 6: approx. 1 μH ferrite wideband choke (Siemens type B 82501-A-C 1, Delevan type 1025-20 or -24)

C 1, C 2, C 5, C 7: 2.0 - 19 pF air-spaced trimmer
C 4, C 8 : 2.5 - 30 pF air-spaced trimmer

Heat sink for TO 5 and TO 39 cases with aluminum oxide disc for insulated and low-capacitance mounting. Heat resistance approx. 26 ^{o}C/W (Souriau type TxP 0503 t),

or Beryllium-oxide washers for TO 5 and TO 39 case,

or Aluminum-oxide washers for TO 5- and TO 39-cases. Heat resistance approximately 1.5 ^{o}C/W.

5.3. ALIGNMENT AND NOTES TO THE OPERATION OF THIS AMPLIFIER

The amplifier is best aligned with the aid of a low-pass or band-pass filter feeding into an artificial or well-matched antenna. A standing-wave meter is placed between the low-pass filter and the antenna (forward power). An operating voltage of 28 V is now connected and the amplifier fed with a 145 MHz signal. The input matching of both transistors is firstly aligned for maximum current and the collector circuit of the final amplifier for maximum indication on the SWR-meter. Since the adjustments have a certain interaction on another, all trimmers should be aligned finally for maximum indication on the SWR-meter. At full power, the collector current of the power amplifier stage amounts to approximately 800 mA; the current of the driver stage is approximately 150 mA. The RF-power will be approximately 1 W down at the band limits. If the power amplifier is adjusted in this manner and operated with a parallel capacitance of 12 pF, it will be able to withstand an open or short-circuit of the antenna or any detuning of the collector circuit, even at full drive. After this alignment at 28 V, the amplifier is completely ready for operation in the CW, FM or AM modes.

Such a power amplifier should not be directly connected to the antenna but via a low-pass filter that attenuates the harmonics by at least 30 dB. Of course, a bandpass filter is even better since it will suppress any lower frequencies that are not sufficiently suppressed in the exciter. According to the regulations valid in Germany, the harmonic suppression of a 25 W transmitter at frequencies above 30 MHz should be at least 60 dB down on the carrier. It cannot be expected that the harmonic suppression at the collector of the final amplifier is any greater than 25 dB, which means that additional filtering is required.

6. MODULATOR CIRCUITS

Even when only amplitude modulation is to be dealt with here, this does not mean that other, popular modes are to be disregarded. It is only that amplitude modulation must be considered during the dimensioning of power amplifier stages.

In conjunction with transistorized transmitters, two modulating methods are mainly used: Collector voltage and collector current modulation. With this type of modulation, voltage and current are both affected, which means that it is immaterial which magnitude is directly influenced. However, it is impossible to obtain a modulation depth of 100% by merely modulating the output transistor since the collector-base diode of the transistor will allow unmodulated drive-power to pass directly from the base into the collector circuit at very low collector voltage. This is the reason why the driver stage is usually modulated at the same time. Even in this case, the carrier power at half the operating voltage is mostly somewhat greater than 1/4 of the peak power level, especially when the transistors are to be driven to their maximum limits. This is dependent on the reduction of the current gain at high currents. The simultaneous modulation of the output stage of the exciter could represent an improvement, however, it is important that a truly positive modulation is obtained or the modulation depth limited in some other manner because otherwise high distortion is to be expected (3). In the circuit in question (Fig. 7), it has been found that it is sufficient for the driver and power amplifier to be modulated.

In this context, the emitter current modulation described in (12) should be mentioned. Since the DC-voltage between collector and base is maintained constant with a Zener diode, only the collector current will vary in the described circuit. This means that similar conditions are obtained as with grid-voltage modulation of tubes. The efficiency of the power output stage is only optimum at the peak power level; it is only half as great at the carrier level. An advantage is that the power dissipation across the modulation transistor is very low. The modulation depth of 100% that can be obtained in this manner is due to the fact that the collector voltage remains constant which means that it is not possible to transfer power via the collector-base diode of the transistor. The difficulty with this modulating method is in the correct bypassing of the emitter of the RF-transistor which, as already explained, becomes even more critical on increasing the power.

All amplitude modulator circuits can be divided into two groups. The first group works at the operating voltage (28 V) necessary for the peak power level and with a series-transistor in the collector current circuit (without transformer). In this manner, it is easy for transmitters to be constructed for several operating modes where the optimum power is always available: 12 - 15 W in FM and CW or 3 - 4 W unmodulated carrier power in AM.

The second group operates at half the operating voltage required for the peak power level (12 to 14 V) and require a "magnetic storage" in order to bring the collector voltage to twice the value in the modulation peaks. With respect to the AM-signal, both methods are equally good. If such a circuit is, on the other hand, also to be used for CW or FM, it will not be possible for a higher power to be operated than was available for the unmodulated carrier in the AM-mode if the alignment is not to be altered. By only altering the matching circuits (less inductivity, more capacitance), it is possible to obtain an output power of 6 to 8 W with the transistor 2 N 3632 at an operating voltage of 12 V to 14 V if transistor types 2 N 3375 or 40 281 (RCA) are used as driver.

6.1. MODULATOR CIRCUITS FOR 28 V

Figure 9 shows a simple modulator output stage without transformer. The NPN-AF power transistor T 7 is connected in the collector current circuit of the RF transistors to be modulated. The PNP transistor T 6 is used as AF driver so that the peak power level of the NPN transistor is obtained. Of course, one could provide an arrangement with a socalled bootstrap capacitor since the same problems occur as with transformer less push-pull audio amplifiers. The given circuit has the advantage that transistor T 7 can be virtually driven into saturation for test purposes in the DC-mode. The total resistance in the base circuit of T 6 is dimensioned for symmetrical modulation so that half the operating voltage is present at the emitter of transistor T 7. The resistance values are dependent on the current gain of both transistors and may have to be varied. The audio preamplifier must provide a drive voltage of approx. 500 mV.

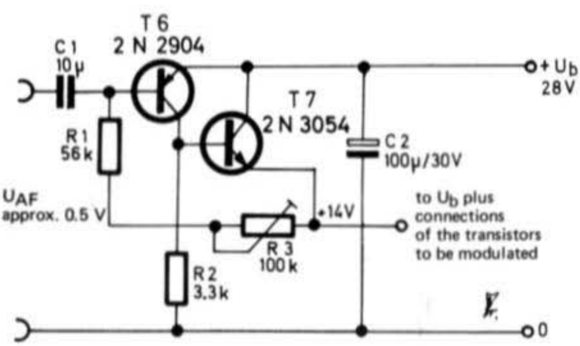


Fig. 9: Transformerless modulator

As can be seen in Figure 10, a PNP power transistor can be used for current modulation. A further PNP transistor (T 8) is connected before the modulator (T 9) to increase the input impedance of the circuit. The carrier level is adjusted by the base current of transistor T 8 so that 14 V is again present between the collector of transistor T 9 and zero. The large feedback (C 3/R 4/ C 2) improves the linearity of the modulation. However, it could lead to AF oscillation if the gain in excess of the modulation frequencies were not sufficiently suppressed with the aid of capacitor C 3.

The demands made on transistors T 7 and T 9 are the same when the polarity is not taken into consideration. The breakdown voltage U_{CE_0} must be at least as high as the operating voltage (28 V), and the permissible collector current must at least be as high as the peak collector current of the modulated RF-transistors (approx. 1 A). The power dissipation in operation is obtained from half the operating voltage and half the current (approx. 7 W).

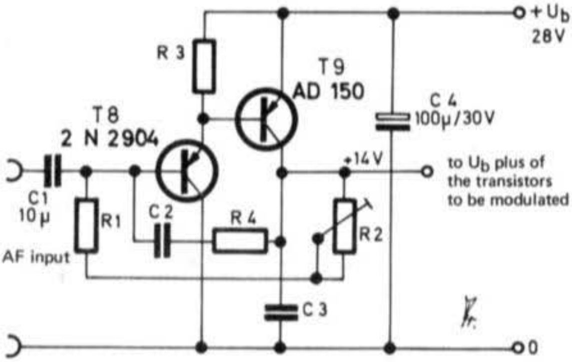


Fig. 10: Transformerless collector-current modulator for 28 V/1 A

6.1.1. SEMICONDUCTORS FOR THE CIRCUITS GIVEN IN FIG. 9 AND 10

T 6, T 8: 2 N 2904, 2 N 2905 (TI) or BC 160 (Siemens, ITT-Intermetall)
T 7: 2 N 3054 (RCA) or 40 250 (RCA)
T 9: AD 150 (Siemens, AEG-Telefunken).

6.2. MODULATOR CIRCUIT FOR 12 TO 14 V

The second group of modulator circuits usually operates with a transformer and a push-pull class B amplifier. This is, without doubt, the most economic manner of generating the modulation power, however, the dimensioning or purchase of a suitable transformer can cause some problems.

If one can accept the lower efficiency of a class A amplifier, it is possible for Heising modulation to also be used in conjunction with transistorized transmitters. In tube technology, it has been said that a modulation depth of 100% cannot bo obtained using this type of modulator, however, this is not valid for transistors since the residual voltage of RF power transistors is greater than that of AF transistors. In addition to this, at least two stages are always modulated. A conventional power choke with an inductivity of approximately 40 to 100 mH will be sufficient as modulator choke since the RF stages of the described transmitter only represent a load of 30 Ω for the modulator. It is important that the current rating is sufficiently great and that the DC-current resistance is as low as possible.

Figure 11 shows a suitable circuit. The NPN audio power transistor T 11 is driven by transistor T 10, which is provided to increase the input impedance. The current flowing via transistor T 11 is adjusted with the aid of potentiometer R 3 via the base current of T 10 to half the peak power value of the modulated transistor (approx. 600 mA). A feedback link (R 5 and C 2) ensures a suitable audio frequency response, and capacitor C 3 connected in parallel to the Heising choke Ch 1 neutralizes any tendency to oscillation via this feedback link.

The power Zener diode D 3 plays an important part. Namely, voltage peaks can be generated across the choke that would destroy the RF transistors. The Zener diode limits these to a value that will not endanger the transistors. The second power diode D 4 in series with the modulator output improves the modulation depth in the region of the zero carrier level. Without this diode (output I) it is not possible to bring the RF power quite to zero in the modulation peaks in contrast to the current modulation modes. However, it will go zero when output II is used since the diode blocks a current in the opposite direction which attempts to flow via the collector-base diode. Since the diode causes approximately 1 V voltage drop at peak power, and the problem is only caused by the driver stage which operates with a constant drive, it is more favourable for output I to be used for the power amplifier stage and output II to be used for the driver. Naturally, this means that a separate connection must be provided for the collector voltage of the driver in the circuit of the power output transistor (Fig. 7). This is also favourable for test purposes.

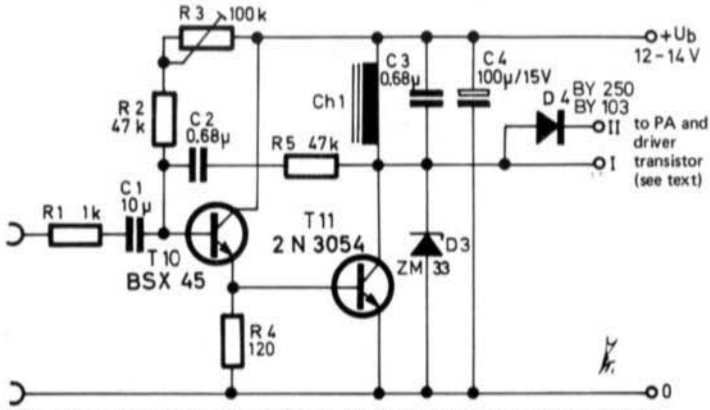


Fig. 11: Collector-voltage modulation with Heising-choke for 12 V operation

The circuit given in Figure 11 also requires approximately 500 mV drive for full modulation. The dropper resistor T 1 can be deleted in practice if the preamplifier has a similar or higher impedance. At a lower resistance, the feedback will not operate correctly. The demands on transistor T 11 with regard to voltage, current and power dissipation are approximately the same as for current modulation. In operation, the current flow of the modulator stage and the modulated RF transistors is slightly higher than the peak power collector current of the modulated stages themselves (approx. 1.1 A). As long as no overmodulation occurs, the current will remain constant. With collector voltage modulation, the final alignment of the RF power amplifier stage and the modulator circuit are most favourably made by observing the envelope of the modulated signal. This method is especially recommended if twice the operating voltage is not available for the peak power alignment of the RF stages. In this manner, the most favourable current for the modulator transistor T 11 will be found quickly.

6.2.1. SPECIAL COMPONENTS FOR THE CIRCUIT GIVEN IN FIG. 11

T 10: BSX 45 (Siemens) or BC 140 (Siemens, ITT-Intermetall)
T 11: 2 N 3054 (RCA) or 40 250 (RCA)
D 3: (33 v/1 W), BZY 92/C 33, ZD 33; D 4: BY 250, BY 103, 1 N 4004
Ch 1: 0.04 to 0.1 H, power choke.

7. NOTES

7.1. MOBILE OPERATION

The modulator circuits for 12 to 14 V are suitable for mobile operation since this is the voltage that is most used in vehicles. This means that no DC-DC converter will be required. In this context, a fact should be pointed out which is relatively unknown, namely, that certain electromagnetic consumers in the vehicle, e.g. the starter relay or the horn, generate peak voltages of up to 100 V which are then present on the power supply of the vehicle. These voltage peaks must not be allowed to reach any transistorized equipment since they would instantaneously destroy the power transistors. Unfortunately, the author only knows of these problems indirectly and it would be very favourable if a specialist in car and mobile radio installations were able to publish details regarding such effects.

The required measures to protect the transistor circuits are often dependent on the type of circuit and the electro-magnetic equipment of the car and also on which point of the car circuit the voltage for the transistorized equipment is taken. The most favourable position is to tap off the voltage directly at the battery using two sufficiently thick wires since the capacitance of the battery will reduce the voltage peaks. For car radios that are connected via the ignition key, a choke is usually provided in the supply current connection which provides the required protection.

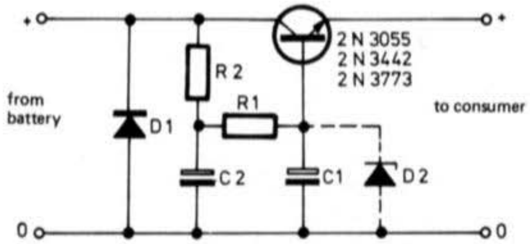


Fig. 12: Protection circuit against overvoltages on the vehicle network

Electronic protection circuits can be dimensioned effectively when the amplitude and the waveform of the voltage peaks is known by measuring them on an oscilloscope. The most simple form of protection is a diode with a breakdown voltage of approximately 15 V. The peak current surge that can be passed through the diode is limited by its data. A somewhat more extensive circuit is given in Figure 12. In this case, a series transistor is used whose breakdown voltage must be higher than the maximum peak voltage that appears. This series transistor is placed in series with the plus connection. The base is bypassed with a large capacitance C 1 to ground. The base current is adjusted with the resistor combination R 1 + R 2 so that the transistor conducts. This is to ensure that only a very small voltage drop occurs in normal operation. If a voltage peak appears on the network of the vehicle, this will firstly cause capacitor C 1 to be charged before the voltage at the consumer can also increase. If the time constant (R 1 + R 2)× C 1 is far greater than the pulse duration, the pulse will not be present at the output. In order to improve the filter effect of the base circuit, the base dropper resistor can be split up and operate with two or more RC-links. A diode connected in parallel to capacitor C 1 could also avoid an increase of the capacitor voltage above a certain value if any sequence of pulses is obtained. Any voltage peaks of opposite polarity can be shorted with the aid of a sufficiently well dimensioned diode D 1.

7.2. TRANSISTORIZED SSB-POWER AMPLIFIERS

In the discussion concluding the lecture, it was discussed what power can be expected from the transistor 2 N 3632 in the SSB-mode. The author has not tried this mode himself but has heard from a reliable source that approximately 5 W PEP are to be expected at a sufficiently acceptable intermodulation ratio (approx. 25 dB).

This may seem disappointing. However, the transistor is not able to directly compete with tubes as a large signal linear amplifier where the efficiency also plays its part. Whereas modern SSB transmitting tubes designed with the aid of computers exhibit intermodulation ratios of 40 to 50 dB at the nominal power (2), it is even difficult with transistors to obtain ratios of 30 dB or more if reliable operation is to be maintained.

With transistors for class C, to which the first Overlay transistor types belong, it is only important for the peak current and the transit frequency to be as high as possible. With SSB transistors, on the other hand, it is necessary for the current gain and transit frequency to remain relatively constant over the whole drive range. This is obtained, for instance, with integrated emitter resistors which also improve the breakdown characteristics (13). Such transistors allow 2/3 the CW power to be expected as SSB-PEP-power with sufficiently good linearity (8).

In contrast to tubes, additional distortions are generated in transistors due to the voltage dependent capacitances. The linearity therefore decreases on increasing the frequency. Feedback links are no longer able to help since the phase shift between input and output voltage will be too great. In the RF-technology, it is not possible for a great ratio to be maintained between the transit frequency and the highest frequency of operation, as is the case with Hi-Fi amplifiers. The linearity is at its best in the common-base circuit as long as the feedback effect is negligible.

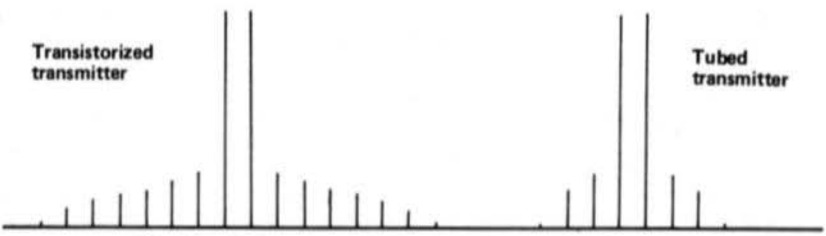


Fig. 13: SSB two-tone test signal with intermodulation spectrum

A special feature of transistor SSB amplifiers is the unusually wide spectrum of intermodulation products (Fig. 13). Whereas it is only third and fifth order products that are suppressed less than 50 dB with most tubed transmitters, all products up to the eleventh or thirteenth order remain above this limit with transistorized power amplifiers. This means that the main advantages of SSB, the low bandwidth requirements, cannot be utilized due to the wide range of distortion around the required signal. A level difference of 30 to 50 dB is very low in comparison to the dynamic range of modern receivers that are able to handle signals between 0.1 μV and 100 mV (120 dB). After considering these facts, it is no wonder that the leading manufacturers only offer very few transistors with guaranteed SSB specifications for the shortwave range and even less for the VHF range.

8. AUXILIARY CIRCUITS

8.1. POWER OSCILLATOR

The exciter stages for the power amplifier shown in Figure 7 were still not developed at the time of the lecture. The author therefore used the power oscillator given in Figure 14 for driving the power amplifier experimentally. The transistor type 2 N 3553 or better still the inexpensive 2 N 5189 (RCA) can be used. This transistor can also be operated at 28 V in the 2 m band but attention must be paid that 150 mA is not exceeded due to the liability of breakdown. This means that it is suitable for driving transistor types 2 N 3553 or 2 N 3375.

The oscillator circuit used oscillates well and can be compared to the Hartley circuit popular in the Twenties and Thirties. It is very suitable for experiments where the power and not the frequency stability are of primary importance. The power can be varied within wide limits by altering the operating voltage.

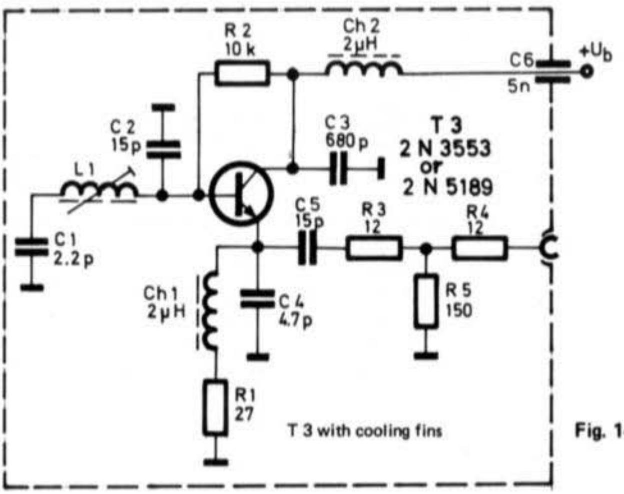


Fig. 14: Transistorized power oscillator for experiments at VHF and UHF

Technically speaking, the circuit represents a modified Clapp oscillator. In the VHF/UHF range, the transistor capacitances are used in the resonant circuits so that external capacitors are not required. In this circuit, however, 15 pF is connected in parallel with the collector-base path in order to limit the dependence of the frequency on the operating voltage (with C 2 = 15 pF approx. 144.5 MHz to 145.5 MHz between 14 and 28 V). The circuit comprising L 1 and C 1 mainly determines the frequency.

The signal is taken from the emitter via a series capacitor into an ohmic load. Since the power amplifier possesses no real input impedance without drive, the oscillator will not commence oscillation in spite of the 3 dB attenuator (R 3, R 4 and R 5). This is the reason why the 4.7 pF is required from the emitter to ground (C 5).

8.1.1. SPECIAL COMPONENTS FOR FIG. 14

T 3: 2 N 3553 or 2 N 5189 (RCA)
L 1: 220 - 315 nH, ceramic inductance
Ch 1, Ch 2: approx. 2 µH, ferrite wideband choke (Delevan)

8.2. CONVERTER FOR OBSERVING THE ENVELOPE OF THE TRANSMITTED SIGNAL

The direct observation of the modulated signal from a 2 m transmitter is only possible with very expensive oscilloscopes that have a bandwidth of at least 150 MHz. For this reason, it is very useful to have a converter which allows the envelope curve to be observed on an oscilloscope with a bandwidth of several MHz. Such an oscilloscope, as is used for television repairs, etc., is more readily available. The circuit of such a converter is shown in Figure 15. It uses the same oscillator circuit as was described in the previous section, which in this case oscillates approximately 2 to 4 MHz above or below the 2 m band. The signal from the modulated transmitter and the oscillator are mixed in a normal RF diode. A low-pass filter comprising Ch 3, Ch 4 and C 5 is used to select the differential frequency which is then fed to the oscilloscope. It can be seen that the frequency stability of the oscillator is not of great importance.

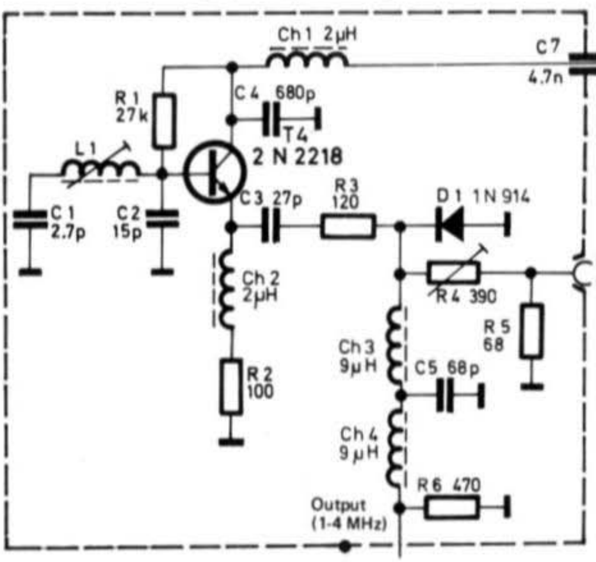


Fig. 15: Converter for observing the envelope of VHF signals with TV service oscilloscopes

However, prerequisite of a good indication is that the converter is not overloaded. The DC voltage measured at the output should only be from the oscillator and should not increase noticeably on injecting the signal to be examined. The permissible drive level can, as long as resistor R 5 is not overloaded (1/4 to 1/2 W), be adjusted with resistor R 4. The value of 390 Ω is for a fixed resistor which was tried out for the lecture. The minimum value is approximately 120 Ω since the diode requires an impressed current via a series resistor. At higher powers, the coupling to the transmitter must be looser. During the lecture a 60 Ω load resistor was used whose attenuated output (1/50 of the input power) was fed to the converter.

8.2.1. SPECIAL COMPONENTS FOR THE CIRCUIT SHOWN IN FIG. 15

T 4: 2 N 2218 D 1: 1 N 914
L 1: 220 - 315 nH, ceramic inductance
Ch 1, Ch 2: approx. 2 µH, ferrite wideband choke (Delevan)
Ch 3, Ch 4: 9 µH, ferrite wideband choke (Delevan)

8.3. AUDIO GENERATOR

Figure 16 shows a simple RC generator which is suitable for providing the sinusoidal modulation required for the envelope test. The waveform of the collector voltage should be checked before operation with the aid of an oscilloscope. The feedback resistor R 6 in the emitter circuit allows the maximum amplitude to be adjusted without limiting effects. Capacitor C 4 suppresses any tendency to RF oscillation. With the exception of this, the circuit does not provide any special features.

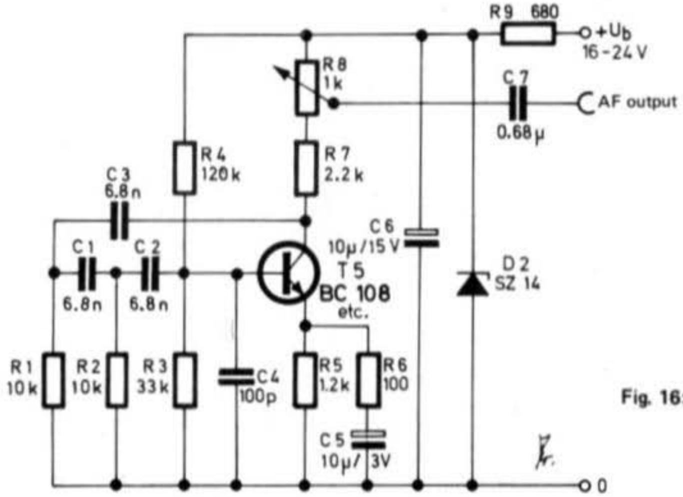


Fig. 16: Simple audio generator for experiments

8.3.1. SEMICONDUCTORS FOR THE CIRCUIT GIVEN IN FIG. 16

T 5: BC 108, BC 183 or 2 N 2926
D 2: SZ 14, BZY 85/D 15 or similar Zener diode for approx. 14 V.

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10. EDITORIAL NOTE

A practical construction of a transistorized linear amplifier using the transistor 2 N 3632 on a PC-board (DC 8 ZX 002) will be described in one of the next editions of VHF COMMUNICATIONS.

FREQUENCY MULTIPLICATION WITH HIGH SPURIOUS SIGNAL REJECTION

by D. E. Schmitzer, DJ 4 BG

With the exception of special circuits, such as varactor triplers, etc., overdriven amplifier stages equipped with a tube or transistor are mainly used for frequency multiplication. A sinusoidal input signal is distorted due to the overloading, which means that a high harmonic content is present. It is then necessary for the required harmonic to be filtered out; all other harmonics and the fundamental wave represent spurious signals and must be suppressed. The required filtering can be kept low when most of the unwanted spurious signals are not generated in the first place.

Symmetrical circuits offer distinct advantages. Such circuits are push-pull circuits for frequency tripling and the push-push circuits for frequency doubling. The push-pull circuit suppresses all even harmonics (2f, 4f, 6f, etc.), whereas the push-push circuit suppresses all odd harmonics and the fundamental wave (1f, 3f, 5f, etc.).

In spite of the clear advantages, these two circuits are seldom used. The reason for this may be that two transistors are required instead of one. According to (1), a push-push frequency doubler circuit can be built up simply and economically using an integrated circuit CA 3028 A. With the exception of the input and output circuit and two bypass capacitors, all the required components are already provided in the integrated circuit.

1. PUSH-PUSH FREQUENCY DOUBLER WITH THE CA 3028 A

Figure 1 shows the circuit of a push-push frequency doubler equipped with the integrated circuit CA 3028 A. The transistor systems T 2 and T 3 operate in the push-push mode, e.g. the input side is in push-pull and the output side in push-push (parallel). If both transistors are provided with enough bias voltage that they just allow collector current to flow (class B), the output circuit will receive a current pulse in both the positive and negative halfwave of the input voltage. The pulses originate alternately from the first and second transistor. This means that the output circuit is fed with a pulse on two occasions during each cycle so that twice the frequency results.

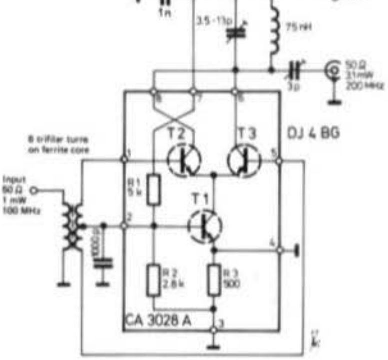

Fig. 1: Circuit diagram of the push-push doubler

The correct bias voltage is adjusted by transistor T 1 in a special circuit: It is saturated due to the fact that its emitter resistor R 3 is short-circuited and its base receives an impressed current via resistor R 1. The base-emitter diode of transistor T 1 represents a reference voltage source which provides transistors T 2 and T 3 with a bias voltage that is well temperature compensated. Since the emitters of the two frequency doubler transistors are biased simultaneously by the residual voltage of T 1, a lower collector current will be adjusted than would be present with R 1 alone. The result is virtually class B operation as is necessary for the planned application.

Some care must be paid to the input circuit: A tuned push-pull resonant circuit with a coupling winding or a wideband transformer can be used. The latter offers a good, stable balance, especially over long periods of time, if the two halves of the secondary winding are made in a bifilar manner, e.g. by winding two wires in parallel on a ferrite rod or toroid core and connecting them in series. The primary winding can be wound at the same time in the form of a third wire. The same winding principle is used in ring mixers.

2. CHARACTERISTICS

The circuit given in Figure 1 only requires a low drive and is very linear. In the original paper (1) it was used to double a frequency of 100 MHz to 200 MHz. The drive level is approximately 1 mW, from which an output power of 3.1 mW is obtained with a very low harmonic content. When one adds the selectivity of the output circuit, the suppression of all unwanted harmonics and the fundamental wave will be between 45 dB and more than 50 dB with respect to the level of the required output frequency. It is true that higher output power levels (approx. 30 mW) are obtained at higher drive levels (up to 10 mW), however, the distortion will increase and the harmonic content will increase to approx. 10%. Since the CA 3028 A is not really designed for such high frequencies, far better results are to be expected at lower frequencies: Higher output powers with lower drive levels together with low harmonic content. The experiments made up till now by the author confirm this. Due to the integrated construction, the frequency doubler circuit is not affected greatly by temperature which means that the suppression values are maintained over a very wide temperature range.

3. OTHER INTEGRATED CIRCUITS THAT CAN BE USED

Besides the CA 3028 A, any other integrated circuits with the same transistor configuration can be used. Integrated circuit types CA 3026 and CA 3054 possess two such stages, unfortunately without resistors, so that a resistor of approx. 4.7 kΩ must be provided externally to replace R 1. This means that it is possible to build up two frequency doubler stages using a single integrated circuit. If an integrated circuit is used whose transistors exhibit a high transit frequency, it is possible to frequency double up to higher output frequencies. An example of this is the CA 3049 which is also provided with two complete transistor systems in the described form so that it is possible to double, for instance, from 108 to 216 MHz and further from 216 to 432 MHz. Since the individual transistors of the CA 3049 are stated to have a transit frequency of typical 1300 MHz, it should be possible for several milliwatt to be generated in the 24 cm band, which would be sufficient for the local oscillator signal in a receiver.

The integrated circuits previously mentioned, e.g. CA 3026, CA 3028 A, CA 3049 and CA 3054 are manufactured by RCA. Other manufacturers produce similar integrated circuits that are also suitable, e.g. LM 371 of National Semiconductor of MFC 8030 of Motorola.

4. GAIN CONTROL OF THE FREQUENCY DOUBLER STAGE

If a lower voltage is connected to point 7 of the CA 3028 A, transistor T 1 will receive less base current and will be brought more and more out of the saturation range. This will mean that the collector-emitter path of this transistor will have a higher resistance and represents an increasingly greater feedback for the doubler transistors T 2 and T 3. It is possible in this manner to control, or modulate the output power of the circuit.

This means that circuits can be built up by which the fluctuations of the output level, which are virtually inavoidable when tuning over a certain frequency range, can be compensated. The often-observed phenomena where a high degree of overdrive is tolerated at the centre of the band in order to obtain sufficient drive at the band limits, is a bad habit which is then no longer state-of-the-art. Since the overdrive at the band centre is avoided in the controlled state, this results in far less harmonic and spurious radiation.

5. REFERENCES

(1) Carl Andren: Low-cost 100-to 200 MHz Doubler has 5 dB Gain, 1% Distort. Electronic Design No. 18, 1 Sep 1970, Page 84

MATERIAL PRICE LIST FOR KITS AND COMPONENTS
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FOR EQUIPMENT DESCRIBED IN VOLUME 3 (1971) OF VHF COMMUNICATIONS

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TRANSVERTERS

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Trimmer set	DL 6 MH 001	DM	21.--
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V X O			
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TRANSMITTER

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DJ 5 HD / DJ 7 JX — SYNTHESIS VFO for 144 MHz or 136 MHz Ed. 1/71

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Kit	DJ 5 HD 002/145	with above components	DM	69.90
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Kit	DL 9 FX 002	with above listed components	DM	43.15

MEASURING, AUXILIARY EQUIPMENT

DJ 4 BG 004 — CALIBRATION-SPECTRUM GENERATOR Ed. 4/71

PC-board	DJ 4 BG 004	(double coated / through contacts)	DM	39.--
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Kit	DL 8 TM 001 with above components	......	DM	43.--

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2 metre MINI-TALKY

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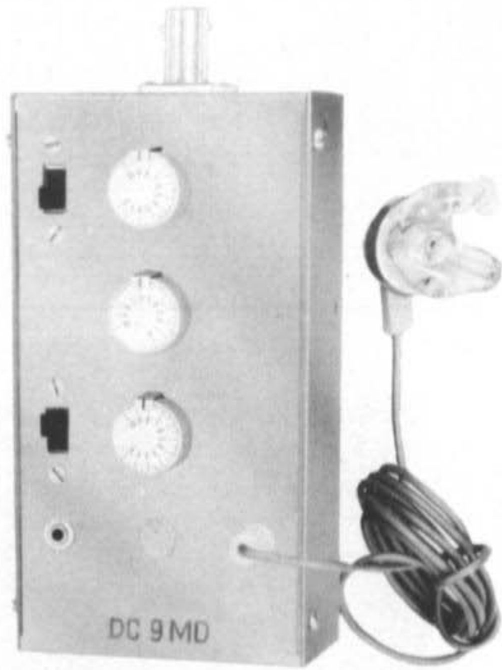
- Ideal for business and vacation trips.
- Kit available complete with all components including all resistors, capacitors, case knobs etc. See Material List for more details.



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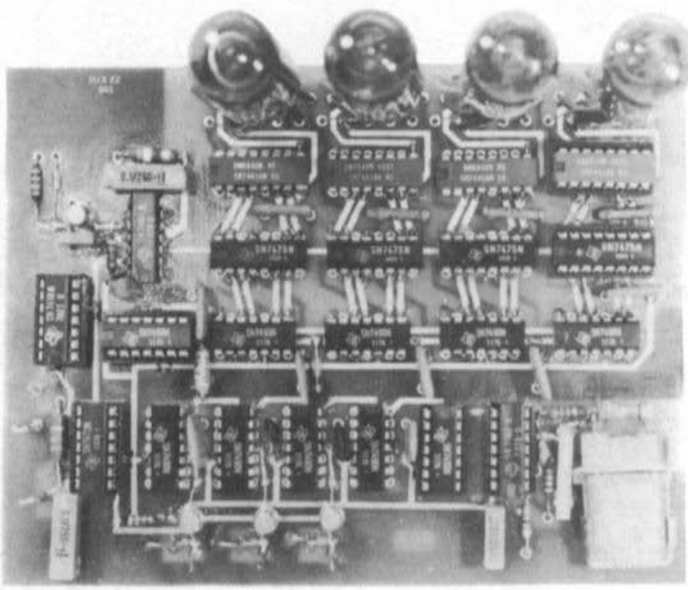

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Selectivity:	4.5 kHz (—6 dB) 12 kHz (—40 dB)
Image and spurious suppression:	60 dB
Operating voltage:	11—15 V/15—22 mA
Dimensions:	200 mm x 83 mm x 32 mm

FET 2 metre converter AC 2: DM 130,68

The matching converter to receiver AR 10. Neutralized FET input stages, push-pull FET mixer. Crystal 38.6667 MHz (HC-25/U).

Specifications:

Input frequency:	144—146 MHz	Noise factor:	1.8 dB
Output frequency:	28—30 or 26—28	Image suppression:	> 70 dB
Gain:	22 dB ± 2 dB	Operating voltage:	12—15 V/15—20 mA
Input frequency:	50 Ohm	Dimensions:	120 mm x 50 mm x 25 mm

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4-stage crystal-controlled transmitter with modulating transformer and antenna relay. Especially suitable for portable and mobile operation from 12 V. Crystals in 72 MHz range (HC-25/U). Connection available for a 24 MHz VFO.

Specifications:

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Output power (unmodulated carrier):	2.2 W at 12 W	Dimensions:	150 mm x 46 mm x 32 mm

Modulator and AF-amplifier AA3: DM 82,50

4-stage modulator and audio amplifier matching the transmitter AT 210 or receiver AR 10. Built-in relay switches the modulator so that it can be used as audio output for the receiver. Output power more than sufficient for mobile operation.

Specifications:

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Output impedance:	3 Ohm
Sensitivity:	2 mV for 2.8 W
Frequency response:	300—3000 Hz (—3 dB)
Distortion:	< 2 % at 2.8 W/1000 Hz
Dimensions:	150 mm x 46 mm x 32 mm

The above mentioned modules can be combined to form a complete twometre transceiver. They are available ex stock in Erlangen.