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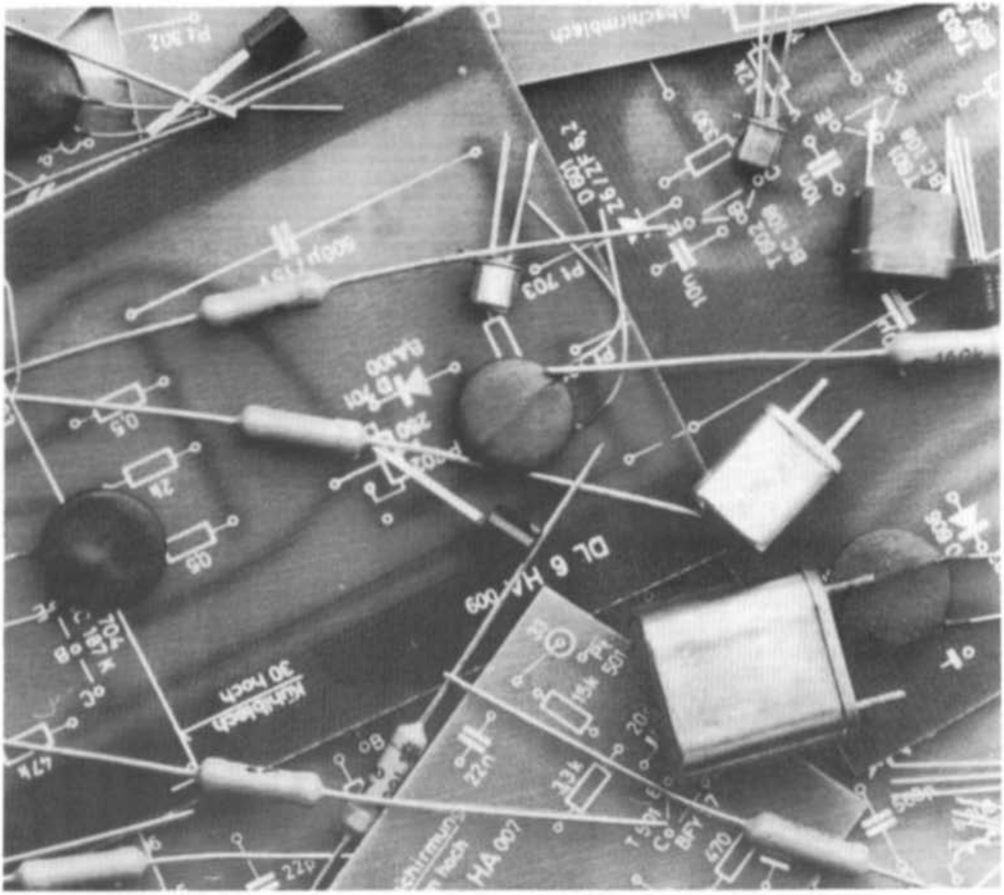

CONTENTS:

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A SSB TRANSCEIVER WITH SILICON TRANSISTOR COMPLEMENT

PART 1: THE 144 MHz CONVERTER WITH DUAL-GATE MOSFET MIXER

by G. Laufs, DL 6 HA

INTRODUCTION

A complete SSB transceiver with a fully silicon complement is to be described; it is equipped with field effect transistors in the critical stages. An intermediate frequency of 9 MHz is used and the selectivity is gained by utilizing a crystal filter. The VFO oscillates in the range of 5 to 5.5 MHz which, when mixed with the 9 MHz IF, results in a frequency range of 14.0 to 14.5 MHz. This is transposed in a second, crystal controlled mixer to the required output frequency in the two metre band.

The system allows a number of modifications to be made: Since the 20 metre amateur band is covered during the frequency synthesizing process, it will be possible after suitable linear amplification of the transmit signal, to also operate on this band. The modification for 80 metres - where the difference frequency and not the sum frequency of VFO and IF is selected - should not be difficult. Using additional mixers, it is possible to cover all shortwave bands.

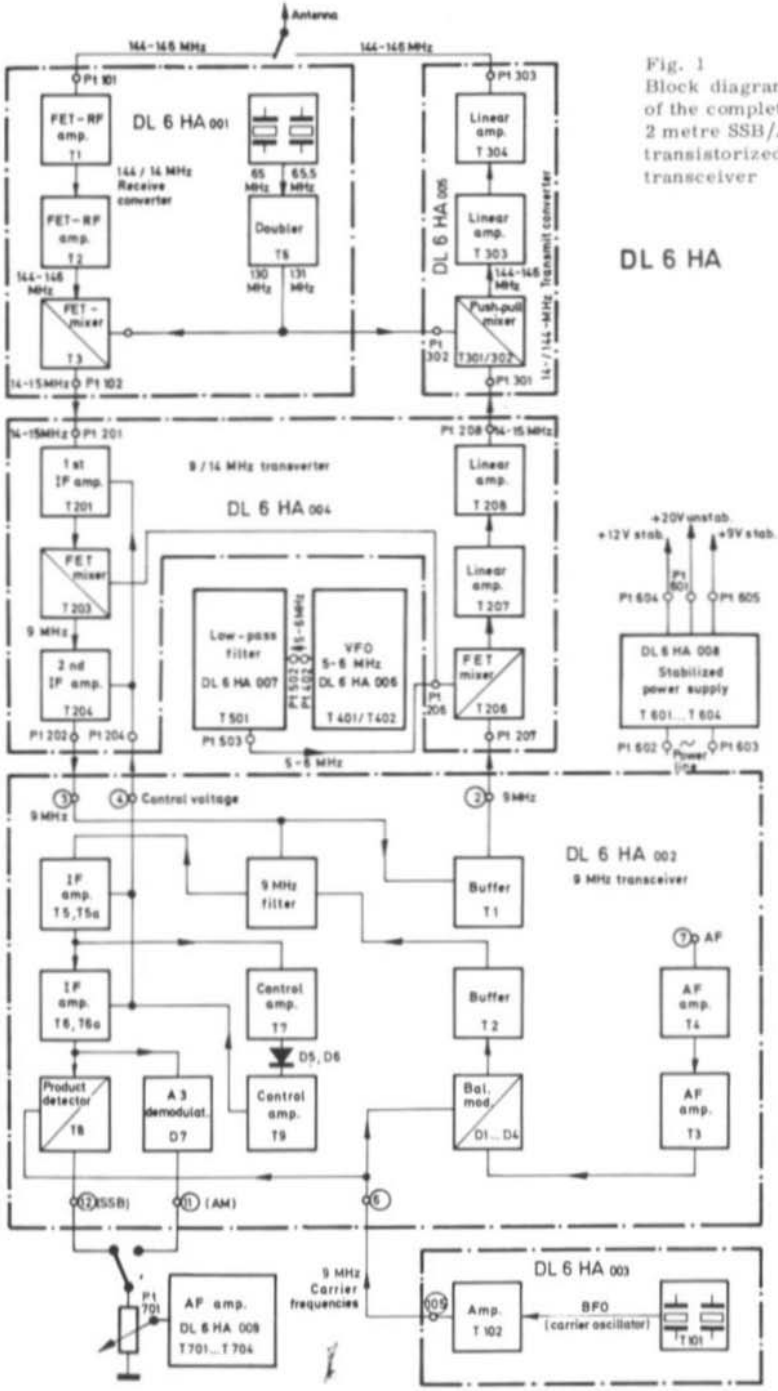
However, our description concentrates itself to the VHF transceiver system and only includes the 14 MHz range as a further intermediate frequency.

Figure 1 shows a block diagram of the complete transceiver and allows the principle of operation as well as the various sub-assemblies to be seen. Since the transceiver was also to be used for mobile operation, an operating voltage of 12 V was selected. The use of field effect transistors in the RF and mixer stages is a further feature of this transceiver.

The transceiver represents the result of a great number of experimental circuits. Special attention has been paid to produce a high degree of reproduceability of each circuit together with a simple alignment. As previously mentioned, the sub-assemblies are just as suitable for use in shortwave transceivers as they are for the VHF station.

Fig. 1
Block diagram
of the complete
2 metre SSB/AM
transistorized
transceiver

DL 6 HA



Part I of this article deals with the VHF portion consisting of a two metre converter with junction field effect transistors in the RF amplifier stages and a MOSFET as mixer.

The description of the converter is to be made with two applications in view: Firstly as a converter with an intermediate frequency of 28 to 30 MHz and secondly as VHF portion of the complete SSB transceiver which will be described in detail in the following editions of VHF COMMUNICATIONS.

1. CONCEPTION OF THE CONVERTER

The design of this converter was commenced after reading several publications discussing the use of field effect transistors in VHF converters (1) and gaining practical experience whilst building and operating the converter described in (1). The task was to develop an efficient converter that was simple to align and whose characteristics were able to satisfy the requirements with respect to a low noise figure, a high selectivity together with a high intermodulation and cross-modulation rejection.

Special attention was paid during the development of this converter to obtain a high degree of pre-amplification. The gain of the stages previous to the mixer must be high enough to ensure that the conversion noise does not significantly add itself to the total noise figure. On the other hand, it should not be too high since a high RF voltage level at the mixer stage could cause cross and intermodulation products. In practice, pre-amplification factors of 13 dB to a maximum of 20 dB are used. The conversion noise must, of course, be kept as low as possible.

Experiments were made using the junction field effect transistors TIXM 12 and BF 245C (TIS 34) in push-push and push-pull mixers; it was found that, although the cross- and intermodulation characteristics were good, the relationship conversion noise to conversion gain was very unfavourable. This meant that a high pre-amplification would be required in order to achieve a satisfactory noise figure. It was not possible, for instance, to obtain the 4 dB noise factor given by the manufacturers of the transistor BF 245C when using an experimental circuit with two RF amplifier stages in a common gate configuration and a push-push mixer. Similar problems are mentioned in the introduction of publication (1).

The dual-gate MOSFET's TA 7150 and TA 7151 were found to be considerably more favourable. Very favourable results were obtained using these two types in multiplicative push-push mixers. Due to the higher transconductance, a considerable improvement is gained in the relationship conversion noise to conversion gain. This in turn means that less pre-amplification is required to achieve good noise figures. The low gain of the RF amplifier stages leads to an improvement of the cross- and intermodulation rejection because, in this respect, the dual-gate MOSFET's possess only slightly poorer characteristics than junction FET's.

The selection of the RF amplifier configuration and complement was also only found after extensive experiment. The result represents a compromise between a low noise figure and high stability. Firstly, two stable common gate RF amplifier stages were built-up which resulted in a noise factor of 4 dB. In comparison, a neutralized common source configuration using the same transistors offered a noise factor of even 2 dB - however the neutralization adjustment was just before the point of oscillation. Naturally, a stable operation under such conditions cannot be guaranteed because even slight variations of the impedance or standing wave ratio at the antenna input could cause the converter to break into oscillation. Since the converter was also to be used for mobile operation, such a circuit could not be considered. Experiments with a dual-gate MOSFET TA 7149 in a non-neutralized common source configuration were not successful; when aligned for stable operation, the noise factor was approximately 4 dB. It is true that the gain is somewhat higher than that of the BF 245C, but because the extra gain is not required and since the MOSFET was twice as expensive, it was not used in the final circuit.

The resulting configuration using a BF 245C in an intermediate common-base circuit and a second BF 245C in a common-gate circuit leads to a sufficiently low noise factor of approx. 3 dB together with stable behaviour.

The mentioned intermediate common-base circuit has its groundpoint somewhere between base and emitter. This leads to a simple means of neutralization and enables noise matching and power matching to coincide. The circuit was developed in Germany and widely used in tubed VHF TV tuners. The original designation is ZWISCHENBASIS-STUFE.

2. CIRCUIT DESCRIPTION

The circuit diagram of the 144 MHz/28 MHz converter is given in Fig. 2. The input signal is inductively coupled to the input circuit of the RF amplifier stage with transistor T 1. The signal is amplified and passed via a bandpass filter to the second RF amplifier stage equipped with the transistor T 2 in a common-gate configuration. A further bandpass filter allows the signal to be fed to gate 1 of the multiplicative mixer comprising the dual-gate MOSFET T 3. The crystal oscillator circuit is basically that described in (1). Interfering shortwave signals in the 9 MHz range that were observed by the author when using the converter described in (1) were suppressed by improving the decoupling of the oscillator and tripler stage to the plus line. The 116 MHz auxiliary frequency is fed via a bandpass filter to gate 2 of the mixer stage (T 3). Gate 2 of this stage is not biased; the mean value of the auxiliary frequency voltage should amount to approximately 0.7 V.

The intermediate frequency of 28-30 MHz is fed out via a π -filter; the advantages of this arrangement over a bandpass filter circuit are the simpler alignment and reduced filter losses. The values given in the circuit diagram correspond to a receiver input impedance of approximately 60 Ω; other impedance values will require changing the value of capacitor C 15 and realignment of inductance L 7.

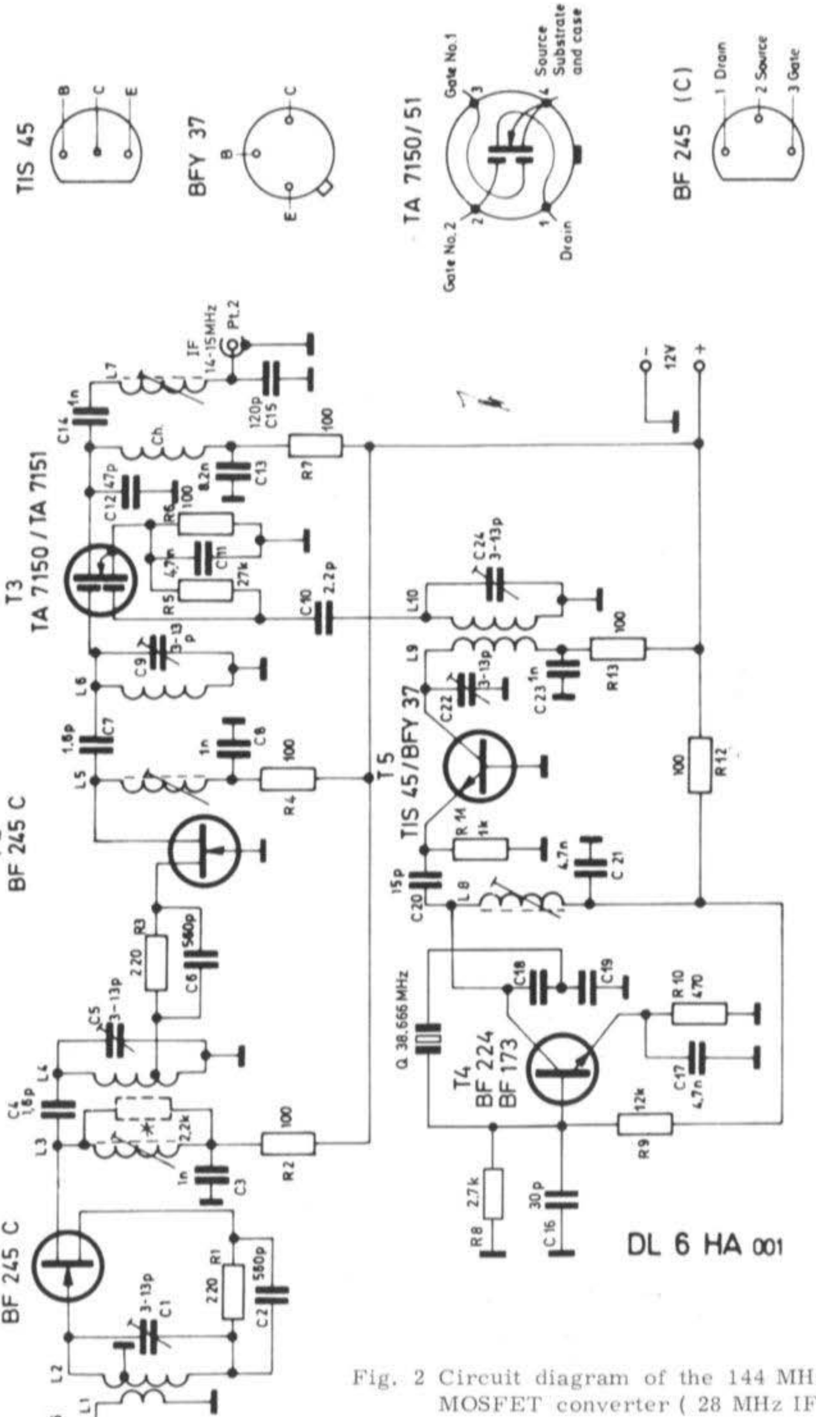


Fig. 2 Circuit diagram of the 144 MHz
MOSFET converter (28 MHz IF)

3. MECHANICAL ASSEMBLY

The converter is built up on a silver-plated epoxy printed circuit board (see Fig. 3a) with the dimensions 113 mm by 63 mm. The location of the components and screening plates can be seen in the component location plan Fig. 3b. The photograph given in Figure 4 shows the completed converter. Attention should be paid when mounting the screening plates - which are grounded using short wire connections - that a good solder connection exists between the screening plate separating the two RF amplifier stages and the ground connection of the trimmer capacitor C 5. This is necessary to avoid any tendency to oscillation of the second RF amplifier stage (T 2).

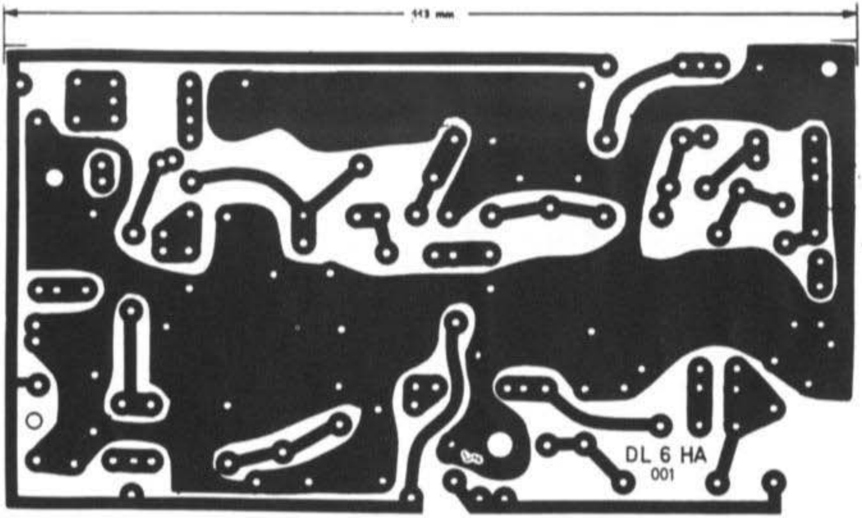


Fig. 3a Printed circuit board of the 144 MHz MOSFET converter

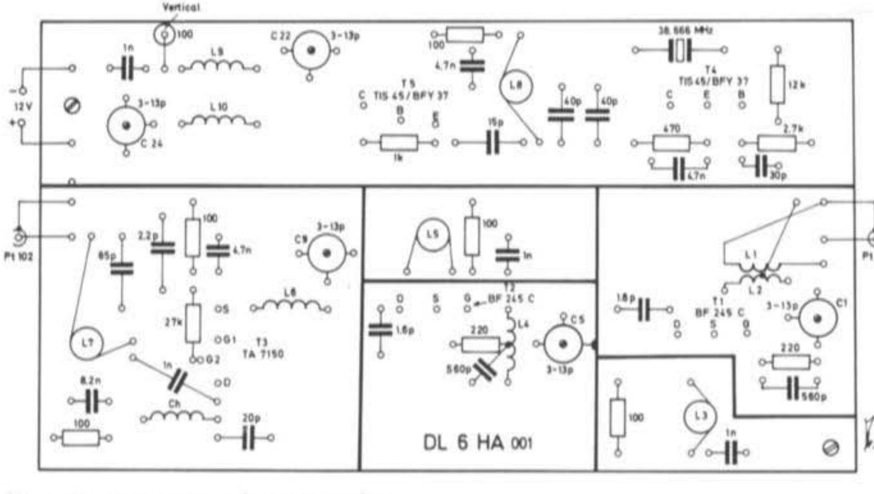


Fig. 3b Component location plan

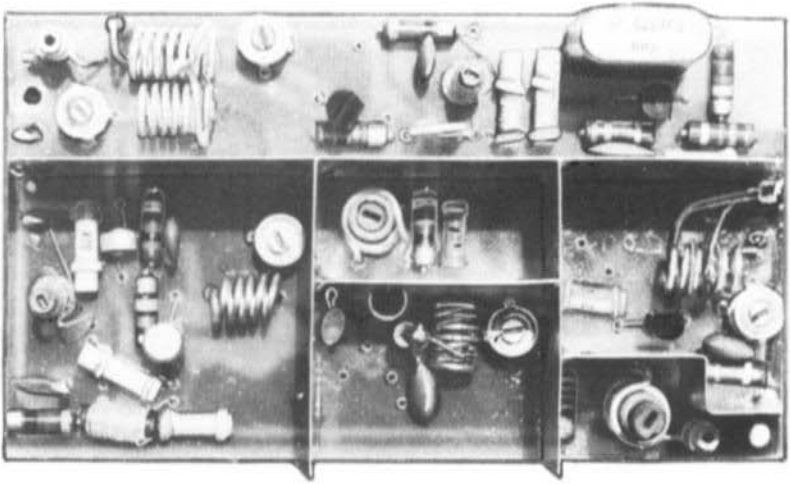


Fig. 4 Photograph of the completed converter

The two 6 mm diameter coil formers for L 3 and L 5 are glued to the PC-board in a vertical position.

Great caution must be paid when soldering the dual-gate MOSFET transistor T 3 into position. Because such transistors exhibit DC input impedances of several hundred megaohm, they are extremely endangered by static charges. This is the reason why they are always supplied with shorted connections (hollow rivit). This rivit should not be removed until the connections have been shorted using a thin copper wire wound just below the transistor case. The dual-gate MOSFET is now soldered in this shorted state into position and the thin wire removed afterwards. This can be avoided when using the double-gate MOSFET 40673 manufactured by RCA, which possesses integrated protective diodes that short out voltages greater that $\pm$ 10 V and thus avoid any damage to the transistor system itself. These transistors are available from the publishers (DM 12.50).

It is advisable to install the converter in a screening cabinet in order to avoid injection of unwanted signals; the power supply connections should be well blocked and filtered.

3.1. COIL DATA

All coils wound on 5 mm dia. formers. With exception of L 1, L 7 and L 8, all coils wound from 1 mm dia. (18 AWG) silver-plated copper wire.

L 1 — 2 turns of 1 mm dia. (18 AWG) insulated copper wire wound onto L 2.

L 2 — 6 turns self-supporting. Coil length 10 mm, coil tap 2.5 turns from the gate connection end.

L 3 — 6.5 turns. Coil length 15 mm.

L 4	6 turns self-supporting. Coil length 10 mm, coil tap 2.5 turns from cold end.
L 5	as for L 3.
L 6	5 turns self-supporting. Coil length 10 mm.
L 7	16 turns of 0.3 mm dia. (29 AWG) silk-covered, enamelled copper wire, with SW core.
L 8	15 turns otherwise as for L 7.
L 9, L 10	7 turns, self-supporting. Coil length 10 mm.
Ch 1	50 turns of 0.1 mm dia. (38 AWG) silk-covered copper wire wound on a 5 mm dia. ferrite coil former.

3.2. COMPONENTS

T 1, T 2 :	BF 245C (TIS 34, 2 N 5284, TIS 88, 2 N 5245)
T 3 :	TA 7150, TA 7151, 40 604, 40 673
T 4 :	BF 224, BF 173, 2 N 918
T 5 :	TIS 45, BFY 37, 2 N 918

The other components have no special features. The resistors have a rating of 0.3 W. All capacitors are conventional disc or tubular ceramic capacitors. The trimmer capacitors are ceramic micro disc types (3.5 - 13 pF). The crystal is a HC-6/U type for 38.6667 MHz.

4. ALIGNMENT

After checking for any short circuits or similar faults, the converter is connected via a mA meter to a voltage of 12 V. The current drain of the converter is approximately 30 mA. It is now possible to check the alignment of the individual resonant circuits with the aid of a dip meter - circuits using FET's can only be checked for resonance when the operating voltage is connected.

This is followed by checking the operation of the crystal oscillator, which is carried out favourably by measuring the RF voltage at the emitter of transistor T 5. If the crystal oscillator is operating, connect an RF probe to the second turn (from cold end) of inductance L 9 and align the stage for maximum RF voltage. This process is repeated for L 10.

The converter should now be connected to the input of a ten metre receiver and provided with a two metre antenna. Align all stages for maximum using a test signal of approximately 145 MHz. The resonant circuits with L 3/L 4 and L 5/L 6 should be alternately aligned because of their interaction. The resonant circuits comprising L 9/L 10 are also aligned for maximum signal. The inductance L 8 should be aligned so that the lowest frequency variation occurs on altering the operating voltage in the range of 11 V to 13 V.

5. RESULTS

The measured values given in Section 5.1. below were measured by the author using the following measuring instruments: two VHF/UHF signal generators (Rohde u. Schwarz); a calibrated coaxial attenuator 0 to 100 dB, Z = 60 Ω (Rohde u. Schwarz); a selective level meter 3 to 30 MHz, measuring range - 112 dB (Siemens) and a noise generator.

The following should be noted with respect to the measurements: The selectivity values as well as the voltage values at which cross modulation occurs cannot be classed as absolute values (see Section 5 in (1)). Measurement of selectivity values greater than 60 dB is not possible when, as is the case of the author, no screened cage is available. This is even more critical during cross-modulation measurements at VHF frequencies which even cause problems to professional engineering teams. The author also did not have modulation depth measuring instrument at his disposal which would allow determination of the 1%-limit. However, the value of the interfering voltage by which the unmodulated required carrier is audibly modulated represents a valuable reference value.

5.1. MEASURED VALUES

Noise factor : 3 dB
Overall gain : 22 dB
Image rejection : appr. 85 dB
Suppression of spurious signals generated by harmonic conversion

$2 \times (f_{osc} - f_{IF})$: approx. 90 dB

Cross modulation:	Required signal = 1 μV; interfering signal is 30% modulated and spaced 500 kHz from the carrier. Audible cross modulation occurs at approx. 50 mV.
Desensitation:	Required signal = 1 μV; interfering unmodulated signal is spaced 500 kHz from the carrier. A desensitation of 3 dB occurs at approx. 70 mV.

6. MODIFICATION FOR AN INTERMEDIATE FREQUENCY OF 14 MHz

This modification is necessary when using the converter in the SSB transceiver project which operates at a first intermediate frequency of 14 to 15 MHz. Two crystal controlled frequencies are required by the receive mixer (on subassembly DL 6 HA 001) and the transmit mixer (on DL 6 HA 005) in order to tune the whole two metre band, namely 130 MHz and 131 MHz. If only the upper half of the band is required, this will simplify the modification since no oscillator switching will be required.

The author suggests building up two crystal oscillators, operating at half the required auxiliary frequencies (65.0 MHz and 65.0 MHz), on a separate screened printed circuit board. The output of both crystal oscillators is connected together and fed to the base of the original oscillator transistor T 4 on PC-board DL 6 HA 001 (see Fig. 3b). Transistor T 4 now operates as a buffer. The originally used 38.66 MHz crystal is, of course, no longer required. The required oscillator switching is made by simply switching the operating voltage from one stage to the other. Figure 5 shows the circuit diagram of the modified converter.

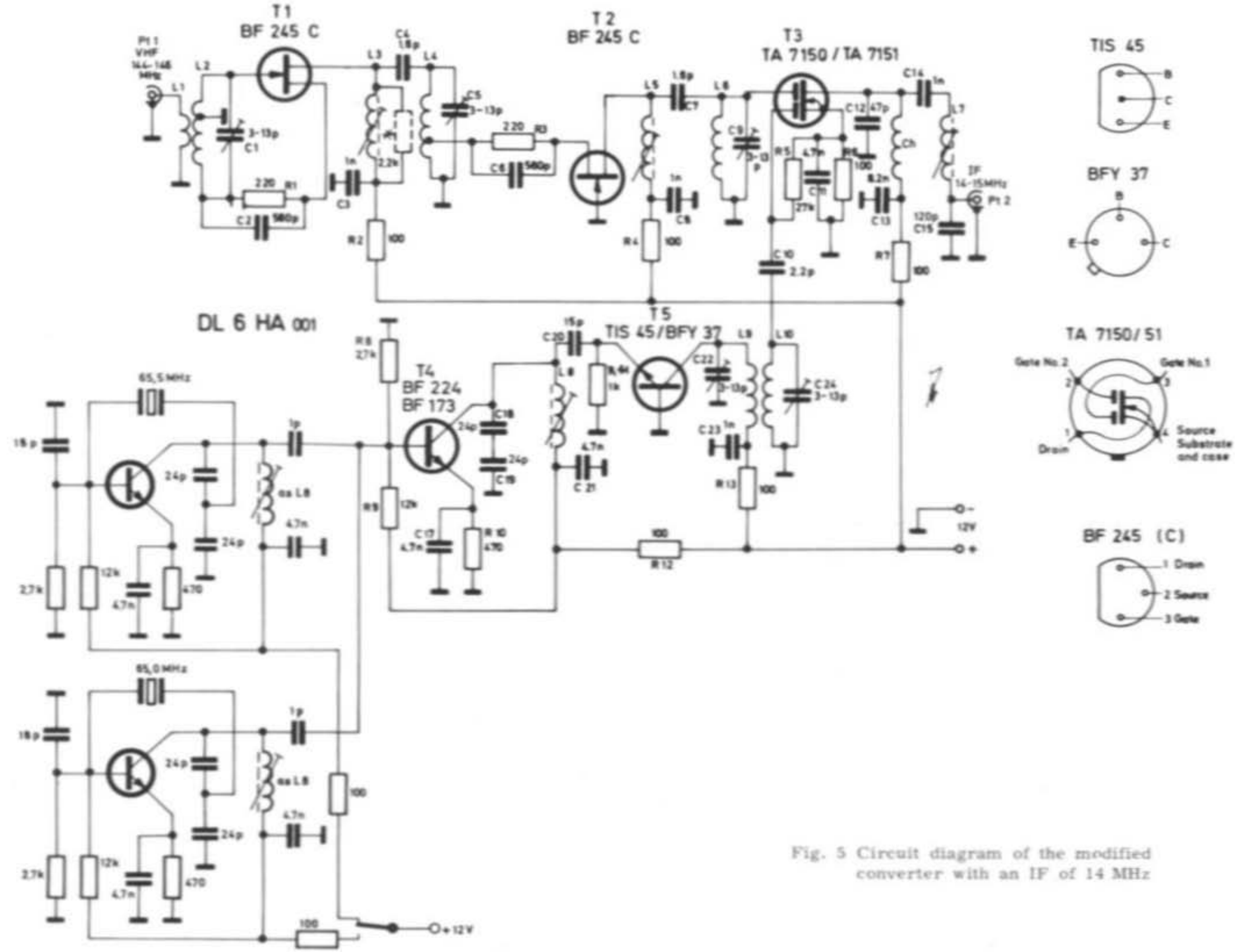


Fig. 5 Circuit diagram of the modified converter with an IF of 14 MHz

The following table shows the components that are affected by the modification:

Component	28 - 30 MHz model	14 - 15 MHz model
C 12	20 pF	47 pF
C 15	85 pF	120 pF
C 16	30 pF	deleted, or if only one crystal is used in the original oscillator circuit = 18 pF
C 18, C 19	4o pF	24 pF
T 4	TIS 45, BFY 37 2 N 706, 2 N 708	BF 173, BF 224, BSX 20 2 N 918
2 oscillator transistors	-	TIS 45, BFY 37, 2 N 918
Quartz crystal	38.66 MHz	65.0 MHz and 65.5 MHz
L 7	16 turns	20 turns
L 8	15 turns	13 turns
L 9, L 10	7 turns	6 turns

7. OPERATIONAL RESULTS

Four prototypes having different intermediate frequencies (two for 28-30 MHz and one each for 14 MHz and 9 MHz) were built up by the author and tested by various amateurs over a considerable period of time. Comparisons made to good tubed converters showed that field effect transistors are by no means inferior in performance. The tests were limited to sensitivity, image rejection and large-signal characteristics (cross-modulation and intermodulation).

Since the characteristics vary greatly between individual field effect transistors of the BF 245C series, too high a reverse capacitance could cause selfoscillation in the first RF amplifier stage. A 2.2 to 2.7 kΩ resistor connected in parallel to inductance L 3 will ensure sufficient neutralization and provide a greater independence to input impedance fluctuations without noticeably imparing the selectivity characteristics.

8. AVAILABLE PARTS

The printed circuit board DL 6 HA 001, trimmer capacitors, coil former set and quartz crystals as well as a kit of all major parts for both the 14 MHz and 28 MHz IF models are available from the publishers or via their national representatives. See advertising pages.

9. REFERENCES

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

A TILTABLE ANTENNA WITH SELECTABLE POLARITY

by E. Reitz, DJ 9 JT

INTRODUCTION

The described antenna array does not represent a high-performance antenna with respect to a high power gain, but more a versatile compromise antenna allowing a great number of experiments to be made. The switchable polarity is not only of interest for operation in conjunction with amateur satellites, but also useful when examining the effect of reflections on VHF propagation. This array allows interesting experiments to be made especially since the author lives in a valley.

The authors antenna array, as shown in Figure 1, consists of two 5 element cross Yagi antennas stacked one beside the other. The antenna can be tilted in the horizontal and vertical planes. Four switchable polarities can be selected: Linear horizontal, linear vertical, clockwise circular and anticlockwise circular. As has been already mentioned, the antenna can be tilted in the vertical and horizontal planes and is equipped with a remote indication of both positions, as well as allowing the antenna to be locked in the horizontal position.

The rather elaborate drive and switching arrangement is, of course, more suitable for larger arrays where it will be truely worthwhile. However, the authors array was only designed for experimental use and this article is only meant to represent an inspiration. Since the mechanical arrangement will be different for each location, only the most important points are to be explained here.

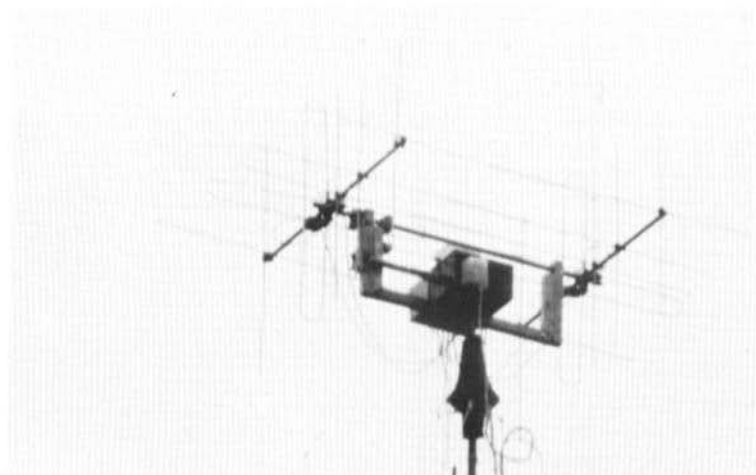


Fig. 1 Photograph of the authors antenna array

1. ELECTRICAL CONSTRUCTION

The author uses four 5 element Yagi antennas which are mounted to form two 5 element cross Yagi antennas stacked one beside the other. The selection of the antennas depends on the required gain and on the required mechanical stability. The standing wave ratio of the individual antennas should be checked before mounting, and adjusted where necessary.

After mounting, each of the four Yagi antennas is provided with a $\lambda/2$ balun transformer and a coaxial cable 1.5 metre in length. These four coaxial cables (cables 1 - 4 in Fig. 4) must be of exactly the same length because a phase shift could be caused within the array if this was not the case.

1.1. POLARITY SWITCHING

Switching to the previous mentioned polarities requires a rather elaborate switching arrangement, especially when exactly matched conditions are required. The following description represents a compromise by which a standing wave ratio of approximately 1.6 : 1 is obtained. The author, however, considers this value to be still acceptable.

Linear polarity is obtained when only the horizontal or only the vertical components of the antenna array are selected. For circular polarity, it is necessary for one plane to be phase-shifted by 90^o with respect to the other plane to which it is connected. The selection of either clockwise or anticlockwise circular polarization depends on which of the two planes leads or lags in the phase sense. The required phase shift of 90^o is obtained by placing a cable having an electrical length of $\lambda/4$ (cable 3 in Fig. 2 and 3) to the feeder of the horizontal plane (cable 1) or that of the vertical plane (cable 2). The phase line can also be an odd multiple of $\lambda/4$. The principle of this arrangement is shown in Fig. 2 (ignoring the matching and cable sheath).

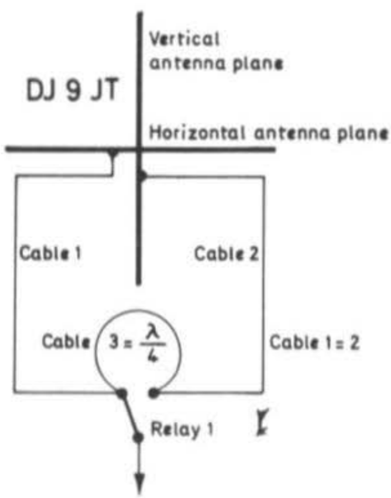


Fig. 2 Principle of the switching for clockwise or anticlockwise circular polarization

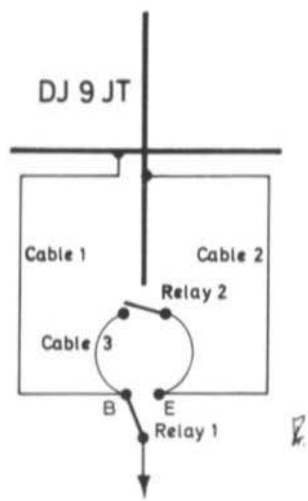


Fig. 3 Principle of the switching for the four different polarities

Since a quarter wave cable with solid dielectric (velocity factor 0.66) is only 33 cm long on the two metre band, the author used a $3/4\ \lambda$ cable (99 cm) to obtain the required phase shift. If cable 3 is disconnected by a second relay, only the vertical or only the horizontal plane will be connected according to the position of relay 1. This principle is shown in Figure 3.

This basic circuit has to be modified in practice due to the necessary transformation links. The circuit used by the author is given in Figure 4. The following considerations led to the development of this circuit.

The horizontal components come from a 5 by 5 antenna array. The feed lines (cables 1 and 3) are connected together a point A. This results in an impedance of 30 Ω at this point. This value is transposed to 120 Ω at point B by means of a $\lambda/4$ transformer (cable 5). The following relationship is valid for the transformation with cables:

$$Z_T = \sqrt{Z_1 \times Z_2}$$

Where: Z_T = impedance of the $\lambda/4$ cable
Z_1 = characteristic impedance to be transformed
Z_2 = required impedance

In our case, the following will result:

$$Z_T = \sqrt{30 \times 120} = 60\ \Omega$$

This means that normal coaxial cable can be used for cable 5. A further cable (cable 6) is connected to point B, which also provides an impedance of 120 Ω at point B - the reason for this will be explained later. Due to the parallel connection, an impedance of 60 Ω will result at point B which means that correct matching occurs when relay 1 is connected to point B.

The $3/4\ \lambda$ phase line for the circular polarization is connected between points B and E of relay 1. This phase line is built up from the three $\lambda/4$ cables 6, 7 and 8. Cable 7 has an impedance of 30 Ω (two 60 Ω cables in parallel), which means that an impedance of 30 Ω is present at point C and D. This impedance is transformed to 120 Ω at point B due to the effect of the $\lambda/4$ cable 6 which has an impedance of 60 Ω. This means - as already mentioned - that an impedance of 60 Ω results due to the parallel connection with cable 5. The same consideration is valid for the transformation from point D to E with the exception that cable 8 consists of two short cables, a T-connector and relay 2, which together represent a $\lambda/4$ line. Since the relay has a different dielectric than the cable, this will not represent a true $\lambda/4$ line. The true resonance can be found using a dip-meter (which is quicker and easier than calculating it). This is made by shorting the cable at point D and soldering a small coupling link to point E. The cabling is completed with cable 9 which has the same function for the vertical polarization as cable 5 has for the horizontal polarity.

Finally the function of the stub connected to relay 2 should be explained. If relay 2 is opened to allow linear polarization, open pieces of cable will be connected to points B and E. Since these cables represent a length of $\lambda/4$ or $3/4\ \lambda$ at the operating frequency, they act as absorption circuits thus shorting the RF voltage. In order to avoid these critical lengths, relay 2 should be connected as near as possible to relay 1 and the path B, C and D increased to more than $3/4\ \lambda$ by means of the stub.

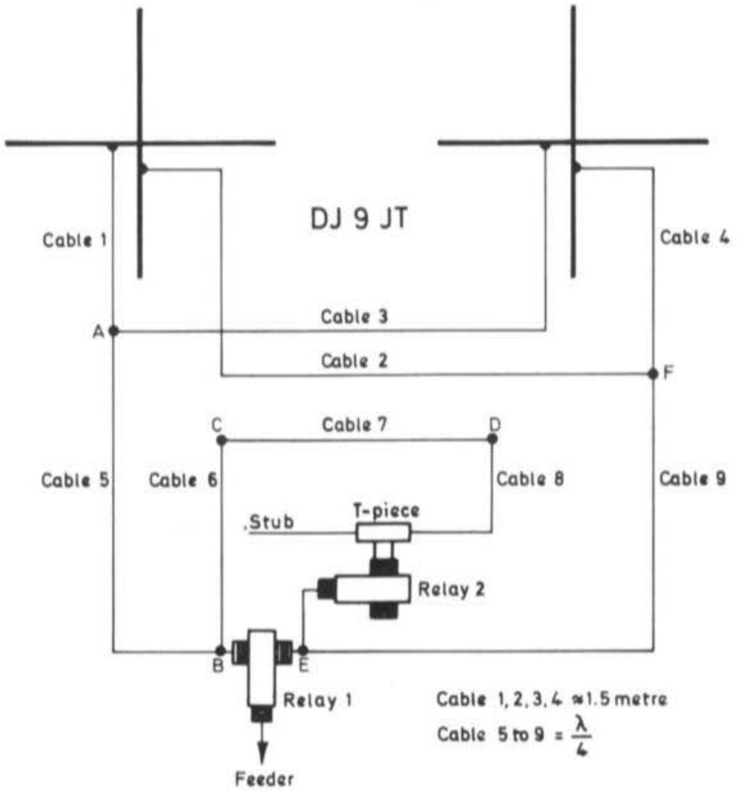


Fig. 4 Matching and polarity switching for two cross Yagi antennas

When relay 2 is closed, it is necessary that the stub also represents an uncritical length. The most favourable method of doing this is to find the stub length by experiment using a reflectometer. The author found a length of 55 cm to be the most favourable compromise.

The following tables summarize these considerations:

Cable No.	Length in cm	Impedance
1	Not important but of the same length	60 Ω
2		60 Ω
3		60 Ω
4		60 Ω
5	$\lambda/4$	60 Ω
6	$\lambda/4$	60 Ω
7	$\lambda/4$	30 Ω (two 60 Ω cables in parallel)
8	$\lambda/4$ incl. relay 2	60 Ω
9	$\lambda/4$	60 Ω
Stub	approx. 55 cm	60 Ω

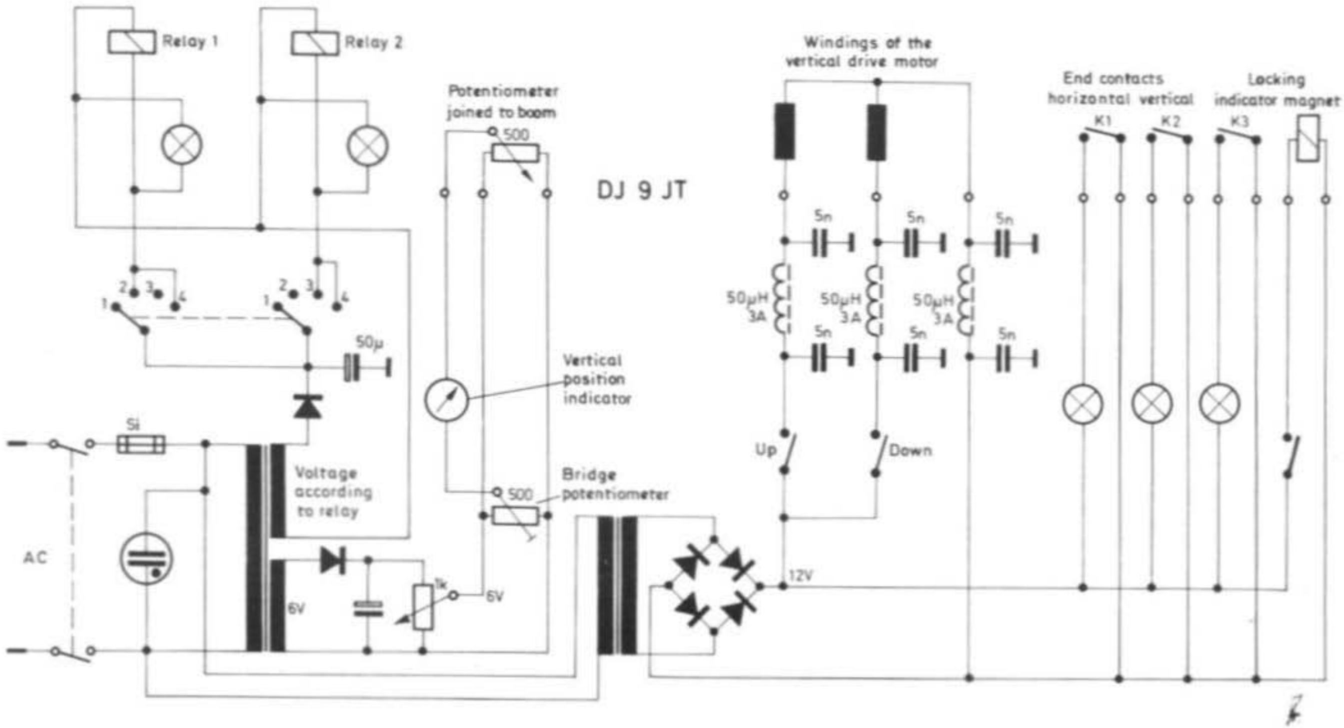


Fig. 5 Vertical drive circuit with indicator panel and power supply for the polarity switching

Relay 1	Relay 2	Polarization	Switch position
off (B)	off	Horizontal linear	1
on (E)	off	Vertical linear	2
off (B)	on	Clockwise circular	3
on (E)	on	Anticlockwise circular	4

1.2. ANTENNA DRIVE AND POSITION INDICATION

Figure 5 shows the circuit of the vertical drive with position indicator, the end contacts, the locking magnet as well as the power supply for the polarity switching. The horizontal rotation of the antenna was made using a commercial rotator having its own position indicator. An additional horizontal position indicator was therefore not required.

A 24 V windscreen wiper motor (from a truck) is used for the vertical drive. This is achieved with the aid of a worm drive and a gear wheel driving a horizontally mounted lever. The motor is operated from 12 V to reduce the speed. So that a vertical position indication is obtained, the lever is fixed to a potentiometer that is connected in a bridge circuit. In order to check the "horizontal" and "vertical" positions, end contacts are provided which are associated to control lamps. The bridge circuit is aligned to zero in the horizontal position. The meter full-scale deflection in the vertical position is adjusted with the aid of the 1 kΩ potentiometer. The meter of approximately 1 mA must be provided with a suitable shunt. The motor leads are filtered by low-pass filters so that no electrical interference is audible in the receiver.

2. MECHANICAL ASSEMBLY

As has been already mentioned, the mechanical assembly of the antenna array will not be explained in great detail. It should be mentioned, that attention should be paid to obtain a robust and waterproof construction.

2.1. THE ANTENNAS

Two of the four, identical Yagi antennas are dismantled and the elements mounted in the same order but 90^{0} from the elements of the other two antennas. It is not important whether these elements are mounted in front of or behind the other elements. It is only important that this should be the same for all elements and that the spacing between the individual elements of each cross should be as small as possible.

2.2. THE ANTENNA FRAME

As can be seen in Figure 1, the two cross Yagis are connected to another using a strong boom. This boom can be rotated and is connected via bearings to the antenna frame. The frame itself is "U" shaped and the author made it from 5 x 5 cm wood. The antenna frame is in turn connected by three screws to the vertical drive. Figure 6 shows a schematic diagram of this construction.

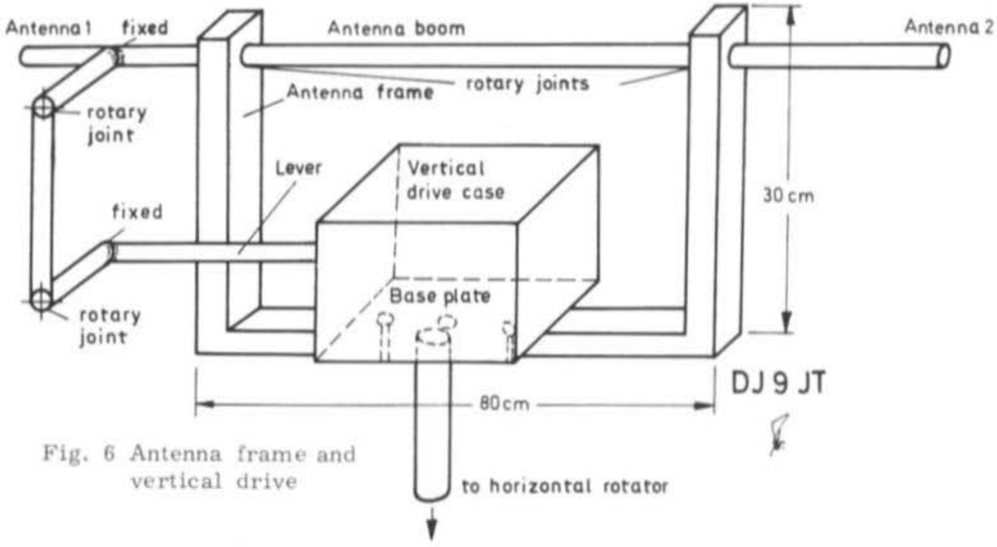


Fig. 6 Antenna frame and vertical drive

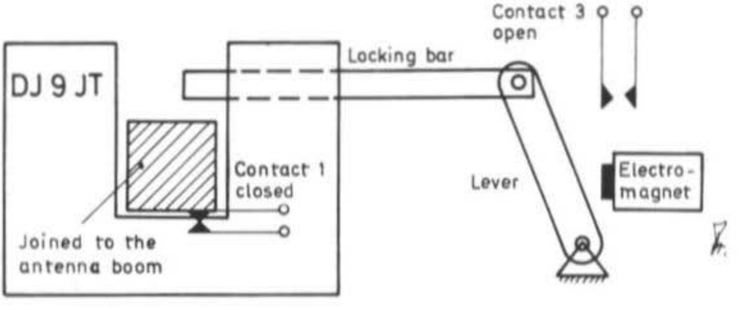


Fig. 7 Vertical locking closed

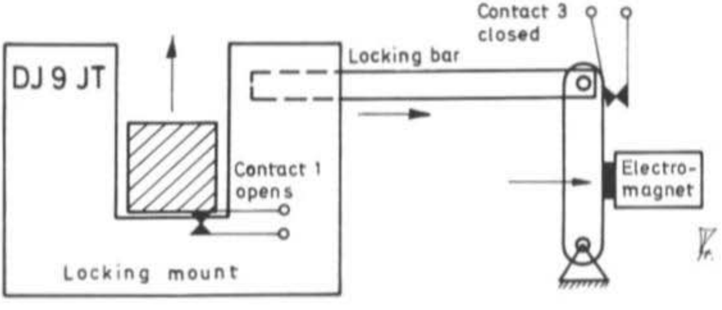


 Fig. 8 Vertical locking released

2.3. THE VERTICAL DRIVE

The rotation of the windscreen motor (driven from 12 V) is transposed - using a worm drive and a gear wheel having a gear ratio of 100 : 1 - to the vertical lever also shown in Fig. 6. The mounting of this lever must be especially stable due to the high force that the long lever arm has at this point. The lever is therefore mounted in two ball bearings which are in turn mounted in a somewhat greater tube that is welded to the base plate of the vertical drive casing. A 30 cm long 1 1/4" pipe is welded to the lower side of this base plate (25 cm x 30 cm x 7 mm). The whole antenna array is connected by means of this pipe to the horizontal rotator.

The transposition of the lever rotation on to the boom is made by means of a wooden brace (Fig. 6). This was found by the author to be simpler and less expensive than a chain or rubber drive.

2.4. LOCKING ARRANGEMENT

Since the whole wind pressure is effective on the vertical drive, a mechanical locking arrangement was provided for the horizontal antenna position (which is the most often used position). In the rest position, a spring presses a strong metal bar or tube into the cut-out on the locking mount so that no movement of the antenna boom can occur. If the antenna is to be tilted vertically, an electro-magnet is energized that removes the locking bar from the cut-out and releases the antenna (see Fig. 7 and 8).

Since a tilting of the antenna without previously unlocking the boom would cause the drive or the motor to be damaged, an end contact (contact 3) has been provided that indicates, by means of an indicator lamp on the control panel, when the magnet has released the boom.

End contact 1, on the other hand, indicates the horizontal position of the antenna. The corresponding contact for the vertical position is contact 2. Contact 1 for the horizontal position indication was found to be very necessary because the position indicator meter did not possess sufficient resolution to indicate whether the locking could been made or not.

3. EXPERIENCE GAINED WITH THE ANTENNA

The author is still not able to give final conclusions with respect to the characteristics and advantages of this antenna when compared to other antennas since it has not been used over a sufficiently long period. Two points, however, should be mentioned: The author was able to observe polarity shift due to reflections on a number of occasions. These were mostly observed on local stations and not on those stations spaced at greater distances. However, the polarity shift due to reflections does not seem to pay such an important part as was assumed at first. The majority of stations maintained their original polarity.

A considerable improvement was found during contacts with vertically polarized stations, such as mobile stations using 5/8 λ antennas. The antenna provided an improvement of approximately 2 S-points over the horizontal polarization. This is a considerable gain and can make the difference between a QSO or not.

This "sideproduct" is, however, in itself, not worth the extensive construction. But this was not the reason for the experimental construction and it is felt that the advantages of such an antenna will show themselves when operating in conjunction with amateur satellites (or MOONRAY on 70 cm).

4. REFERENCES

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73 Magazine, 1966

VOLUME 1 (1969) OF VHF COMMUNICATIONS

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IS FM ADVANTAGEOUS ON THE VHF-UHF BANDS?

by D.E. Schmitzer, DJ 4 BG

INTRODUCTION

Three main modulation modes are permissible for amateur radio voice transmissions. These are: double sideband amplitude modulation A 3 (AM), single sideband amplitude modulation A 3 J (SSB), and (narrow band) frequency modulation F 3 (NBFM). Of these three modes, frequency modulation has not been used to any great extent. The following consideration compares all three modulating modes, determines what advantages NBFM could have on the VHF/UHF bands and examines why previous experiments with FM have not been successfully able to prove its suitability. The parameter chosen for the comparison is the AF signal-to-noise ratio as a function of the signal level at the receiver input; the required complement at the transmitter is also considered.

1. COMPARISON OF THE MODULATION MODES

The three modes are to be observed at the receive end. The criterion for each of the modulation modes is the signal voltage required at the antenna input connector to obtain a certain AF signal-to-noise ratio. It is necessary to lay down certain characteristics of the receiver such as:

Input impedance :	$Z_{in} = 60\ \Omega$
Noise figure :	2 (3 dB)
IF bandwidth switchable to :	
SSB :	2.5 kHz
AM :	5 kHz
FM 1 :	10 kHz (for a modulation index of M = 1)
FM 2 :	15 kHz (for a modulation index of M = 2)
AF bandwidth	2.5 kHz (300 Hz to 2.8 kHz)

Figure 1 shows the family of curves for each of the modulating modes. They show the AF signal-to-noise ratio as a function of the input signal when taking the fundamental characteristics into consideration. An additional curve is included for comparison: it is valid for telegraphy at a bandwidth of 500 Hz.

As can be seen in Fig. 1, the single sideband mode (SSB) only offers a 3 dB improvement on the double sideband modulation (AM). This is because the receive bandwidth is only half as great for SSB than for AM which means that the receiver exhibits only half (thus 3 dB less) the noise power for SSB than for AM. Telegraphy at a bandwidth of 500 Hz correspondingly results in a 10 dB better signal-to-noise ratio than AM at a bandwidth of 5 kHz.

Frequency modulation requires a greater transmission bandwidth which means that a higher input voltage level will be required until the signal appears above the noise level. However, the AF signal-to-noise ratio increases more rapidly than with the other modes on raising the input voltage. This results in signal-to-noise ratio values that are better than those of SSB if a modulation index of only M = 1.5 is assumed. This superiority of FM is, however, only noticeable at higher signal-to-noise values ($\gtrsim$ 10 dB). At lower input voltages, FM is somewhat inferior to SSB and approximately equal to AM. The difference is not very great as long as the modulation index is not greater than M = 1. This value should therefore prove to be the most favourable compromise.

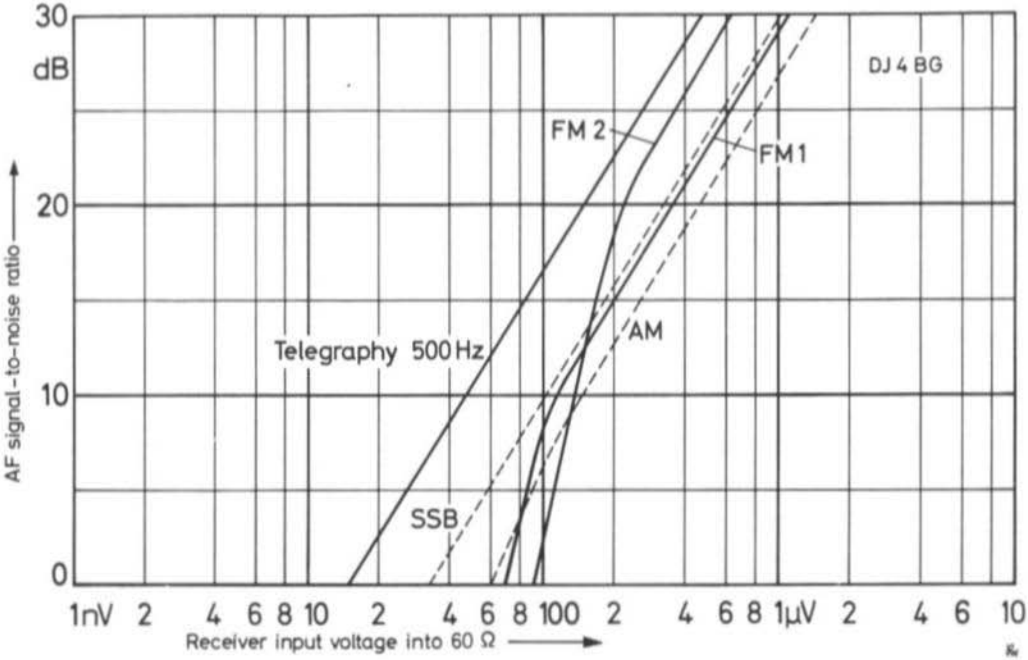

Fig. 1 AF signal-to-noise ratio as a function of the receiver input voltage for the different modulating modes

These simple comparisons differ when also considering the transmitter complement required for each of the three modes. The superiority of SSB when using tubed power amplifiers is that modern tubes allow peak power levels (pulse power) far higher than that permissible for continuous carrier operation. The advantage of FM over AM in tubed power amplifiers is that FM can be operated at the peak carrier value allowed for continuous carrier operation whereas the PA tube must be operated at reduced power in the AM mode.

Virtually the same considerations are valid for transistor transmitters. However, the difference between the carrier power of AM and the peak-power of FM is even greater due to the limited reverse voltage; with 100% AM, the operating voltage must not exceed 1/4 of the permissible reverse voltage, whereas it may be 1/2 the peak voltage. This means that four times the carrier power is available with FM than with AM. SSB is not so advantageous with transistorized PA stages because transistors cannot be overloaded (peak current and voltages) during pulse operation to such an extent as tubes. Furthermore, an additional phase modulation occurs at high levels due to the drive - dependent impedance values of transistors. This will be noticeable as signal distortion and splatter.

The comparison between tubed and transistor PA stages is, of course, not the only consideration. The frequency stability required for SSB operation is increasingly more difficult to achieve on raising the operating frequency. If the permissible frequency difference is assumed to be 43 Hz, this will represent a frequency stability of 10^{-7} at 430 MHz. Even if only the short-time stability is considered, the complicated complement to achieve this will keep a great number of amateurs from using the UHF bands. FM, on the other hand, requires far less frequency stability and allows automatic frequency control to be used easily.

An additional advantage of FM is the amplitude limiting at the receiver which suppresses the amplitude components of ignition and similar interference. An automatic gain control is also not required in the receiver. Since the signal amplitude of the transmitter remains constant, interference caused by unwanted envelope demodulation in audio amplifiers of neighbouring television, radio and audio equipment will be avoided as will the modulating bars sometimes visible on television receivers when using AM.

The author would therefore like to suggest that FM should be used on the 70 cm and higher bands with a modulation index of between 1 and 2. The most favourable mode for the two metre band should continue to be SSB due to the limited bandwidth and prevailing conditions.

2. REASONS FOR UNSATISFACTORY EXPERIMENTS WITH FREQUENCY MODULATION

The question is now why FM has not been extensively used if the already described advantages are true. The explanation is simple: it is necessary for the receiver to fulfil a number of requirements if the advantages of frequency modulation are to be realized. This is not the case with present equipment, mainly due to the excessive accentuation of SSB.

A primary factor is for a sufficiently good amplitude limiting of the FM signal to be made. Furthermore, it is necessary that a frequency discriminator is used for demodulation. Finally, the bandwidth of the audio signal must be limited. This is because the low-frequency noise voltages appearing after the demodulation of FM are not equally distributed but increase toward higher audio frequencies. The signal-to-noise ratio curve shown in Fig. 1 is only valid in conjunction with a suppression of all frequencies in excess of 2.5 to 3 kHz and realization of the other demands.

FM demodulation using the skirt of a passband curve or with wideband IF/AF portions of VHF broadcast receivers is therefore only a makeshift arrangement which cannot be used for comparison with the other modulating modes.

3. APPENDIX : FORMULAR APPERTAINING TO FM

The following formula has been used to calculate the required bandwidth B for FM :

$$B = 2 \times (\text{Dev.} + f_{mod})$$

where: Dev. is the frequency deviation in kHz and f_{mod} is the highest modulation frequency. The bandwidth B results in kHz.

The modulation index M should also be explained:

$$M = \frac{\text{greatest frequency deviation Dev.}}{\text{highest modulating frequency } f_{mod}}$$

Therefore if the highest modulating frequency is $f_{mod} = 2.5$ kHz and the permissible modulation index is M = 2, the following frequency deviation will result :

$$\text{Dev.} = M \times f_{mod} = 2 \times 2.5 = 5 \text{ kHz}$$

and the required bandwidth B will be :

$$B = 2(\text{Dev.} + f_{mod}) = 2(5 + 2.5) = 15 \text{ kHz}$$

At the Region I Conference of the IARU in Brussels, May 1969, agreement was reached regarding the modulation index for NBFM transmissions. A modulation index of 1 and a maximum modulation frequency of 3 kHz were laid down. The maximum frequency deviation is therefore 3 kHz and the required filter bandwidth is 12 kHz.

FREQUENCY MODULATION OF CRYSTAL-CONTROLLED OSCILLATORS BY USE OF RESISTOR DIODES

by G. Damm, DM 2 AWD

INTRODUCTION

Most crystal-controlled oscillators operate in the narrow frequency range between the series and parallel resonance of the crystal. In this range, the crystal represents a highly frequency-dependent impedance. According to the circuit and the characteristics of the feedback link, the resulting oscillation frequency will be where the reactance is nearly or completely compensated for by the parallel capacitance.

If this parallel capacitance is partly formed by a variable network, such as a combination of capacitors and diodes, it should be possible to modulate the frequency of the crystal-controlled oscillator so that the frequency deviation and linearity are sufficient for amateur transmissions.

1. ASSESSMENT OF THE REQUIRED PULL-RANGE

The deviation between the transmit frequency obtained in the modulation peaks and the mean value of the transmit frequency is the frequeny deviation Δf max; the highest audio frequency to be transmitted is designated f_m max.

The modulation index M of the transmitter is obtained as:

$$M = \Delta f\,\mathrm{max}/\mathrm{fm\ max}$$

The recommended modulation index for amateur transmissions is 1.0.

If audio frequencies of up to 3 kHz are to be transmitted this will mean that the maximum frequency deviation should not exceed 3 kHz.

If the transmit frequency is obtained by a frequency multiplication of factor n from the crystal frequency f_q, then a correspondingly lower frequency deviation will be required from the crystal:

$$\Delta f_q\,\mathrm{max} = \Delta f\,\mathrm{max}\,/\,n = \Delta f\,\mathrm{max} \;\;x\;\; f_q\,/\,f$$

Example: If a 16 MHz crystal is used in the oscillator circuit of the 144 MHz transmitter, a frequency multiplication of factor 9 will be required. In this case, it is necessary for the crystal, to be pulled by approximately ± 330 Hz. This can be achieved easily when using the described circuit as long as the fixed parallel capacitance is not too great.

2. OPERATION AND ALIGNMENT OF THE CIRCUIT

In Fig. 1, the RF voltage from the oscillator is fed to a voltage doubler circuit comprising the coupling capacitor C 1 and diodes D 1 and D 2; capacitor C 2 is the charge capacitor and R_L represents the load impedance. On the DC link, the audio voltage obtained from the modulation amplifier causes an additional increase or decrease of the load. This in turn increases or decreases the angle of current flow through the diodes and will effect the value of the coupling capacitance C 1 parallel to the crystal.

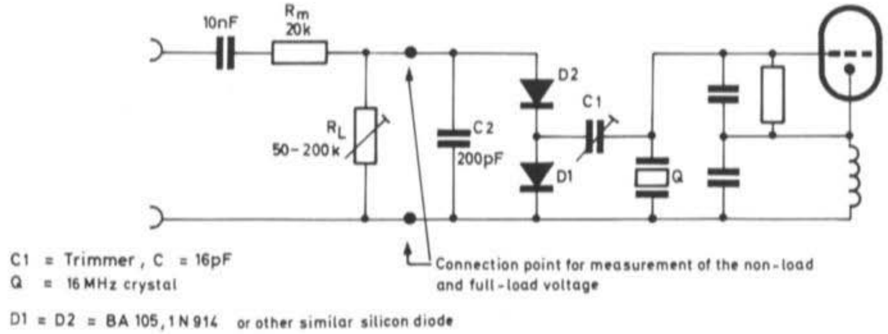

Fig. 1 Circuit diagram for frequency modulation of a crystal oscillator

It should be noted that the peak RF voltage across this circuit should amount to at least twice that of the forward voltage of the diodes if a low-distortion modulation is to be achieved, i. e. approximately 0.6 V for silicon diodes. If the RF voltage is lower, it will be necessary to vary the operating points of the diodes by using an additional voltage as shown in Fig. 2.

The most favourable operating point of the circuit for frequency modulation of quartz crystals is roughly where the non-load voltage of the diode circuit falls by half due to the loading by R_L. It is advisable to start with this setting before checking the frequency deviation and linearity and trimming coupling capacitor C 1 and R_L.

a) Unsolder the load resistor R_L and measure the non-load voltage between the two connection points with a VTVM ($Z > 5\ M\Omega$).

b) Resolder R_L and align it so that the voltage across R_L falls to half the non-load voltage.

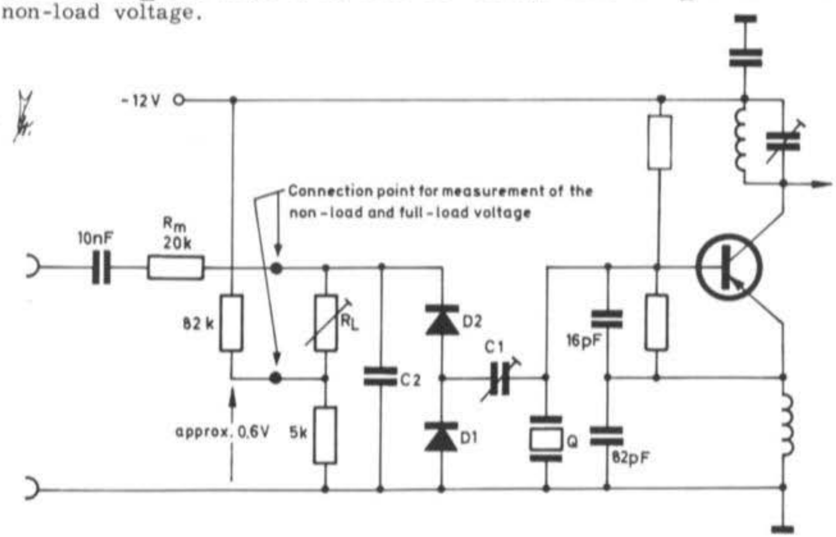

Fig. 2 Circuit similar to Fig. 1 but with additional DC bias for the diodes

3. THE MODULATOR

The required AF power at the input is very small because the modulator represents a very high impedance at the AF input; the impedance amounts to approximately half the load impedance of R_L plus the matching resistor R_m. A peak voltage of about 3 V (2 V_{rms}) was found to be sufficient during prototype measurements. An amplification factor of 100 - 1000 will be required when using crystal or moving coil microphones.

4. EXPERIMENTAL CIRCUITS

The diode circuit given in Fig. 1 has been tested on a number of different crystal oscillator circuits, whereby various types of diodes - from the cheapest IF types to professional diodes - were used. The circuit operated satisfactorily in all cases and it was only the required AF voltage that was altered. The matching resistor R_m can be exchanged for a 10 μH choke if RF-injection should occur.

5. RECEPTION REPORTS

The reports differed greatly: some amateurs with receivers not equipped for FM were not able to demodulate the FM transmission on the slope of the IF characteristic even when the frequency deviation was reduced at the transmitter. On the other hand, some amateurs reported that reception on the AM receiver was still possible even when the input signal was reduced to S 3.

However, receivers equipped with an additional FM demodulator indicated good results. When using a moving-coil microphone at the transmitter, extremely good speech quality was reported on a receiver equipped for FM. This quality could not be achieved when using AM with the same transmitter.

6. CONCLUSIONS FROM THE EXPERIMENTS

The described circuits can be simply assembled from cheap components and do not require any critical alignment. Since the RF output voltage remains constant, unwanted spurious envelope modulation in TV and broadcast receivers will not occur which means that TVI and BCI will be greatly reduced. If the transmit frequency is to be multiplied to higher frequencies, the operating points of subsequent stages can be aligned for highest gain of the tube or transistor since no demands need be placed on the linearity between input and output amplitude. It is only necessary to reduce the frequency deviation - by reducing the modulation voltage - so that the required value is obtained at the higher frequency.

A great advantage is obtained when the PA stage is transistorized, since the operating voltage can be increased to a higher level than with amplitude modulation; this should result in a higher output power.

A modification of the receiver for FM reception is very advisable. This can be achieved by use of an ratio or digital detector.

Editorial Note:
Two frequency demodulation circuits are to be described in later editions of VHF COMMUNICATIONS. These will be available in kit form for the modification of existing AM receivers.

NARROW BAND FREQUENCY MODULATION OF OVERTONE CRYSTAL OSCILLATORS

by E. Harmet, OE 6 TH

INTRODUCTION

When certain conditions are fulfilled, the same advantages obtained by commercial operators when using narrow band frequency modulation (NBFM) will be available to the amateur. These advantages were discussed in detail in (1).

Frequency modulation is especially advantageous for the UHF amateur bands. A circuit is now to be described that allows narrow band frequency modulation of overtone crystal oscillators. The frequency deviation that can be obtained using this circuit is also sufficient for operation on the two metre band.

1. PRINCIPLE OF OPERATION

As was explained in (2), most crystal-controlled oscillators operate in the narrow frequency range between the series and parallel resonance of the crystal. In this range, the crystal represents a frequency-dependent impedance. This is also valid for the circuits where the oscillator is said to oscillate at parallel resonance frequency of the crystal, e.g. Pierce-circuit. In this range, the oscillation frequency can be slightly pulled. The VXO circuit given in (3) uses this principle.

Some form of controlled capacitance can be used to pull the oscillation frequency. This means that the frequency of a crystal-controlled oscillator can be modulated in the same manner so that the frequency deviation - after frequency multiplication - is sufficiently great for amateur applications. With circuits where the crystal represents a parallel resonant circuit, the controlled capacitance should be connected in parallel. The circuit in (2) used this principle for the narrow band frequency modulation of a 16 MHz crystal-controlled oscillator.

In overtone oscillator circuits, the crystal represents a series resonant circuit. In order to pull or modulate the frequency, it is necessary for the controlled capacitance to be connected in series with the crystal. Such an arrangement is to be described in conjunction with a 72 MHz crystal-controlled oscillator.

2. CIRCUIT DESCRIPTION

Figure 1 shows the circuit diagram of a 72 MHz crystal oscillator that can be frequency modulated. The overtone circuit used, is so well known that the operation is not explained in detail. A varactor diode is utilized as the controlled capacitance. The DC operating point and thus the quiescent capacitance is adjustable with the aid of a potentiometer. Since the value of the diode capacitance in series with the crystal has an influence on the pull range, it is possible to select the maximum frequency deviation by varying the bias voltage. The circuit given in Fig. 1, allows the maximum frequency deviation to be varied with the aid of the potentiometer to values in the order of 2 to 5 kHz (at 144 MHz). This means that the modulation index M suitable for amateur transmissions between M = 1 and M = 2 can be easily achieved. For lower frequency deviation values, it is only necessary to reduce the amplitude of the AF voltage.

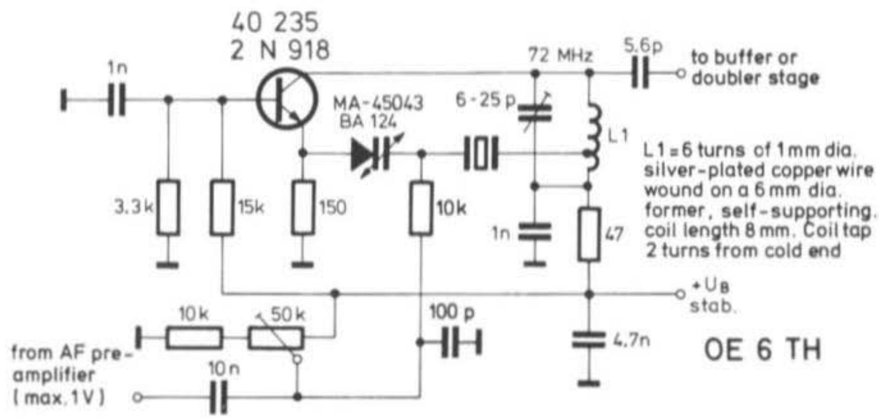

Fig. 1 NBFM of a 72 MHz crystal oscillator

3. AF DRIVE

Since the maximum bandwidth of a frequency modulated signal is dependent both on the highest modulation frequency and the maximum frequency deviation, it is necessary that both quantities should not exceed a certain maximum value. The highest modulation frequency can best be limited by use of an active low-pass filter with a cutoff frequency of approximately 3 kHz or by an active audio bandpass filter as described in (4).

The frequency deviation is proportional to the amplitude of the AF signal. In order to ensure that a certain maximum frequency deviation is not exceeded, an amplitude limiting device (clipper) or a dynamic compressor with limiter characteristics will be required. The principle of an amplitude limiter for voice applications is given in Fig. 2. When using active AF filters - for instance as given in (4) - in conjunction with an integrated circuit, a very high quality NBFM modulator can be constructed without difficulty.

Another advantage is that this AF preamplifier with its frequency and amplitude limiting is just as suitable for use with AM modulators as it is for FM. The circuit will ensure that the signal bandwidth in the AM mode will be limited by the given circuit; further advantages of the circuit are that overmodulation can be avoided and the mean modulation level increased due to the amplitude limiting process.

A dynamic compressor can be used instead of the limiter. Since compressor circuits usually result in a lower harmonic distortion factor, it is only necessary for a voice bandpass filter or a 3kHz low-pass filter to be provided at the input. A low-pass filter at the output is not required.

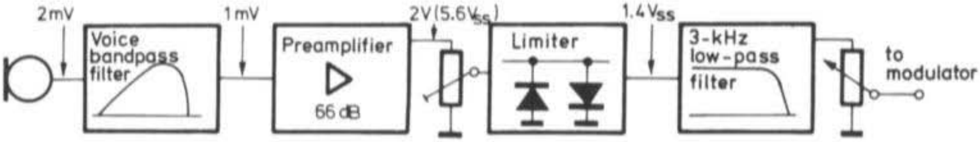

Fig. 2 Principle of a voice frequency and amplitude limiter

4. EDITORIAL NOTES

All the dynamic compressors known to the editorial staff possess the same, unfavourable characteristics, since they were mostly developed for other applications:

a) The rise time is too great which means that sudden amplitude peaks are not controlled and can cause overmodulation in the AM mode.

b) Their fall time is too great which means that the weak tones subsequent to strong tones disappear due to the previous reduction of the gain.

Such circuits represent, generally speaking, an AGC amplifier and only keep the mean level constant; this can be achieved just as well by ensuring a constant spacing from the microphone. To be effective, a dynamic compressor is required whose rise time is less than 5 ms and whose fall time is a maximum of 100 ms. This must be achieved without the DC pumping effect obtained when using biased tubes or transistors as control links.

Finally, it should be stated that a limiter as shown in Fig. 2 is more effective even if the distortion factor or harmonic content is greater.

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INDEX TO VOLUME 1 OF VHF COMMUNICATIONS 1969

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A 48 MHz VFO FOR 144 MHz TRANSMITTERS

by H. Matuschek, DJ 3 MY

1. INTRODUCTION

The following article describes a simply constructed VFO which can be connected to a two metre transmitter instead of a 48 MHz crystal. It is then possible to tune over the whole two metre amateur band.

The tuning of the oscillator - which oscillates in the range of 24.0 MHz to 24.334 MHz - is carried out using a variable capacitor. This is followed by a frequency doubler stage to multiply the oscillator signal to the output frequency of 48.0 MHz to 48.668 MHz. The operating voltage in the range of 12 V to 14 V is stabilized at 9 V using a pass transistor and zener diode; the plus pole was grounded in the prototype. The dimensions of the complete oscillator are 80 x 80 x 60 mm; the weight when constructed using double-coated epoxy board is approximately 120 g.

Special attention was made during the development of this VFO to obtain a compact and robust construction and a very low LC-ratio of the oscillator resonant circuit; the latter is especially important with regard to the frequency stability. The oscillator transistor has been loosely coupled so that oscillation only just occurs. This means that the transistor must exhibit a high gain, which is obtained at a collector current of approximately 2 mA. The Q of the oscillator resonant circuit is dependent on the degree of coupling and thus indirectly on the gain of the transistor. When the frequency is known, the LC ratio of the resonant circuit is limited by the required frequency variation which in turn requires a certain capacitance variation.

These demands lead to the design of the following circuit where the oscillator transistor is loosely coupled so that its dynamic capacity variations have practically no effect on the oscillator frequency. This results in a good short term stability and low voltage fluctuation dependence.

2. CIRCUIT DESCRIPTION

The VFO comprises an oscillator, amplifier and frequency doubler stage. The operating voltage is stabilized with a pass transistor having a zener diode in the base lead. The circuit diagram is given in Fig. 1.

The 24.0 MHz to 24.334 MHz Colpitts oscillator operates in a common-emitter circuit. The relatively high emitter resistor R 4 affects a strong DC feedback and thus a stable operating point of the transistor. This feedback is inoperative in the RF sense because resistor R 4 is bridged by capacitor C 10.

Transistor T 1 is loosely coupled to the resonant circuit via the two capacitive voltage dividers C 5, C 9 and C 6, C 7, C 8. The relatively large capacitance values of the voltage divider are directly parallel to the input or output capacitance of transistor T 1, which means that its capacitance variations can have virtually no effect on the oscillation frequency. The effective capacity of the resonant circuit is formed from the parallel connection of C 1, C 2, C 3, C 4 and the series connection of C 5, C 6, C 7, C 8, C 9. The capacitors C 1 to C 4 contribute approximately 220 pF, whereas the capacitive voltage divider comprising C 5 to C 9 only provides about 15 pF.

The variable capacitor C 1 comprises two gangs with a capacitance variation ΔC of 11 pF each. Since both stators are connected to the resonant circuit, the effective capacitance variation is only approximately 5 pF which provides the required frequency variation. The coupling to the subsequent amplifier stage (T 2) is made at a very low impedance point of the oscillator circuit; this means that load variations will have no great effect on the oscillator frequency.

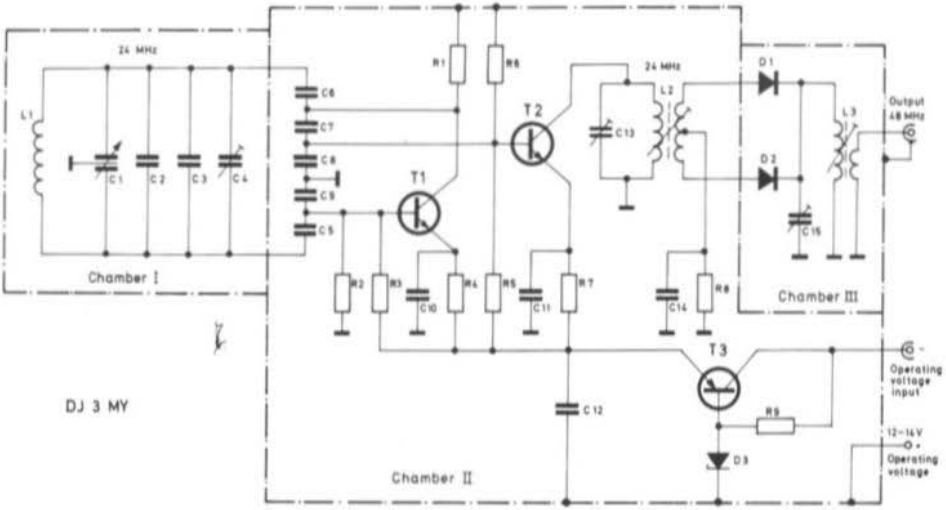


Fig. 1 Circuit of the 48 MHz VFO

The amplifier stage equipped with transistor T 2 is in a common emitter configuration and operates in class A. The output circuit is tuned to 24 MHz; it possesses a push-pull output winding which feeds two diodes whose outputs are joined and fed to a 48 MHz resonant circuit. This configuration represents a true frequency doubler. A good suppression of the fundamental and odd harmonics will occur, if the secondary winding of inductance L 2 is balanced and the diodes are well paired. The diodes operate in class C; the operating points are automatically adjusted by the voltage drop across resistor R 8. This resistor is connected between the centre tap on the secondary winding of inductance L 2 and ground; it is bypassed in the RF sense by capacitor C 14. The most favourable value with respect to the frequency doubling is dependent on the drive voltage.

In the authors prototype, the VFO output signal was tapped off at the "hot" end of the 48 MHz resonant circuit (L 3). This was possible due to the very short connection to the subsequent stage in the transmitter; however, this connection will affect the alignment of inductance L 3. A coupling link, as shown in the circuit diagram, can be used. In this case, it is possible to use longer cables; the disadvantage is, however, that the output voltage is lower.

3. MECHANICAL ASSEMBLY

The cabinet of the VFO is built up from 1 mm thick double copper-coated epoxy board. This material has the advantage that the only tools necessary for construction are a fret-saw and a normal soldering iron. After the soldering process, the cabinet will provide a good screening and be very light. The epoxy also provides a thermal isolation.

The cabinet with the dimensions 72 mm x 52 mm x 62 mm is provided with two partitions (see Fig. 2) so that three chambers are formed. The distribution of the various components in the three chambers is shown by the dashed lines in Fig. 1. The oscillator resonant circuit including the variable capacitor C 1 is accommodated in chamber I separately from the transistor and other circuitry so that external effects are avoided. The 24 MHz balancing transformer L 2, is located below partition I (chassis) in chamber II, where the zener diode and all resistors are also to be found. Chamber III accommodates the three transistor sockets, the doubler diodes, the 48 MHz resonant circuit and the output connector Pt 1. The interconnections between the three chambers are made by means of teflon (PTFE) insulated feedthroughs.

The location of the individual components can be clearly seen in the photographs given in Fig. 3a and 3b. The dimensions of the cabinet pieces and the positions of the main components are given in Figures 2 and 3. The prototype was, however, not equipped with the trimmer capacitors C 4 and C 13; they were replaced by fixed capacitors by the author. Trimmer capacitors are, however, more favourable for the alignment.

3.1. CABINET PIECES AND OTHER PARTS

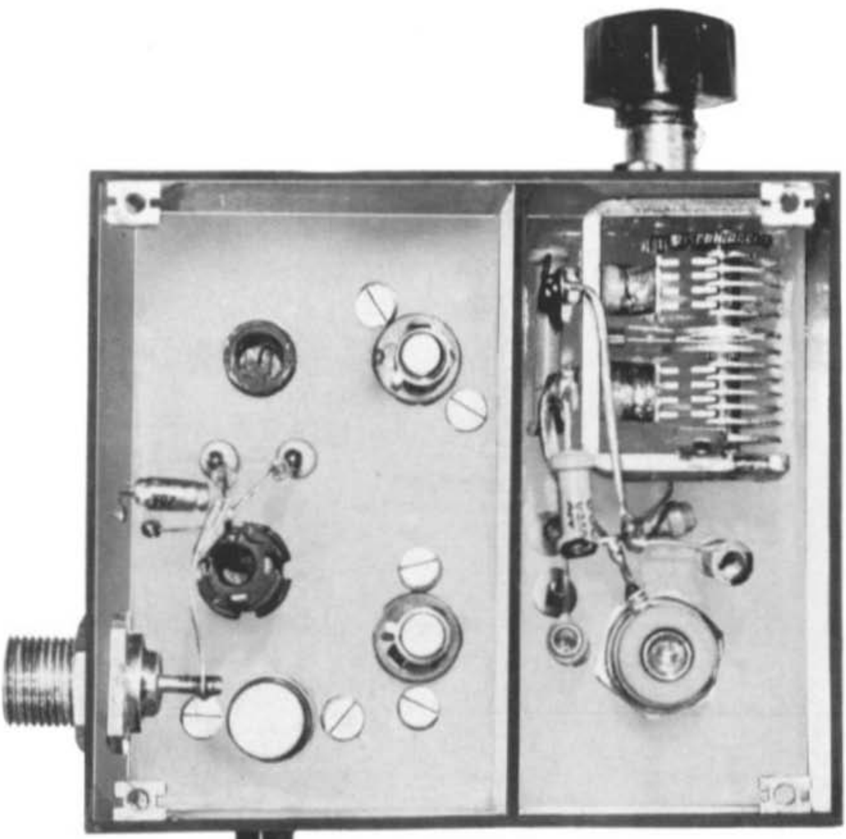
Required number	Item	
1	Output socket (Pt. 1)	
1	Socket for the operating voltage (Pt. 2)	
3	Transistor socket TO 5	
5	Teflon (PTFE) feedthroughs	
1	Side piece I, double-coated epoxy	1 x 72 x 52 mm
1	Side piece II, " " "	1 x 72 x 52 mm
1	Side piece III, " " "	1 x 60 x 52 mm
1	Side piece IV, " " "	1 x 60 x 52 mm
2	Covers, " " "	1 x 72 x 62 mm
1	Partition I, " " "	1 x 70 x 60 mm
1	Partition II, " " "	1 x 60 x 28 mm

3.2. COMPONENTS

T 1, T 2	2 N 918, 2 N 708	R 1 1.5 kΩ	R 5 2.2 kΩ
T 3	AC 122 or similar Ge-pnp	R 2 4.7 kΩ	R 6 8.2 kΩ
	AF-Trans. 30 V/200 mA/130 mW	R 3 2.7 kΩ	R 7 560 Ω
D 1, D 2	OA 90, AA 112, 1 N 54 A	R 4 1.2 kΩ	R 8 2.2 kΩ
D 3	Z 9, OA 126/9, 1 N 960 B		R 9 1 kΩ

All resistors have a rating of 0.1 W

C 1	Two-gang variable capacitor $\Delta C = 2 \times$	11 pF
C 2, C 3	Ceramic tubular capacitor	100 pF
C 4	Trimmer capacitor	20 pF
C 5, C 6	Ceramic tubular capacitor	39 pF
C 7, C 8	Styroflex capacitor	125 pF
C 9	Styroflex capacitor	200 pF
C 10, C 11, C 12	Ceramic capacitor	4.7 nF
C 13	Trimmer capacitor	25 pF
C 14	Ceramic capacitor	1 nF
C 15	Styroflex or ceramic capacitor	25 pF

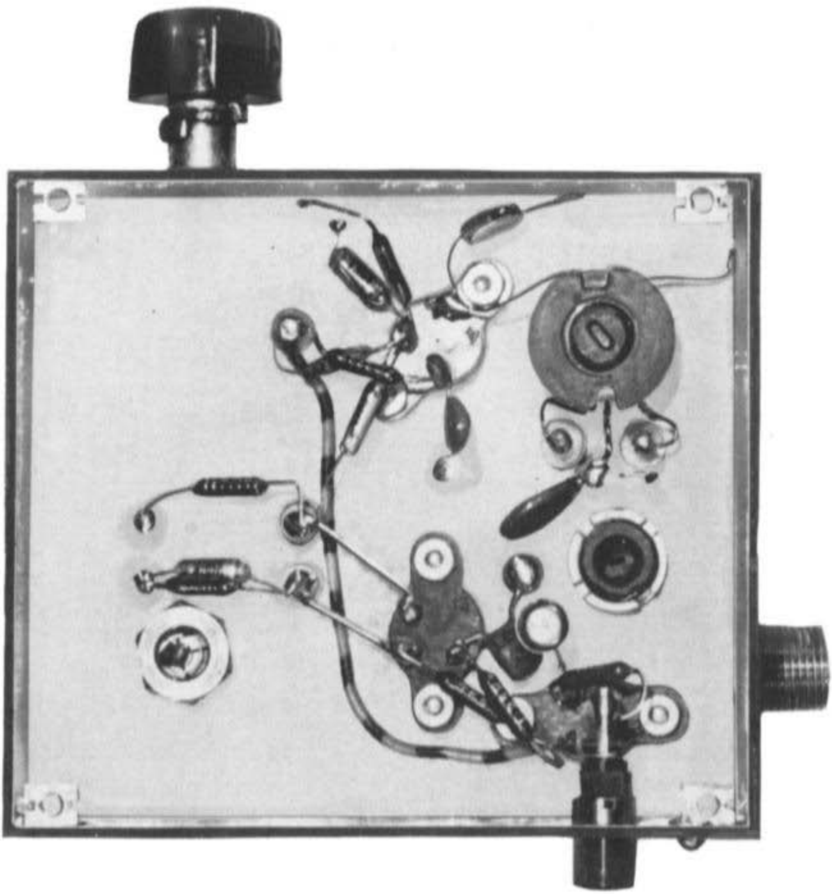


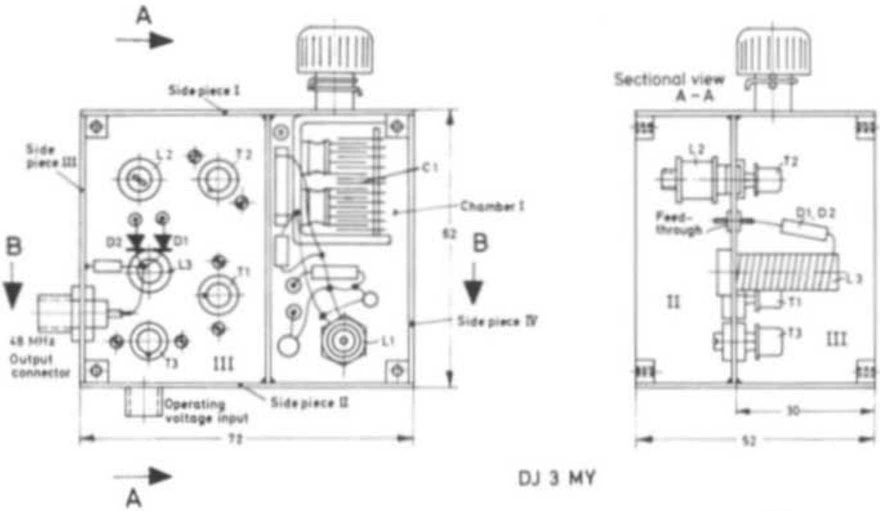
3.3. COIL DATA

L 1 5 turns of 0.5 mm dia. (24 AWG) silver-plated copper wire wound on a ceramic 9 mm diameter former with variable (copper) core. Turn spacing approx. 1 mm.

L 2 Primary: 16 turns of 0.3 mm dia. (29 AWG) enamelled copper wire on a 8.6 mm diameter coil former (height 5.2 mm). Secondary: 40 turns of 0.3 mm dia. (29 AWG) enamelled copper wire with centre tap insulated from the primary with plastic foil. The secondary must be symmetrical to the primary.

L 3 11 turns of 0.4 mm dia. (26 AWG) silver-plated copper wire on a 6 mm dia. coil former. Turn spacing approx. 1 mm. Core: $\mu_R = 12$.





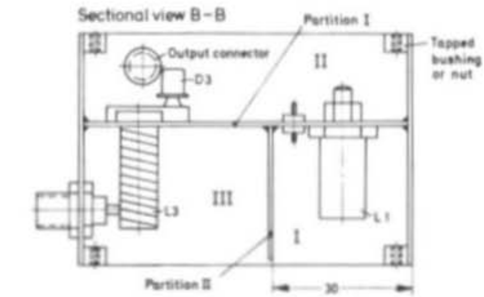
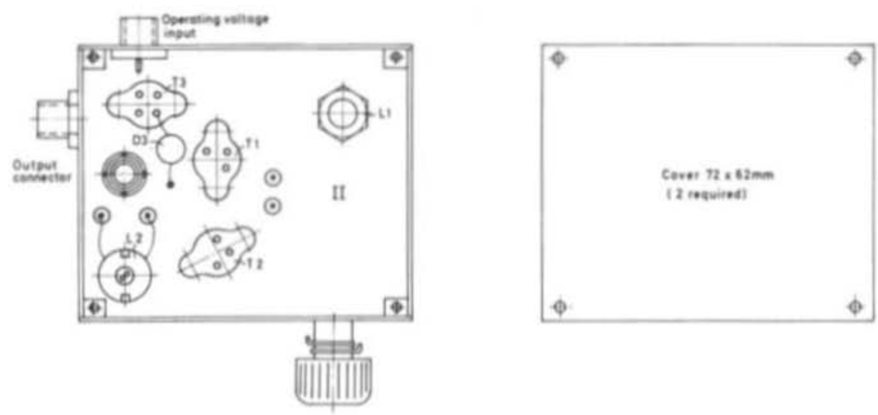


Fig. 2
Mechanical build-up and locations of the main components in the 48 MHz VFO



4. ALIGNMENT

The oscillator frequency must be variable in the range of 24.0 MHz to 24.334 MHz. The authors prototype had a tuning reserve of ± 50 kHz, i.e. the oscillator frequency range was 23.950 MHz to 24.390 MHz. If the nominal frequency greatly differs from this, it will be necessary to reduce or increase the inductivity of L 1; small deviations can be compensated for using the trimmer capacitor C 4.

The temperature-dependent frequency drift noticed during the preliminary measurements was compensated by placing a 1 pF capacitor having a negative temperature coefficient of $750 \times 10^{-6}/^{o}$ C in parallel with the resonant circuit.

After aligning the oscillator to cover the required frequency range, the band centre frequency should be selected and trimmer capacitor C 13 of the amplifier stage aligned for maximum output. This is followed by aligning the output circuit for maximum at 48.33 MHz, which may be made by varying the inductivity of inductance L 3 with the core or by altering the capacitance of capacitor C 13. It is important that the subsequent transmitter is connected when aligning this circuit so that the load capacitance is also taken into consideration during the alignment process.

5. MEASURED VALUES

The following frequency variations were measured on the prototype at 144 MHz under the following conditions:

In the first 10 minutes after switching on :	850 Hz
In the first 30 minutes after switching on :	2.15 kHz
On reducing the operating voltage by 15% :	206 Hz
On shorting the output to ground :	30 kHz
After switching on after a period of 24 hours :	12 kHz

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CALIBRATION-SPECTRUM GENERATOR FOR TWO METRES

by H. Götting, DL 3 XW

INTRODUCTION

Harmonics of crystal oscillators are often used for the calibration of short-wave receivers. Usually, a frequency of 100 kHz is chosen and the crystal pulled to the correct frequency after comparison against a standard frequency, such as Droitwich (200 kHz) or one of the WWV transmitters (5 MHz, 10 MHz, 20 MHz).

The harmonics of such spectrum generators are, however, not usually audible on the two metre band. The described 1 MHz spectrum generator therefore comprises a selective amplifier stage for the frequency range of interest, i. e. the two metre band.

1. CIRCUIT PRINCIPLES

The circuit consists of a 1 MHz oscillator whose 144th, 145th and 146th harmonics fall into the two metre band. A subsequent amplifier is provided to amplify that part of the spectrum which falls between 144 and 146 MHz.

A frequency spectrum of 1 MHz was chosen for this spectrum generator because it was considered that a 100 kHz spectrum could cause some doubt as to which harmonic the receiver was tuned to; this is especially true when calibrating new homebuilt or somewhat unstable receivers. The spacing between the individual lines of the main spectrum therefore corresponds to 1 MHz and is audible at 144 MHz, 145 MHz and 146 MHz. This provides an easy means of determining the band limits.

Since a 1 MHz spectrum is too coarse for a number of calibration applications, a 100 kHz spectrum can also be selected. This second spectrum is obtained using an astable multivibrator which is synchronized by the 1 MHz oscillator. This synchronization ensures that the frequency of the 100 kHz spectrum has the same accuracy and stability as that of the 1 MHz spectrum.

The subsequent amplifier stage can, of course, be tuned to any other frequency range of interest. If the spectrum generator is only to be used in the short-wave range, the amplifier stage may be deleted.

2. CIRCUIT DESCRIPTION

The circuit diagram of the calibration spectrum generator is given in Figure 1. The astable multivibrator comprising tube V 1 can be switched on by closing switch S 1. The oscillation frequency of approximately 100 kHz is determined by the RC values and can be somewhat varied by adjusting potentiometer P 1.

The 1 MHz crystal oscillator comprises the tube V 2; the screen grid of this tube represents the anode of the oscillator circuit. The 1 MHz oscillation also appears at the suppressor grid of this tube; from here it is fed to the multivibrator circuit where it actuates each tenth deflection of the multivibrator. This represents the synchronization process. The 100 kHz pulses from the multivibrator circuit are fed to the suppressor grid of the oscillator tube V 2 where they are mixed with the 1 MHz spectrum. The harmonic content is formed by operating this tube in class C (grid current).

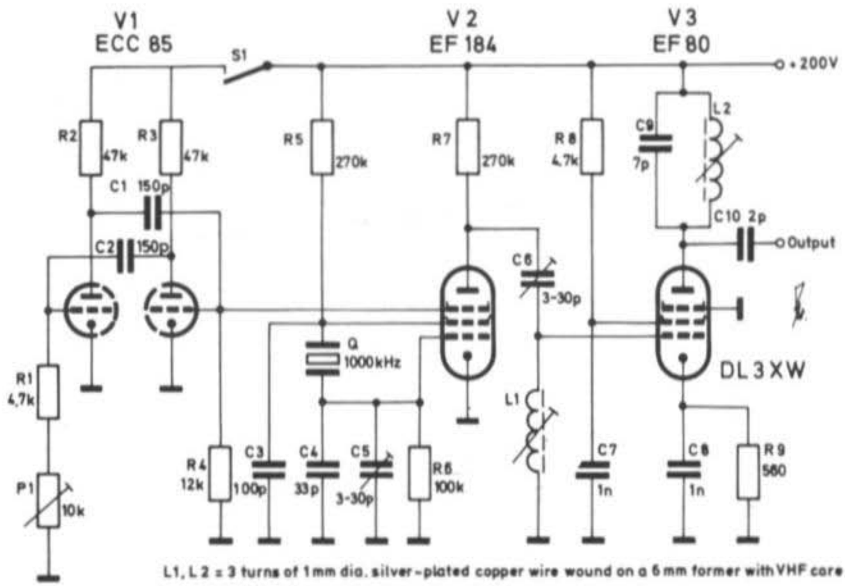


Fig. 1   A calibration spectrum generator for two metres

The class A amplifier stage comprising tube V 3 possesses a resonant grid circuit which is formed by the inductance L 1 and the input capacitance of the tube. The whole harmonic spectrum is coupled at high impedance to the amplifier via trimmer capacitor C 6. A second resonant circuit comprising inductance L 2 and capacitor C 9 is to be found in the anode lead of tube V 3. These two resonant circuits are tuned to the frequency range of interest - in our case, the two metre band. The spectrum is fed out at high impedance via capacitor C 10.

3. CONSTRUCTION DETAILS AND ALIGNMENT

The values of all components are given in Fig. 1; the assembly is not critical and can be built-up as desired. The only points that should be noted are that all wiring subsequent to the anode of tube V 2 should be made low-capacitively and inductively in "VHF manner". Since tube V 3 is a VHF amplifier, it is necessary to provide the same amount of screening as for converters etc.

The alignment process is commenced by pulling the crystal frequency to zero-beat against a standard frequency (e.g. WWV). This is achieved by aligning capacitor C 5 until zero-beat occurs between the standard frequency and the corresponding harmonic of the spectrum generator. This is followed by tuning a two metre receiver to the 145th harmonic (145 MHz) of the 1 MHz spectrum. Inductance L 2, and subsequently, inductance L 1 and capacitor C 6 are aligned alternately for maximum S-meter reading on the receiver.

The astable multivibrator is now switched into circuit and the two metre receiver tuned over the band. At first, alternating crystal-controlled, and self-excited oscillations will be heard between the 1 MHz harmonics at 144, 145 and 146 MHz. Trimmer potentiometer P 1 is now adjusted so that only 9 crystal-controlled oscillations can be heard between each 1 MHz line of the spectrum. At this point the multivibrator is synchronized. Potentiometer P 1 is now carefully varied to determine the limits of the synchronization range, and adjusted to a central position between these limits.

4. TUBE COMPLEMENT AND EQUIVALENTS

V 1	=	ECC 85, 6 AQ 8 or 6 L 12
V 2	=	EF 184, 6 EJ 7
V 3	=	EF 80, 6 BX 6

MODIFICATION OF THE DJ 9 ZR 001 5 W SSB TRANSMITTER

by H. U. Beitz, DJ 8 XR

During construction of the transistorized 5 W SSB transmitter described in (1), some improvements were found which are now to be described.

1. IMPROVEMENT OF THE AUXILIARY FREQUENCY SUPPRESSION

The 9 MHz SSB signal is converted to the 145 MHz range in the push-pull mixer comprising the transistors T 5 and T 6. The suppression of the auxiliary frequency (135-137 MHz) is dependent on the balance of the push-pull mixer. This balance can be improved if the base circuit inductance L 4 is wound onto the resonant inductance L 3 in a bifilar (double wound) manner. No other modifications are required.

2. INCREASING THE GAIN

It was not only the author who found it difficult to achieve the claimed output power. This seemed, at least partly, to be caused by the great fluctuations between individual transistors of type BF 224. After selection of those BF 224 transistors having the highest gain, it was possible to obtain the given output power from the transmitter. A further improvement in this direction resulted by close winding inductance L 5.

3. OTHER PA TRANSISTORS

At the moment, transistors type 2 N 3632 are available at a modest price. They are suitable for use in the PA stage of this transmitter. The author observed linearity difficulties when this transistor was aligned to the operating point given in (1) for the 2 N 3375. This was cured by replacing the 30 Ω resistor between the base of transistor T 9 and ground by a 75 Ω resistor.

4. REFERENCES

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

A CALIBRATION-SPECTRUM GENERATOR FOR TWO METRES

by H. Götting, DL 3 XW and D. E. Schmitzer, DJ 4 BG

1. INTRODUCTION

Simple calibration-spectrum generators usually consist of a crystal-controlled oscillator for 1 MHz or 100 kHz and a stage having a high harmonic distortion for the generation of the harmonic spectrum. This harmonic spectrum can reach into VHF-range and thus be used to calibrate the scales of VHF receivers. The frequency or corresponding harmonic of the crystal-controlled oscillator is compared to that of a standard frequency such as the longwave transmitter Droitwich (200 kHz) or one of the frequency standard transmissions, for instance WWV (5 MHz, 10 MHz, 20 MHz). The resulting superheterodyne tone is pulled into zero-beat by means of a trimmer.

1 MHz crystal-controlled oscillators are usually used when higher demands are made on the frequency stability of the calibration spectrum. This also offers the advantage that the frequency of an harmonic can be determined more easily. This is especially true for high-order harmonics such as those used for the calibration of VHF equipment. However, such a 1 MHz spectrum is too coarse for a number of applications. In order to achieve calibration points having a spacing of 100 kHz, the described calibration spectrum generator is provided with an additional frequency divider that can be switched on as required. The frequency divider generates a 100 kHz oscillation from the 1 MHz crystal-controlled signal. The 100 kHz signal is also fed via the stage having a high harmonic distortion factor in order to generate the required harmonic spectrum. The 100 kHz signal and harmonics thereof possess the same frequency accuracy and stability as the 1 MHz signal since they are obtained by frequency division from the crystal-controlled frequency.

The generator is also provided with a tuned amplifier stage for the amplification of the calibration spectrum in the frequency range of interest. Reference (1) described a calibration-spectrum generator for the two metre band and explained this principle in more detail. The described generator (1) was equipped with a 1 MHz crystal and three tubes. One of the tubes (a double triode) operated in a multivibrator circuit which was synchronized from the crystal-controlled oscillator. The multivibrator divided the 1 MHz crystal-controlled frequency by ten which resulted in a frequency of 100 kHz. A similar, silicon transistorized calibration-spectrum generator is now to be described.

2. CIRCUIT DESCRIPTION

The circuit diagram of the transistorized calibration-spectrum generator is given in Figure 1. The crystal-controlled oscillator equipped with transistor T 1 is followed by the common-collector buffer T 2. The calibration spectrum is then amplified in two tuned amplifier stages that are designed to accentuate the signal in the frequency range of interest. The collector circuits of transistors T 3 and T 4 are tuned to the required frequency range. The high-pass filter comprising inductance L 3 further accentuates the calibration spectrum harmonics in the VHF range so that no unwanted injection into the intermediate frequency range can occur.

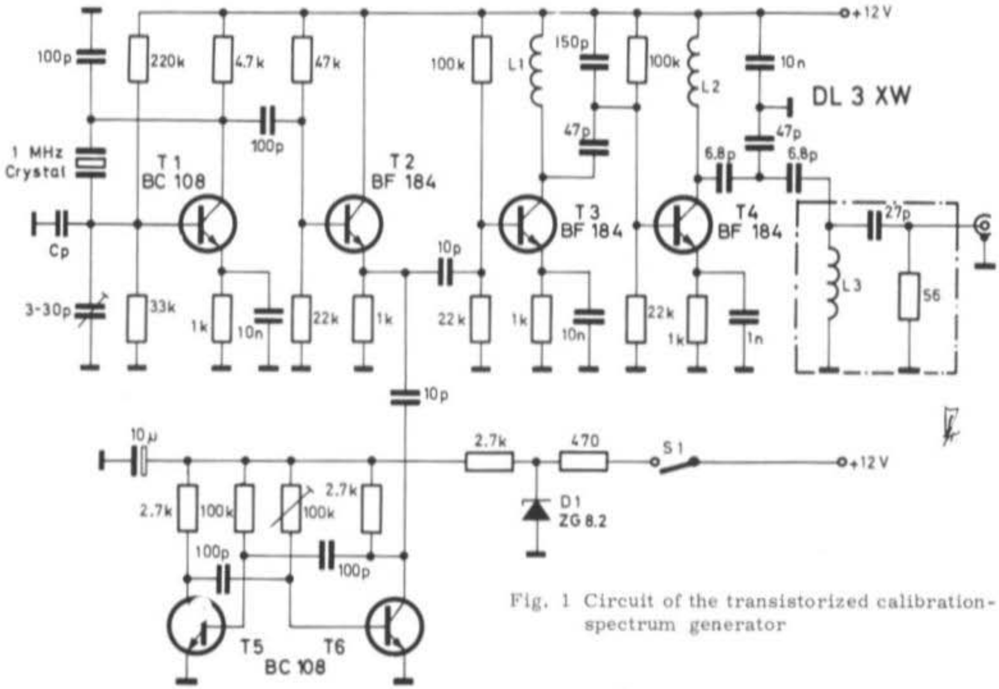


Fig. 1 Circuit of the transistorized calibration-spectrum generator

The 100 kHz calibration lines are generated when the multivibrator circuit comprising T 5 and T 6 is switched into operation. The time constant of the RC-links results in a 100 kHz deflection frequency. Each tenth cycle of the multivibrator is synchronized by the 1 MHz crystal-controlled oscillator signal. This means that the 100 kHz spectrum possesses the same stability as the crystal-controlled frequency.

In order to synchronize the frequency, the frequency of the multivibrator can be varied using a trimmer potentiometer in a feedback link. The operating voltage of the multivibrator circuit is stabilized using a 8.2 V zener diode so that operating voltage fluctuations cannot affect the synchronization.

The parallel capacitor C_p connected in parallel with the crystal alignment trimmer capacitor (of 3 - 30 pF) should be added if the nominal frequency is still not obtained at the maximum capacitance of the trimmer. Most amateur crystals are designed for a load capacitance of 30 pF at the parallel resonance point. However, it is possible that a 30 pF trimmer capacitor in series with the 100 pF capacitor at the other side of the crystal will represent too low a load capacitance. This can be avoided in two ways: firstly by exchanging the 30 pF trimmer capacitor for one of 10 - 45 pF or by adding the parallel capacitor C_p whose value should be in the order of 10 - 40 pF.

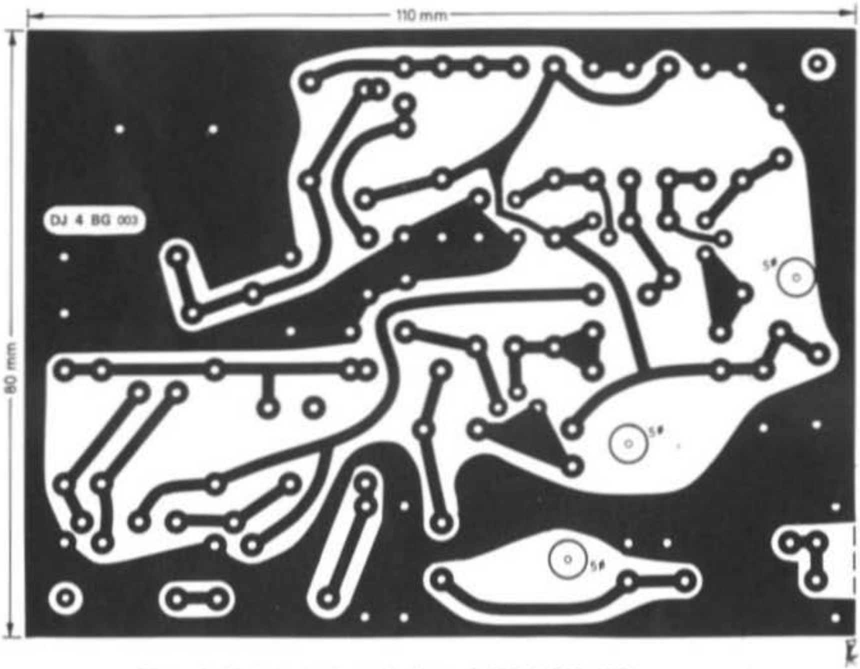


Fig. 2 Printed circuit board DJ 4 BG 003

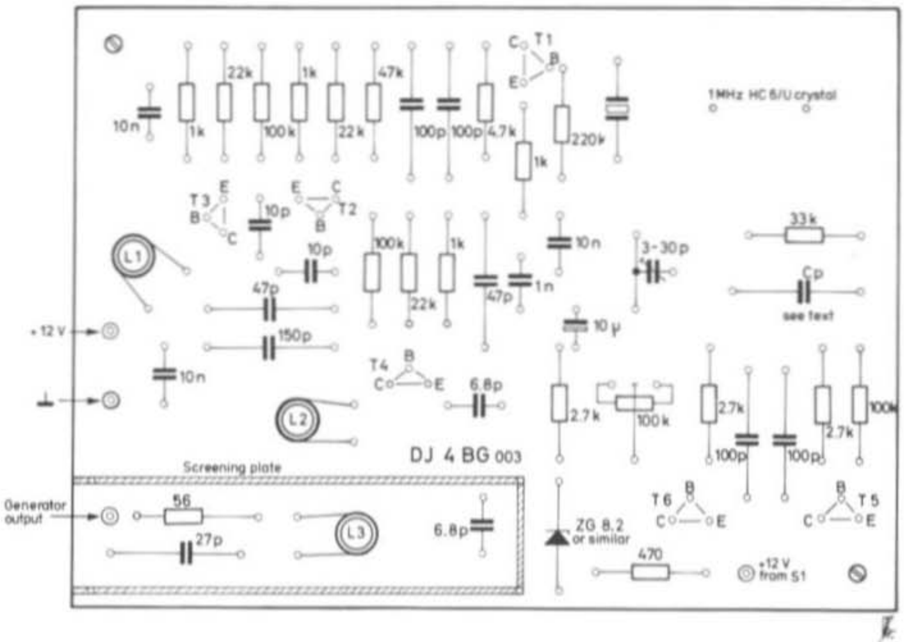


Fig. 3 Component location plan of the calibration-spectrum generator

3. MECHANICAL ASSEMBLY

The transistorized calibration-spectrum generator can be built up on the 80 mm by 110 mm printed circuit board DJ 4 BG 003 shown in Figure 2. The corresponding component location plan is given in Figure 3. High quality components should be used for the crystal oscillator circuit. No screening is necessary for the two resonant circuits. It is only the output high-pass filter that should be screened so that unwanted injection is avoided. A spacing of 15 mm was selected for the resistors so that normal types could be used. Greater hole spacings have also been allowed for the capacitors which means that standard Styroflex and ceramic capacitors can be used. The author used ceramic disc bypass capacitors whereas the small bead type capacitors were more suitable for use as coupling capacitors (6.8 pF and 10 pF). A tantal 10 μF/10 V capacitor was used to bypass the multivibrator.

3.1. SPECIAL COMPONENTS

T 1 : BC 108, BC 183, 2 N 2926, 2 N 3903, 2 N 708 or similar
T 2 to T 4 : BF 184, BF 115, AEG-Tfk, or BF 224 (TI)
T 5, T 6 : BC 108, BC 183, 2 N 2926, 2 N 3903, 2 N 708 or similar
D 1 : ZG 8.2, ZF 8.2, OA 126/8, BZY 85/C 8 V 2, 1 N 959 B
L 1 : 35 turns of 0.2 mm dia. (32 AWG) enamelled copper wire wound on a 5 mm dia. coilformer
L 2 : 5 turns of 1 mm dia. (18 AWG) silver-plated copper wire, otherwise as L 1
L 3 : 3 turns, otherwise as L 2

4. ALIGNMENT

After having pulled the crystal-controlled frequency to the nominal value, it is only necessary to synchronize the multivibrator circuit. This is made in a similar manner to that described in (1): The calibration spectrum is monitored on a two metre receiver and the trimmer potentiometer adjusted until only nine synchronized spectrum lines are audible between the 1 MHz calibration points at 144 MHz, 145 MHz and 146 MHz. If synchronization has not occured, additional unstable signals will be observed.

5. EDITORIAL NOTE

The author (DJ 4 BG) is developing a versatile calibration-spectrum generator equipped with integrated circuits that will allow switchable spectrum lines with a spacing of either 1 MHz, 500 kHz, 100 kHz, 50 kHz, 10 kHz, 5 kHz and 1 kHz. This spectrum generator will, of course, be far more expensive than the generator described here.

6. AVAILABLE PARTS

The printed circuit board DJ 4 BG 003 as well as the 1 MHz quartz crystal, the trimmer, coilformers and the semiconductor complement are available from the publisher (See advertising page).

7. REFERENCES

(1) H. Götting: A Calibration-Spectrum Generator for Two Metres
VHF COMMUNICATIONS 2 (1970), Edition 1, see contents

SIMPLE, COMPACT PA STAGES FOR TWO METRES

by D. Grossmann, DJ 4 RX

INTRODUCTION

The advantages and disadvantages of LC and resonant line circuits at higher frequencies are to be considered. Special attention has been paid so that the PA stages exhibit a sufficient bandwidth. As a result of these considerations, construction details will be given that allow two metre PA stages to be built-up with a modest amount of metalwork.

The described designs refer to a single-ended final amplifier stage equipped with a tube of the 4 X 150/4 X 250 family.

The adjustment of the output coupling is described in detail.

GENERAL REQUIREMENTS

A two metre PA stage should possess the following characteristics:

a. A high efficiency of the PA circuit
b. Sufficient bandwidth; it should not be necessary to retune to cover the whole two metre band.
c. A high harmonic suppression.
d. The lowest possible dimensions together with the least possible metalwork.

Of course, all of these requirements cannot be fulfilled simultaneously. However, satisfactory designs can be obtained according to which of the given requirements is considered to be most important. The PA circuit should possess certain safety features. For instance, the screen grid supply should be designed so that the screen grid is not damaged or destroyed should a failure of the anode voltage occur.

1. CIRCUIT EFFICIENCY AND Q

The Q of a resonant circuit is defined as the relationship between reactive power P_r and active power P_a (Equation 1).

$$\textbf{Equation 1} \qquad Q = \frac{P_r}{P_a}$$

Whereas the unloaded Q (Q_o) only takes the losses of the circuit itself into consideration, the operative Q_{op} results from the reactive power and the sum of the losses P_l and the output power P_{out} (Equations 1a and 1b).

$$Q_0 = \frac{P_r}{P_l} \quad (1a) \qquad\qquad Q_{op} = \frac{P_r}{P_l + P_{out}} \quad (1b)$$

The efficiency η of the circuit accords to equation 2 :

$$\eta = \frac{Q_0 - Q_{op}}{Q_0} = \frac{\frac{Q_0}{Q_{op}} - 1}{\frac{Q_0}{Q_{op}}} \qquad (2)$$

Equation 2 shows that Q_o/Q_{op} should be as great as possible. However, since the unloaded Q of a resonant circuit cannot be increased beyond a certain point, one must attempt to obtain a very low operative Q. It will be seen that internal tube capacitances will place a lower limit for the operative Q. The circuit is loaded with the output impedance of the PA tube.

As an example, let us consider a linear amplifier equipped with a 4 X 150 tube operating in class AB_1. It is assumed that the anode voltage U_a is 1250 V and the anode current I_a is 200 mA.

The characteristic curves show that the most favourable output impedance Z_{out} is 3.5 kΩ.

The following would result for a LC circuit:

$$Q_{op} = \frac{Z_{out}}{X_C} \qquad (3)$$

With: $C = C_{out} + C_1$

Where: C_{out} is the output capacitance of the tube
C_1 is the tuning capacitance.

If the tuning capacitance is not taken into consideration, the following will be valid for $C = C_{out}$:

$$Q_{op\,min} = Z_{out} \times \omega \times C_{out} = 15 \text{ (at } \omega = 2\pi \text{ 145MHz)} \qquad (4)$$

In order to achieve a circuit efficiency of 95%, the inductance of the PA circuit must have an unloaded Q of at least $Q_o = 300$. This clearly shows the limits of LC circuits since it is very difficult to construct such a coil which does not only have to possess the required Q but also has to provide a good heat dissipation.

Unfortunately, a resonant line circuit also possesses additional capacitances, which means that the operative Q will be increased. Let us now make the simplified assumption that the shortening capacitance of the line is only formed by the tube output capacitance C_{out}. The following is then valid for resonance:

$$\omega_r C = \omega_r C_{out} = \frac{1}{Z} \times \cot(\beta l) \qquad (5)$$

Where: ω_r is the resonant frequency.

For a given resonant frequency, it is possible to determine the impedance Z as a function of the electrical length 1 of the resonant line, or vice versa, from equation 5. A LC circuit is now to be calculated that possesses the same characteristics as the resonant line circuit directly adjacent to its resonant frequency.

The equivalent circuit is described by:

$$\omega_r^2 = \frac{1}{L'C'} \qquad (6)$$

In order to obtain the second conditional equation, it is necessary to derive susceptance (B) as a function of the frequency. The following is valid for the LC circuit:

$$\frac{dB_1}{d\omega} = \frac{d}{d\omega}\left(\omega C' - \frac{1}{\omega L'}\right) = C' + \frac{1}{\omega^2 L'} \qquad (7)$$

If the resonance condition is considered, the following will be valid:

$$\left.\frac{dB}{d\omega}\right|_{\omega_r} = 2\,C' \qquad (7a)$$

The same calculation process is now to be carried out for a capacitively shortened resonant line circuit. The susceptance of the resonant line circuit is:

$$B_2 = \omega C - \frac{1}{Z}\cot\left(\beta_0 l\,\frac{\omega}{\omega_r}\right) = \omega C - \frac{\omega_r C}{\cot(\beta_0 l)}\cot\left(\beta_0 l \times \frac{\omega}{\omega_r}\right) \qquad (8)$$

After an intermediate calculation, the following results:

$$\text{Equation 9} \quad \left.\frac{dB_2}{d\omega}\right|_{\omega_r} = C\left(1 + \frac{2\beta_0 l}{\sin(2\beta_0 l)}\right) \qquad (9)$$

The operative Q of the resonant line circuit is directly proportional to the equivalent capacitance C'. In Fig. 1, the equivalent capacitance C', the 1 dB bandwidth and the operative Q of a 4 x 150 tube are drawn as a function of the electrical length of a resonant line circuit. It will be seen that the bandwidth of the resonant line will decrease on increasing the length of the line.

The greatest bandwidth is achieved with $\beta l = 0$, e.g. with $Z = \infty$.
This corresponds exactly to a simple LC circuit.

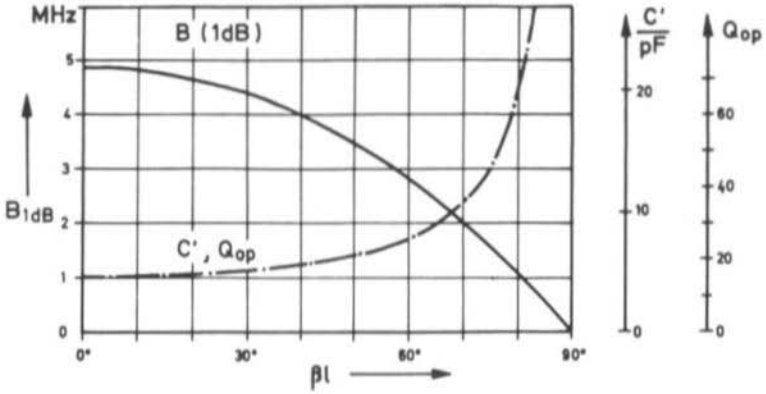


Fig. 1
Equivalent capacitance C, 1 dB bandwidth and Q as a function of the electrical length of the coaxial line circuit.

2. THE BANDWIDTH PROBLEM

It may seem paradox to demand the highest possible bandwidth or a low Q from a power amplifier stage. However, it has been seen that a PA stage is not suitable for providing the adjacent selectivity - even at an operative Q of, for instance, $Q_{op} = 100$. In this case, the 3 dB bandwidth would still be 1.5 MHz on the two metre band and the transmitter would transmit any distortion products that occur within this band with virtually no rejection. This means that all spurious signals must be rejected in the previous stages where losses of a few dB are not important. In these stages, it is not too important if the circuit efficiency is 50% or less.

It is also important that the PA stage is not overdriven in the single sideband mode. Even the best PA circuit cannot suppress splatter spaced only 200 or 500 kHz from the carrier frequency. This is also valid for antenna filters etc.

As an example, a PA circuit is now to be calculated that exhibits an attenuation of 20 dB 500 kHz from the resonant frequency. The standardized frequency Ω is:

$$\Omega = \frac{\omega}{\omega_0} = \frac{f}{f_0} \qquad (10)$$

and the impedance of the PA circuit is :

$$Z = \frac{Z_{out}}{1 + jQ\left(\Omega - \frac{1}{\Omega}\right)} \qquad (11)$$

For a frequency f_o of 145 MHz and $f = 0.5$ MHz, Ω is $1 + 1/290$. When solving for Q, the following will result:

$$Q = \frac{10}{\Omega - \frac{1}{\Omega}} = 1450 \qquad (12)$$

An operative Q of 1450 corresponds, at a circuit efficiency of 90%, to an unloaded Q (Q_0) of 14,500, which is unattainable in practice.

According to these considerations, it will be seen that the most important demands for the construction of resonant line PA circuits is a short electrical length or, in other words, the highest possible impedance. A short electrical length also leads to compact mechanical dimensions which in turn satisfies point "d" of the general requirements. If should be noted that the 3 dB bandwidth is not suitable for classifying a PA stage since the power loss of the PA tube would amount to non-permissible values at the bandlimits. A demand for a power loss of 1 dB at the bandlimits is more favourable. This 1 dB bandwidth corresponds to approximately half the 3 dB bandwidth.

3. HARMONIC SUPPRESSION

If a PA stage is only to be used in the A 1 or A 3 mode, it would be possible for the harmonic suppression to made in a subsequent low-pass filter as long as one is willing to accept the virtually unimportant decrease in the anode efficiency and if the PA stage is "RF-tight". A linear amplifier on the other hand, is far more sensitive with respect to the efficiency and linearity.

Figure 2 shows the simplified I_a - U_a characteristic curves of a tetrode or pentode tube. Under full-drive conditions, the tube is driven up to the immediate vicinity of the bend in the characteristic curve. If the AC voltage at the anode contains noticeable harmonic components, unwanted conversion products (intermodulation) will be generated by the non-linearity of the characteristic curves; this is the main cause of splatter. The only remedy of this is for the drive to be reduced, which means decreasing the efficiency by the same value.

The PA circuit should, therefore, be designed for the highest harmonic rejection. It is possible, for a final amplifier that works satisfactorily in the telegraphy mode, to be completely unusable as a linear amplifier. Since resonant line circuits possess a great number of resonant frequencies, it is important that a check should be made to find out if one of the resonances also corresponds to a harmonic of the wanted frequency.

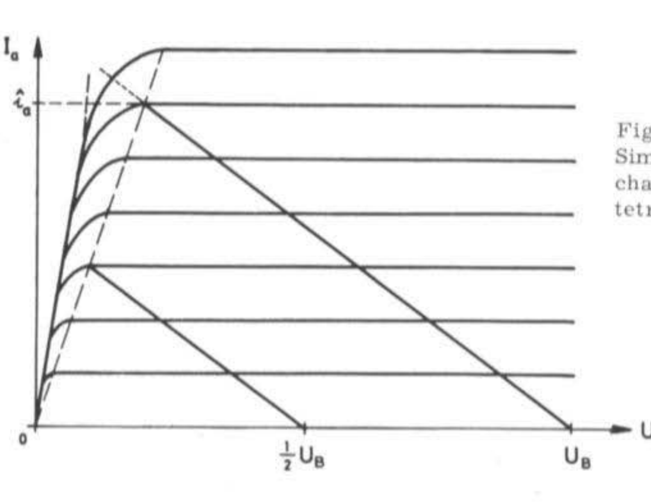
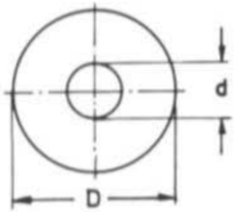
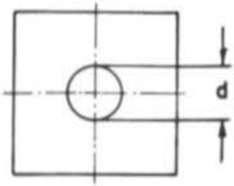


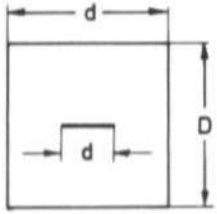
Fig. 2
Simplified I_a - U_a characteristics of a tetrode or pentode tube



$$Z = 60 \ln \frac{D}{d}$$

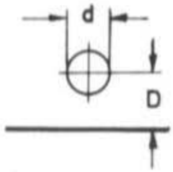


$$Z = 60 \ln\left(1.08 \frac{D}{d}\right)$$

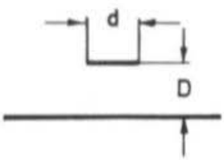


$$Z = 60 \ln\left(2.16 \frac{D}{d}\right)$$

$$\frac{D}{d} > 2$$



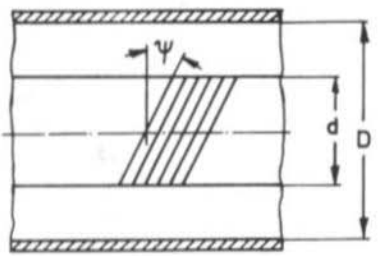
$$Z = 60 \ln \frac{2D}{d}$$



$$Z = 60 \ln \frac{7D}{d}$$

$$\frac{D}{d} > 2$$

Table 1

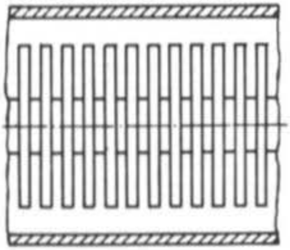


$$Z_h = Z \times F_h$$

$$F_h = \left(1 + \frac{1 - \left(\frac{d}{D}\right)^2}{2 \ln \frac{D}{d}} \cot^2 \psi\right)^{\frac{1}{2}}$$

$$\cot \psi = n \pi d$$

Radial slottet lines



Meander lines

Table 2

4. SELECTION OF A SUITABLE RESONANT LINE

It is true that higher impedance values can be achieved easier with balanced lines, however, the following considerations are limited to unbalanced (coaxial) lines.

The impedance values Z of various unbalanced RF lines are given in Table 1.

At impedance values of higher than approximately 120 Ω, the inner conductor is very small in comparison to the outer conductor which means that the current density and losses are too great. It is, however, still possible to use inner conductors having a diameter of approximately 10 mm. Such a PA circuit is to be described. In order to ensure that the inner chamber is accessible, a square outer conductor is more favourable than a tubular type. The tuning disc and the coupling loop are mounted on a metal plate that can be unscrewed. The rest of the outer conductor can be soldered, providing that it is made out of copper plate and as long as provision has been made to allow replacement of the PA tube.

There are, of course, further resonant line types which simultaneously allow a high impedance and a reduced phase velocity to be realized. Helical lines, and other constructions that are often used as delay lines, such as coaxial lines with radially slotted inner and outer conductors, similar strip and meander lines, also belong to this category (See Table 2). The impedance of a helical line Z_h is given by equation 13a; the helical factor F_h can be calculated according to equation 13b:

$$Z_h = Z \times F_h \qquad (13)$$

$$F_h = \left(1 + \frac{1 - \left(\frac{d}{D}\right)^2}{2 \ln \frac{D}{d}} \cot^2 \psi \right)^{\frac{1}{2}} \qquad (13a)$$

$$\cot \psi = n \times \pi \times d \qquad (13b)$$

Helical lines are the most interesting types for the construction of coaxial PA stages since they allow the highest impedance and lowest phase velocity to be achieved. Due to the helical form, it is possible for the inductance component of the line to be increased virtually at will without altering the capacitance component noticeably.

In the same way that the impedance is increased with respect to a normal inner-conductor, the propagation speed of the electrical waves along the helical line will be reduced. A helical line is thus longer in the electrical sense than a normal line having the same mechanical length. A helical resonant line circuit that is capacitively loaded, represents, to a certain extent, a link between the simple LC circuit and the normal resonant line circuit. Figure 3 indicates the increase of impedance and phase component as a function of the helical pitch. Each individual operating frequency corresponds to a certain optimum helical pitch. Unfortunately, the author has not been able to find a source giving the information with regard to the most favourable dimensioning of such resonant line circuits.

Detailed information about resonant lines is given in references (1), (2), (3); information about helical lines is given in references (1), (2) and especially in (4) to (5).

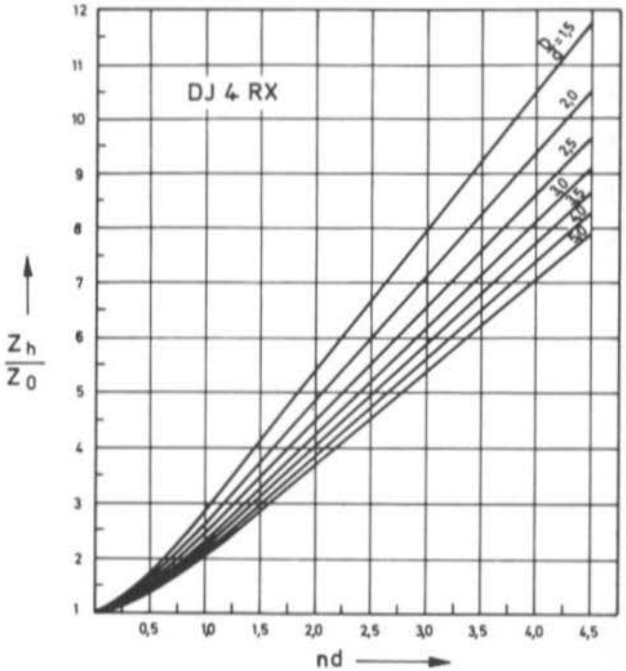


Fig. 3
Increase of the impedance and phase value as a function of the of coaxial line circuits

D = dia. of outer conductor
d = dia. of inner conductor
n = number of turns per length unit

5. CONSTRUCTION OF RESONANT LINE PA STAGES

5.1. A NORMAL RESONANT LINE PA STAGE

The first constructional example is the 145 MHz PA stage given in Fig. 4. This final amplifier does not require a great deal of mechanical construction. The resonant line circuit itself consists of the line, shown as piece 4a in Fig. 4, having a high impedance (inner conductor: 10 mm diameter, outer conductor 80 x 80 mm^2, Z approx. 130 Ω) and a short strip line (Fig. 4b) having a low impedance which is additionally capacitively loaded with a tuning disc (Fig. 4c). The bypass arrangement shown in Figure 4f is mounted on the anode side (hot end) of the circuit. If copper plate is used for the casing shown in Fig. 4d, it is possible, with the exception of the covering plate shown in Fig. 4c, for all parts to be soldered. The anode voltage is fed through the 10 mm diameter inner conductor (Fig. 4a) using a piece of coaxial cable.

The casing or outer conductor of the $\lambda/4$ resonant line circuit (see Fig. 4d) is made from 1 mm soft copper plate. The cabinet is tightly soldered at the edges. The tube base for the 4 X 150 or 4 X 250 is mounted in the 56 mm diameter cutout on the base of the casing. The inner conductor extension (Fig. 4b) is screwed to the flattened end of the inner conductor (Fig. 4a), as is the bypass arrangement comprising two pieces of part 4fa, one piece each of part 4fb and 4fc (see Fig. 4g). Plate 4b is screwed into the hole located above the tongue of the inner conductor (Fig. 4a). This is followed by placing one teflon (PTFE) disc (part 4fa), plate 4fb (after bending up the solder tag T and the three tongues), a second teflon disc 4fa and plate 4fc one above the other and screwing them together with three screws. Plate 4fb

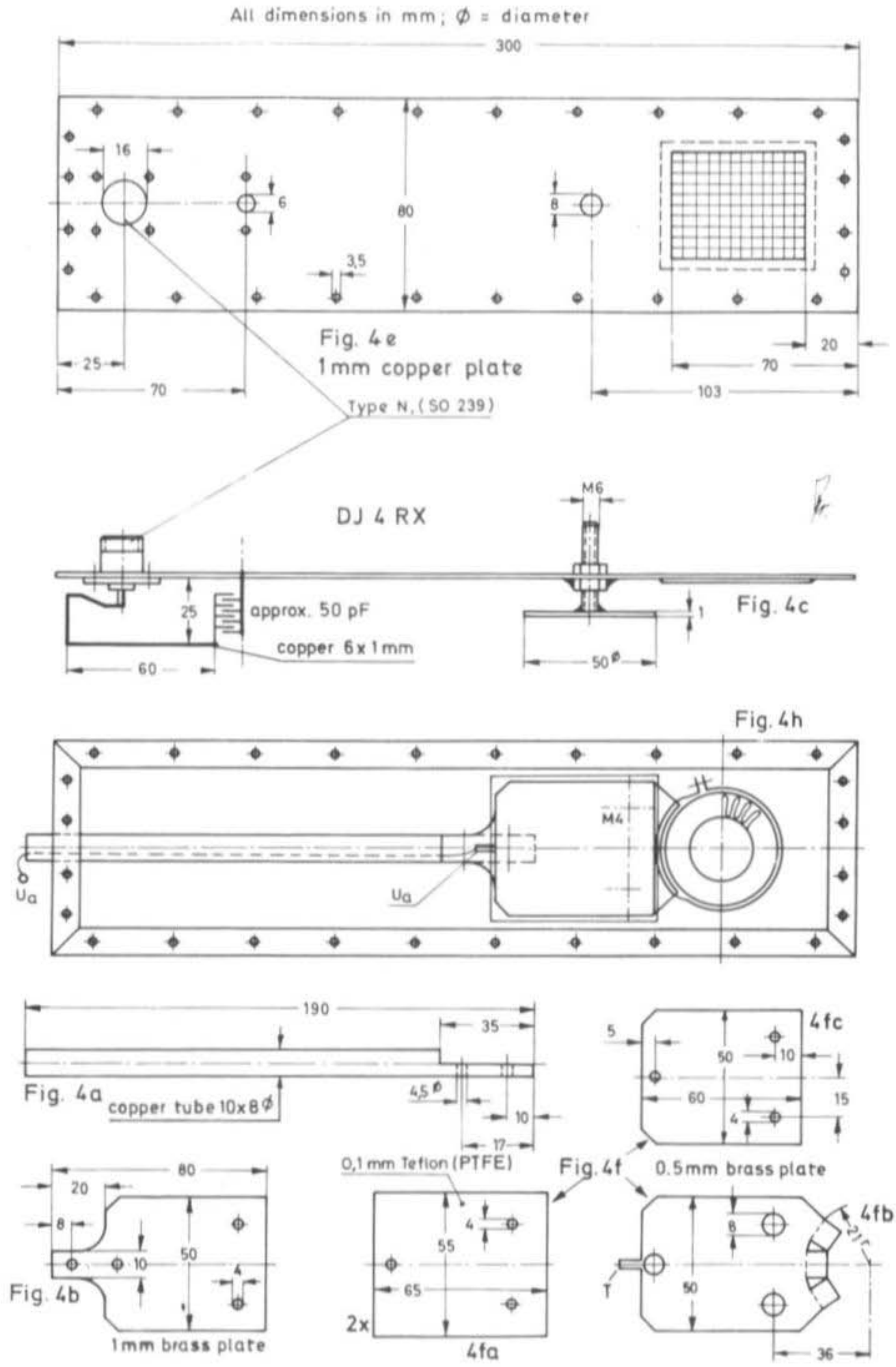


Fig. 4 Individual pieces of the 144 MHz power amplifier equipped with a 4 x 150 A tube

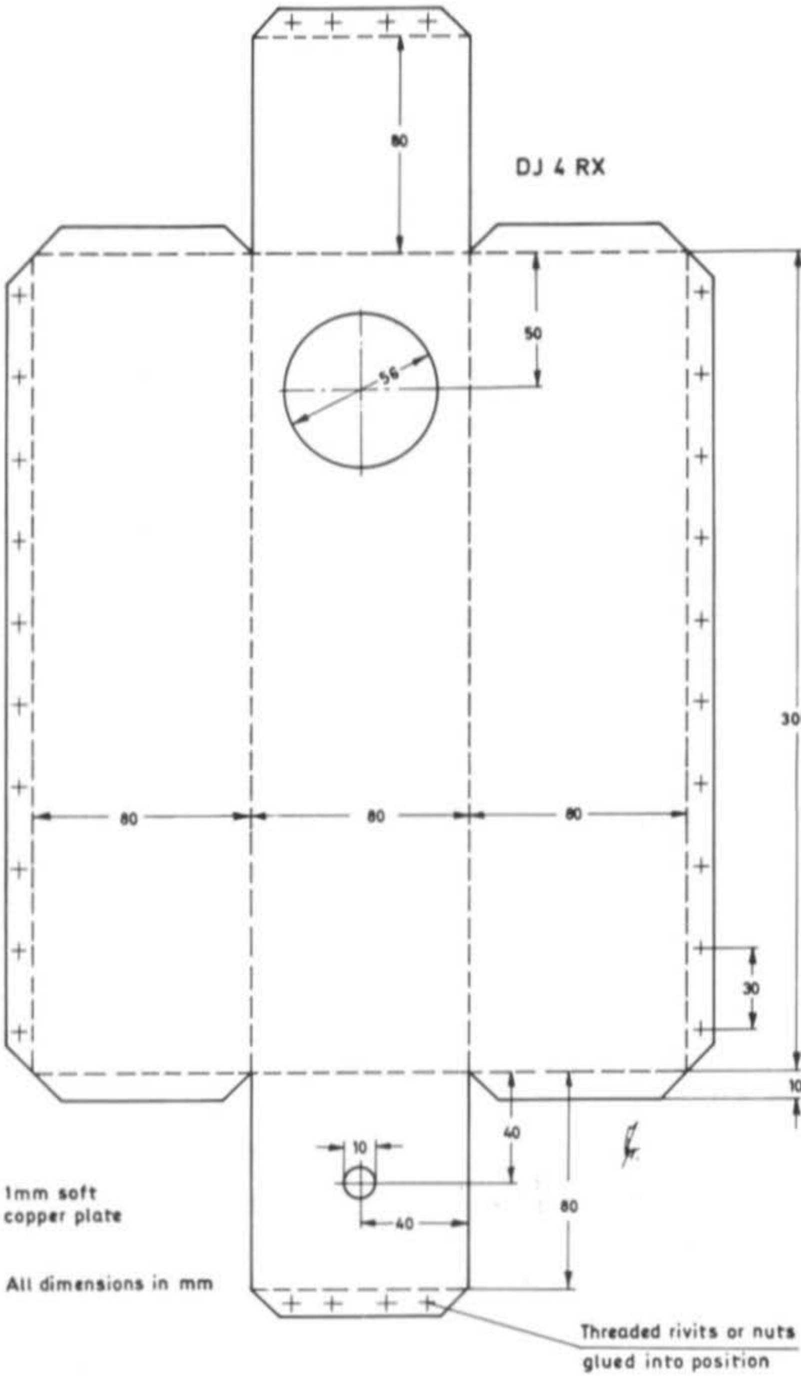


Fig. 4d Casing for a 144 MHz power amplifier with λ/4 coaxial line circuit

possesses greater cutouts than the other pieces; an insulation piece of 4 mm inner and 8 mm outer diameter (0.5 mm thick) is placed into these cutouts to keep it in place. The anode voltage is connected to the solder tag T. Only plate 4fb is connected to the anode voltage (the others are insulated by the teflon discs) which is then fed via a coaxial cable through the inner conductor (Fig. 4a). The three tongues of part 4fb are connected by means of a bracket to the anode radiator of the tube. The other (cold end) of the inner conductor can be directly soldered to the cold end of the outer conductor case (Fig. 4d). The mounted arrangement is shown in Fig. 4h.

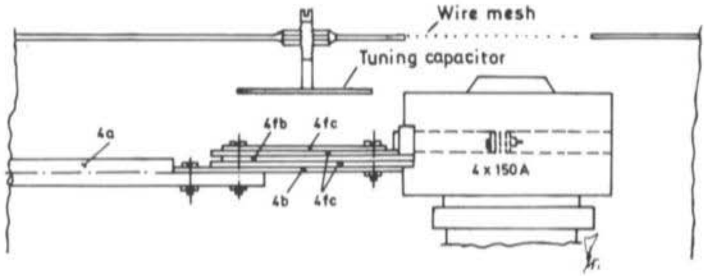


Fig. 4g Bypass arrangement at the hot end of the $\lambda/4$ anode circuit equipped with a 4 X 150 A tube

For optimum cooling of the 4 X 150 pressed glass base, the ventilation is fed from the grid side. A ceramic chimney is not absolutely necessary; a suitable chimney could be made from thickish drawing paper. After having cooled the tube, the air leaves the PA circuit by means of the wire mesh window that is to be found on the covering plate (Fig. 4e). With the exception of the inner conductor, all dimensions are not critical. The resonant frequency of the completed circuit can easily be checked using a dip-meter. The adjustment of the coupling loop is somewhat critical. Since the correct output coupling is very important for correct operation of a linear amplifier, this will be explained in detail in Section 8.

This article is to be continued in the next edition of VHF COMMUNICATIONS. Besides giving full and exact constructional details for a helical PA stage, the problem of the grid circuit, neutralization and the coupling loop will be discussed in detail.

-- to be continued --

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AN IF DIPLEXER FOR 28 - 30 MHz

by E.H. Reitz, DJ 9 JT

This diplexer was designed to connect several shortwave receivers(IF strips) to a single VHF converter. It allows the band to be monitored on a number of receivers without loss of sensitivity. This is of advantage during contests as well as when monitoring the beacon signal, transmit and receive frequencies of active translators such as Oscar and balloon carried types (ARTOB, BARTOB).

In contrast to using several VHF receivers on one antenna, it is not difficult using the diplexer to maintain the same sensitivity and large-signal characteristics as would be obtained using a single receiver.

1. CIRCUIT DESCRIPTION

The diplexer (see Fig. 1) consists of an RF amplifier equipped with the low-reactance pentode EF 183 (6EH7) (V 1) and six double triodes ECC85 (6AQ8) in a common plate configuration for six outputs.

A bandpass filter is to be found before and after the RF amplifier stage which is adjusted for a bandwidth of 2 MHz (28 - 30 MHz) by damping each resonant circuit with a 5.1 kΩ resistor. The six common plate stages are tightly coupled to the "hot end" of the last resonant circuit of the bandpass filter. They do not possess any frequency dependent elements since they are merely provided for the impedance conversion to approximately 60 Ω.

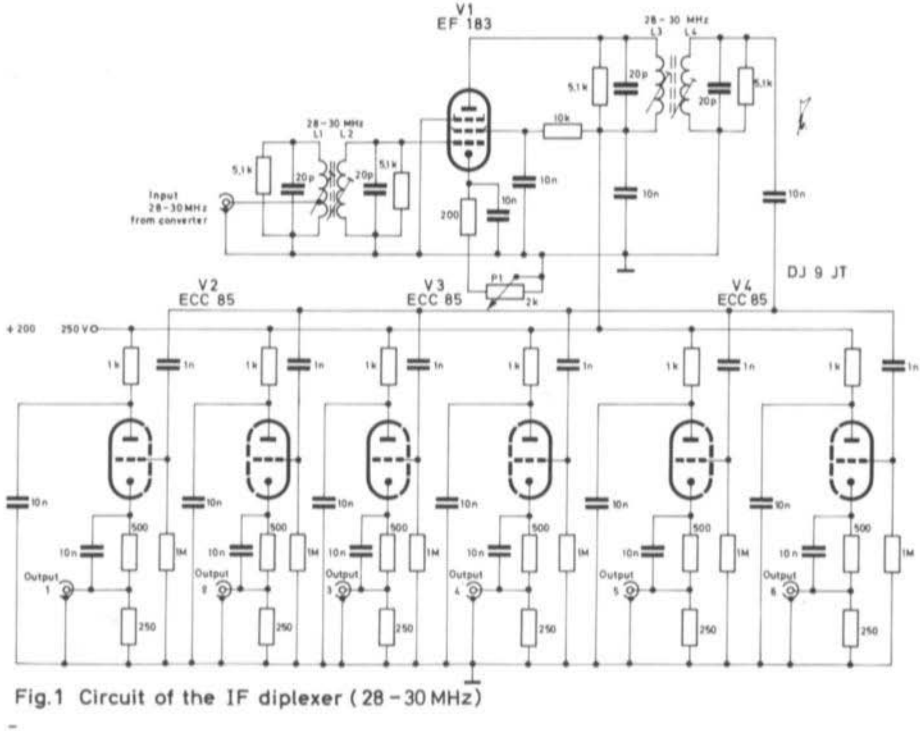


Fig.1 Circuit of the IF diplexer (28 - 30 MHz)

Since the bandpass filter and the common plate stages induce very little loss, the gain of the RF amplifier need not be so high. The potentiometer in the cathode circuit allows the gain to be varied and simultaneously to reduce the cross modulation on short wave receivers sensitive to such interference.

2. ASSEMBLY AND ALIGNMENT

Since assembly of the IF diplexer is not critical, no detailed description is given. It is important that the input and output are well screened from another. This means that it is advisable to enclose each bandpass filter in a screening can and to filter all operating voltages by using feed-through capacitors.

The alignment process is limited to the two bandpass filters. It is, of course, best to carry this out using a sweep measuring set but it can also be carried out easily using a 10 m VFO.

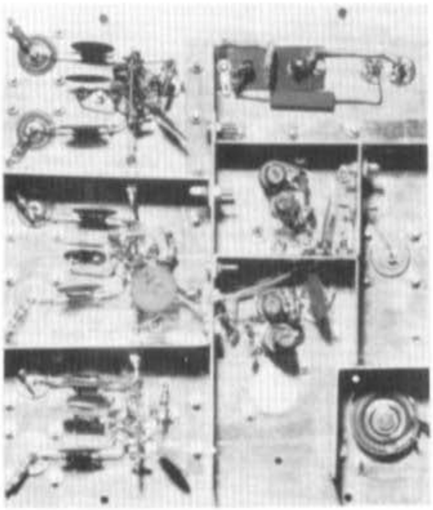


Fig. 2 and 3 show the authors prototype who's construction may be copied

STANDARD FM FREQUENCY ON TWO METRES

The modification of the channel spacings for commercial users in the 150 MHz band has led to a great number of such transceivers to be offered to amateurs at very reasonable prices. Since the receiver of such equipment is also crystal-controlled, it is advisable for a certain channel to be agreed for such transmissions so that QSOs can be made when travelling outside the home town. Since reciprocal licencing also allows one to take the mobile station abroad, what would be better than to know the frequency of the local stations. Another advantage would be for emergency communications or even when your car breaks down one dark rainy night.

A frequency of 145.150 MHz is used as the standard FM frequency for Southern Germany and is rapidly spreading. How about making this the standard FM frequency for Europe? It would be extremely advantageous.

CASCODE IF STAGES

by D. E. Schmitzer, DJ 4 BG

This article was inspired as it became known that certain integrated circuits (CA 3005, CA 3028) could be operated in a cascode circuit (1), (2). The following short description has resulted after studying the matter in more detail. The same results can be achieved using discrete transistors (3).

1. CIRCUIT PRINCIPLE

When using transistors, a cascode circuit usually comprises a common-emitter stage which is galvanically coupled to a common-base configuration (Fig. 1). This arrangement exhibits the same amount of voltage amplification as a conventional common-emitter stage. The advantage is, however, that the so-called Miller Effect is no longer present at the input. This is because no voltage amplification takes place previous to the collector of the common-emitter stage. The input impedance and input capacitance therefore correspond to a low-reactive or completely neutralized common-emitter stage. Due to the common-base stage, the output impedance is great enough to ensure that the connected filter circuits are damped far less than would be the case with a common-emitter circuit. The main advantage of this configuration is, however, that the additional common-base circuit reduces the reaction from the output on to the input so that neutralization is not required, even at high amplification factors.

To summarize, the cascode circuit combines the amplification and input impedance of a low-reactive common-emitter circuit with the low-reactivity of the common-base circuit. This is made with very few additional components and without increasing the current requirements.

2. PRACTICAL CIRCUITS

The circuit diagram of an uncontrolled cascode stage is given in Figure 2. The additional component requirements consist of one transistor, one resistor and one capacitor. This is compensated for by the fact that no neutralizing elements are required. It is possible, using one additional capacitor, to build up the controlled cascode stage shown in Fig. 3. It should be noted that a cascode stage can, due to the lower reaction, be far better controlled than a common-emitter stage.

As can be seen in Figures 4 and 5, simple cascode configurations can be built up when two operating voltages are used; Fig. 4 shows the uncontrolled, and Fig. 5 the controlled configuration.

3. APPLICATIONS

Perhaps the best possible application of the described cascode amplifier would in the IF portion of a receiver where cascode stages complete favourably with FETs.

Field-effect transistors, especially MOSFETs, possess good overload characteristics. This will no doubt result in them being increasingly employed in the input and mixer stages of VHF receivers, even though their current requirements are greater than that of bipolar transistors. The advantages of FETs and dual-gate MOSFETs are not utilized to the full in intermediate frequency amplifier stages. This is especially true if the major part of the selectivity is made directly after the mixer in single-conversion superhets.

The described cascode stages with their inherent stability and considerably lower current requirements are therefore more favourable for IF applications.

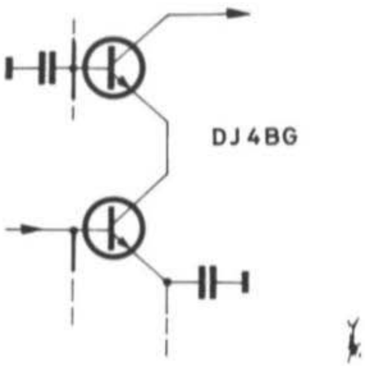


Fig. 1 Principle of a cascode stage

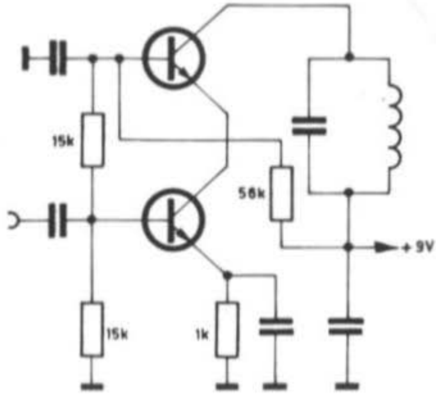


Fig. 2 Uncontrolled cascode IF stage

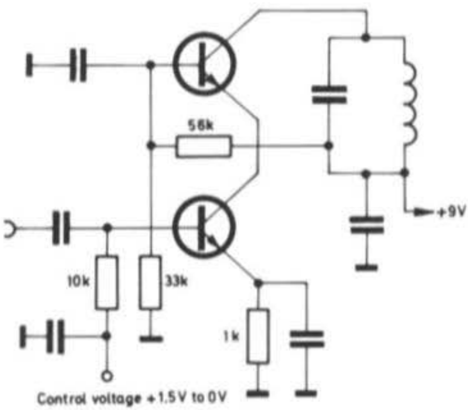


Fig. 3 Controlled cascode IF stage

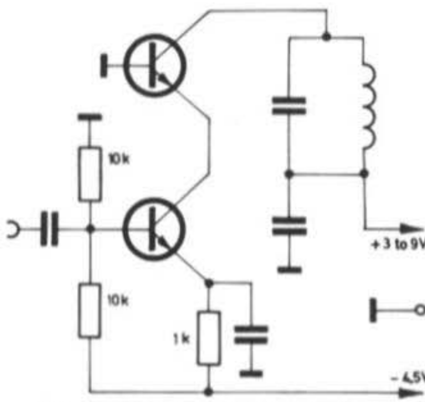


Fig. 4 Uncontrolled cascode IF stage
with two operating voltages

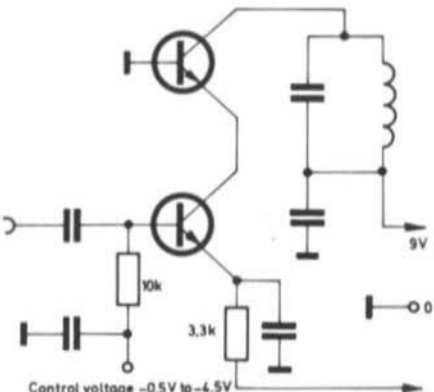


Fig. 5 Controlled cascode IF stage
with two operating voltages

(1) RCA-Application Notes for Types CA 3004, CA 3005, CA 3028 Integrated RF-Amplifiers ICAN-5022

(2) RCA Linear Integrated Circuit Fundamentals IC 40

(3) Mc Aleer: Emitter Peaking Pushes Bandwidth to 500 MHz Electronics 40 (1967), Edition 18, pages 96 and 97.

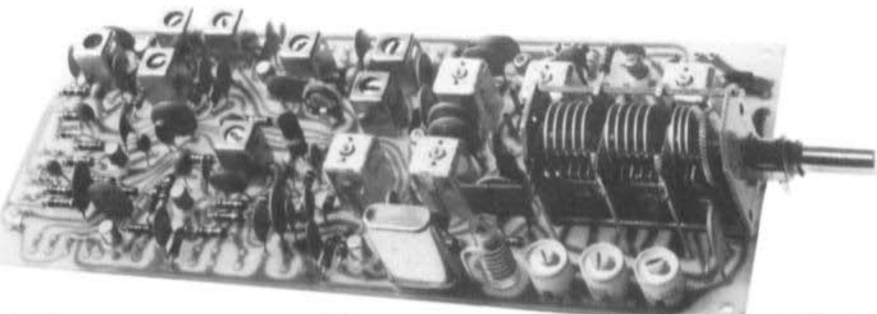
Inexpensive VHF-modules, manufactured by the STE-company of Milan, who are well-known in professional telecommunication circles.
European components, European manufacture. First-class commercial construction and quality. Available from the publishers of VHF COMMUNICATIONS or the manufacturers. Please request further information from Verlag UKW-BERICHTE, H. Dohlus oHG, D-852 Erlangen, Gleiwitzer Str. 45.

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

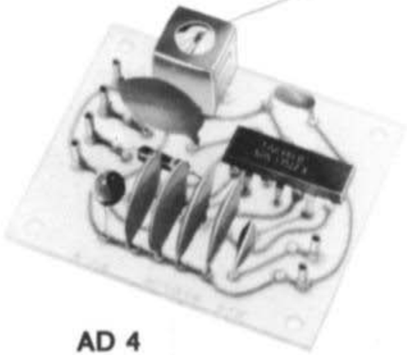


MOSFET 10 metre receiver AR 10: DM 208,45

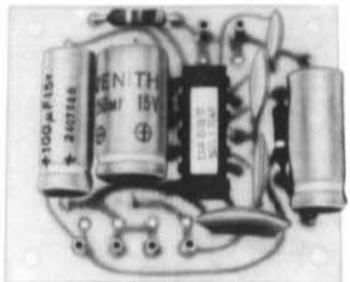
Available for 28—30 MHz as IF strip for 2 m converters, or for 26—28 MHz for reception of the citizen band or for use with the 26—28 version of the AC 2 converter below. High sensitivity ensures excellent shortwave reception. The good large-signal behaviour makes it suitable for use as an IF strip for converters. Double superhet with crystal controlled second conversion. 7-stage second IF of 455 kHz. Built-in SSB and CW demodulator. Connections for S-meter, noise limiter/squelch. Prepared for installation of a piezo-ceramic or mechanical filter. A discriminator module is available so that the receiver can be switched to AM, FM, CW or SSB. A matching AF-module AA 1 is also available.

Specifications:

Input impedance:	50 Ohm	Image and spurious	
Sensitivity:	1μ V for 10 dB S/N	suppression:	60 dB
Selectivity:	4.5 kHz (—6 dB)	Operating voltage:	11—15 V/15—22 mA
	12 kHz (—40 dB)	Dimensions:	200 mm x 83 mm x 32 mm



AD 4



AA 1

FM Limiter and Discriminator AD 4: DM 30.35

FM discriminator for the AR 10 and other receivers with an intermediate frequency of 455 kHz. Advantage over FM demodulation using the IF-slope: the receiver need not be tuned away from the signal, and ignition interference is suppressed by the limiter (AM-suppression: 40 dB, limiter threshold: 100 μV).

Audio Amplifier AA 1: DM 29.70

Miniature integrated AF-amplifier with an output power of 1.5 W at 12 V. Ideal for many applications.

Transistor transmitter AT 210: DM 152,90

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:

Frequency range:	144—146 MHz	Operating voltage:	12 V/400 mA (max. 15 W)
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:

Output power:	2.8 W at 12 W
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 two metre transceiver. They are available ex stock in Erlangen. Each module is accompanied by a description including circuit diagram, list of components and connection details. These modules can also be purchased from the representatives of VHF COMMUNICATIONS in the following countries: Denmark & Sweden, Norway, Spain & Portugal, South Africa. Other countries please inquire directly to

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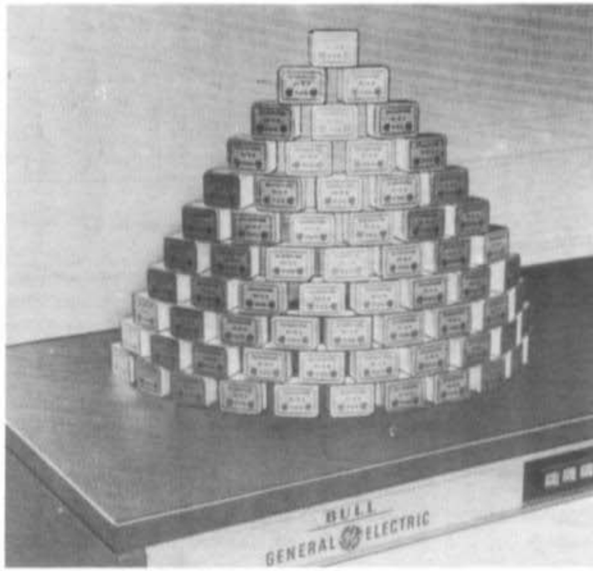
Listed is our well-known series of

9 MHz crystal filters for SSB, AM, FM and CW applications.

In order to simplify matching, the input and output of the filters comprise tuned differential transformers with galvanic connection to the casing.



Filter Type		XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9M
Application		SSB-Transmit.	SSB	AM	AM	FM	CW
Number of Filter Crystals		5	8	8	8	8	4
Bandwidth (6dB down)		2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz
Passband Ripple		< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 1 dB
Insertion Loss		< 3 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 3 dB	< 5 dB
Input-Output	Z_t	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω	500 Ω
Termination	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
Shape Factor		(6:50 dB) 1.7	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:40 dB) 2.5
			(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:60 dB) 4.4
Ultimate Attenuation		> 45 dB	> 100 dB	> 100 dB	> 100 dB	>90 dB	> 90 dB

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