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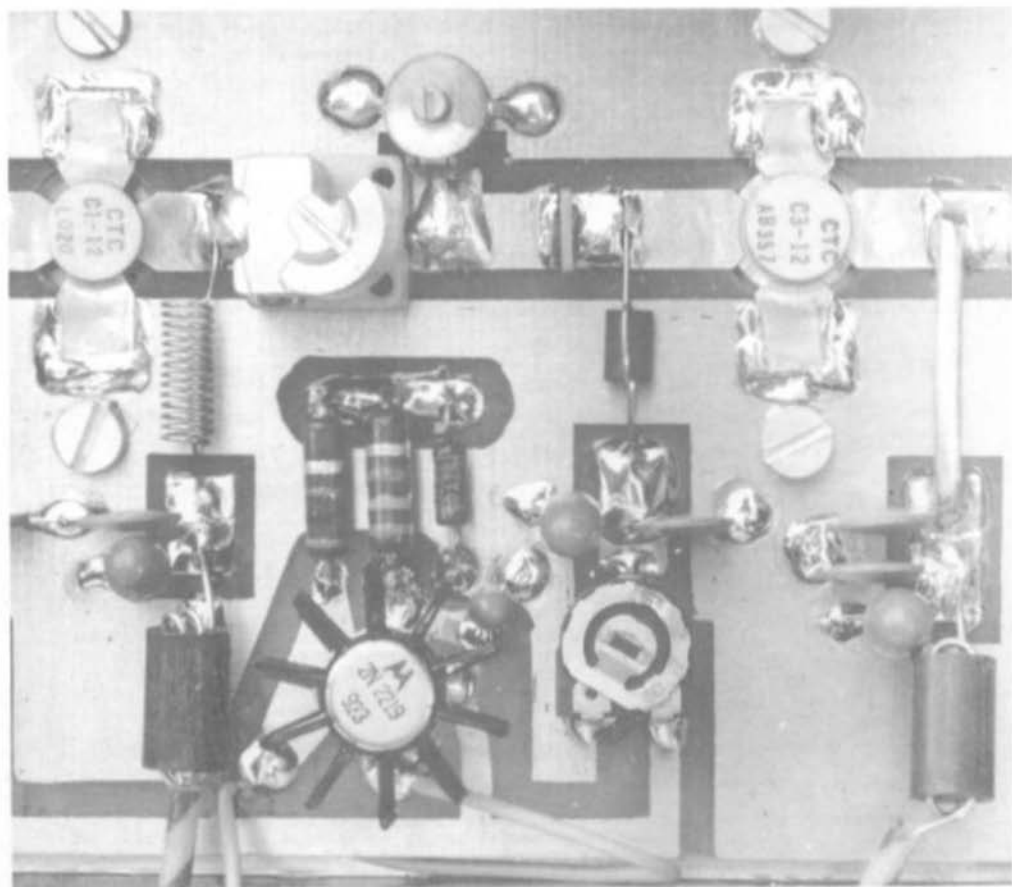
A PUBLICATION FOR THE RADIO AMATEUR
ESPECIALLY COVERING VHF, UHF AND MICROWAVES

VOLUME No. 9

SPRING

1/1977

DM 4.50





VHF COMMUNICATIONS

Published by:

Verlag UKW-BERICHTE · Hans J. Dohlus oHG · Jahnstraße 14 · D-8523 BAIERSDORF
Fed. Rep. of Germany · Telephones (0 91 91) 9157 / (0 91 33) 855, 856.

Publishers:

T. Bittan, H. Dohlus.

Editors:

Terry D. Bittan, G 3 JVQ / DJ 0 BQ, responsible for the text and layout
Robert E. Lentz, DL 3 WR, responsible for the technical contents

Advertising manager:

T. Bittan.

VHF COMMUNICATIONS,

the international edition of the German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is published in Spring, Summer, Autumn, and Winter. The subscription price is DM 16.00 or national equivalent per year. Individual copies are available at DM 4.50, or equivalent, each. Subscriptions, orders of individual copies, purchase of P.C. boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative.

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Printed in the Fed. Rep. of Germany by R. Reichenbach KG · Krelingstr.39 · 8500 Nuernberg

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

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TRANSISTOR LINEAR AMPLIFIERS FOR ATV OPERATION

Experiments made on the DJ 3 SC 435 MHz linear amplifier

by G. Sattler, DJ 4 LB

The described measurements were made in order to check the suitability of the transistor linear amplifier DJ 3 SC 001 (1) in the ATV mode on the 70 cm band, and to examine the special considerations that must be made with respect to this mode.

The linear amplifier was fed from an ATV-exciter comprising modules DJ 4 LB 001 to 004 as described in (2). An additional circuit, similar to the video/sound combining stage DJ 6 PI 002 described in (3) allows a continuously variable IF level adjustment at the input of mixer DJ 4 LB 004. In this manner, it is possible to vary the output power of the mixer, and thus the drive power for the linear amplifier DJ 3 SC 001. This means that the linear range of this amplifier is utilized to the full. Various test signals were taken from a pattern generator instead of a TV video signal. In order to compare the characteristics of transistor linear amplifiers to that of tube amplifiers, measurements were also made on a linear amplifier equipped with the EC 8020 tube as described in (4).

With the modifications that were established experimentally, and which are given at the end of this article, the DJ 3 SC amplifier which was originally designed for SSB operation, is also suitable for ATV operation.

1. CHOKES

1.1. The Collector Chokes

Most linear amplifiers operate in class AB. This means that the collector current is dependent on the value of the RF drive voltage. If the control voltage is amplitude modulated, the collector current will vary in time with the modulation frequency. Whereas only frequencies of several kHz are to be expected with voice modulation, modulation frequencies of up to 5.5 MHz are present in the ATV mode.

Figure 1 gives the circuit diagram of the 70 cm transistor amplifier as described in (1). Whereas the collector chokes Ch 7, 8 and 9 do not provide any resistance for audio frequencies and thus ensure that the operating voltage on the bypass circuits remains constant, they cause collector currents and voltage drops in time with the pulsing video signals. At higher RF output powers and correspondingly high currents, it is possible for the voltage drops to correspond to several Volts and to be increased even more by voltage peaks which often appear on switching off inductances. One of the prerequisites of satisfactory operation of the linear amplifier, namely a constant collector DC voltage independent of the modulation is no longer fulfilled even when the operating voltage is stable.

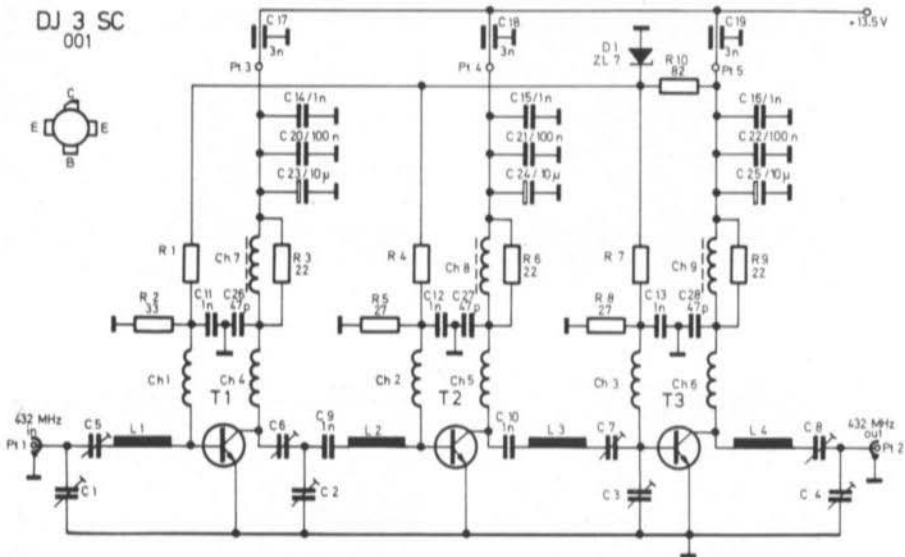


Fig. 1: Three-stage transistor linear amplifier for the 70 cm band

Figures 2 to 5 show oscilloscope traces made at stage T3 of the amplifier equipped with the transistor C 25-12 (see Figure 1). Since this stage takes a high collector current, the voltage drop across the ferrite ring choke will be especially large. The collector DC voltage at the intersection of chokes Ch 9 and Ch 6 is given in the lower half of the illustration (10 V/division). The demodulated UHF output signal obtained from module DJ 4 LB 005 is shown in the upper half of the illustration. In the case of Figures 2 and 3, the video carrier is modulated with 1 MHz sinewave signal, and in the case of Figures 4 and 5 with a 500 kHz square-wave signal.

Figures 2 and 4 show the oscilloscope traces measured with stage T3 in the original circuit. It will be seen that voltage peaks of over 40 V are present instead of the stable DC voltage, and such voltages can endanger the transistor. In addition to this, the output signals are very distorted due to collector voltage modulation.

Figure 3 and 5 were made with the same test signals, and it is only the voltage drop and voltage overshoot across choke Ch 9 that has been suppressed with the aid of an electrolytic capacitor.

A TV video signal consists of a complete mixture of many modulation frequencies in contrast to the test signals shown here. In addition to this, a sound carrier with a spacing of 5.5 MHz is also added. If these frequencies cause a voltage drop or voltage overshoot across the collector choke, the collector DC voltage will be superimposed with a mixture of the modulating frequencies. This will be visible on a TV receiver in the form of a noisy picture inspite of a high input voltage.

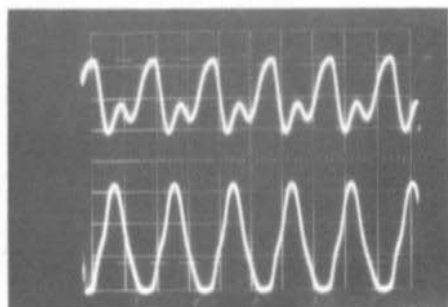


Fig. 2: Original configuration
sinusoidal drive

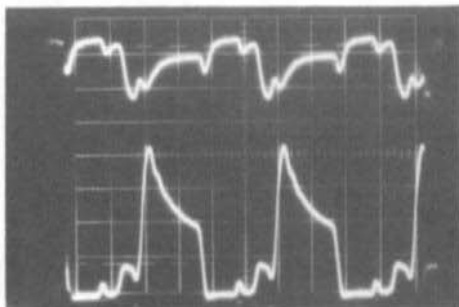


Fig. 4: Original configuration,
square-wave drive

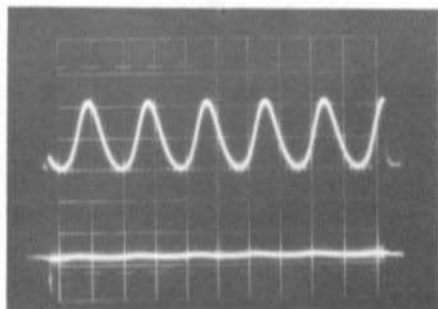


Fig. 3: After modification

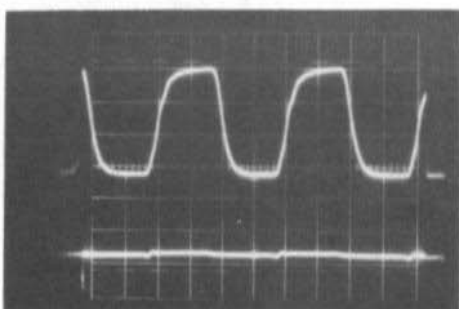


Fig. 5: After modification

This means that the demands made on the chokes of the collector DC-circuit are for a sufficiently high resistance for the transmit frequency, and the lowest possible resistance for all modulating frequencies in the signal. In the case of ATV amplifiers, only UHF-chokes with a few turns are suitable. The detrimental effects in the ATV mode caused by ferrite chokes can be eliminated by using sufficiently large capacitance values.

Of course, the other demands made on the chokes and bypassing of the amplifier stages are also valid in the ATV mode. Attention should be paid to avoid resonances at low transmit frequencies. Otherwise self-oscillation could occur, e.g. in the shortwave range, since the gain of the transistors increases on decreasing the frequency.

1.2. The Base Chokes

Basically the same is valid for the base voltage as for the collector voltage. This voltage should also remain constant and not be affected by the modulation. This means that the base chokes should also only represent a noticeable resistance with respect to the transmit frequency, but not for the modulation frequency. Due to the fact, that the base current is considerably lower than the collector current, the voltage drop across the chokes will be far less. Therefore also conventional ferrite wideband chokes can be used in the base current circuit up to the impedance of a 6-hole-core.

Chokes Ch 1, Ch 2, and Ch 3 of the DJ 3 SC amplifier (18 turns on a resistor) form resonant circuits together with the other components, which can be brought into oscillation, for instance from the TV synchronizing pulses. The TV signal would then be additionally modulated with approximately 13 MHz. In order to avoid this, it is recommended that these chokes should be exchanged for 6-hole-core types.

The base voltage at bypass capacitors C 11, C 12, and C 13 will only remain constant at high modulating frequencies. It is therefore necessary for electrolytic capacitors to be connected in parallel with these capacitors so that load variations of the base voltage divider are compensated for at low modulation frequencies.

However, occasional voltage oscillations in the kHz range are observed across these capacitors when collector currents of over 100 or 200 mA have been selected. This can be avoided by connecting real resistors of approx. 10 to 22 Ω in series with the base chokes.

2. BANDWIDTH

In the case of the ATV concept described in (2), the modulated video carrier is passed via the linear amplifier together with the sound carrier which is spaced 5.5 MHz higher. This means that the bandwidth of such a linear amplifier is of particular importance. The flat top of the passband curve should amount to at least 6 MHz.

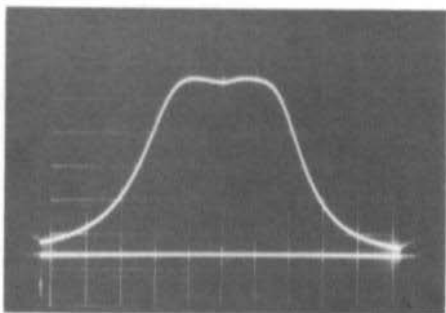


Fig. 6: Swept frequency response,
DJ 4 LB 004

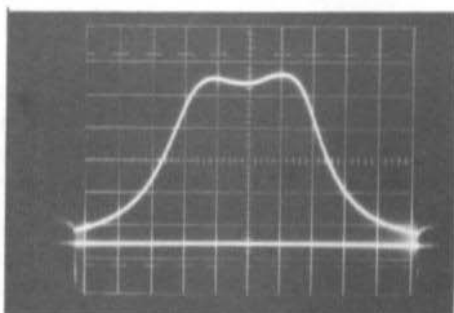


Fig. 7: Over all frequency response,
DJ 4 LB + DJ 3 SC

Figure 6 gives the passband curve of module DJ 4 LB 004 (with 2.5 MHz/division), which fulfills the given demand. After connection to the DJ 3 SC linear amplifier, the overall passband curve as shown in **Figure 7** results, which is practically the same as that of the exciter. This indicates a wideband width of the linear amplifier in comparison to the exciter. The alignment of the linear amplifier is therefore uncritical. It is sufficient when the trimmer capacitors are aligned for maximum output power at the video frequency. It is only the input circuit comprising the C 1 and C 5 that has a considerable effect on the passband curve, because they represent the terminating impedance for the exciter. Generally speaking, the resonant circuits in the transistor amplifier will become lower in impedance and thus wider in bandwidth on increasing the RF power. This means that the DJ 3 SC 001 linear amplifier is therefore suitable for ATV operation with respect to the bandwidth and requires no modification.

Figure 8 shows the passband curve (also with 2.5 MHz/division) of a tube stage equipped with the EC 8020 as described in (4). The bandwidth is not sufficiently wide to ensure a good video and sound transmission.

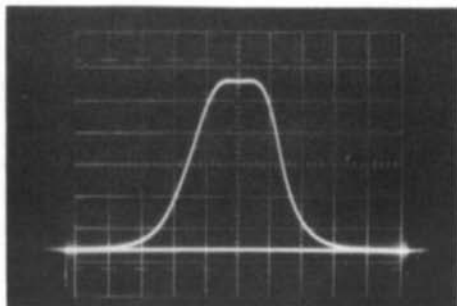


Fig. 8: Over all frequency response, DJ 4 LB + EC 8020

3. LINEARITY

Especially high demands are placed on the linearity of RF amplifier stages in the ATV mode. The RF-TV signal contains between 10 % and 75 % of its total amplitude the video information and between 75 % and 100 % the synchronizing pulses. This means that a limiting of the maximum amplitude by only 25 % will suppress the synchronizing pulses completely, and the TV picture can no longer be synchronized.

Delta-wave modulating signals are suitable for measuring the linear relationship of the RF amplifier stages. They allow a limiting of the peak amplitude as well as non-linearities at the zero-transition to be established, such as can be generated with too low a quiescent current of the amplifier stages.

The following oscilloscope photographs show test signals that are obtained by demodulation of the transmit frequency using the UHF demodulator DJ 4 LB 005 (2).

In **Figures 9, 10, and 11**, the waveform at the output of exciter DJ 4 LB 004 is shown at 20, 40 and 80 mW PEP. The differing output powers are obtained by a variation of the IF level, as was mentioned at the beginning of this article. If ATV operation is made according to the CCIR standard, negative going modulation will be used, and a high RF amplitude will cause a correspondingly large negative voltage in the demodulator which will be orientated in a downward direction on the oscilloscope. The limiting of the signal at the peak power output is clearly seen in Figure 11.

These measurements were carried out to find the maximum IF-voltage at the input of the mixer in order to obtain a non-limiting UHF output signal. This allows one to establish whether the limiting is taking place in the exciter or in the linear amplifier to be measured.

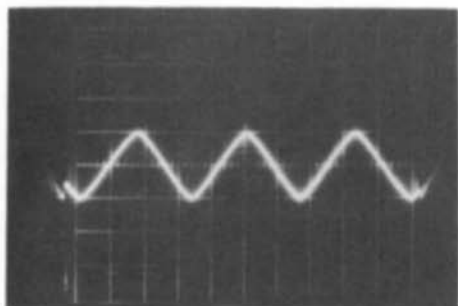


Fig. 9: 20 mW PEP

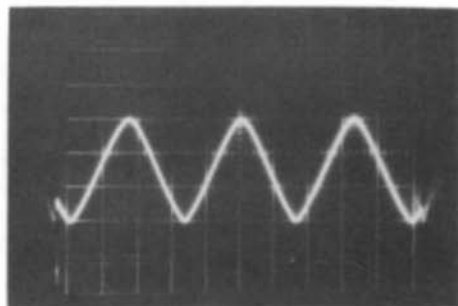


Fig. 12: 0.5 W PEP

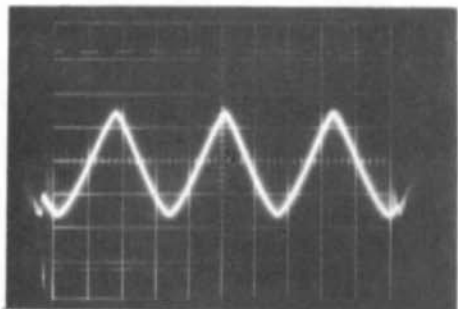


Fig. 10: 40 mW PEP

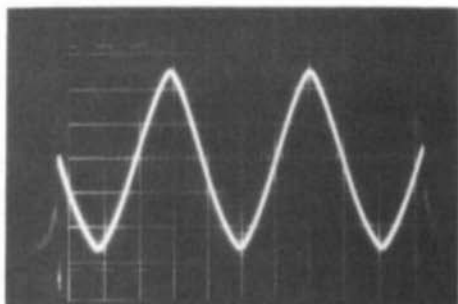


Fig. 13: 1.5 W PEP

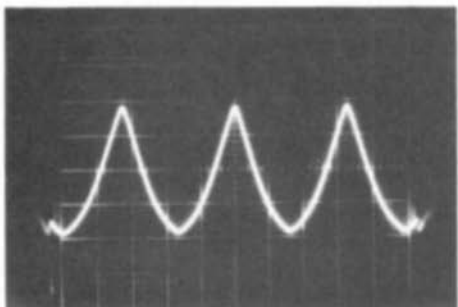


Fig. 11: 80 mW PEP,
demodulated UHF signal from DJ 4 LB 004

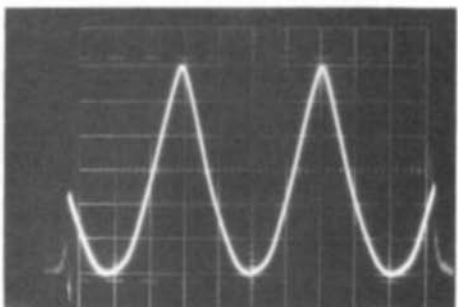


Fig. 14: 2.1 W PEP
demodulated UHF signal from
DJ 3 SC amplifier, two stages

Firstly a two-stage version of the linear amplifier DJ 3 SC 001 equipped with transistors C 1-12 and C 3-12 was tested. **Figures 12, 13, and 14** give the output signals of this amplifier at 0.5 W, 1.5 W, and 2.1 W PEP. For comparison: A EC 8020-stage as described in (4) resulted in a maximum of 0.9 W PEP when using the same exciter, and exhibited too low a bandwidth. Due to the far easier construction of the two-stage transistor amplifier when using the PC-board DJ 3 SC 001, this technical article is to be followed by another article giving full constructional details of a two-stage ATV amplifier.

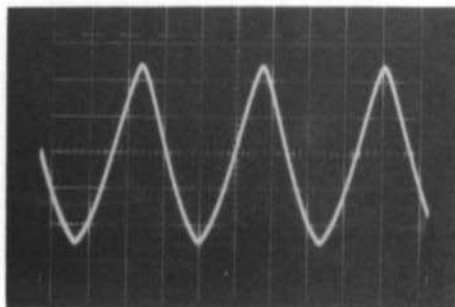


Fig. 15: 8.5 W PEP

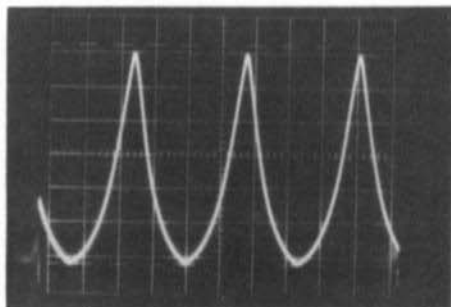


Fig. 16: 12 W PEP

demodulated UHF signals from modified 3-stage DJ 3 SC amplifier

Figures 15 and 16 finally show the demodulated UHF output signals at 8.5 and 12 W PEP. They result from the modified 18 W version of the DJ 3 SC amplifier with the following complement: C 3-12, C 12-12, C 25-12 and were measured at a 13.5 V operating voltage.

3.1. Collector Quiescent Current

The collector quiescent current has mainly an effect on the signal components in the vicinity of zero carrier. It is possible when using delta test signals to notice improvements of the commencement linearity when the quiescent current is increased over 50 mA per stage. The signal distortions are especially low when the first stage (C 1-12) of the amplifier is virtually operating in class A, which is the case at approximately 70 - 110 mA. The following stages can be operated at 100 to max. 200 mA quiescent current as a compromise between low distortion and thermal stability.

In order to align the quiescent current, it is advisable for at least parts of the resistance of R 1, R 4, and R 7 to be in the form of a trimmer resistor.

3.2. Tuning

The lowest limitation of the modulation peaks results when all resonant circuits of the amplifier are aligned for maximum output power. In order to achieve this, the highest possible, unmodulated signal is fed from the exciter to the amplifier and the amplifier trimmers are aligned several times for maximum output power. For linear operation, the drive level should be reduced until a linear relationship exists between the drive and output power.

4. SUMMARY of the MODIFICATIONS on the AMPLIFIER DJ 3 SC 001 for ATV OPERATION

The following drop-type tantalum electrolytics should be installed:

- parallel to C 26: 3.3 μ F, 16 V
- parallel to C 27: 10 μ F, 16 V
- parallel to C 28: 10 or 22 μ F, 16 V
- parallel to C 11: 10 μ F, 3 V
- parallel to C 12: 10 μ F, 3 V
- parallel to C 13: 22 μ F, 3 V
- parallel to D 1 (ZL 7): 22 μ F, 16 V

Ch 1, Ch 2 and Ch 3 should be exchanged for 6-hole ferrite core types.

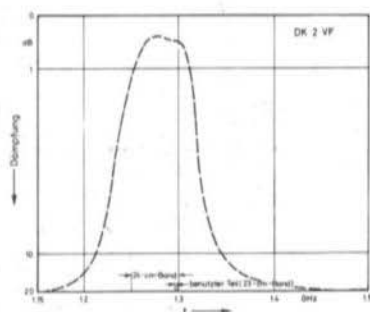
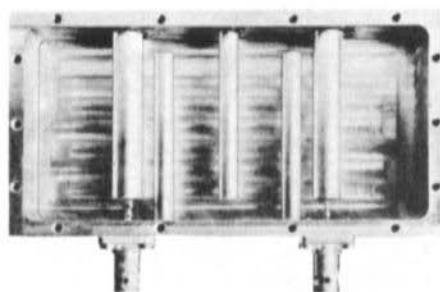
Provide trimmer resistors of approximately 220 Ω in series with R 1, R 4 and R 7 for alignment of the collector quiescent current.

5. REFERENCES

- (1) G. Freytag: A Transistorized Linear Amplifier for 70 cm
VHF COMMUNICATIONS 6, Edition 1/1974, Pages 30 - 37
- (2) G. Sattler: A Modular ATV Transmitter
VHF COMMUNICATIONS 5, Edition 1/1973, Pages 2 - 15
- (3) J. Grimm: ATV Information
VHF COMMUNICATIONS 8, Edition 2/1976, Pages 90 - 95
- (4) K. Eichel: A Stripline Transverter for 70 cm
VHF COMMUNICATIONS 2, Edition 4/1970, Pages 225 - 239

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TWO-STAGE ATV LINEAR AMPLIFIER FOR 435 MHz

by G. Sattler, DJ 4 LB

The described two-stage amplifier is similar to that described in (1). It provides an output power of 1.5 W with an operating voltage of 12 V over the whole of the 70 cm band without external tuning. The power gain is in the order of 50 to 60 times, corresponding to 17 - 18 dB; a drive power of approximately 25 mW is required.

If the mixer module DJ 4 LB 004 is used as exciter, a maximum of approximately 4 W of RF power output (FM) can be obtained with an operating voltage of 12 V. This means that sufficient power reserve is available for linear operation with an output power of 1.5 W (SSB, ATV). This module can easily be run from batteries, and closes the gap between exciters and higher power amplifiers (C 12-12 / C 25-12, 2 C 39, 4 CX 250 B).

1. CIRCUIT DESCRIPTION

The circuit diagram of the two-stage amplifier is given in **Figure 1**. It is formed from the two first stages of the DJ 3 SC amplifier. The modifications for ATV operation given in (2), have been taken into consideration. The output power is coupled out where the base of transistor T3 is connected in the case of the three-stage DJ 3 SC amplifier. The original trimmer capacitor C7 in series with L3 can be deleted in our application without loss of output power, since the output impedance is 50 Ω .

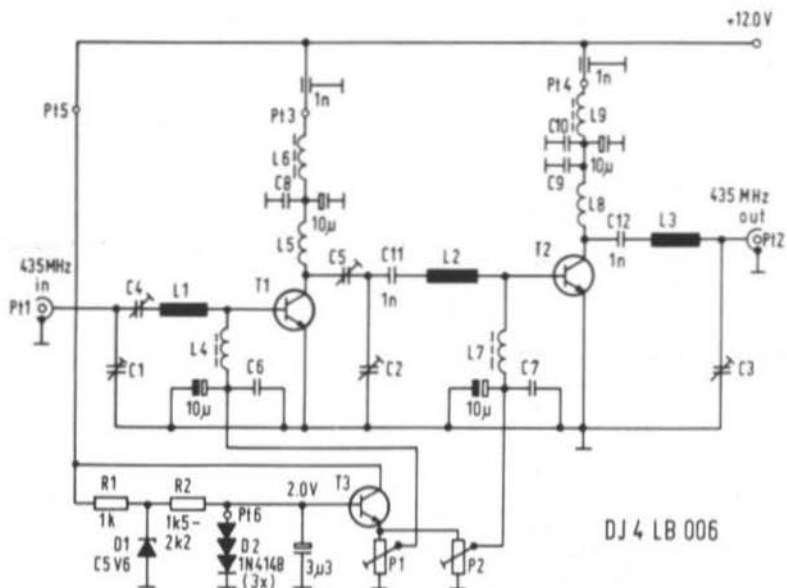


Fig. 1: 2-stage power amplifier for 70 cm ATV

1.1. Generation of the Base Bias Voltages and Temperature Compensation of the Collector Quiescent Currents

The base voltages of the two RF transistors are adjustable with the aid of potentiometers P 1, or P 2. Transistor T 3 is connected in an emitter follower circuit and its base bias voltage is formed as the sum of the forward voltages of three silicon diodes connected in series. The dropper resistor R 2 feeds these series-connected diodes with a stable, temperature independent voltage (R 1, C 5 V 6) at a low current (approx. 1 to 2 mA). Since the intrinsic heat is negligible, the diodes will be practically at the ambient temperature. They can be used as temperature probes when they are in direct heat-contact with the heat sink of the transistors.

The forward voltages of the diodes will be reduced when they are heated, and this will reduce the base bias voltages of the RF-transistors via the circuit comprising T 3. This means that the collector quiescent currents, which would increase considerably as a function of temperature if no compensation was made, will remain relatively stable.

Temperature	- 20°C	+ 20°C	+ 60°C
Quiescent current T 1	75 mA	70 mA	60 mA
Quiescent current T 2	95 mA	90 mA	75 mA

It will be seen that the temperature response of the quiescent currents is slightly over-compensated, which has a favorable effect: When the case is heated, the quiescent currents will slightly fall. The values given in the table at 20°C were found to be most favorable experimentally, using delta-wave test signals. Unwanted oscillation in the kHz-range does not appear until quiescent currents of over approx. 150 mA for T 1 and 300 mA for T 2.

On heating transistor T 3, the base-emitter voltage will decrease, which will cause the output voltage of the circuit to increase. Since this works in an opposite manner to the required temperature-voltage behaviour of the circuit, it is not possible for the transistor to be used as temperature probe. This transistor should therefore have the most constant operating temperature possible, and is therefore provided with cooling fins. The described circuit cannot avoid the collector currents increasing shortly after switching on. According to the thermal coupling, a certain time is required until the heat from the RF transistors is passed to the diode probe, and the compensation circuit can be effective.

1.2. Component Details

- T 1: C 1 - 12
T 2: C 3 - 12
T 3: 2 N 2219, 2 N 1613 (NPN transistor with B min. 60 to 70) with cooling fins
D 1: Zener diode 5.6 V (BZY 83 C5V6)
D 2: Three silicon diodes, e.g. 1 N 4148 (1 N 914) connected in series
P 1, P 2: 220 Ω , trimmer potentiometer, for horizontal mounting, spacing 10/5 mm
C 1, C 2, C 3: plastic foil trimmer 2 - 22 pF
C 4, C 5: air-spaced trimmer 34 pF with two connection pins

- C 6 - C 10: ceramic disc capacitors, value uncritical, between 47 and 100 pF
 C 11, C 12: Chip capacitors of approx. 1 nF
 L 4, L 6, L 9: Ferrite wideband chokes (6-hole core)
 L 7: 1.5 turns of 0.4 mm dia. (26 AWG) enamelled copper wire passed through a ferrite bead, 5 mm long
 L 5: Approx. 17 cm enam. copper wire on 3 mm former, self-supporting
 L 8: 20 to 25 mm length of 2 mm dia. (12 AWG) silver-plated copper wire, slightly bent to ensure that no short-circuit is made to the ground
 Electrolytic capacitors: Tantalum drop-types for the required voltage.

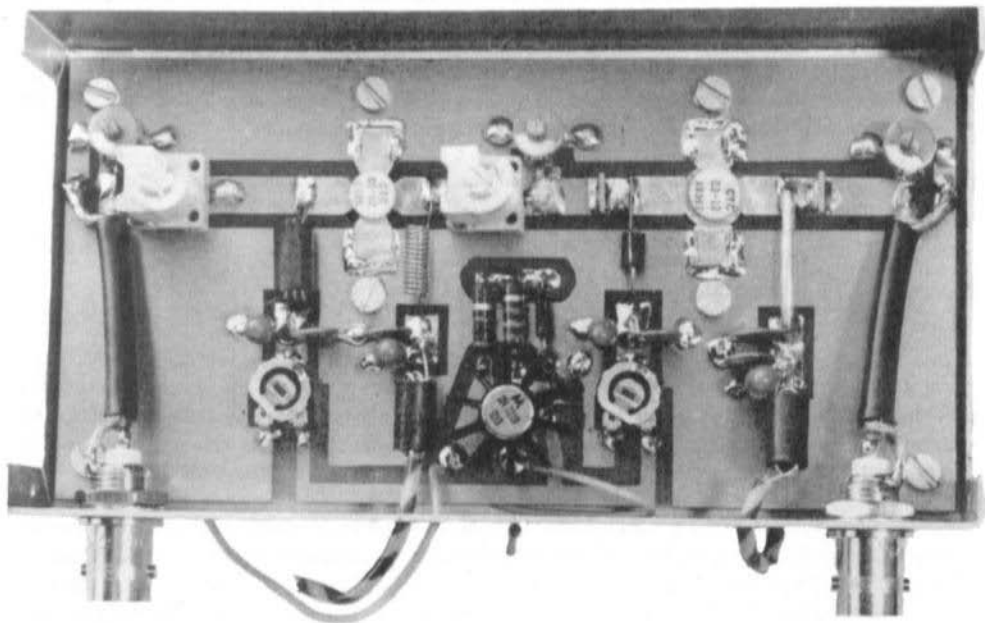


Fig. 2: Author's prototype

2. CONSTRUCTION

The photograph shown in **Figure 2** shows the author's prototype of the two-stage amplifier mounted in a TEKO box 4 A. The components locations and conductor lanes of the 135 mm x 65 mm PC-board DJ 4 LB 006 are shown in **Figure 3**. The PC-board is only single-coated and the components are mounted on the conductor side. The three temperature-probe diodes are mounted underneath the PC-board and are glued to the metal plate of the TEKO box. One diode is in the direct vicinity of T 1, another near T 2 and the third diode is located between T 1 and T 2 interconnecting the two other diodes. The cathode connection of the diode is screwed to the box near transistor T 2. A flexible wire connects the anode of the diode placed in the vicinity of T 1 and connects it to Pt 6 on the other side of the board.

An M 3 (3 mm) screw has been soldered to the lower side of the transistor T 1 so that it is possible for it to be directly screwed to the TEKO box in a similar manner to T 2 for heat dissipation. No heat sink is required in addition to the aluminium case (TEKO 4 A).

Two M3 screws are required in addition to the four mounting screws at the corners of the board. These screws are provided in the vicinity of the emitter connections of the transistors and are required for a good mechanical and electrical stability. M3 nuts are provided on each of the mounting screws between the lower side of the PC-board and the case, so that a spacing of approximately 2.5 mm results between the board and the box.

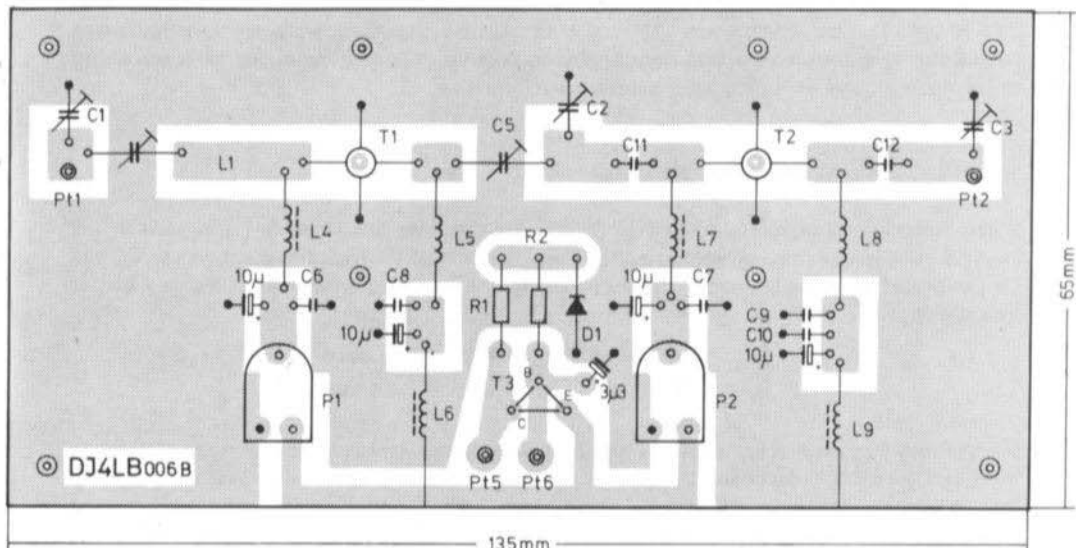


Fig. 3: PC-board for the ATV power amplifier DJ 4 LB 006

3. ALIGNMENT

The collector currents of the two RF transistors T1 and T2 are firstly aligned without RF drive by slowly and carefully adjusting potentiometers P1 and P2.

A TV video carrier, e.g. from DJ 4 LB 004, or another signal at the center of the band are suitable for alignment of the resonant circuits. All trimmers should be aligned for maximum output power, at the highest drive voltage. An exact alignment of the input circuit is possible with the aid of a sweep generator. It is possible using slight corrections on the adjustment of trimmers C1 and C4 to obtain the same passband curve at the output of the linear amplifier, as was present at the output of the exciter.

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A VESTIGIAL SIDEBAND FILTER FOR ATV

by J. Grimm, DJ 6 PI

A printed circuit board has been developed for the ATV vestigial sideband filter described in (1), which greatly simplifies the construction of the filter. This PC-board is designated DJ 6 PI 004. The dimensions are 110 mm x 45 mm. No screening walls are required since neighbouring inductances are at right angles to another. It is only necessary for a screening to be made around the board as is the case with any filter.

1. APPLICATIONAL DEMAND AND CIRCUIT

If the video-sound coupling stage DJ 6 PI 002 that was also described in (1) is used in the DJ 4 LB ATV-transmitter, unwanted conversion products of the video and sound carrier can be generated by non-linearities. These are mainly the sum and difference of the two carrier frequencies:

$$f_{M1} = 33.4 \text{ MHz} + 38.9 \text{ MHz} = 72.3 \text{ MHz}$$

$$f_{M2} = 38.9 \text{ MHz} - 33.4 \text{ MHz} = 5.5 \text{ MHz}$$

Since these frequencies can produce further unwanted conversion products in the UHF mixer module DJ 4 LB 004 described in (3), it is necessary for them to be suppressed previous to the UHF mixer. This means that the vestigial sideband filter must be connected between the video-sound coupling stage and the mixer module. Since the filter provides more than 30 dB attenuation in excess of 40.2 MHz, the conversion product $f_{VC} + f_{SC} = 72.3 \text{ MHz}$ will reach the UHF mixer suppressed 1000 times (30 dB).

In order to suppress the conversion product $f_{VC} = f_{SC}$, the vestigial sideband filter has been extended to include a highpass filter. Since the demands on the slope of this part of the filter are not too great due to the large frequency spacing between 33.4 and 5.5 MHz, a five-stage filter circuit is sufficient.

The filter was calculated according to data given in (4). The designation of the lowpass filter is C 09 15 81, and C 05 15 31 for the highpass filter. **Figure 1** shows the circuit diagram of the complete vestigial sideband filter.

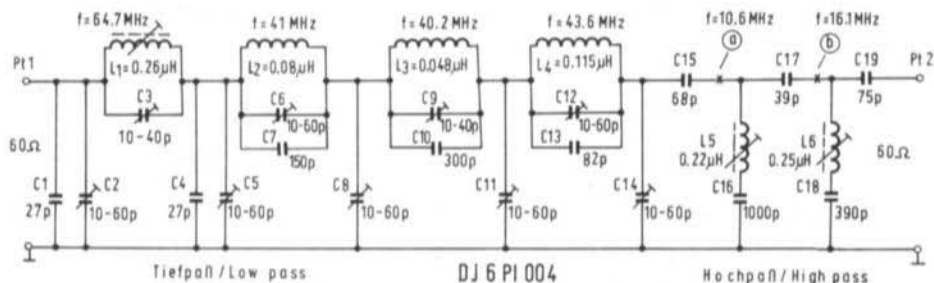


Fig. 1: Circuit diagram of a vestigial sideband filter for ATV

2. MEASURED VALUES

Figures 2 and 3 give the passband curves of the filter. The curve of the vestigial sideband slope (Figure 3) is slightly different from the normal CCIR curve. However, complete coincidence with the nominal CCIR curve would require a considerably higher extent of filtering. Figure 4 shows how the described filter is able to suppress ATV signal components in the frequency range used for voice communications (432 to 433 MHz). Figure 4 shows the spectrum of an ATV transmitter modulated with a video and sound signal in the 70 cm band, using the described vestigial sideband filter.

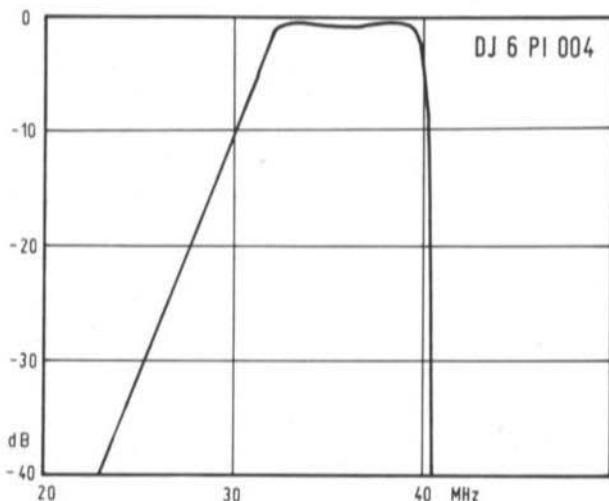


Fig. 2: Passband curve of the vestigial sideband filter DJ 6 PI 004

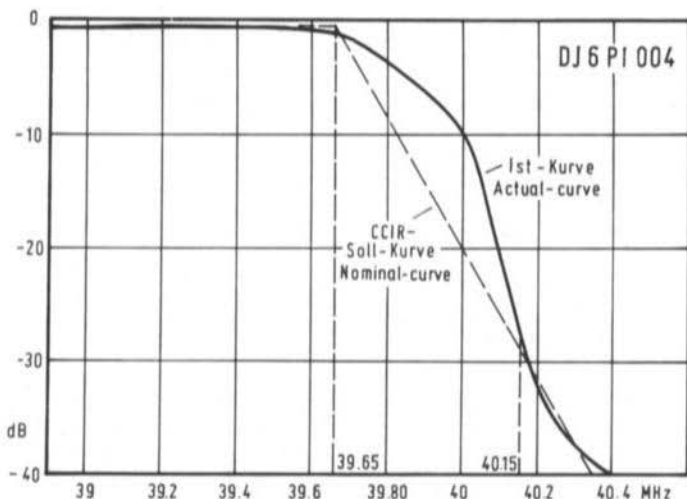


Fig. 3: The upper slope of the passband curve compared to the normal CCIR curve

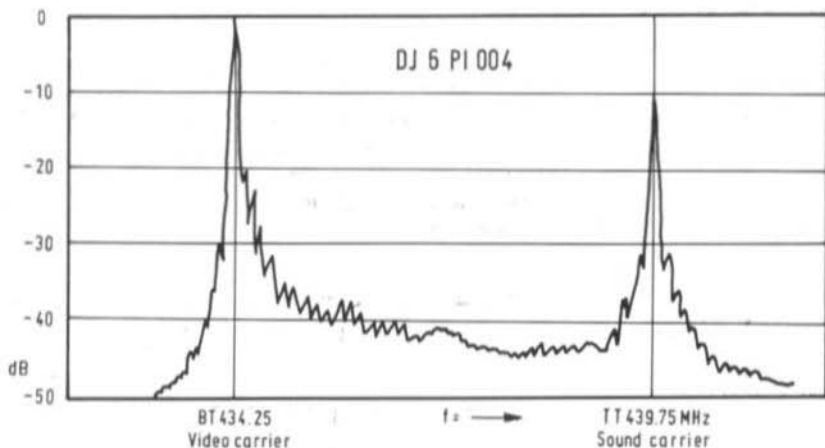


Fig. 4: Spectrum of an ATV transmitter in the 70 cm band using the described vestigial sideband filter

3. COMPONENTS

All inductances are made from 1 mm dia. (18 AWG) silver-plated copper wire.

- L 1: 9.5 turns wound on a 5 mm coil former with a core (green), coil length 15 mm
- L 2: 5 turns wound on a 5 mm former, self-supporting, coil length 7.5 mm
- L 3: 3 turns, otherwise as L 2
- L 4: 6 turns, otherwise as L 2, coil length 10 mm
- L 5: 8.5 turns wound on a 5 mm coil former with core (green), coil length 15 mm
- L 6: 9.5 turns, otherwise as L 5

All trimmer capacitors: ceramic disc types of 10 mm diameter, or foil trimmers of 7.5 mm dia.: 7 pieces 60 pF max. capacitance, 2 pieces 40 pF max. capacitance.

C 16: 1000 pF styroflex capacitors

All other fixed capacitors: ceramic disc or tubular capacitors; styroflex capacitors can be used for the larger values.

4. CONSTRUCTION AND ALIGNMENT

Figure 5 shows the component locations and conductor lanes on the board DJ 6 PI 004. A photograph of the author's prototype without screening is given in Figure 6. The unobtainable capacitance values that result during the filter calculation must be provided as trimmer capacitors in the lowpass filter. Since the demands on the slope are not so high in the case of the highpass filter, fixed nominal values can be used that roughly coincide with the calculated values.

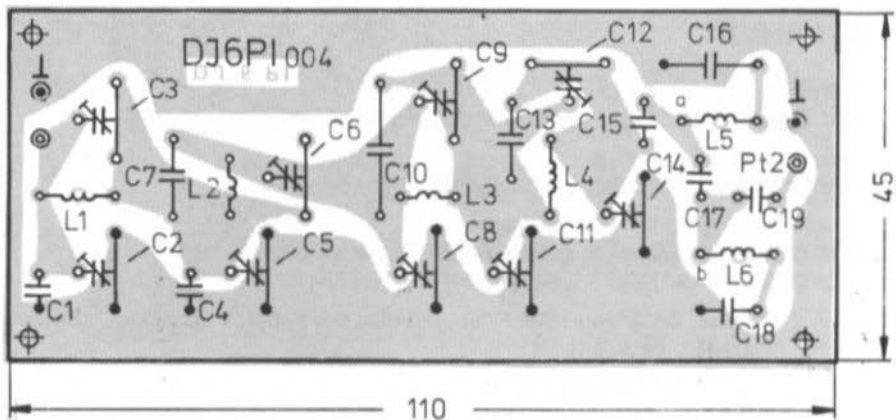


Fig. 5: Component locations and conductor lanes of the vestigial sideband filter

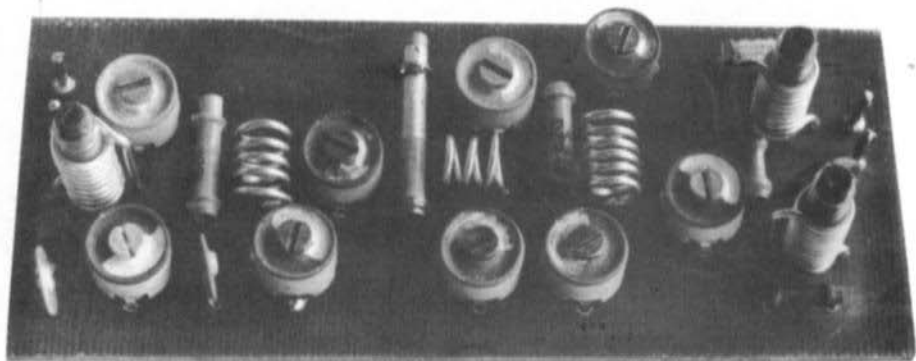


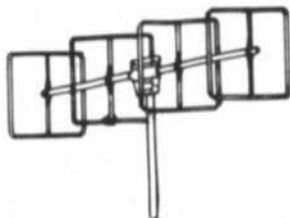
Fig. 6: Author's prototype of the vestigial filter DJ 6 PI 004

The filter should be roughly aligned with the aid of a dipmeter before the alignment is made with the aid of a sweep-measuring system. Firstly, only the following inductances and capacitances of the six resonant circuits comprising L 1 to L 6 should be mounted: C 3, C 6, C 7, C 9, C 10, C 12, C 13, C 16, C 18. Points «a» and «b» should be temporarily grounded so that parallel resonant circuits result. The required resonant frequencies are now coarsely aligned by aligning the L and C values in conjunction with the dipmeter.

After this, the two bridges at «a» and «b» are removed and the remaining capacitors are soldered into place. After this, the screening made from brass or tin plate, is soldered around the board. It is then possible for the fine alignment to be made on a sweep-measuring system. The lowpass filter only affects the upper vestigial sideband slope, and the highpass filter only the lower slope of the filter.

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Firma Telefunken, Backnang



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Weight (kg)	2.7	3.5
Wind loading (kg) at 160 km/h	22	33
Boom diameter (mm)	32	32

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GETTING STARTED ON THE 10 GHz BAND

by Dr. D. Evans, G 3 RPE

The information given in this article is probably well-known to British amateurs. However, since UK amateurs only represent a small percentage of our readers, we think that this information is too interesting just to be published in UKW-BERICHTE in German.

1. INTRODUCTION

Until recently, virtually all regular amateur activity on 10 GHz throughout the world was by the 50 or so UK amateurs who operate this band. Reference to **Figure 1** shows that they have been quite successful: some of the country-to-country contacts worth noting are GM-G over a 521 km path (the current world record). GM-GW at 322 km, G-PAØ at 306 km, with contacts from G to F, to GC and to ON being regarded by UK amateurs, at least, as nothing very special. Many contacts were made under genuine contest conditions, the best performance so far being that by G 3 KSU who worked 12 stations from the Isle of Wight in one event.



Fig. 1: Some of the paths worked on 10 GHz by UK amateurs. The 521 km contact probably is the world record.

One of the most gratifying aspects of these efforts is that almost all of the contacts made so far have been done with very simple low-power equipment usually generating only 1 - 20 mW and entirely home-made; however, operated with careful planning and much perseverance and skill. A striking feature is that some of the more successful operators have no previous experience whatsoever in the microwave area.

It is worth asking how this situation came about. During the late 1960's, UK amateurs became concerned about the slow development of the microwave bands, and tried to understand the reasons for it. It appeared to be mainly due to regarding microwaves as being simply an extension of VHF, from which viewpoint the higher frequencies could only appear progressively more and more difficult to cope with. Once this approach was questioned, it became obvious that microwaves are no more an extension of VHF than VHF is an extension of HF. It was realized that it was pointless to handicap ourselves by, for example, thoughtlessly insisting on using the narrow-band techniques which are essential on the crowded lower frequencies when there is over 1000 MHz of bands available above 1 GHz, and especially so when the conventional crystal oscillator/multiplier approach so common at lower frequencies may in fact be a quite inferior technique at higher microwave frequencies. Wideband techniques suddenly became of practical significance.

A second reason for the slow development was the presumption that the range of equipment became smaller and smaller the higher the frequency, eventually reducing to perhaps only a few kilometres. However, on asking the question: What equipment is easy for amateurs to build without elaborate facilities, and what is the best performance that can be expected, calculations for once gave a most encouraging answer. It appeared that with antennas of moderate gain, the amount of power required at any frequency to work any unobstructed path in the world is usually at the microwatt or milliwatt level. Since almost all practical transmitters generate very much greater powers, **however inefficient they are**, this means that success over unobstructed paths was virtually guaranteed. Since these paths could be up to several hundred kilometres in length, depending on the geography of the country, it meant that this order of long distance communication (DX) was just waiting to be taken. In other words, the DX that could be worked by this mode was not primarily set by the size and efficiency of equipment, but rather by how much skill and effort could be put into finding long unobstructed paths. Further calculations showed that to transmit via perfect ducts required even less power

These observations, which apply to all frequencies, were the starting point for the rapid development of interest in 10 GHz in the UK. Why 10 GHz? There were several reasons: there was the challenge of new techniques, the size of the equipment was convenient and some »surplus« components were already available. It was also far enough away from VHF not to be unduly influenced by presumptions about techniques, but the best reason of all was that it was difficult to find a good reason why not.

The intention of this article is to convince readers who might be interested in trying 10 GHz that, if the experience of UK amateurs is to be believed, then it is much simpler than they fear, certainly very different, and for many people the most rewarding amateur radio that they have ever done. The article will attempt to summarise UK experience with regard to both equipment and operating techniques. Much of the available information has been published by the RSGB in **Radio Communication** from 1970 onwards. Information available up to the end of 1974 is given in the third edition of the RSGB VHF-UHF Manual published in May 1976.

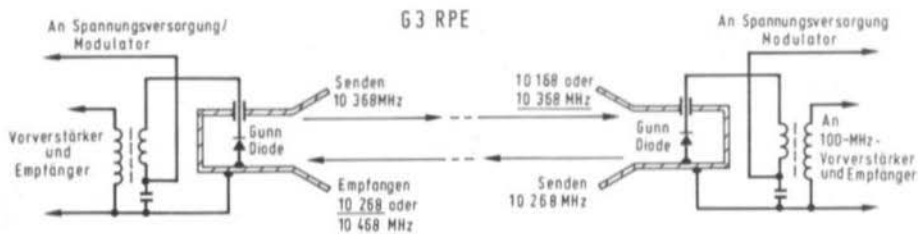


Fig. 2: Use of a Gunn oscillator as both the transmitter and self-excited mixer in a duplex system

2. A SIMPLE TRANSCEIVER FOR 10 GHz

Figure 2 shows schematically what is probably the simplest form of equipment possible. It is not particularly efficient, but nevertheless it enables a useful start to be made without too much expense or effort. All the components later can be used with others to be described to form a more efficient system. The 10 GHz part consists simply of a low-power Gunn oscillator which acts as both the RF source for the transmitter and also as a self-excited mixer, two similar units using a common IF giving the full duplex operation as shown. The maximum range to be expected over unobstructed paths will be up to about 50 km if antennas of 20 dB are used, and up to a few hundred kilometres with dishes 0.5 m diameter.

The RF section of a complete transceiver is illustrated by Figure 3. Constructional details for individual components are given below.

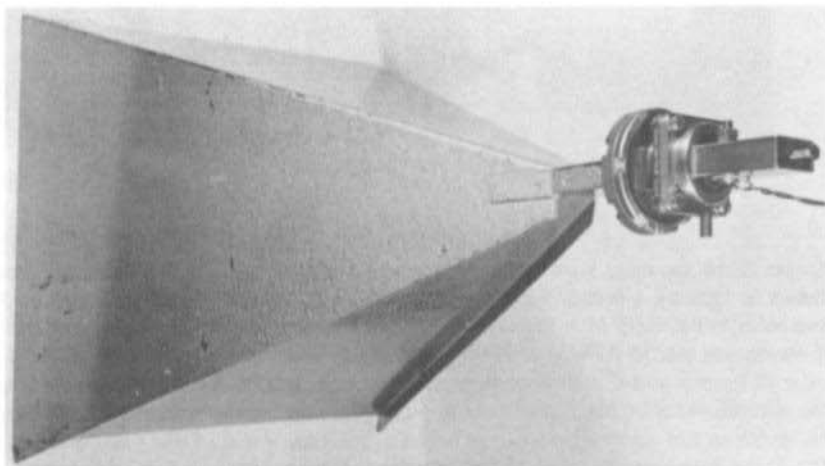


Fig. 3: Complete 10 GHz RF section using a Gunn oscillator and large horn antenna

2.1. A simple Gunn oscillator for 10 GHz

A Gunn oscillator consists of a resonant cavity into which a device called a Gunn diode is fitted. The diode is not a rectifier, but has the property that when typically 7 - 10 V DC is applied a steady current of about 150 mA flows, superimposed on which are current spikes which occur at roughly 10^{10} /s for a device intended for operation at 10 GHz. When operating within a cavity, the pulse rate becomes set by the resonant frequency of the cavity, and the pulses are converted into sinusoidal oscillations. For the input power levels quoted, which are characteristic of a low power diode, the output power will usually be in the range 1 - 20 mW. High power diodes generating up to a watt are available, but here are a number of problems in their use, such as dissipating the heat associated with their low efficiency.

The construction of the cavity does require care but seems to be within the capacity of most who attempt it. The most important factor is not accuracy as such, but mechanical reliability. It must be remembered that if an IF bandwidth of 100 kHz is in use, then a change of frequency of the oscillator by more than 1 part in 100,000 (1×10^{-5}), will take the signals out of the passband of the receiver. In practical terms, this means that the mechanical stability of the cavity itself, and especially that of the tuning mechanism, must be of a high order. Commercial oscillators often are simply not good enough in this respect and therefore cannot always be recommended for use by amateurs.

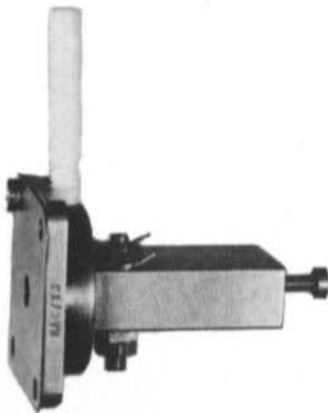


Fig. 4:
A simple but reliable Gunn oscillator with dielectric tuning

What appears to be the most successful of a large number of designs produced for amateur use is shown in **Figures 4 and 5**. While it operates to the highest standards, it can be fabricated even with hand tools in a few evenings. The cavity is based on a short length of a standard waveguide No. 16 (WR 90 or R 100), which has outside dimensions of 1.0 x 0.5 inch (25.4 mm x 12.7 mm) and a wall thickness of 0.050 inch (1.27 mm). The diode is mounted across the narrow width of the guide from a pillar. The maximum frequency of oscillation is set by the length of the cavity as measured between the centre line of this pillar and the front face of the fixing flange against which the iris plate is bolted. This length is very close to half the wavelength in the guide, i.e. $\lambda_{wg}/2$, at the frequency of operation, and selected values are given in Figure 5.

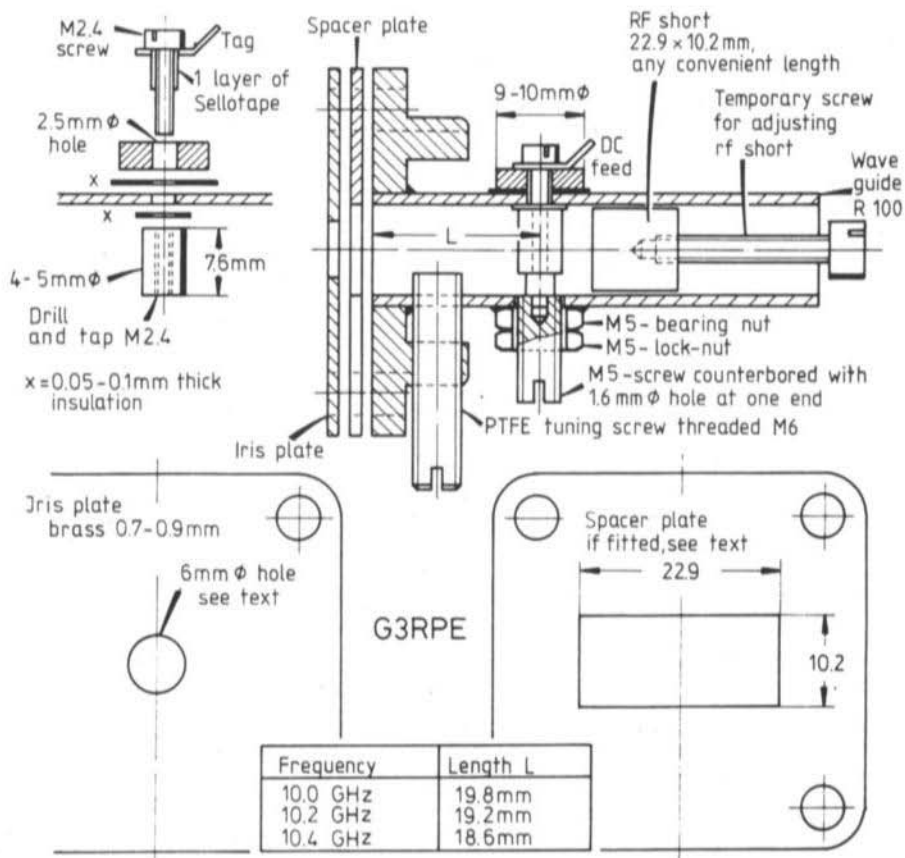


Fig. 5: Constructional details of oscillator. Suitable Gunn diodes are CXY 11 series (Philips), DC 1251 series (AEI)

2.1.1. Construction Details

On the centre-line of the broad face of the guide, at a point about 22 mm from one end, drill a 3/32 inch (2.4 mm) hole through both top and bottom faces in one operation. Enlarge the lower hole to 4 mm and tap with a partial 2 BA or M 5 thread. Fit a standard waveguide flange to give the appropriate cavity length, and using the tap to locate the 2 BA or M 5 bearing nut, solder both flange and nut in a single operation. The tap is then used to clear the thread. The excess waveguide in front of the flange can be removed by careful filing followed by rubbing on silicon carbide papers. If the latter are used with a sheet of glass as backing, and are lubricated with water, then a very flat surface can be achieved easily.

The RF short preferably should be a tight sliding fit in the guide. This can be done by filing to nearly the final dimensions and again finishing with carbide papers. If the short is a less tight fit, then it should be secured in place by screws through the walls of the guide when its optimum position is known.

It is important to remove all traces of flux residues and excess solder from inside the cavity as these are lossy at frequency. Burrs should also be removed especially from regions in contact with the insulation, the contact faces of the diode, and those of the iris plate.

It is important that the hole in the 2 BA or M 5 diode fixing screw is central otherwise there is a risk that the Gunn diode will be sheared when the screw is rotated. Determining the centre of a threaded rod is difficult and if this cannot be done in a lathe, then its position can be marked out by temporarily fitting a nut (and lock nut) to the screw and using this as a reference.

2.1.2. Alignment

The setting up the oscillator is as follows. With this size of diode there is no heat-sinking problem and the diode may be fitted with either end grounded. Care should be taken not to reverse the supply to the diode as this may damage it. With the RF short spaced from the diode pillar by a few millimetres, apply about 7 V from a power supply capable of supplying 200 mA. The voltage should be increased progressively (when the current should fall) until the oscillator will tune reliably and will restart immediately each time on applying the voltage. The position of the RF short can then be optimised and then clamped in place.

The tuning screw is preferably made from PTFE (teflon), but nylon, perspex or even glass can be used. This use of a dielectric for the screw is highly recommended; in terms of reliability of tuning, it appears to be far superior to any using a metal as the tuning screw no matter how well engineered. With the thread specified (which is about 6 mm diameter, 1.2 mm pitch), the tuning rate is about 100 MHz/turn, which is ideal if a 6 : 1 slow-motion drive is used. Alternatively, a smaller screw with a finer thread may be used with some form of turn-counting mechanism. In either case, as much care should be taken with the reliability of the drive mechanism as at lower frequencies – the precision required is about the same. The tuning range can be several hundred MHz, and the tuning rate is fairly linear except when the dielectric screw is just entering the guide or just approaching the opposite wall. Metal screws, by contrast, produce very non-linear tuning. This wide tuning range means that the length of the cavity is rather less critical than at first sight. However, if the cavity tunes too low, its maximum frequency may be increased by grinding back the front face of the fixing flange. If it will not tune low enough in frequency, then a spacer plate can be fitted between the flange and the iris plate as shown in Figure 5. In either case, the change in frequency is in the region of 300 MHz/mm.

The oscillator may also be tuned over a few megahertz by varying the voltage applied to the diode. The frequency shift depends on the Q of the cavity and the operating point of the diode but usually is in the range 1 - 20 MHz/V. This characteristic is used to modulate the output to produce a mixed AM/FM signal which can be of very high quality. It can also be used for fine tuning and the means by which AFC can be applied. Alternatively, varactor

tuning can be used for these purposes, but so far has not been widely used by amateurs because of the extra complication and because the maximum tuning range with high Q cavities may be restricted to a few tens of MHz.

The diameter of the hole in the iris plate is a compromise. Enlarging it generally will increase the output power, although the stability and tuning range of the oscillator may suffer. On the other hand, the stability of an erratic oscillator can often be improved by reducing the size of the iris. It is worth noting that an unstable oscillator is of little use even if it is generating much power, whereas reducing the output even drastically to achieve stability may have little effect in practice on the DX performance of the equipment. Stability, not output power, is top priority.

2.2. Power supply / modulator

The following points can be made regarding this unit:

- a) The output voltage should be adjustable over the range 7 - 10 V, and it should be able to supply at least 200 mA. It is an advantage if the output voltage is stabilized, although this is not essential except with unreliable oscillators which often are unduly sensitive to variations of the operating voltage.
- b) It should be possible to modulate the supply with tone or speech at a level up to a few hundred millivolts. It is useful to be able to tone modulate the local oscillator in a receiver as this allows unmodulated signals from harmonic generators, for example, to be detected.
- c) The psu/modulator/oscillator combination should be carefully checked for parasitic oscillations which can also modulate the signal besides causing other problems. Stabilized power supplies seem to be particularly prone to oscillate at VHF. Gunn diodes, being negative resistance devices, tend to oscillate with stray inductances and therefore it is a good practice, wherever possible, to fit a suppressor across the DC input terminals of the Gunn oscillator. This can consist of a 0.01 μ F capacitor in series with a 10 - 100 Ω resistor. A zener diode rated at about 0.5 V above the maximum working voltage of the oscillator, including modulation and AFC voltages, is a useful safety feature.

For preliminary testing, any suitable variable power supply can be used. The simplest method is probably a zener diode selected to be of the optimum working voltage connected across the Gunn diode, with a dropping resistor from a 12 V car battery, for example. A typical comprehensive power supply/modulator which operates from an unstabilized 12 V DC supply is shown in **Figure 6**.

2.3. Preampifier and IF strip

As has already been emphasized, it is unlikely that the communication that can be achieved will be affected by deficiencies in the system amounting to a decibel here and there. One consequence of this is that the choice of IF can be made in terms of convenience rather than strict technical considerations. There is one exception: if two stations operate a duplex system of the type shown in **Figure 2**, then they must use the same IF if continual retuning is to be avoided.

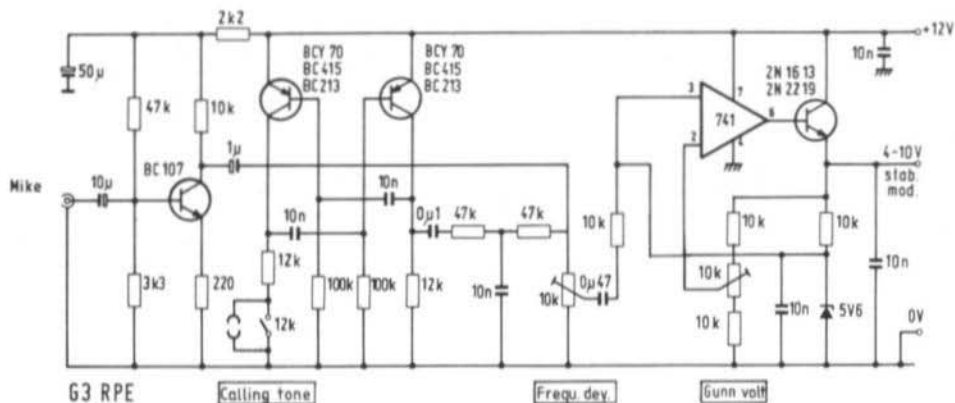


Fig. 6: Typical power supply/modulator for a low-power Gunn oscillator

The development of reliable oscillators which can be tuned over a wide range, together with equipment rarely being operated remotely, has meant that simple single conversion receivers are most common. The alternative system, that of a fixed frequency first local oscillator used with a tunable IF is much less common mainly because it is more difficult to produce a low noise IF tunable over a wide range (at least 100 MHz) and to avoid IF breakthrough particularly when operating from elevated sites. The most common forms of IF amplifier are as follows:

- Standard FM broadcast receiver commonly tuned to a quiet frequency near 100 MHz. Sometimes the circuitry is removed and fitted into a metal box to improve screening. The AFC voltage, if available, is sometimes used to control the frequency of the Gunn oscillator.
- A 30 MHz IF using standard 10.7 MHz FM components, usually built up from parts but sometimes produced by modifying the signal frequency and local oscillator circuits of an FM broadcast receiver. This IF seems to be particularly popular amongst those who use balanced mixers.
- A 10.7 MHz IF is sometimes used but the confusion caused by the small spacing between signal and image channels and oscillator noise problems make this IF less popular.
- TV IF strips retuned to 30 MHz by adding extra capacitance across the tuned circuits. This modification also reduces the unnecessarily large bandwidth.
- Conventional **narrow-band** AM receivers can be used to detect tone modulated wideband signals with little loss in efficiency provided that the IF bandwidth exceeds the frequency of the tone modulation, usually 1 kHz.

It is usual to use a preamplifier mounted close to the mixer (whatever its form) to deal with all the DC and RF problems associated with the mixer circuitry and also the possibly uncertain input characteristics of the main IF strip which may be located some distance away. Typical tuned and untuned preamplifiers are given in **Figures 7 and 8**.

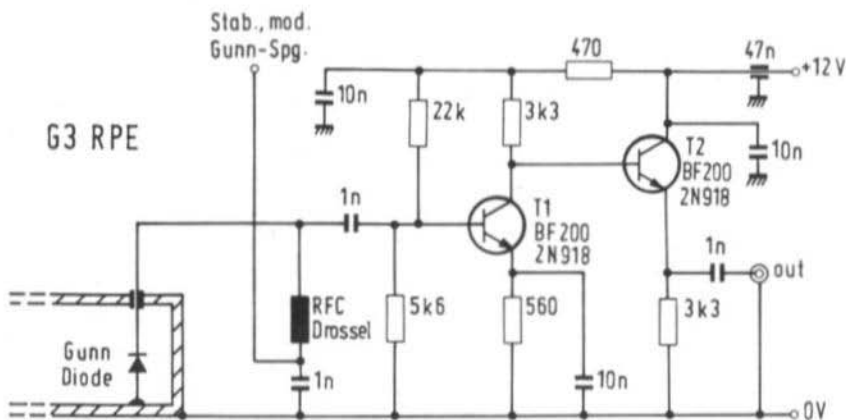


Fig. 7: Simple «untuned» preamplifier
 RFC should range from about 100 μ H at 10 MHz to 3 μ H at 150 MHz

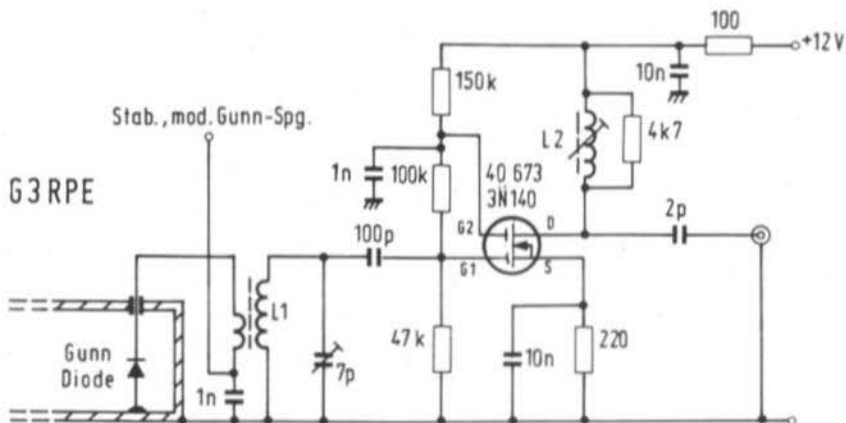


Fig. 8: Simple tuned preamplifier for 100 - 150 MHz
 L 1 5 turns 20 swg copper wire, 0.25 inch inside diameter,
 0.5 inch long, primary 2 turns thin pvc-covered wire.
 L 2 6 turns 20 swg enamelled copper wire on 0.25 inch former

2.4. Antennas

An interesting feature of antennas on 10 GHz is that, probably for the first time, amateurs can produce antennas of such high gain that they are not able to cope with the problems associated with the corresponding narrow beamwidths. Indeed, it has been suggested that more 10 GHz contacts probably have been lost by having antennas of too high a gain than too low for this reason. Fortunately, most operating can be done with relatively small antennas, the simplest of which surely must be the open end of the waveguide. This has a gain of about 5 dB, very low by 10 GHz standards, but nevertheless sufficient with equipment such as described above to give a range of a few tens of kilometres. Note that horizontal polarisation, which has been adopted as standard in IARU Region I, is obtained when the broad face of the waveguide is vertical.

What is probably the most useful form of antenna for initial operation is the large horn, details of which are given in **Figure 9**. The waveguide feed is expanded in both planes to produce a rectangular funnel-like shape. This type of antenna has a number of advantages. Firstly, its performance is substantially unaffected by quite large errors in construction. Secondly, it presents a well-matched load without any setting up, which is especially valuable when the transmitter uses free-running oscillators such as Gunn oscillators or klystrons as their stability is sometimes dependent on the match. A third advantage is that the gain of the horn can be predicted from its dimensions within a dB or so. This is useful in checking the overall performance of the complete system, and it also means that the horn can later be used as a reference against which other forms of antennas can be judged.

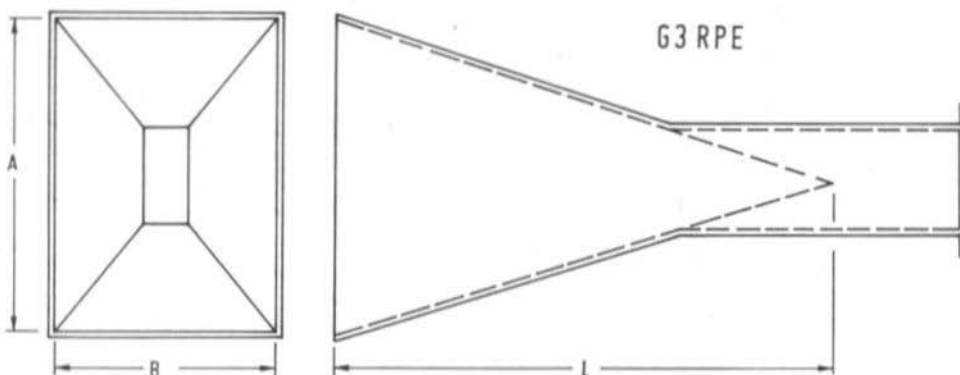


Fig. 9: Dimensions of a horn antenna for use at 10 GHz

Nominal Gain at 10.368 GHz	3 dB Beamwidth	Dimensions (mm)		
		A	B	L
15 dB	$\pm 15^\circ$	70	57	57
20 dB	$\pm 8^\circ$	128	104	189
25 dB	$\pm 5^\circ$	220	180	568

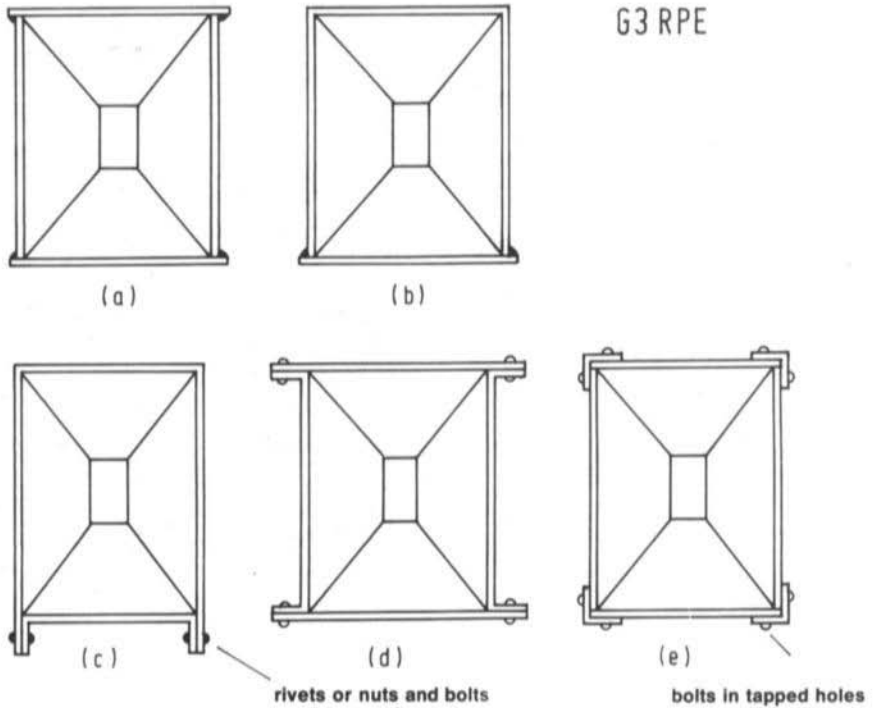


Fig. 10: Various methods for constructing horns

The construction of horns is not too difficult once the general principles are grasped: one of these is that for a body that looks as though it should be full of right angles, there are precious few. Consequently, most people get into an awful mess with their first attempt, and it is therefore worthwhile doing a practice run with perhaps cardboard before moving on to materials such as tinplate, brass or copper. Aluminium sheet is rarely used in this application because of the difficulty in soldering this material reliably.

An awkward point in construction is the junction between the horn itself and the waveguide feed. For all but the smallest horns, some strengthening is necessary, and this can be done by strips bolted using brass screws into holes tapped into the walls of the horn and the guide which subsequently are soldered in place. These strips also have the function of jiggling the parts prior to soldering. It is important that any of the screws projecting into the inside are carefully filed away, and that the transition between the guide and the horn is smoothly rounded. Various methods of making up the horn are summarised in **Figure 10**. It is sometimes a convenient method of fabrication to braze on two or three sides of the horn in one operation, clean up the joint, and then solder the remainder. Note that it is important to have good electrical contact at the corners but at the same time excess solder and any flux should be removed as these materials are rather lossy.

A POWER AMPLIFIER FOR THE TWO METER BAND USING THE TUBE QQE 06-40

by H.J. Dierking, DJ 6 CA

Tubes still have their applications in the higher power range even in our modern semiconductor age. The advantages of tubes such as the double tetrode QQE 06/40 (two tubes with four electrodes each in a single glass envelope) are able to take short over-current and over-voltage conditions far better than semiconductors. Furthermore, they normally provide a narrower output spectrum, that is an output signal with less intermodulation products when correctly aligned for linear operation, and usually provide more gain than a transistor stage.

The following article is to describe a simple power amplifier for the frequency range of 144 to 146 MHz equipped with the tube QQE 06/40 (**Figure 1**). The basic design and circuit can, of course, also be used for other tubes using a common-cathode circuit. This is not to be a scientific article but more a practical description to allow simple construction.

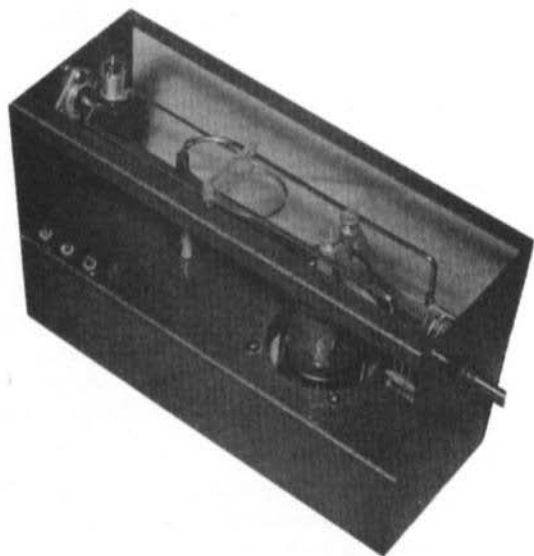


Fig. 1:
2-meter power amplifier
equipped with the tube
QQE 06/40

1. MECHANICAL CONSTRUCTION

The panels made from tin or brass plate should be cut and drilled according to the dimensional drawings given in **Figure 2**, and soldered to the U-pieces as shown. After construction, the surface should be cleaned with sandpaper and/or painted.

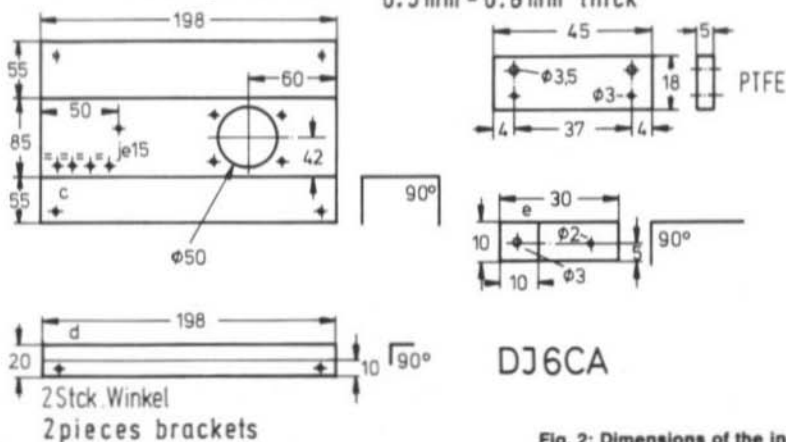
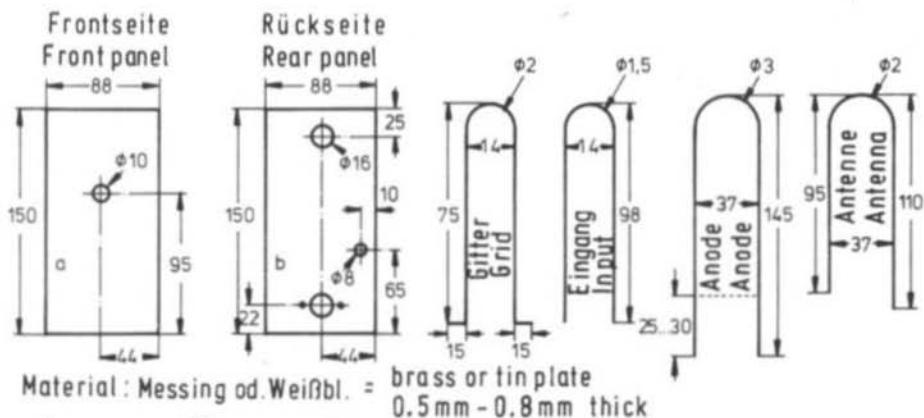


Fig. 2: Dimensions of the individual parts

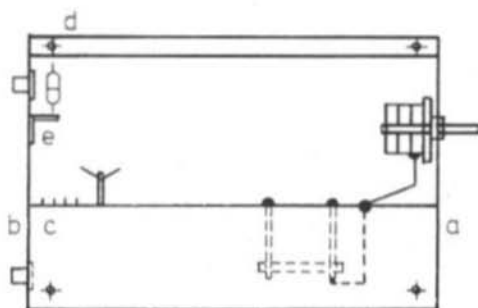


Fig. 3: Most important parts of the chassis

The individual components given in the parts list should be mounted as shown in **Figures 3, 4 and 5** and connected together as shown in the circuit diagram given in **Figure 6**. The tube socket is held in place with four M 4 - screws and is mounted below the chassis so that the metal plate of the tube system is at the same level as the chassis. The grounding point on the variable capacitor and at the cathode should be as short as possible and made with a wide metal tape to the chassis where it is soldered into place.

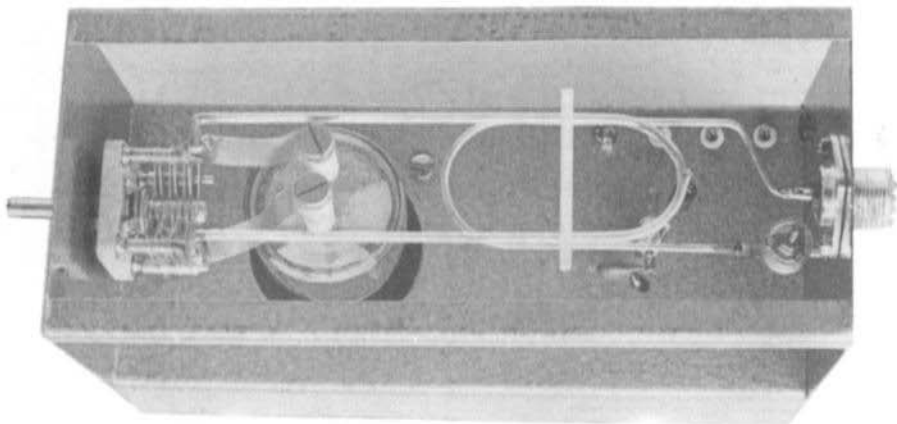


Fig. 4: Power amplifier from above

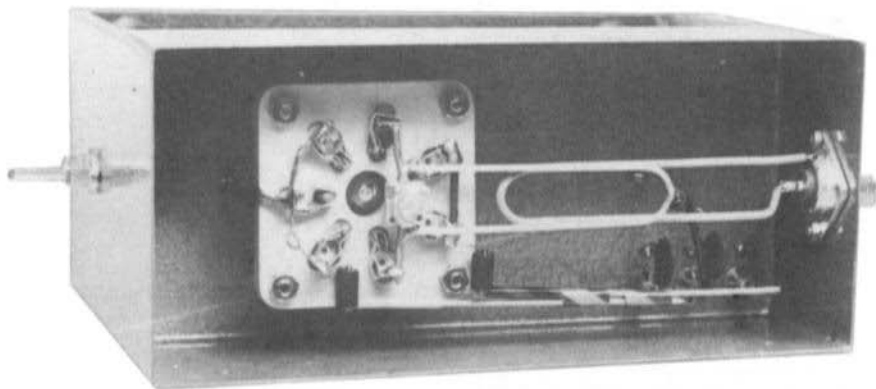


Fig. 5: Grid side of the QQE 06/40

The input coupling of the 144 MHz drive power is made with the aid of a link. It can, as required, also be provided with a parallel or series trimmer of 3 - 30 pF at the coaxial input socket and tuned for resonance. The drive can be varied within certain limits by altering the spacing of the coupling link to the grid circuit .

The grid circuit is a capacitively shortened $\lambda/4$ -circuit and is directly soldered to the grid connections of the tube base. A ceramic air-spaced trimmer for resonance alignment is provided between the two 90° bends of the grid lecher line. No neutralization of this power amplifier will be required since the anode to grid capacitances are neutralized within the tube itself.

This is very easy in push-pull stages by using capacitances that are just as large as the anode to grid capacitance and are placed from the anode of one tube to the grid of the other tube. These two neutralizing capacitances can easily be seen in the upper part of the envelope.

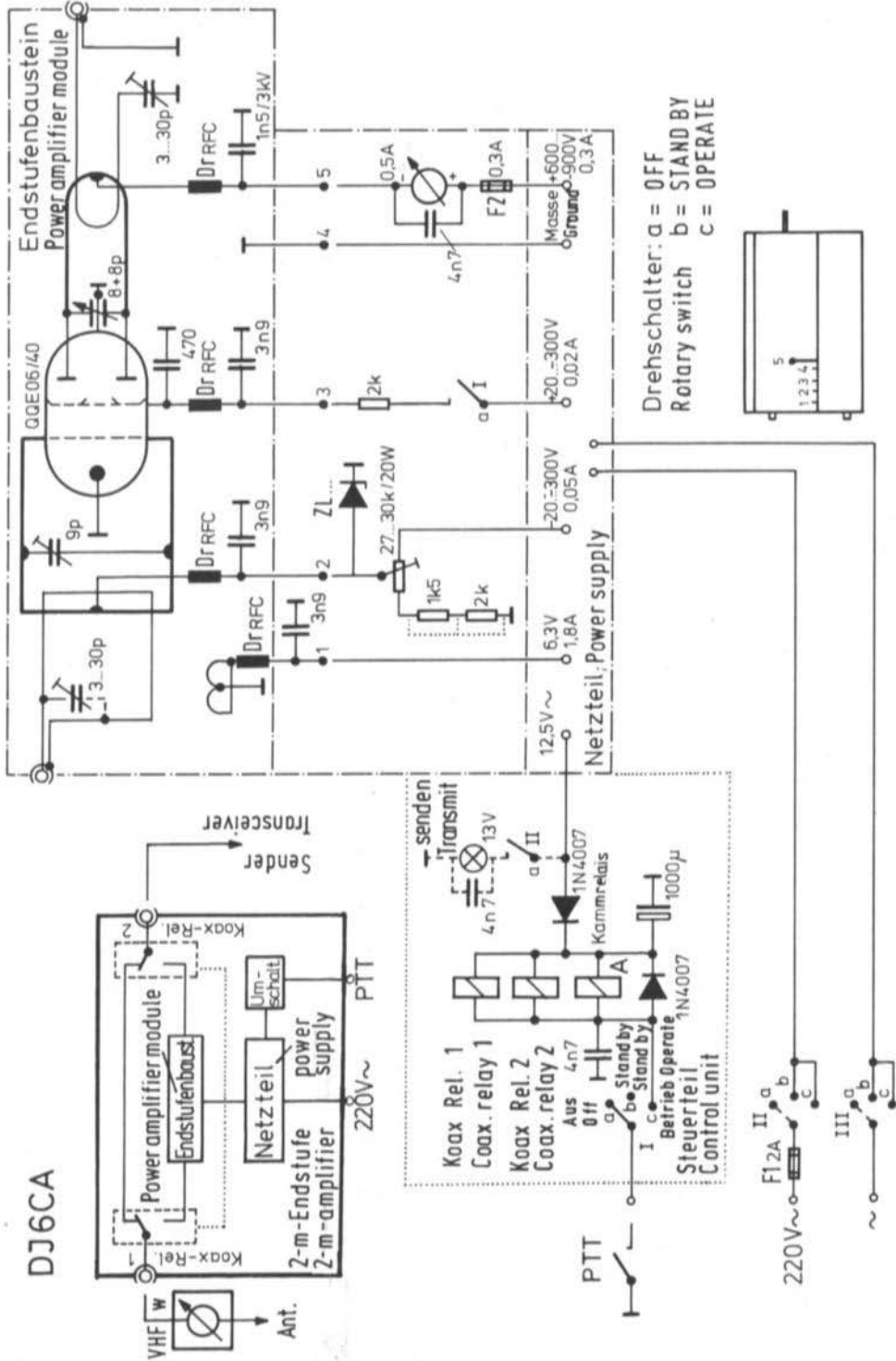


Fig. 6: Block diagram and circuit diagram

The anode circuit is also a shortened $\lambda/4$ -circuit. This circuit is held mechanically with the aid of the variable capacitor that is used to bring this circuit to resonance. It is necessary to provide a spacer manufactured from PTFE (teflon) or some other heat-proof isolating material before soldering this lecher line to the variable capacitor. Only the original anode clamps and silver-plated brass tapes should be used for the interconnection between the anode circuit and the anode pins of the tube. These anode clamps also help to cool the tube, and it is sometimes necessary to drill out their collars in the case of some tubes so that they fit tightly onto the pins.

The output power is taken from the amplifier with the aid of another lecher line that is also shown in Figure 2. The PTFE spacer of the anode circuit also supports the output coupling link and ensures that the required spacing is maintained. A trimmer of 3 - 30 pF is used to resonate the output coupling link. It has been found that it is not necessary for this trimmer to be accessible externally since it has been found that an alignment at the centre of the band is sufficient.

1.1. PARTS

The following parts are required in addition to the tube QQE 06/40 and the mechanical parts given in Figure 2:

- Tube socket, seven pin (Philips 40202)
- 2 anode clamps (Philips 40623)
- 1 tubular trimmer 3 - 30 pF (Philips)
- 1 air-spaced trimmer approx. 13 pF
- 1 variable capacitor, butterfly type, 2 x 8 pF, 1.5 mm spacing
- 3 chassis feedthroughs, ceramic or PTFE
- 1 ceramic support, approx. 20 mm high
- 4 screws M 4 x 35 (4 mm x 35 mm) with three nuts each
- 4 six-hole cores (Philips 4312 02036700)
- 3 ceramic disc capacitors 3.9 nF / 400 V
- 1 ceramic disc capacitor 1.5 nF / 3 kV
- 1 ceramic disc capacitor 470 pF / 400 V

2. CIRCUIT

The power amplifier module is connected to the power supply and exciter as shown in the circuit diagram and in Figure 6. The power amplifier is controlled via the PTT-line since this is cheaper and more reliable than a RF-vox circuit. Two coaxial relays are used for switching the input and output. A small relay is used to switch the screen grid voltage (switching current approx. 20 mA), and any pilot lamps if required.

In the receive mode, the tube is blocked by switching off the screen grid voltage (relay contact a I). This is more favorable than blocking the tube with a high negative control grid voltage since the tube could possibly generate a noise level which would be audible in the receiver.

The bias voltage for the control grid is variable so that the power amplifier can be used in class A, B, or C. Class A can be used for all modulation modes. The following values are valid for the QQE 06/40 independent of the anode and screen grid voltage:

Class A	30 - 40 mA
Class B	1 - 5 mA
Class C	0 mA

The plate voltage should always be fed via a fuse of approximately 0.3 A and a mA-meter. The fuse protects the power supply in the case of flashover within the tube.

At a plate voltage of $U_a = 900$ V, the screen grid voltage U_{g2} should not be in excess of 240 V. At $U_a = 600$ V, it is possible for U_{g2} to be 280 V. The control grid bias voltage U_{g1} need not be stabilized. If the power amplifier is only to be used in one mode (class), it is possible for U_{g1} to be stabilized with the aid of a power zener diode (ZL ...). The voltage value of the zener diode should be found experimentally. Usually, a diode ZL 32 is suitable. Of course, nearly every power supply of a shortwave transceiver using a tube PA can also be used for this power amplifier.

3. ALIGNMENT

A VHF wattmeter with terminating resistor (dummy-load) able to handle approximately 200 W is required for alignment of the described power amplifier, as is a multimeter. The anodes of the tubes represent a good alignment indicator since they start to glow at high dissipation power levels.

After connecting the power amplifier to the power supply, dummy-load, exciter and control line, it is then possible to check the individual voltages without drive at the connection pins of the tube. In order to protect the tube, a resistor of approx. $4 \text{ k}\Omega / 2 \text{ W}$ is placed in the grid lead.

The PTT contact is grounded at $U_{g1} = -50$ V and the quiescent plate current is aligned with the control grid potentiometer to 30 - 40 mA. Commencing at a low drive level, the grid and anode circuits are aligned as well as the input and output coupling. It should now be possible to read off some output power indication on the wattmeter.

The following values are given for orientation in the case of the tube QQE 06/40:

Heater:	6.3 V / 1.8 A
$U_a = 900$ V,	$I_a = 250 - 300$ mA ($I_{a0} = 30 - 40$ mA)
U_{g2}	= 225 V / approx. 20 mA
U_{g1}	= -30 to -32 V

In the case of a lower plate voltage, the screen grid voltage may be increased slightly. The anode current will be somewhat lower in this case. The efficiency should be at least 50 % if the given voltage and current values are approximately achieved. This means that the output power is in the order of 125 to 130 W with a DC input power of 250 W. If this is not the case,

a miss-match condition will exist and the anodes of the tube will glow. If the maximum output power is obtained with the antenna trimmer (30 pF) at maximum capacitance, it will be necessary to shorten the output coupling link by 1 to 2 cm so that the trimmer is at roughly half capacitance in the case of maximum output power. Length and capacitance values are dependent on the selected plate voltage since they are dependent on the power impedance R_i of the tube which must be transformed to the cable impedance R_{out} .

4. FINAL DETAILS

A good standing wave ratio should exist between the exciter and power amplifier. This is achieved by carefully optimizing the input coupling. In the case of linear operation (class A), the required drive power should be made with the aid of a potentiometer built into the exciter (see separate article from the author regarding this point). In the case of class C (possible for FM and CW) the efficiency will be higher, but a higher drive power will be required (approx. 6 to 10 W).

No forced air cooling of the tube is necessary, but air should be able to freely flow around the tube and exit the case. This means that holes should be provided on the lower and upper panels of the case.

With the input and output terminated, no tendency to self-oscillation should be observed. If any tendency to self-oscillation is assumed, this will no doubt come from the fact that RF voltage is present at the input of the modulator. This can be cured by placing a carbon resistor of 1 to 5 k Ω in series with the input, and placing a ceramic disc capacitor of 4.7 nF to ground directly at the base of the first transistor of the modulator.



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REDUCING THE OUTPUT POWER OF TRANSISTORIZED SSB TRANSMITTERS AND TRANSVERTERS

by H.J. Dierking, DJ 6 CA

Linear amplifiers or UHF transverters often require far less drive power than provided by the available transmitter. Since it is not economical to burn up the RF power in a well matched power attenuator, two methods of adjusting the output power in the low-signal part of the transmitter are to be recommended that do not affect the modulation quality.

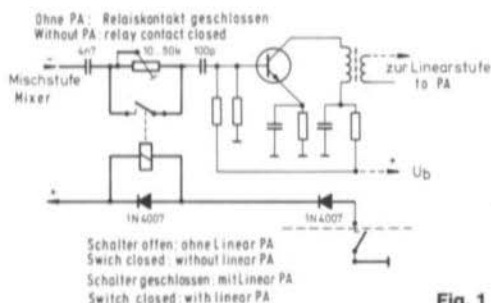


Fig. 1

The thick lines shown in **Figure 1** show the modifications that should be made between the last mixer stage and the first linear amplifier stage. The principle used is that a trimmer potentiometer is placed into the circuit that can be bridged when no linear amplifier etc. is used. The additional isolating capacitors, the trimmer resistor and the relay contact should be connected with the aid of short leads and mounted in the direct vicinity of the actual stage. The relay should have low-capacitance contacts (wire contacts, Reed relays etc.). It should be possible with the aid of the trimmer resistor to reduce the output power of the transmitter by approximately 20 dB. The relay can be switched using an additional contact on the relay of the linear amplifier.

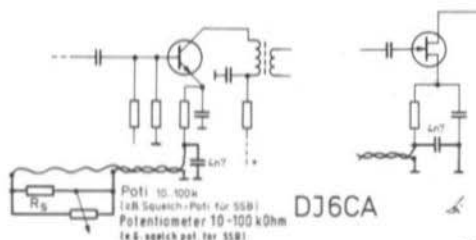


Fig. 2

Figure 2 shows how the power can be reduced in a buffer or amplifier stage of the SSB exciter. It is usually only necessary to disconnect the ground lead of an emitter resistor of such a stage and to connect this with the aid of a twisted wire. This wire is fed to a potentiometer with a value of between 10 and 100 kΩ. It is often possible to use the squelch potentiometer, which means that no additional holes must be drilled, and no external modifications need be made to the equipment. However, it may be advisable to connect in parallel with the potentiometer a carbon resistor as shown in **Figure 2**. The resistance value must be found experimentally.

OVERTONE CRYSTAL OSCILLATORS in Series and Parallel Resonance

by H.J. Brandt, DJ 1 ZB

Although overtone crystal oscillators are well used in amateur radio technology, their exact operation seems still not to be completely known. A large number of amateurs are of the opinion that all overtone crystals operate at series resonance (1, 2), mainly because industrial companies virtually only use series resonance for their equipment.

When viewed on a frequency counter, many readers will have noticed that the same crystal will oscillate at different frequencies in different circuits, without this being able to be corrected sufficiently with a trimmer. This difference can amount to several kHz and can cause difficulties in a converter for instance, where the transposed frequency on the HF receiver is not correct. Such a frequency deviation results by energizing the crystal in a different resonance than was planned.

On studying popular overtone oscillators and their operation, it was found that there are several proved circuits available since the tube days that operate in parallel resonance. Due to the large number of overtone oscillators known today, it seemed important that the difference in circuitry should be discussed both with respect to series and parallel resonance.

1. BASIC OSCILLATOR CIRCUITS

Figure 1 shows four basic circuits of oscillators (3). A Hartley oscillator is shown in (a) with inductive division, and a Colpitts oscillator with capacitive division in (b). If the inductance components of the Hartley oscillator are extended to provide two inductively tuned resonance circuits this will result in the Huth-Kühn oscillator shown in (c). If only one inductance is used with a Hartley oscillator which is then tuned with a capacitor, one will obtain the classic Meissner oscillator with inductive feedback as shown in (d).

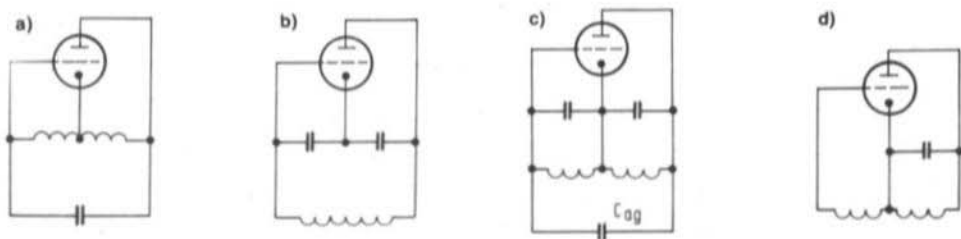


Fig. 1: Basic oscillator circuits a) Hartley b) Colpitts c) Huth-Kühn d) Meissner

Only the RF paths have been shown in these diagrams. The ground can be connected as required to one of the three tube electrodes, without interfering with the operation of the oscillator. In this manner, it is possible for the various oscillator circuits to be easily modified. Semiconductor technology has brought a further flexibility due to the deletion of a heated cathode.

2. CRYSTAL OSCILLATORS

2.1. Meissner Oscillator

In the case of the Meissner oscillator, it is virtually only possible to place a crystal in the grid lead in series with the feedback inductance (**Figure 2a**). This means it is the series resonance of the crystal that is used in this case. Its main characteristic is that the crystal can be replaced by a coupling capacitor without altering the operation of the circuit, of course, with the exception of the stability.

A transistor version of this circuit is shown in **Figure 2b**. In this case, the holder capacitance of the crystal is neutralized in a bridge circuit comprising a balanced feedback winding. This is to ensure that no spurious oscillation of the circuit can take place at higher frequencies in addition to that of the crystal resonance. The Meissner oscillator is just as suitable for fundamental wave and overtone crystals since the crystal resonance to be used can be set clearly with the aid of the resonant circuit tuning.

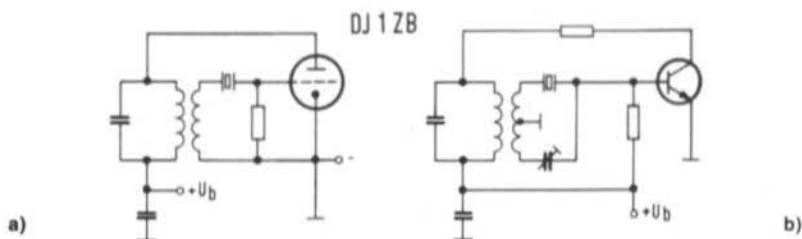


Fig. 2: Meissner oscillator with crystal control

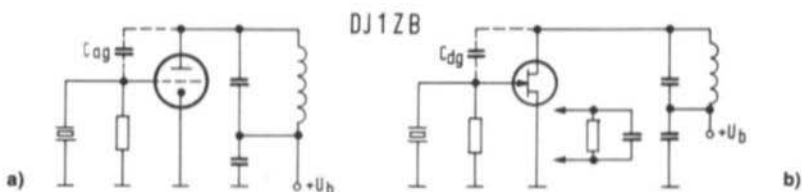


Fig. 3: Huth-Kühn oscillator with crystal control

2.2. Huth-Kühn Oscillator

In the case of the Huth-Kühn oscillator, the grid resonant circuit is replaced by a crystal (see **Figure 3a**), and the crystal will oscillate at parallel resonance. If the crystal was replaced by a capacitor in this circuit, the input of the tube would be short-circuited and no oscillation would be possible. This simple experiment allows one to differentiate between series and parallel resonant circuits.

It would be difficult to construct a Huth-Kühn oscillator using a bipolar transistor circuit, but a field effect transistor can be used since its input impedance is sufficiently high (see **Figure 3b**). However, the feedback capacitance of the FET must be sufficiently high, or it will be necessary to provide additional capacitance. This circuit is also just as suitable for operation with fundamental wave and overtone crystals since it possesses a resonant circuit.

2.3. Colpitts Oscillator

In the case of the Colpitts oscillator, the crystal is placed in the grid or cathode lead as shown in **Figures 4a and 4b**. Both these circuits can easily be obtained using bipolar (**Figures 4c and 4d**) or field-effect transistors (**Figures 4e and 4f**). Series resonance is used in all cases. At higher frequencies, an inductance is usually provided parallel to the crystal for neutralizing the crystal holder capacitance. This is aligned so that it produces a parallel resonance circuit at the oscillating frequency together with the unwanted capacitance. If a capacitor is provided in series with this inductance, it is only provided for DC-blocking.

If an RC-combination is placed in the source line of the circuit shown in 3 b in order to provide an additional method of adjusting the operating point, it will then be difficult to differentiate from the circuit given in Figure 4e. The operation of the circuit, immaterial whether Huth-Kühn or Colpitts oscillator, is only dependent on the value of the source bypass capacitor, in other words whether the source is hot or cold for RF. This example shows clearly that it is often very small differences on which it is dependent whether the crystal operates at series or parallel resonance.

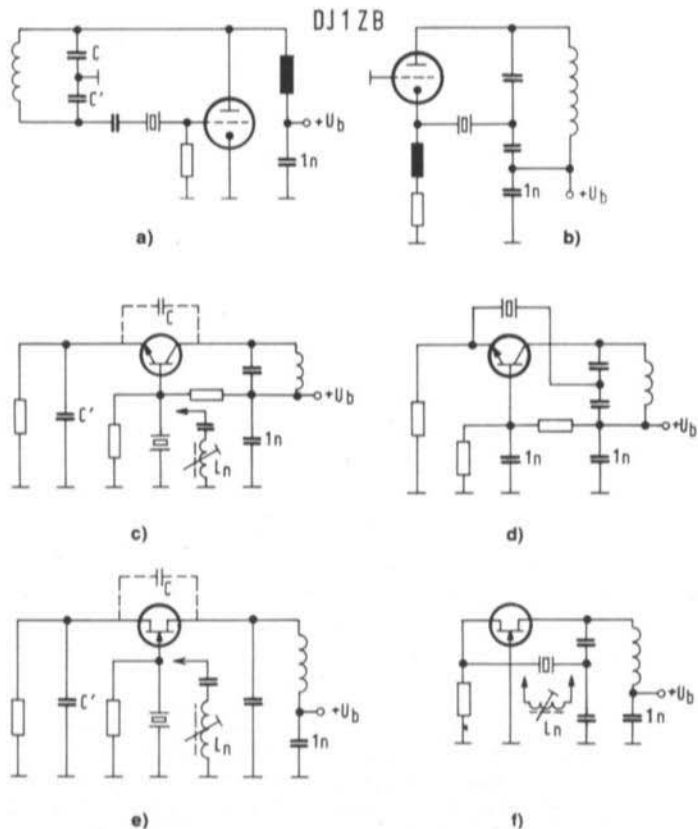


Fig. 4: Colpitts oscillator with crystal control at series resonance

2.4. Colpitts Oscillator for Parallel Resonance

As in the case of the Huth-Kühn oscillator, it is also possible with the Colpitts oscillator to replace a complete parallel resonant circuit with the aid of a crystal. In this case, it is once again the parallel resonance that is used (see Figure 5a). This circuit can also be used in conjunction with a bipolar or a field-effect transistor. However, it is not suitable for use with overtone oscillators since these would normally oscillate at their fundamental wave.

In order to avoid this, it is necessary to provide a resonant circuit for the required overtone frequency in parallel with one of the capacitors (if possible at the high-impedance output). This is shown in Figure 5b. In practice, capacitors C_2 and C_3 are usually combined but the fact still remains that the resulting resonant circuit should be tuned to the capacitive side of the overtone frequency, in contrast to the Huth-Kühn oscillator whose output circuit is aligned to the inductive side of the crystal frequency. This circuit can also be used just as well with a bipolar or field-effect transistor as shown in Figures 5c and 5d. Attention must be paid when using bipolar transistors since they possess a distinct tendency to parasitic VHF-UHF oscillation when resonant line circuits result from the connection inductances of capacitors C_1 and C_2 when constructing the circuit. For this reason, resistors of several tens of Ohms (R_1 , R_2) are used in Figure 5c which are placed in the collector and base leads in order to dampen these parasitic circuits. This neutralization can also be necessary in the case of other overtone circuits.

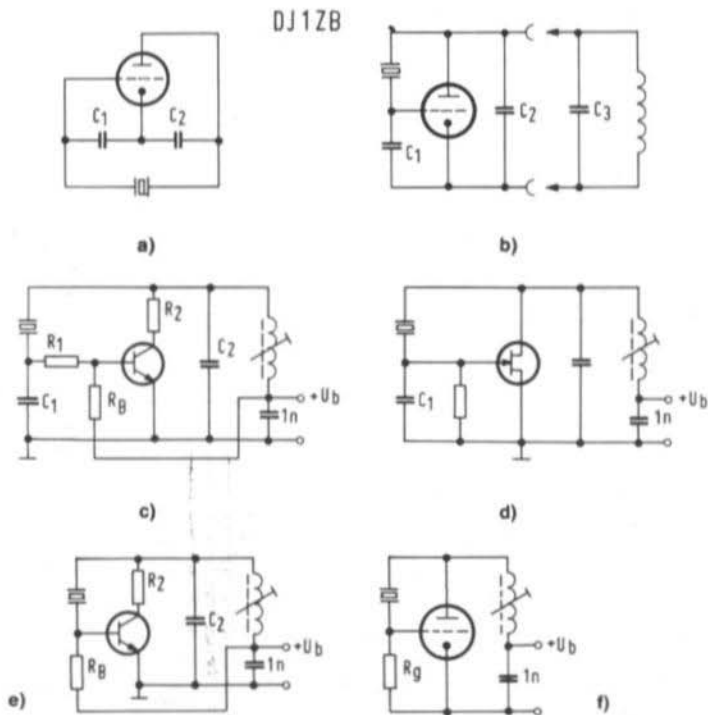


Fig. 5: Colpitts oscillator circuits with crystal control at parallel resonance

At higher frequencies, the input capacitance (Figure 5a) and the output capacitance (Figure 5f) of the amplifying element can be sufficiently great in order to take over the feedback function of capacitor C_1 and C_2 . In this case, external fixed capacitors and damping resistors can be deleted since it will not be possible for spurious resonances to exist.

The operation of the circuits shown in Figures 5b to 5f remains principally the same if the crystal is not placed at the hot end of the circuit but to a coil tap, or placed at another position in the circuit capacitance.

3. COMPARING THE CIRCUITS

3.1. Advantages of the Overtone Parallel Resonant Circuit

With parallel resonance, the crystal is operated in the same way as the resonant circuit with a very high Q. The unwanted crystal holder capacitance in the series resonance mode now forms part of the circuit and need no longer be neutralized. Even when the external circuit or capacitance C_1 is greatly detuned, it will still not be possible for any uncontrolled oscillation to take place in addition to the crystal frequency. This clean operation is especially an advantage when the measuring instruments of a radio amateur are limited, and probably indicates the reason for the popularity of the oscillator circuits shown in Figures 3 and 5.

If a large variation range of the external resonant circuit is provided, the oscillator will directly jump from one overtone to the next. It is possible in this manner to establish the highest overtone of an unknown crystal. This is dependent on the plane parallelity of the contact surfaces.

3.2. Advantages of Series Resonance

In the case of series resonance, variations of the oscillator frequency due to external effects are far smaller than would be the case with parallel resonance. This means that such circuits are of advantage when a high frequency stability is required. Of course, series resonance will always be required when the crystal is specified for series resonance and should really oscillate at its nominal frequency.

4. NOTES

4.1. Practical Differentiation

One must often consider whether a circuit is oscillating at series or parallel resonance. In order to differentiate between these two modes, the following experiment can be made: The crystal should be replaced by a ceramic capacitor of approximately 4.7 nF. If the oscillator then operates with approximately the crystal frequency and with approximately the same amplitude, this will indicate that it is a series-resonance circuit. If, on the other hand, the oscillator no longer operates, or operates wildly in another frequency range, this will indicate that it is a parallel-resonance circuit. The circuit can then be compared with the circuits given in this article for further details.

Two-stage oscillator circuits such as the Heegener and Butler circuit, as well as oscillators operating with logic-gates, always operate at series resonance. Such circuits are not considered here.

4.2. A further Reason for Deviations of the Crystal Frequency

Finally another cause why a crystal should oscillate at a frequency different than its nominal frequency. In amateur radio technology, one often finds overtone oscillators (4) in which, for instance, a 38.666 MHz crystal is energized in series resonance, whereas the third harmonic of 116 MHz is coupled out. Since this circuit is either not provided with a resonant circuit for the frequency 38.666 MHz or only with a resonant circuit which is capacitive for 116 MHz, the crystal will probably be energized at its ninth overtone, and not as planned at its third overtone. This means that it cannot be guaranteed that the frequency generated in this manner is really the third harmonic of 38.666 MHz, since the overtone resonances are not exact multiples of another. The loading of the crystal disc by the contact surface and electrodes has a greater effect at some overtones than others; the crystal is aligned in the factory only to the required overtone (nominal frequency).

In order to obtain a clean spectrum to avoid spurious reception points, it is a great advantage when the local oscillator frequency required in the converter is directly generated. If an accurate oscillator frequency is required, it is either necessary for a 38.666 MHz crystal to be used at its nominal overtone and provided with a subsequent tripler, or for a more expensive 116 MHz overtone crystal to be used.

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INTERESTING LINEAR INTEGRATED CIRCUITS

by E. Schmitzer, DJ 4 BG

Several companies have brought new linear integrated circuits onto the market that are mainly designed for broadcast receiver use. However, many of these integrated circuits also offer very interesting applications for amateur radio equipment. This is to be discussed with the aid of integrated circuits manufactured by Siemens and Telefunken.

A short functional description, specifications and circuit diagrams taken from the data sheets are to be shown, which is to be followed by several recommendations on how to use these integrated circuits in amateur radio equipment. If further information is required, this can be taken from the data books and data sheets of the manufacturer. The integrated circuits are to be considered in the order of their type numbers.

1. SOME INTERESTING INTEGRATED CIRCUITS

Siemens TDA 1037

AF power amplifier

Specifications:

Operating voltage range:	4 - 28 V
Output current:	2.5 A
Output power (distortion = 10 %):	$U_b = 24 \text{ V}, Z = 16 \Omega: P_{out} = 5.5 \text{ W}$ $U_b = 18 \text{ V}, Z = 8 \Omega: P_{out} = 5.0 \text{ W}$ $U_b = 14 \text{ V}, Z = 4 \Omega: P_{out} = 5.0 \text{ W}$

The internal diagram of the integrated circuit is given in **Figure 1** together with the test and applicational circuit. The new case (single-in-line) allows a very compact construction.

Siemens TDA 1046

AM receiver circuit with demodulator

This integrated circuit is designed as AM receiver module for car radios, and AC broadcast receivers. Provided are a controlled RF-stage, a push-pull mixer, amplitude-controlled oscillator, controlled IF amplifier, dual-path demodulator, active lowpass filter, as well as an amplifier for indicating the field strength.

Specifications:

Operating voltage range:	8 - 18 V
Oscillator frequency:	0.5 - 31 MHz
Input frequency (RF-module):	0 - 30 MHz
Input frequency (IF-module):	0.2 - 1 MHz
AF-output voltage ($m = 80\%$):	800 mV (RMS)
Total control range:	85 dB
Input sensitivity:	60 Ω , $m = 30\%$, $(S+N)/N = 6 \text{ dB}$: 2.5 μV

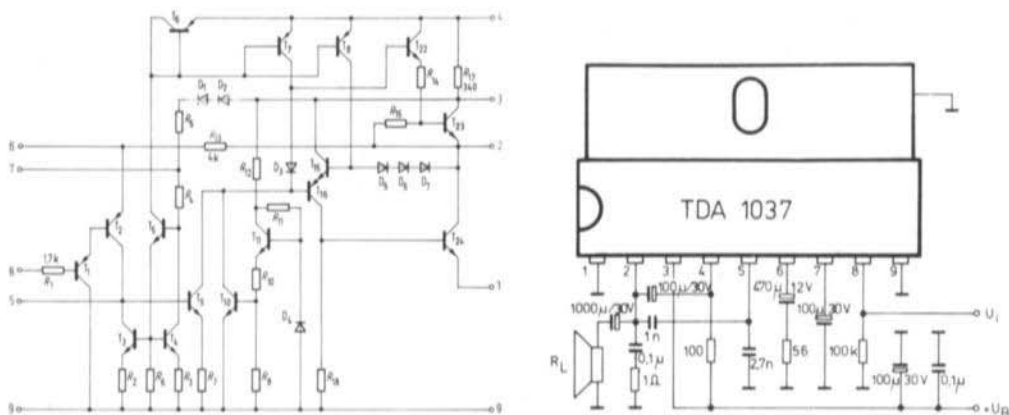


Fig. 1: Internal circuit of the AF amplifier TDA 1037, as well as test and applicational circuit

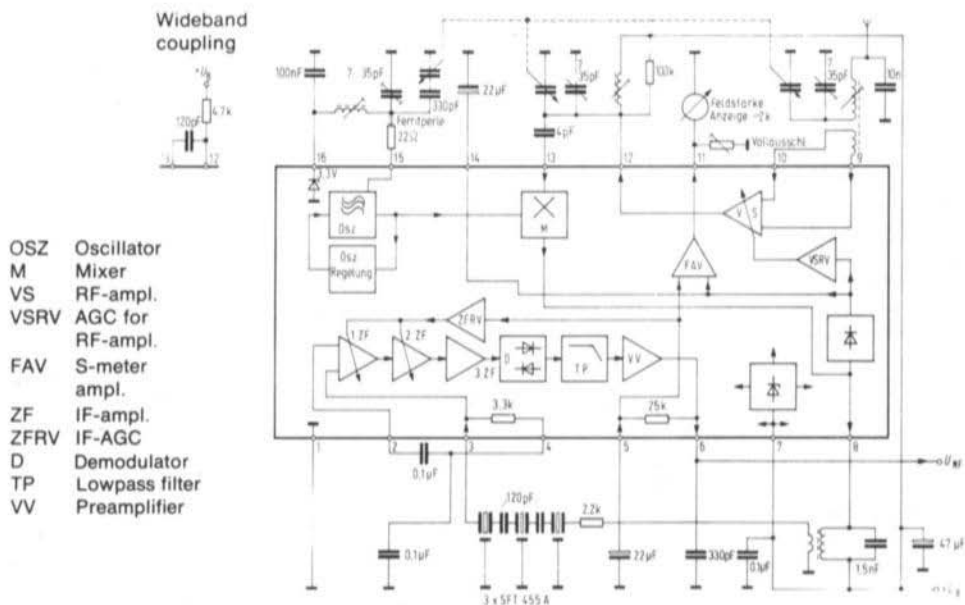


Fig. 2: Block diagram and applicational circuit of the TDA 1046

Further Details:

Since the output of the RF-stage and the input of the mixer are virtually separate, it is possible for a bandpass filter to be used here instead of a single resonant circuit, for instance, to improve the image rejection. In order to use the available capacitance between the two pins, it is recommended that capacitive coupling at the hot end of the bandpass filter circuits is used.

Siemens TDA 1047
FM-IF-amplifier and demodulator

This integrated circuit contains an eight-stage balanced limiter amplifier with balanced coincidence demodulator, signal strength indicator, and squelch. The mono-stereo switching, as well as the AF-muting under detuned conditions, are not of interest for amateur applications but could provide a possibility for automatic frequency control.

Specifications:

Operating voltage range:	4 - 18 V
Frequency range:	0 - 15 MHz
Operating current (typ. at 12 V):	12 mA
Limiting threshold:	30 μ V

Other data given in the data sheet refer to a special application as VHF-FM-IF circuit and are therefore not given here.

The block diagram of this integrated circuit with applicational circuit is given in **Figure 3**.

Stereo-Schaltsschwelle
 = Stereo threshold

AFT-Abschaltzeitkonstante
 = AFC - time constant

AFT-Verstärker
 = AFC-amplifier

Verstimm-Indikator
 = Tuning indicator

NF-Verstärker
 = AF amplifier

Mute-Schalter
 = Mute switch

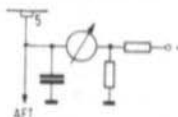
Schmitt-Trigger
 = Schmitt trigger

Phasenumkehr
 = Phase reversal

NF = AF

Tuner = Tuner
 von Abstimmspannung
 = from tuning voltage

Connection of
 center-zero meter



$$k \left(\frac{R}{C} \right) \approx 1 \text{ sec}$$

$k = 10^3$

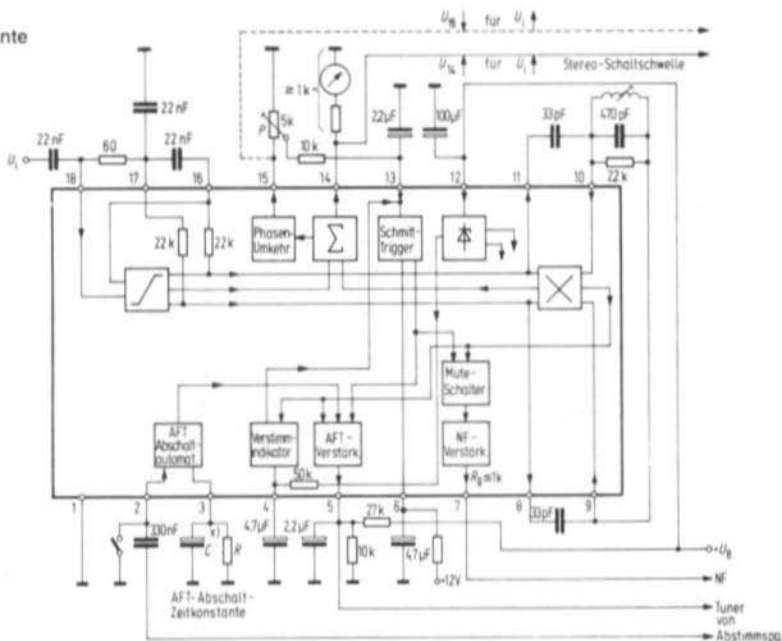


Fig. 3: Block diagram and applicational circuit of the TDA 1047

Siemens TDA 1048

Controlled AM amplifier, demodulator and AF level adjustment

Specifications:

Operating voltage range:	10 - 15 V
Operating current (at 12 V):	37 mA
Operating frequency (nominal 40 MHz):	approx. 0 - 50 MHz
Input voltage for control threshold:	200 μ V
AF-output voltage ($m = 80\%$):	max. 1 V (RMS)
AF volume range:	min. 70 dB

Figure 4 shows a block diagram of this IC together with applicational circuit.

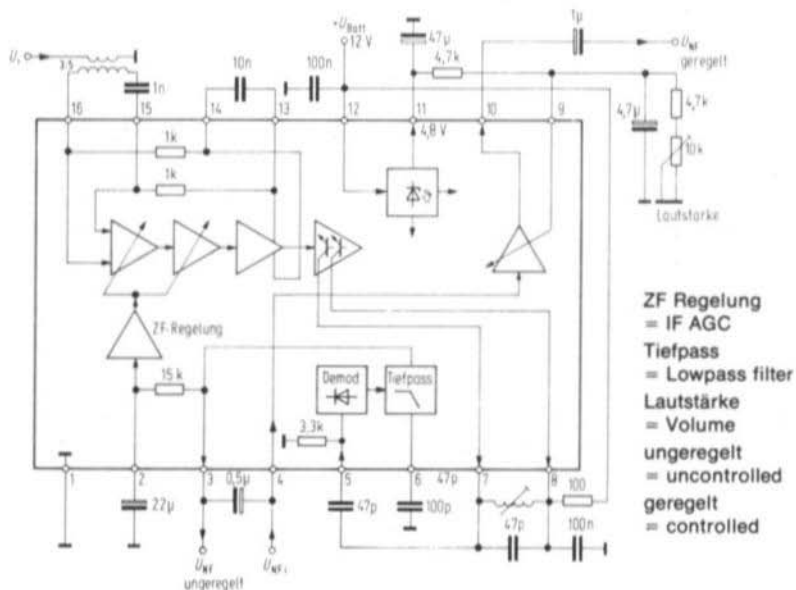


Fig. 4: Block diagram of the TDA 1048 with applicational circuit

Notes

Since the intermediate circuit previous to the demodulator is accessible externally (in contrast to the TDA 1046), it is possible when using this integrated circuit to connect an external SSB-demodulator (product detector) and to obtain the control voltage from the AF output as is often the case in the SSB mode. It may seem a disadvantage at first that the input of the control amplifier accessible via connection 2 is connected internally via 15 k Ω to the AF-output. This means that an external control voltage at connection 2 should be fed in at as low an impedance as possible, in order to override the control voltage coming from the AM-demodulator. It seems, however, possible for the AM-demodulator to be used for control voltage generation using the AF generated in the SSB-demodulator.

Telefunken TDA 1062

VHF tuner for AC and car radios up to 200 MHz

This integrated circuit possesses a preamplifier in a common base circuit, push-pull mixer, oscillator and buffer, IF lowpass filter, and control amplifier for pin-diode RF control.

Specifications:	Operating voltage range:	9 - 15 V
	Operating current (at 10 V):	30 mA
	Noise figure (at 95 MHz):	5.5 dB
	Power gain:	30 dB
	Intermediate frequency (typical):	10.7 MHz

All further specifications given on the data sheet refer to an application as VHF-FM broadcast tuner and cannot be used as they are for amateur applications.

Figure 5 shows a block diagram and applicational circuit for this IC.

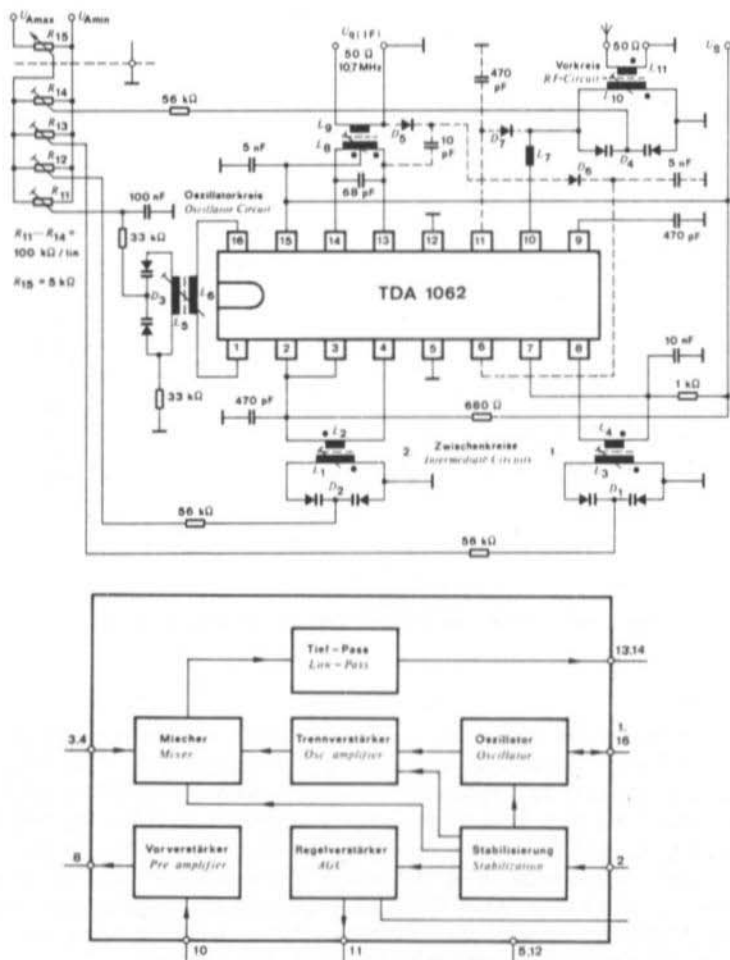


Fig. 5: Block diagram and applicational circuit of the TDA 1062

2. SUGGESTED APPLICATIONS

Several suggestions are to be made how the described integrated circuits can be used for amateur applications. This information is only provided to stimulate interest, and cannot be classed as being proved circuits. The author is sure that far more interesting circuits could be obtained in conjunction with these ICs, and possibly lead to some new descriptions of equipment in this magazine, so that other readers can profit from this.

If the applicational circuit given in the data sheet is used for the IC TDA 1046, and this is then designed for the 80 m band, for example, it would be possible to easily construct a foxhunt (DF) receiver. In this case, the AF output voltage should be sufficient to allow direct connection of a high-impedance earphone ($2 \times 2 \text{ k}\Omega$). Of course, low-impedance earphones will require a matching transformer.

Of course, CW and SSB signals will not be audible, however, a field strength indication (S-meter) is provided. The current drain of approx. 18 mA for the total receiver is relatively low, but it does not seem advisable to operate the receiver with 9 V. A voltage of 12 V or 13.5 V should be better (car battery or 3 series-connected flat batteries).

Since the integrated circuit TDA 1046 can be used for applications up to 30 MHz, it is possible for the receiver to also be designed for 28 to 30 MHz and used as IF for a 2 m converter. Since only a very low intermediate frequency can be processed, a bandpass filter should be provided between the preamplifier and mixer stage in order to improve the image rejection.

Of course, the applications of such a combination are limited since only AM reception is possible. However, an FM module (e.g. TDA 1047) could be connected after the selective circuitry directly in front of the IF input of the TDA 1046 (pin 3) in order to provide FM reception.

It is, of course, possible for a number of AF amplifiers to be connected such as the TDA 1037.

An interesting application for the TDA 1062 would be for it to be used as a PLL oscillator. The provided oscillator stage could, for instance, be used in the frequency range of 135 to 137 MHz (local oscillator for the 2 m band when using an intermediate frequency of 9 MHz), whereas the transistor provided for the RF-stage could be used as crystal oscillator. This means that the signal of the variable oscillator can be converted to a frequency range that can be processed with conventional PLL circuits. The control voltage of the PLL then controls the 135 MHz oscillator with the aid of varactor diodes. This suggestion is shown in **Figure 6** in the form of a block diagram.

Since the TDA 1062 can be used up to 200 MHz, it is possible for it to be used as 2 m tuner. This tuner circuit can be VFO, or crystal-controlled, if a crystal is provided in the oscillator circuit. This could be followed by a TDA 1046 as has been mentioned previously. A TDA 1062 could be used as crystal-controlled converter from 144 to 28 MHz, and a further TDA 1062 used as variable IF, for instance at 9 MHz. This could be followed by a TDA 1048 as controlled IF amplifier, which can be provided with a product detector for SSB and CW as has already been mentioned.

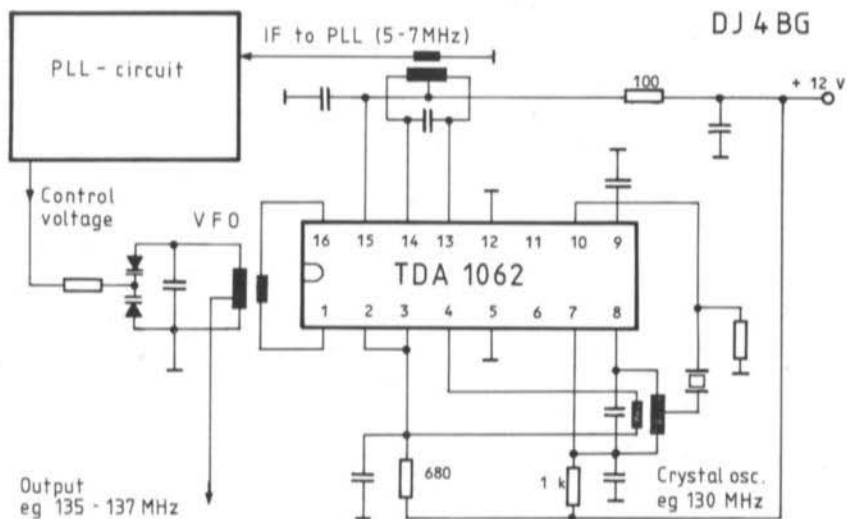


Fig. 6: Circuit of a PLL-oscillator for 135 - 137 MHz

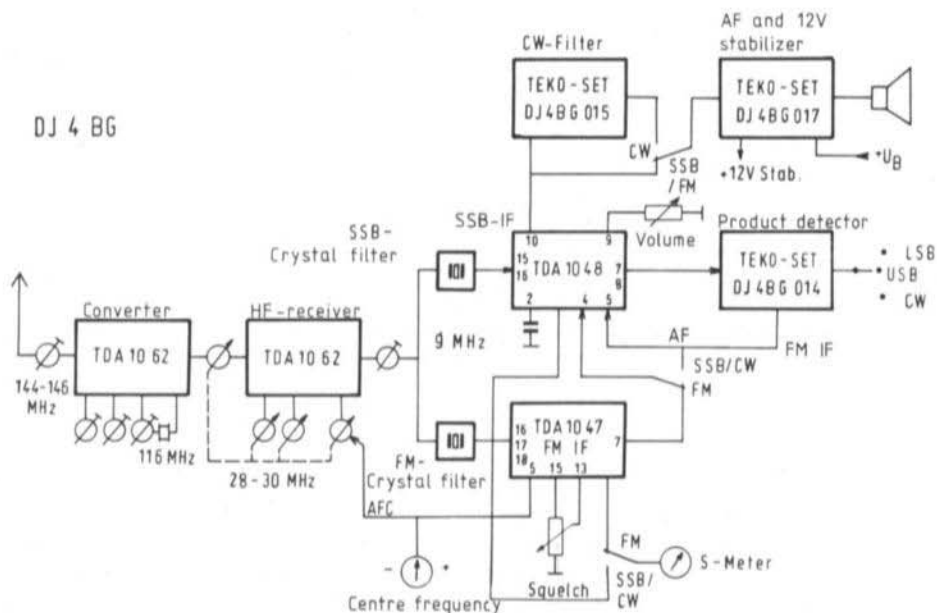


Fig. 7: Suggestion for 2 m receiver for SSB, CW and FM, using the described ICs and some TEKO-modules

If a TDA 1047 is used as FM-module, it should be connected directly at the output of the second mixer in order to ensure the required bandwidth. **Figure 7** shows a combination constructed in this manner together with CW-filter, AF-module and voltage stabilizer as described in the modular receiver system using TEKO boxes. If the sensitivity of approx. 6 dB is not sufficient, a preamplifier stage can be provided, however, this will cause a deterioration of the large-signal behaviour.

Since each manufacturer develops his integrated circuits for special applications, there are a large number of types available on the market. However, this allows integrated circuits of many companies to be combined, which would not be possible when integrated circuits of one single company were to be used.

The author hopes that he has been able to give some stimulation for further development. He has to also be able to give the results of his own developments at a later date. The data and application sheets are available from the manufacturers or their distributors.

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ANTENNA NOTEBOOK

by T. Bittan, G 3 JVQ / DJ 0 BQ

1. ROTATORS

This article is not going to discuss antennas themselves, but moreover their rotating systems. Uptil now, the rotator market has been virtually monopolized by one American company. However, there is now further competition in this field and competition is usually a good thing especially for the customer and user since there is often no incentive to change and redevelopment if only one company is able to monopolize the world market. But this is not the reason for this article but a study of which specifications are important to the user, and what means can be used to improve these characteristics further.

The vertical load is often used as criterion, but is hardly important with exception of with the smaller rotators since amateur antenna arrays are usually far lighter and other factors will limit the size of the array before this vertical load can be obtained.

The braking moment is more important with larger antenna arrays, and is important both for keeping a stationary antenna in the selected position, and also stopping the rotation at the required point. This means that it is important that the rotator has a brake system, not only for the above reasons but also to protect the gearing. It is also important with larger high performance VHF/UHF arrays that this braking can really take place at the angle required and not approximately every 10° or so.

The rotating torque is usually only important at the start of each rotation in order to overcome the mass of the array. Very little torque is required during rotation itself if the array is well balanced both with respect to wind leverage and weight.

The speed and rotation is usually approximately 60 to 70 s for all types. An exact setting of the rotator will be easier if a slow speed is used; however, one may miss several contest stations in this manner. If a faster speed were used, this would lead to a far higher loading of the rotator during the braking process and cause a considerable torque on the mast or tower.

The rotation time of 60 s to 70 s is therefore a good compromise both for rotator and mast.

The mast diameter is in the opinion of the author an important criterion for a rotator. The greater the diameter of the lower mast, the better will be the stability under storm and high wind conditions. There are considerable differences on the market in this respect. For years, only rotators for support masts of up to 2" or approx. 50 - 52 mm diameter were available. Now there are several models allowing mast diameters of up to 62 mm diameter, and even more, both for the lower and rotating mast.

Electrical circuitry: Here there are some differences which affect operation. There are a number of rotators that get extremely hot during continuous operation and often cause the thermo-cutout to actuate. If continuous operation is required, e.g. during contests, this must be taken into consideration.

Although not of importance technically, the noise made by a rotator is of great importance to those amateurs active during the evening and night hours. The author knows a large number of amateurs that are not able to operate during this period for that reason.

The noise made by a rotator can be split into two categories: Firstly the mechanical noise of the gearing and ballbearings. There is quite a difference here between the various types. The second category is the noise made by the braking solenoid. However, this only seems to be a problem with one rotator type.

But now the most important criterion that is often not given, or not considered during selection of a rotator: This is the bending moment, usually the limiting factor of the whole array. However, it is very difficult to get a common system for this, because many manufacturers only give wind surface area of the antenna without giving the length of the mast above the rotator (leverage). Some manufacturers give the calculated breaking point or only a safety factor of say 1.5, whereas other manufacturers use safety factors of 5 or more.

It would be possible for the author to give calculated mechanical values, but this could cause legal consequences.

To summarize, the most important specifications for selection of a rotator are:

- Has it a brake ?
- Does it have sufficient torque for my array ?
- What is its bending moment or wind surface area at what mast length ?
- What maximum mast diameters can be used ?
- Can the electrical circuitry handle continuous operation without thermal overload of the transformer ?

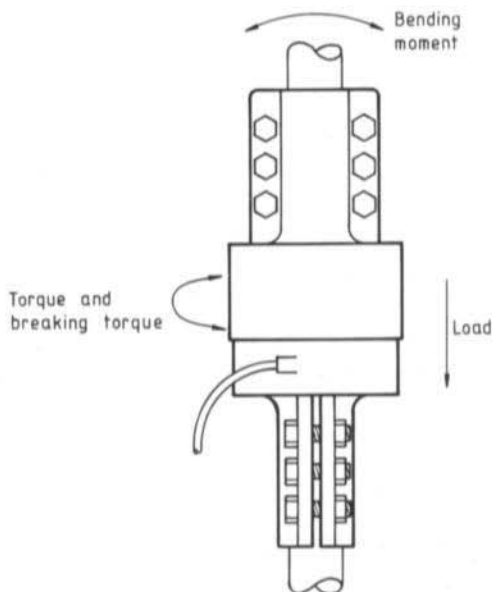


Fig. 1: Forces active on a rotator

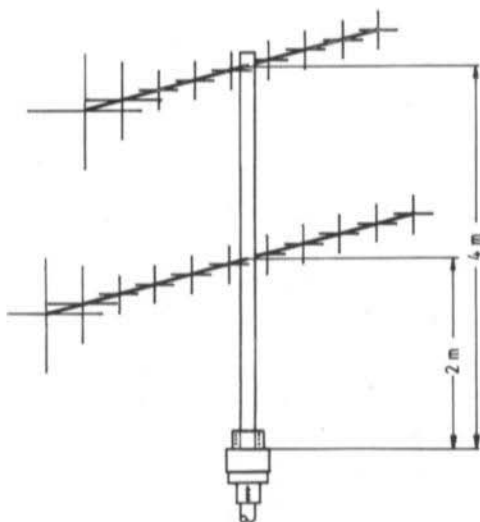


Fig. 2: A typical simple array for 2 m and 70 cm

2. INTEGRATING THE ROTATOR INTO THE ANTENNA SYSTEM

The rotator is usually the most misused and overloaded unit of an amateur radio station. The author has seen so many rotators, designed for rotating a small TV or VHF-FM antenna, used for rotating large amateur arrays. In fact this is not only technically bad practice and dangerous under storm conditions, but also not economical in the long run, since such an overloaded rotator will not have a long life and will have to be replaced far more often than when using a larger rotator. A life factor of five-to-one would not be uncommon.

Let us make a small calculation for a typical antenna system and see what forces are present at the rotator. Assumed is an antenna array comprising a crossed Yagi antenna for 2 m and 70 cm. The 2 m crossed Yagi ($10 \text{ XY} / 2 \text{ m} = 36 \text{ kg} / 160 \text{ km/h}$) is spaced two meters above the rotator and the 70 cm antenna ($12 \text{ XY} / 70 \text{ cm} = 21 \text{ kg} / 160 \text{ km/h}$) at the top of the mast, a total of 4 m above the rotator. Assumed is also a windspeed of 160 km/h (100 mph). The following leverage is provided:

$$\text{Ant 1} = 36 \text{ kg} \times 2 \text{ m} = 72 \text{ kg}$$

$$\text{Ant 2} = 21 \text{ kg} \times 4 \text{ m} = 84 \text{ kg}$$

Total leverage at rotator: $72 + 84 \text{ kg} = 156 \text{ kg}$

It will be seen that this rather modest array already overloads nearly all rotators on the market by more than two times, assuming that the construction given in **Figure 2** is employed. This is improved if a construction, as shown in **Figure 3**, is used in conjunction with a tower. This will convert the bending moment to a sideways pressure in the opposite direction with reduced force according to the leverage.

This horizontal force is not so damaging to a rotator as the bending moment, and if two thrust bearings are used, this horizontal force will be virtually removed from the rotator and transferred directly to the mast structure. The thrust bearings can also support the weight of the antenna array, and it is only the turning and braking torque of the rotator which remains unchanged.



Fig. 3: Mounting of a rotator in a tower

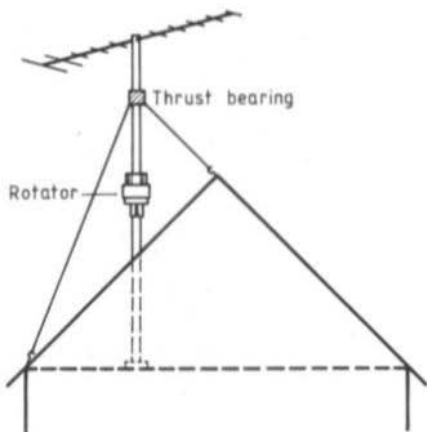


Fig. 4: Reducing the bending moment using a thrust bearing and guys

However, the majority of radio amateurs are probably not able, or are not allowed to use a tower. In this case, how can the antenna system be constructed to ensure that the rotator is not overloaded? It is assumed that only a tubular mast can be mounted either on the chimney side of the house or through the roof.

The easiest method is to use a thrust bearing and guys as shown in **Figure 4**. This will considerably reduce the bending moment on the rotator but not so well as the arrangement given in **Figure 3**. A further improvement will be provided if two thrust bearings and two sets of guys are used.

Another method would be to mount the rotator underneath the roof and use a thrust bearing where the mast passes through the roof (**Figure 5**). The strength of such a system is mainly dependent on the strength of the roof beams. Of course, guys could also be used here.

A better method is shown in **Figure 6**. The rotator is again mounted on the floor but a second guiding outer mast is used together with a thrust bearing. This outer mast is provided with a three or four legged base mount which is bolted to the floor, and securely fixed at roof level. Such an arrangement will accept virtually all the sideward forces and the rotator only has to provide the turning and braking torque.

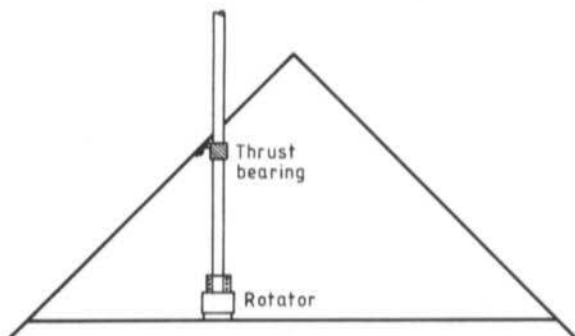


Fig. 5: Another method using a thrust bearing and floor-mounted rotator

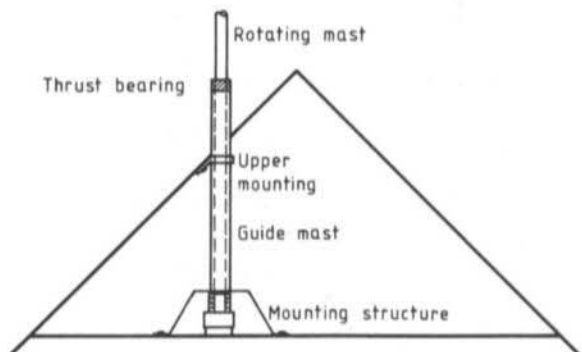


Fig. 6: A combination using a guiding and rotating mast

3. SUGGESTED METHOD FOR ROOF-MOUNTED MASTS

The author prefers a different approach, which does not require guys except for very large arrays and is able to provide a considerable reduction in the rotator loading.

The arrangement shown in **Figure 7** allows a large diameter stationary mast to be used. The diameter of the rotating mast depends on the rotator and the thrust bearing used, but can be upto 60 mm. The latest construction by the author uses a 60 mm diameter reinforced mast as stationary mast and a similar mast of 60 mm diameter for rotation. This system is so successful that manufacture of the platforms for mounting rotator and thrust bearing has been commenced and it is hoped that these items can be marketed at least in Europe. A close-up of such a platform is shown in Figure 6. It is hoped that this method will be able to ease construction of larger antenna arrays that have not been possible upto now, at least not without considerably overloading the rotator.

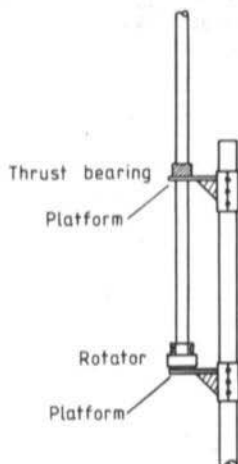


Fig. 7: Suggested system using two side-mounted platforms and thrust bearings

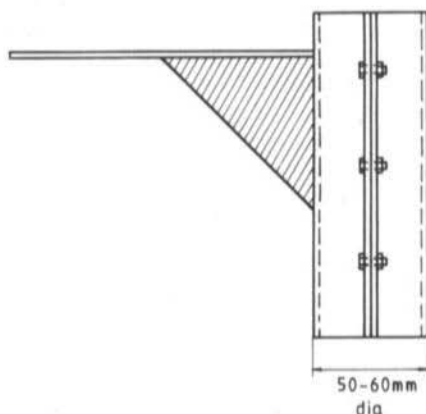


Fig. 8: Close-up of such a rotator platform

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CORNER REFLECTOR ANTENNAS

by R. Lentz, DL 3 WR

Corner reflector antennas comprise two flat reflector surfaces having a certain angle to another, and a dipole element (see Figure 1). The ease of construction and the obtainable gain values of 10 to 14 dB make corner reflector antennas very suitable for home construction.

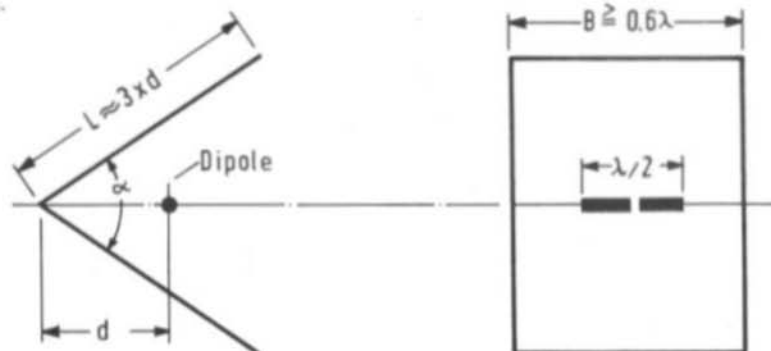


Fig. 1: Design and main dimensions of a corner reflector antenna

Whereas the dimensions of the reflector are very uncritical, the spacing of the dipole (d) determines both the gain and impedance of the antenna. This spacing and the corresponding gain was given in computer-calculated diagrams in (1). These diagrams were not easily readable and had therefore been redrawn and the gain values given in dB over dipole (Figure 2). The maximum gain values obtainable coincide very well with the data given in (2), however, the spacings between dipole and apex of the corner reflector are somewhat different.

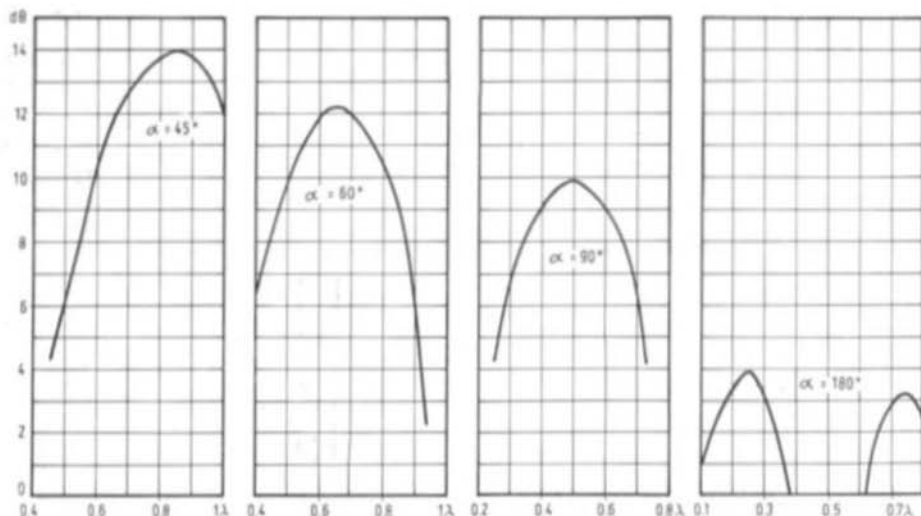


Fig. 2: Gain referred to a dipole as function of the spacing between dipole and reflector apex

As has already been mentioned, also the feed point impedance of the dipole changes, dependent on the spacing to the apex of the corner reflector, and this dependence is different for each angle. Since no values were given in (1), the values given in (2) had to be used. The values of the feedpoint impedance of a dipole (not folded dipole), at the spacing for maximum gain, are given in **Table 1**:

Angle	45°	60°	90°	180°
Spacing d_{opt}	0.85 λ	0.68 λ	0.50 λ	0.25 / 0.75 λ
Max. antenna gain (dipole)	14 dB	12 dB	10 dB	4 dB
Feedpoint impedance	150 Ω	130 Ω	120 Ω	75 / 70 Ω

Table 1: Nominal values for the most important corner reflector antennas

Since the feed point of the dipole is symmetrical, preferably a $\lambda/2$ balun transformer is used, which will provide an impedance of 38 Ω unbalanced in the case of the corner reflector antenna with the highest gain (angle 45°). This value can be transformed up to 60 Ω with the aid of an electrical $\lambda/4$ length of 50 Ω cable which would allow 60 Ω feeder cable to be used. In the case of 50 Ω feeders, it will be necessary to use a $\lambda/4$ transformation piece at an impedance of $Z = 43.3 \Omega$.

Of course, spacings can also be selected which provide a feedpoint impedance of 50 or 60 Ω (corresponding to 200 or 240 Ω in the case of a folded dipole).

However, this will not coincide with the maximum antenna gain as will be seen in Figure 2. The editors would be very grateful if any of our readers were able to measure such an antenna under laboratory conditions, at least at corner reflector angles of 45° and 60°. It would then be possible to clear up the conflicting information of the two references given.

Table 2 gives the dimensions of corner reflector antennas having a reflector angle of 45°. The dimensions have only been given for those bands where reasonable dimensions are possible.

Amateur band	435 MHz	1275 MHz	1296 MHz	2304 MHz
Reflector length L	1200-1700 mm	400-600 mm	400-600 mm	220-330 mm
Reflector width B	min. 420 mm	min. 140 mm	min. 140 mm	min. 80 mm
Dipole spacing	586 mm	200 mm	197 mm	111 mm
Dipole length	320 mm	106 mm	104 mm	58 mm

Table 2: Dimensions of 45° corner reflector antennas for the UHF amateur bands

A shortening factor of 0.9 has been taken into consideration when calculating the dipole length since the dipoles are relatively thick with respect to the wavelength.

REFERENCES

- (1) Proctor D.: Graphs simplify corner reflector antenna design
MICROWAVES, July 1975, Pages 48 and 52
- (2) Rothammel, K.: Antennenbuch
Telekosmos-Verlag Stuttgart 1976, Edition 4, Pages 397 - 400

NOTES and MODIFICATIONS

The following information on the 24 cm converter DJ 5 XA 004 is quite extensive. However, this is not because there are so many errors in the original description, but more because it provides information of general interest with respect to UHF frequency generating. The modification of the frequency plan and use of the stripline transistor BFW 92 are not absolutely necessary but provide a power reserve.

The same is valid for the overrange indication in the case of the DK 1 OF digital multimeter. This modification is an interesting example on the transition between analog and digital technology, but hardly important for the operation of the equipment.

1. 24 cm CONVERTER DJ 5 XA 004

Examination on a spectrum analyzer has shown that the input and output circuit of the last oscillator stage (T 4: BF 479) is not correctly dimensioned. No maximum is observed for the required frequency on rotating trimmer C 4, immaterial what type and capacitance is used. It has been seen that a large number of spurious lines are generated that cause a higher diode current and thus seem to indicate a resonance condition. This also results in too good a measured noise figure; the more spurious reception points, the more noise will be accepted from the wideband noise spectrum. The sensitivity at the required frequency will also be deteriorated. The output circuit L 6 / C 5 can also not be resonated.

The operation of the local oscillator chain can be checked by only installing one mixer diode and measuring its current. The antenna input is terminated with 50 Ω , and the current of D 1, and then of D 2 is measured. If there is a great difference in current (1:20 or more) this will indicate that the oscillator frequency is far from the nominal value (Attention: the current flow in D 1 is opposite to D 2). This could be, for instance, 840 MHz instead of 1260 MHz. The difference in current will be present because the hybrid-ring is not designed for the other frequency. The following modifications ease the alignment to the correct harmonic and improve the spectrum purity of the local oscillator signal.

- 1.1. The BF 223 used for T 3 is replaced by another type with a higher transit frequency, and is then used as a frequency tripler. The subsequent stage (T 4) then doubles to the required frequency. The frequency plan after modification is approximately: 70 - 210 - 630 - 1260 MHz. The transistors BFR 90, BFR 34 A or BFW 92 are suitable, and can be directly soldered to the conductor lanes on the lower side of the PC-board.
- 1.2. Inductances L 4 and L 5 should be as follows: Bend from 0.8 mm silver-plated copper wire as shown in **Figure 1a** and place solder to the original holes in the PC-board. The position of the inductances given by the original holes provides the correct coupling.
- 1.3. Modify the series circuit L 5 / C 4 to a parallel circuit as shown in **Figure 1b**. Disconnect the conductor lane from C 4 to L 5, and connect the hot end of C 4 to the hot end of L 5 with a short piece of metal foil, and the cold end of L 5 to ground.

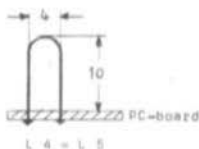


Fig. 1a

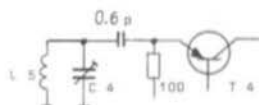


Fig. 1b

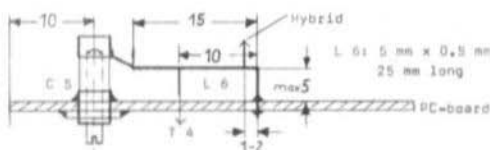


Fig. 2

The emitter connection should only be connected to the 100 Ω resistor and an additional coupling capacitor of 0.3 to 0.6 pF. The other end of this capacitor should be connected to the top of the U-shaped inductance of the L 5 / C 4 circuit.

- 1.4. Modify the resonant line circuit comprising L 6 according to **Figure 2**. Reduce the total length to approx. 25 mm before bending. Saw a slot in the board at the base of L 6, and after installing this inductance solder above and below the board.

A low-loss trimmer (preferably with glass dielectric) should be used for C 5. A hole is drilled 10 mm from the edge of the board for this trimmer, and the surface around the hole carefully tinned to ensure a good ground connection of the trimmer on both sides of the board.

- 1.5. Modern plastic foil trimmers have a large variation range which means that it is possible to tune the frequency multiplier to the wrong harmonic. For those constructors without frequency or wavemeter, it is important that the inductance dimensions are accurate and the following trimmer positions result:

Capacitors C 1 and C 2 are inserted by over half at 210 MHz (at 1/3 insertion, frequency quadrupling will occur). C 3 is approx. 1/3 inserted and C 4 about 1/4.

All resonant circuits and the emitter resistor of T 3 are aligned for maximum diode current of one mixer diode (approx. 1.5 mA is obtainable). The second diode can then be installed and Pt 6 and Pt 7 interconnected with respect to UHF.

A weak 23 cm signal is now received on the converter and C 5 aligned for minimum noise. The emitter resistor of T 3 is adjusted towards higher values until minimum noise is obtained.

- 1.6. After modification, the spurious lines at 630 MHz and 1890 MHz will be at least 26 dB down. The selectivity at these frequencies is only made in the resonant circuit L 5 / C 5. The spurious lines spaced 70 MHz or 210 MHz in the local oscillator spectrum are 46 dB down on the required signal due to the bandfilter coupling.

1.7. For ATV operation with direct conversion to the standard TV-IF of 38.9 MHz, which is recommended when strong TV-Band I signals are present in the area, the following modifications are required:

Crystal frequency: $(1252.5 \text{ MHz} + 38.9 \text{ MHz}) \div 18 = 71.7444 \text{ MHz}$

L 7, L 8: 5/13 turns; L 9, L 10: 13/3 turns

2. IMPROVED CIRCUIT FOR THE OVERRANGE INDICATION OF THE DK 1 OF DIGITAL MULTIMETER

In the previous circuit for the overrange indicator (flasher), the negative input voltage of the analog-digital converter is compared with a positive reference voltage. In this case, the negative voltage (amplified measuring voltage) is connected via a voltage divider to +15 V. The current flow via the resistors for adjustment of the full-range (R 277, R 278, and R 279) causes various different voltage drops that have an effect on the readouts.

These errors must be corrected by the zero-correction. However, since the voltage-divider remains unchanged the overrange indication can start blinking too soon or too late. If the multimeter is to be calibrated, it will be necessary to correct the zero-point after each adjustment of these potentiometers. This makes the alignment both time-consuming and difficult.

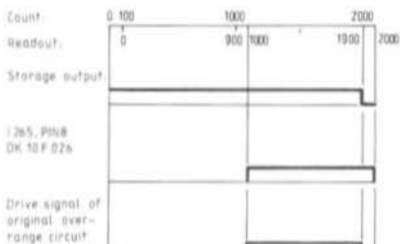


Fig. 1

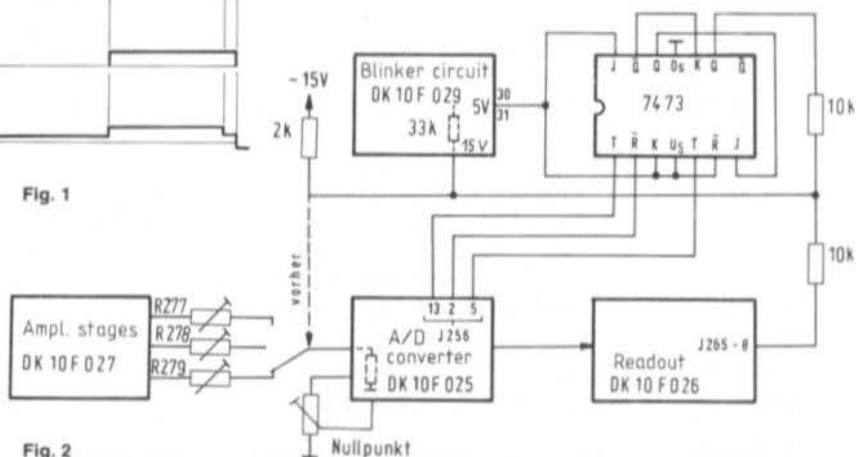


Fig. 2

The circuit given in **Figure 1** operates digitally and does not exhibit these disadvantages. A divide-by-two and an SN 7473 storage are added and connected to the output of the analog-digital converter. Low-level is present at the output of this storage as soon as the counter indicates more than 1999. Since the readout 0100 is valid for the zero-point, 1900 will be indicated instead of 2000. As can be seen from the pulse plan in **Figure 2**, the level at the output of the storage is not a suitable criterion on its own. The readout of the one at the

thousands digit must also be evaluated. The range is first exceeded when it is not driven, and a low level is present at the output of the storage at the same time. This means that zero must also be present at connection 8 of I 265 (DK 1 OF 026).

The original overrange circuit can be driven from these two outputs so that - 10 V will not be exceeded at their inputs when low level is present.

The additional IC SN 7473 and three resistors can easily be accommodated on a Vero-board, and connected to PC-board DK 1 OF 025. Attention should be paid when aligning potentiometer R 291 of the original overrange circuit that the overrange warning actuates after exceeding a readout of 1999, and not at 1900.

It would also be possible for a lamp to be driven via a PNP transistor. However, a blinking indication would then not be possible.

3. PROGRAMMABLE FOX-HUNT RECEIVER DL 9 FX 006/007

Since some tendency to self-oscillation was noticed by a number of constructors, the author has experimented to determine improvements. These are available from the publishers, and can be brought in VHF COMMUNICATIONS if there is sufficient interest.

MATERIAL PRICE LIST OF EQUIPMENT described in Edition 1/1977 of VHF COMMUNICATIONS

DJ 4 LB 006	TWO-STAGE LINEAR AMPLIFIER for 70 cm ATV (1.5 W)	Ed. 1/1977
PC-board	DJ 4 LB 006 (with printed plan)	DM 12.—
Semiconductors	DJ 4 LB 006 (3 transistors, 4 diodes)	DM 75.—
Minikit	DJ 4 LB 006 (4 tantalium caps., 5 trimmers, 2 chip caps., 5 ceramic caps., 1 cooling fins, 1 ferrite bead, 3 ferrite chokes, 2 BNC sockets, 1 TEKO box, 2 trimmer potentiometer)	DM 43.—
Kit	DJ 4 LB 006 with above parts	DM 128.—
TOTAL KIT	DJ 4 LB 001 - 006 - Price only:	DM 520.—
PC boards	DJ 4 LB 001 - 006 - Price only:	DM 60.—
DJ 6 PI 004	VESTIGIAL SIDEBAND FILTER FOR ATV TRANSMITTERS	Ed. 1/1977
PC-board	DJ 6 PI 004 (with printed plan)	DM 11.—
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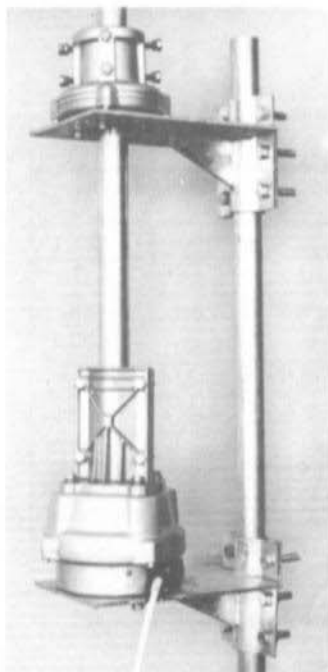
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