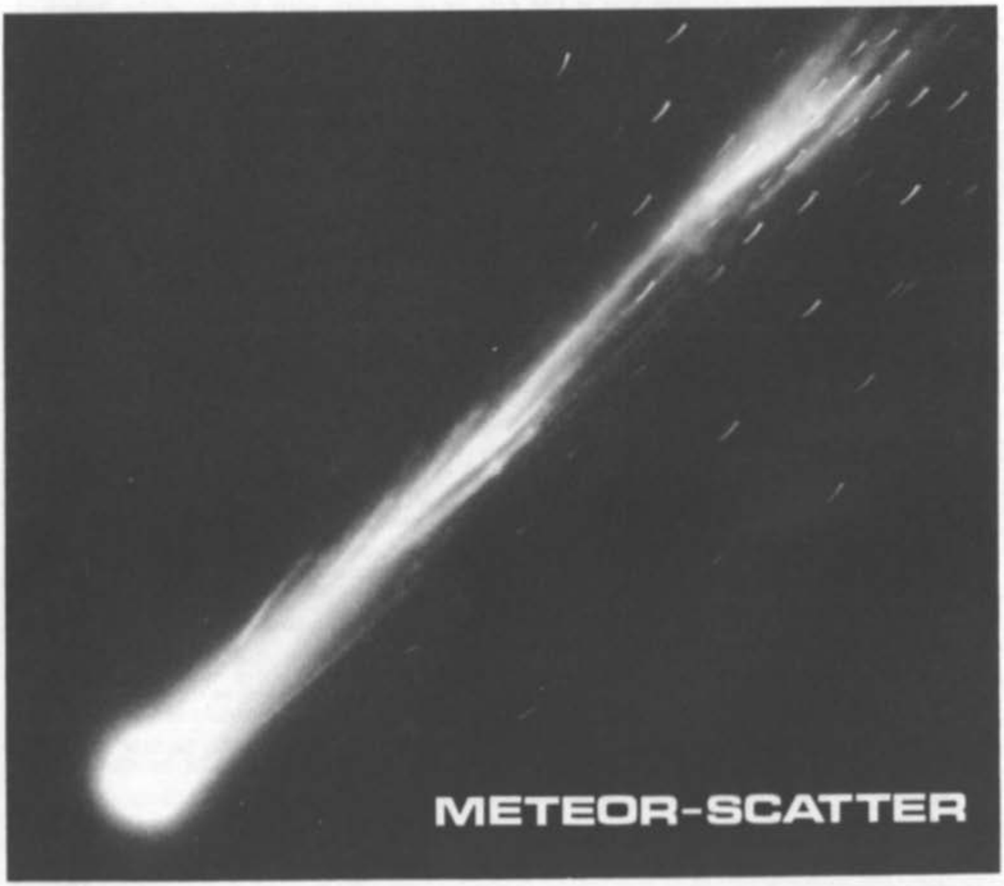
J 20 419 F

VHF COMMUNICATIONS

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

VOLUME NO. 6 WINTER EDITION 4/1974 DM 4.00



VHF COMMUNICATIONS

Published by:
Verlag UKW-BERICHTE, Hans J. Dohlus oHG, 8521 Rathsberg/Erlangen, Zum Aussichtsturm 17
Fed. Rep. of Germany. Tel. (0 91 91) 91 57, (0 91 35) 4 07, (0 91 31) 2 28 80

Publishers:
T. Bittan, H. Dohlus, R. Lentz at equal parts

Editors:
Terry D. Bittan, G3JVQ/DJ0BQ, responsible for the text and layout
Robert E. Lentz, DL3WR, responsible for the technical contents

Advertising manager:
T. Bittan, Tel. (0 91 91) 91 57

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 14,00 or national equivalent per year. Individual copies are available at DM 4.00, 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.

Printed in the Fed. Rep. of Germany by R. Reichenbach KG, 85 Nuernberg, Krelingstraße 39

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

VERTRETUNGEN: **REPRESENTATIVES:**

Austria	Hans J. Dohlus, Creditanstalt Bankverein WIEN Kto. 17-90.599; PSchKto. WIEN 1.169.146
Australia	WIA PO Box 150, TOORAK, VIC. 3142, Tel. 24-8652
Belgium	see Germany, PSchKto. 304 55-858 Nürnberg
Canada W. Prov.	Otto Meginbir, VE 6 OH, 1170 Bassett Cres. NW. MEDICINE HAT, Alb., Tel. 403-527 6349
Denmark	Sven Jacobson, SM 7 DTT, Gamlakommungården 68, S-23501 VELLINGE, Tel. Malmö 420430 Postgiro: Köpenhamn 149 85
France	Christiane Michel, F 5 SM, F-89 PARLY, les Pillés
Finland	Eero Valio, OH 2 NX, 04740 SÄLINKÄÄ, Postgiro 4363 39-0 und Telefon 915/86 265
Germany	Verlag UKW-BERICHTE H. Dohlus oHG, D-8521 RATHSBERG/Erlangen, Zum Aussichtsturm 17 Tel. 0 91 35-4 07, 0 91 31-2 28 80, 0 91 91-91 57, Konten PSchKto. 304 55-858 Nbg. Commerzbank Erlangen 820/1154, Deutsche Bank Erlangen 76/403 60
Holland	see Germany, PSchKto. 304 55-858 Nürnberg
Italy	STE s.r.l. (I 2 GM) via maniago 15, I-20134 MILANO, Tel. 21 78 91, Conto Corrente Postale 3/44968
Luxembourg	P. Wantz, LX 1 CW, Télévision, DUDELANGE, Postscheckkonto 170 05
New Zealand	E. M. Zimmermann, ZL 1 AGQ, P.O. Box 56, WELLSFORD, Tel. 80 24
Norway	H. Theg, LA 4 YG, Ph. Pedersensv. 15, N 1324 LYSAKER pr. Oslo, Postgiro 31 6000
South Africa	Arthur Hemsley ZS 5 D, P.O. Box 64, POINT, Durban Tel. 31 27 27
Spain+Portugal	Julio A. Prieto Alonso, EA 4 CJ, MADRID-15, Donoso Cortès 58 5°-B, Tel. 243.83.84
Sweden	Sven Jacobson, SM 7 DTT, Gamlakommungården 68, S-23501 VELLINGE, Tel. 040-420430 Postgiro: 43 09 65-4
Switzerland	Hans J. Dohlus, Schweiz. Kreditanstalt ZÜRICH, Kto. 469.253-41; PSchKto. ZÜRICH 80-54.849
USA	VHF COMMUNICATIONS Russ Pillsbury, K 2 TXB, & Gary Anderson, W 2 UCZ, 915 North Main St. JAMESTOWN, NY 14701, Tel. 716-664-6345

A PUBLICATION FOR THE RADIO AMATEUR
ESPECIALLY COVERING VHF, UHF AND MICROWAVES
VOLUME NO. 6 WINTER EDITION 4/1974

As you know prices have increased terribly in 1974. With the exception of the crystals, we have kept all prices stable throughout the year. The new postal rates for printed matter valid from January 1975, and the increase in the cost of paper and envelopes etc., have forced us to increase the price of the 1975 subscription to DM 16.--. The subscription price has been DM 14.-- for three years now.

We hope that you have enjoyed reading VHF COMMUNICATIONS this year and wish to renew your subscription to include 1975. We assure you that we will be doing our best to offer you the finest constructional designs and theoretical articles. If you have suitable designs that would be of interest to other readers, please do not hesitate to write to us giving us full details. We have some really outstanding articles waiting to be published.

We would like to take this opportunity of wishing all readers a happy Christmas season and prosperous New Year 1975.

METEOR SCATTER: THEORY AND PRACTICE

by T. Damboldt, DJ 5 DT

1. INTRODUCTION

Meteorites glow on entering the Earth's atmosphere. Usually, such meteorites burn up completely before reaching the ground. However, some are large enough that they hit the surface of the Earth before they vapourize completely. Meteorite observers usually note the position of such visual meteorites on celestial maps and list location and time of the observation, as well as duration and brightness. These observations are gathered on a worldwide basis by the British Astonomical Association, Burlington House, LONDON W1V ØNL, England, who evaluate and summarize them. Such summaries are to be mentioned later in this article.

2. THE RADIANT

It will be noticed when entering several meteors onto a celestial map that the extension of the visible tracks lead to a common point in space. This point is called the radiant. The radiant is therefore the central source of a meteor shower. It does seem that the meteors are scattered in space, but this is only due to the perspective of the observer on the Earth.

3. METEOR ACTIVITY

If the meteor activity is observed as a function of the time of day, it will be seen that the greatest period of activity is around sunrise. The cause of this that the Earth is orbiting the sun and the direction of the Earth's path is 90° from the direction of the sun (Midday). This position is also the point on the Earth's surface where sunrise is occuring.

It is this position of the Earth that gathers the most of the interplanetary particles (meteorites), since it is only meteorites that are travelling faster through space than the Earth, which are able to enter the atmosphere on the sunset (evening) side of the Earth.

Meteor activity also varies from day to day because the density of particles is not homogeneous in interplanetary space through which the Earth passes on its orbit around the sun. This results in a seasonal variation of meteor activity which has a maximum in July and a minimum in February (see Fig. 2).

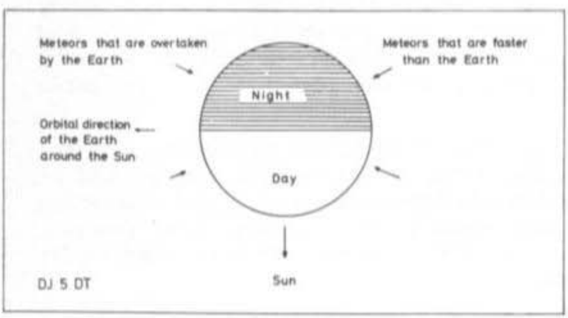


Fig. 1 : Orbit of the Earth around the Sun: The majority of the meteorites enter the atmosphere at sunrise

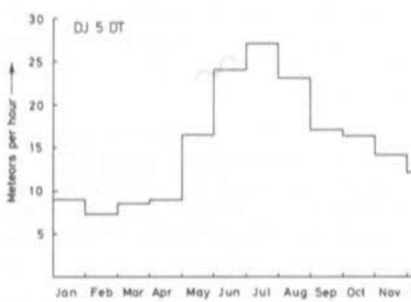


Fig. 2 : Number of sporadic meteors per year

4. SPORADIC METEORS AND METEOR SHOWERS

If meteor activity is recorded over a period of several years, it will be noticed that these vary from day to day, as previously mentioned, and also that an extraordinarily high activity is present during several days each year. On observing the times of maximum meteor activity on celestial maps, it will be noticed that a large number of meteors originate from the same radiant. These periods of high meteor activity are called meteor showers, and the individual showers are named after the constellation in which their radiant lies: e.g. Orionids from the Orion constellation, Lyrids from Lyre etc. At least some of the meteor showers are the remains of one time comets. In this case the appearance of a meteor shower will only take place when the Earth's orbit crosses the track of such a comet. Of course, it is possible that such comet particles have distributed themselves along the whole path. In this case, the meteor activity will be low but possesses a longer duration of upto several days such as is the case with the Taurids in November. If the particles from the comet are still relatively near to the original comet, the meteor activity can be very great but of short duration, as with the Perseids in August. Extremely active showers can be observed when the Earth passes or is passed by comets. An example of this are the Drakonids in October 1946 in which a shower of several thousand meteors lasted approximately 30 minutes. In the past, various of the meteor showers have been extremely active, e.g. the Leonids in November, 1967. Unfortunately, no forecast can be made for the future. The following is to mainly deal with meteor showers rather than sporadic meteorites, since it is the showers that are of interest to radio amateurs. The number of sporadic meteors observed is approximately 10 per hour whereas a value of 100 per hour can be observed during a meteor shower.

A table with data for the major meteor showers are given in many of the astronomical almanacs. Attention must, however, be paid that the Earth does not pass through the same positions in interplanetary space at an identical calendar time on each annual orbit around the sun. This is due to the fact that our year varies in length (leap-year) and that the orbital velocity of the Earth is not constant. For this reason such astronomical almanacs usually only list the times of maximum activity for each shower (Table 1).

The time of maximum activity, the period and activity of the meteor shower are very important in arranging meteor scatter schedules. The British Astronomical Association has determined the activity distribution as a function of time for the various meteor showers. Curves are given in Figures 3 and 4 for the two largest showers: The Quadrantids on January 4th and the Perseids around August 12th. It can be seen that the Quadrantids shower only has a duration of nine hours, whereas the Perseids shower lasts three to four days. This means that it is useless to arrange schedules for Quadrantids shower on January 2nd or 5th.

5. PATH OF THE RADIANTS IN SPACE

Another basic fact in astronomy must be mentioned before practical meteor scatter work can be discussed: Since the Earth rotates around its own axis approximately every 24 hours, all celestial points (stars etc.) possess a circular movement around the celestial pole (around the North Star for the northern hemisphere). Depending on the spacing of these points from the pole, all points (stars, meteor radiants, Sun etc.) will make a smaller or larger

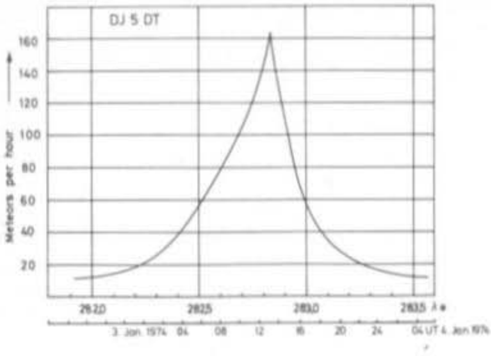


Fig. 3 : Activity of the Perseids shower

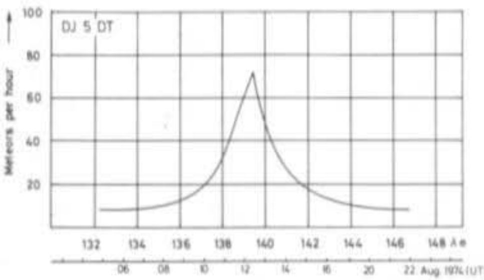


Fig. 4 : Activity of the Quadrantids shower

circle around the pole star. If the spacing is small, a completely visible circle will be drawn. If the spacing is greater than the geographical latitude of the observer (Germany approximately 50° N), the star, radiant, Sun etc. will drop below the northern horizon. If this is the case with a meteor shower, this will mean that no visible meteors will be seen in the atmosphere, and therefore no meteor scatter will be possible from the observer's location. It is therefore necessary to know the celestial position of the radiants in order not to attempt observation of meteors (or MS-schedules) at unfavourable times (when they are below the horizon).

Table 1: Maximum activity of several Meteor showers in 1974

Shower	Maximum activity	usable period
Quadrantids	Jan. 3rd. 13.00	9 hours
Lyrids	Apr. 21st. 22.00	2 days
η Aquarids	May 5th.	5 days
Arietids	June 7th.	8 days
ζ Perseids	June 9th.	8 days
June Lyrids	June 16th. 00.00	2 days
δ Aquarids	July 28th.	2 days
Perseids	August 12th. 10.00	4 days
Drakonids	Oct. 9th.	1 hour
Orionids	Oct. 21st.	2 days
Taurids	Nov. 8th.	20 days
Leonids	Nov. 17th. 11.00	3 hours
Geminids	Dec. 14th. 07.00	3 days
Ursids	Dec. 22nd.	12 hours

In the astronomical tables, the coordinates of the radiants are given as ascension and declination values for each of the meteor showers. With the aid of spherical trigonometry and the sidereal time, it is possible to calculate the horizontal coordinates with the aid of a few formulas. These coordinates are then given as elevation above the horizon (0^o to 90^o) and direction (0^o to 360^o). However, these horizontal coordinates are only valid for a certain geographic latitude. Figure 5 shows the path of the radiants in the sky for the Quadrantids shower (January 3rd or 4th). The numbers given on the curve indicate the time at which the radiant is at the given direction. It can be seen that the radiants do not fall below the horizon at a latitude of 50^o N. Other showers such as the Geminids do go below the horizon. The Geminids shower rises (as Sun and Stars) in the East (at approx. 17.00) and sets in the West (at approx. 11.00). All times given for the meteor showers except for the time of maximum activity are to be given in local time. They are valid for all locations having a latitude of 50^o N.

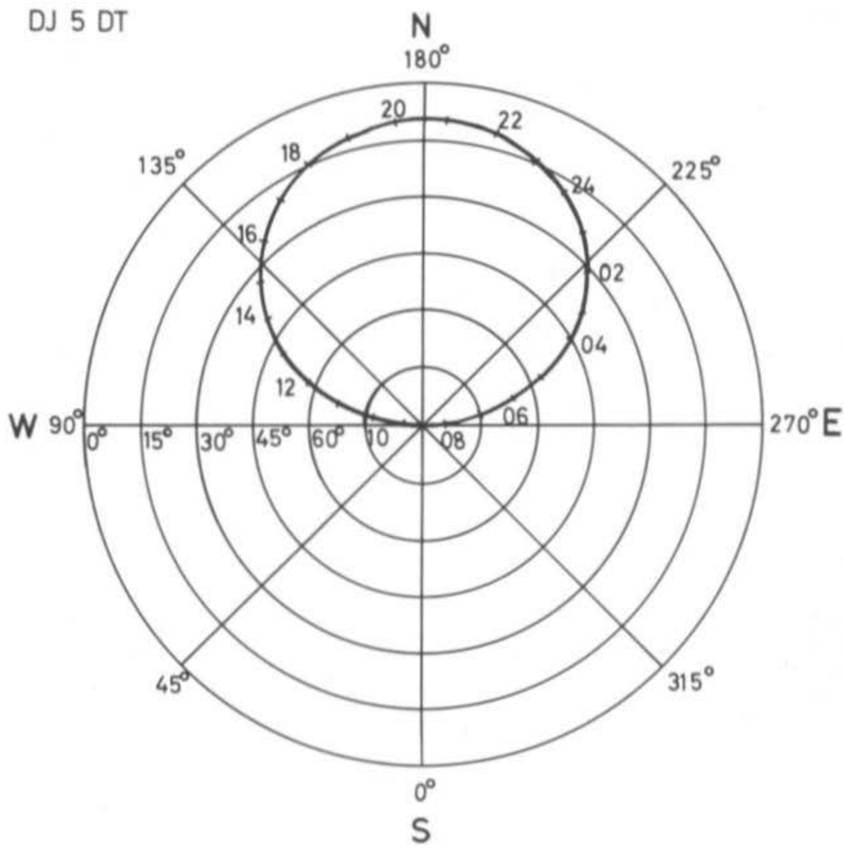


Fig. 5 : Path of the radiants in the sky (Quadrantids)

6. REFLECTION CHARACTERISTICS OF METEOR TRAILS FOR RADIO WAVES

Readers will now have sufficient knowledge of astronomy in order to understand the theory of the reflection of radio waves on meteor trails. Radio waves will be reflected by electrically conductive matter. An electrically conductive condition exists when free charge carriers (electrons or ions) are present in the matter. When meteorites enter the atmosphere they do not only leave a visible trail but the vapourizing particles also leave an ionized trail. This trail is soon diffused in the atmosphere but is able to reflect radio waves as long as the ionization is sufficiently great. This is valid as long as conditions for the reflection are given. In the case of a radar station where transmit and receive locations are identical, the radio wave must hit the trail at right angles (Back scatter). In the case of forward scatter where transmitter and receiver are located at a considerable distance from another, the reflection condition is given when the trail is tangential to an ellipsoid focused both on the transmit and receive station.

7. THE DIFFERENCE BETWEEN PINGS AND BURSTS

When studying the physics of the reflection of radio waves on meteor trails, one must differentiate between underdense and overdense trails. In the case of underdense trails, the electron density is very low so that the radio wave passes through the trail and is only scattered by the individual electrons. The received energy is the sum of all discrete reflections after combining the individual directions and phase angles. Since the diffusion of the trail takes place relatively quickly, the phase differences increase and the discrete reflection of the electrons and the amount of receive energy is rapidly reduced. The signals received in this manner are of extremely short duration (fraction of a second) and are designated pings. The duration of such pings is proportional to the square of the wavelength. This means that the pings will have 25 times the duration on the 10 m band than on the 2 m band. The duration on the 70 cm band is only one tenth of that of 2 m. The received energy is proportional to the third power of the wavelength. This means that the received energy is 120 times greater on 10 m than on the 2 m band, and 30 times weaker on 70 cm. The electron density can amount to 10^{14} electrons per metre with underdense trails.

If the electron density is greater, the radio wave will not be able to penetrate into the ionized matter, but will be totally reflected on the surface. In this case one speaks of overdense trails. The reflected energy lasts far longer and the reflected energy is called a burst. The duration is once again proportional to the square of the wavelength, and the intensity is directly proportional to the wavelength. The same trail that provides a 10 dB signal over noise for a burst duration of 10 s at a wavelength of 2 m, will produce a 14 dB signal with a duration of 250 s at 10 m and only 6 dB for 1 s at 70 cm. The ionization of overdense trails often remains until it is diffused by the very strong winds of the upper atmosphere. The echos from the various parts of the trails then interfere with another and cause considerable fading.

The ionization caused by meteor trails occurs at an altitude of 80 km to 120 km. The mean altitude is approximately 95 km. This means that the normal maximum range for meteor scatter communications is in the order of 2000 to 2200 km. This distance is greater than is possible with Aurora communication, and can be compared to E-layer propagation since the reflection takes place in the E-layer.

8. INCIDENCE OF THE TRAIL TO THE PATH ANGLE

A favourable incidence of the meteor trail to the radio wave between transmitter and receiver is very important. As was mentioned previously, the trail must provide a tangent on an ellipsoid in whose focus both the transmit and receive stations must be found. In practice, virtually every trail offers some sort of ellipsoid that fulfils this demand. However, there are a number of limitations so that only a small part of the meteor trail really reflects a signal to the receive location. As one can imagine, a meteorite entering the atmosphere horizontally covers a long path in the atmosphere. Since it only has limited amount of energy it is not possible for an infinite number of electrons and ions to be generated along this path. This means that the electron density is very low.

A meteorite entering the atmosphere vertically possesses a far shorter path and generates a higher electron density. There are very few suitable ellipsoids from vertical meteorites to fulfil the reflection conditions. Furthermore, these ellipsoids are located far from the centre of the communications path. Of course, if antennas are used that have a narrow beamwidth in the horizontal plane that are directly beamed to the partner station, such off-beam reflections will not be useable. Many meteor scatter stations therefore use stacked arrays having a greater horizontal beamwidth (The editors are of the opinion that circular-polarized antennas should provide the best polarisation for MS-propagation and would like to hear from MS-stations using this polarisation mode).

The most favourable height h of a radiant is an angle of 45^{o} since the effectivity is proportional to sine h and cosine h. As can be shown theoretically, it is also more favourable when the angle of the radiant is perpendicular to the transmission path. The effectivity is proportional to sine (p - a) where "p" is the azimuth of the path and "a" is the azimuth of the radiant. The azimuth is measured from South via West, North, East from 0^{o} to 360^{o}. The most favourable path for each shower is when it is perpendicular to the radiant when the latter possesses a height of 45^{o}. All other times or directions are at a disadvantage. The effectivity of a meteor shower as a function of the azimuth angle can be given as:

$$\text{Eff.} = \text{sine } h \times \text{cosine } h \times \text{sine } (p - a)$$

The effectivity can now be calculated for each hour of the day as long as the height and azimuth of the meteor shower are known for each hour. This allows the most favourable time for MS-communication to be calculated for a given direction. Due to the selection of a sine or cosine function, a value of maximum 0.5 can be calculated. Also this value will only be achieved when the radiant has a height of 45^{o}, and when the path angle is directly perpendicular to the azimuth. Generally, a higher efficiency is obtained in the required direction at greater values of effectivity, than with lower values. Figure 6 gives an example of the effectivity as a function of the time of day. The Fortran program was provided by DK 1 KW and kindly offered to the author. The effectivity has been determined for four path angles (North-South, East-West, NW-SE and NE-SW and for the 14 large meteor showers).

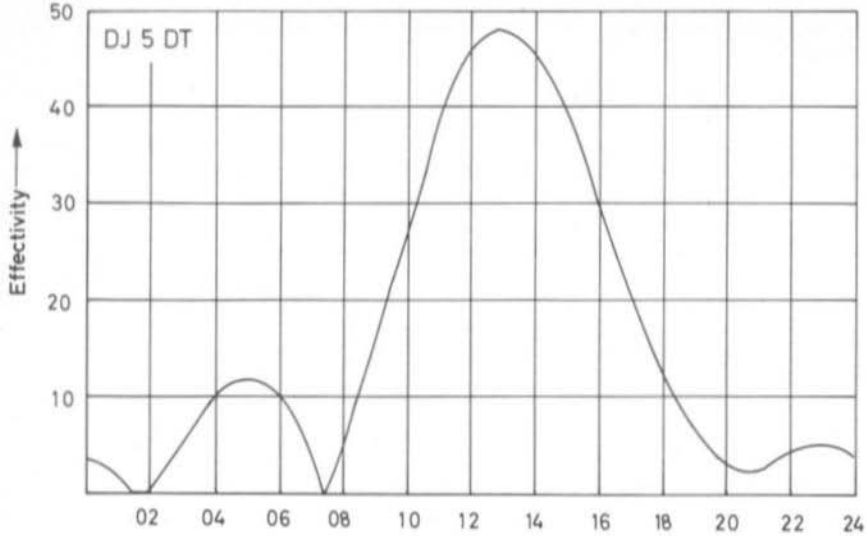


Fig. 6 : Effectivity of the Quadrantids shower for a path angle of 45° (SW-NE)

A table can be prepared from the total of 40 curves which allows one to establish the most favourable time for each direction during each of the showers (Table 2). The first column gives the date of the shower, the second the name of the shower, the third the rise and setting time or whether it is a pole-circular shower; the fourth gives the number of pings or bursts per hour (as a relative number approximating an ERP of 200 W at 2 m), the fifth the duration of the shower (take the time of max. activity for short showers from an astronomical almanac). The sixth column finally gives the most favourable time for the various directions.

The following note is necessary at this point: Very long bursts of over 10 s are produced by large meteorites of more than 1 gramme in weight whose ionized trails are able to last for a certain period. As was mentioned in Section 7, very strong winds are present at an altitude of 100 km that soon diffuse the trail. This means that the relationship between the direction of the path angle with respect to the trail is soon lost. This means in practice that Table 2 is no longer valid for long bursts.

It is important to remain near to the maximum activity of the shower. If the information given in Table 2 is followed, this will considerably increase the number of pings and short bursts.

Table 2

Date	Shower	Rise-Set time	Activity Echos/h	Usable period (days)	Most favourable direction and time (local) SW-NE	E-W	SE-NW	N-S
Jan. 3rd	Quadrantids	C	100	9 hours	1030-1600	(0000-0430, 1230-1730)	0000-0600	0100-0630 1030-1530
Apr. 21st	Lyrids	19/13	15	2	0000-0230 0700-1000	(0300-0500)	0430-0830 2130-2400	0800-1030 2130-0230
May 4th	η Aquarids	01/14	20	5	0330-0730	0500-1000	0730-1100	(0300-0600, 0900-1200)
June 7th	Arietids	02/18	60	8	0600-0930 1330-1500	0830-1130	1000-1430	0430-0800 1200-1530
June 9th	ζ Perseids	03/19	40	8	0630-1030 1430-1600	0930-1230	0530-0630 1100-1530	0530-0900 1300-1630
June 16th	June Lyrids	19/13	10	2	0000-0230 0700-1000	(0300-0500)	0430-0830 2130-2400	0600-1030 2130-0230
July 29th	δ Aquarids	20/08	15	2	2200-0230	2330-0430	0200-0530	(0400-0630, 2200-0030)
August 12th	Perseids	C	60	4	0700-1400	1000-1500 2000-0100	1900-0300	0800-1230 2230-0300
Oct. 9th	Drakonids	C	10	1 hour	1800-2400	0830-1030	0600-1330	0900-1400 1830-2300
Oct. 21st	Orionids	21/12	20	2	0000-0400	0200-0630	0400-0900	0600-0930 2330-0230
Nov. 9th	Taurids	17/08	10	20	2000-2400	2200-0300	0000-0500	0230-0530 1930-2230
Nov. 17th	Leonids	22/14	10	3 hours	0200-0530	0500-0800	0630-1100	0030-0430 0800-1200
Dec. 14th	Geminids	17/11	60	3	0500-0800	(0030-0330)	0300-0600 1930-2230	0400-0800 1930-2400
Dec. 22nd	Ursids	C	15	12 hours	0800-2000	0000-2400	1900-0700	nil.

9. PRACTICAL METEOR-SCATTER OPERATION

A special type of communication is necessary to transfer information by meteor scatter due to the short period of reflection offered by the pings and bursts. In principle both single sideband (SSB) and morse (CW) transmissions can be used for meteor scatter work. However, CW has a great advantage over SSB in that small frequency deviations from the nominal frequency are not so important. SSB would have the advantage of a faster flow of information. Since the duration of the pings and bursts only last either a fraction of a second or several seconds, this means that the information to be transmitted must be kept to a minimum. For this reason, a different method of reporting signal characteristics is used by indicating the length of the burst and the signal strength. The length of the burst is given as follows:

1 Only pings (seldom used)
2 Bursts of upto 5 s
3 Bursts of 5 s to 15 s
4 Bursts of 15 s to 30 s
5 Bursts of over 30 s

The actual signal strength is given in S-units. This means that the signal report comprises two figures. In Europe, it is usually arranged that one of the stations only sends both callsigns for a period of 5 minutes (example: UR 2 BU DJ 5 DT UR 2 BU DJ 5 DT etc). This is followed by the second station also transmitting both callsigns for a further period of 5 minutes. After both callsigns have been heard during one or more bursts, the callsigns are given again together with the report (UR 2 BU DJ 5 DT 27 27 27 UR 2 BU DJ 5 DT etc.). If one of the stations has received both callsigns and report, he will know, that the partner has already received the callsigns and transmits only the report together with the confirmation code "rr" (rr 27 rr 27 rr 27 etc.). If the second station then receives, for instance, only ... 27 ... but has already received both callsigns he must once again send the signal report together with the confirmation "rr" since he has not received the confirmation "rr" from the other station and does not know whether the first station has received his report. As soon as both stations have received callsigns, report and confirmation "rr", they confirm completion of the MS-QSO by transmitting "rrr rrr rrr".

The 5 minute cycle is relatively standard practice in Europe although shorter intervals of say 1 minute or 30 seconds would often be more favourable. In some cases it would be possible to complete the whole MS-QSO within this period of 5 minutes. The following method is used in the USA: Station A transmits for a period of 5 minutes using a cycle of 5 s transmit and 2 s receive. After this period of 5 minutes, station B also transmits for a period of 5 minutes using the same cycle of 5 s transmit and 2 s receive. This method allows a complete MS-QSO to be completed during a long burst, since the receiving station can answer immediately a suitable burst appears. This method is not popular in Europe since it is not too suitable for use with automatic keyers. Furthermore, it is necessary for a completely automatic break-in switching to be provided. Automatic keyers are used extensively since high speeds in excess of 45 wpm allow even relatively short bursts to be utilized for communication. The receive station then uses a tape recorder with at least two speeds in order to slow down the speed to a readable level. The transmission is recorded at the higher speed and replayed at the lower speed (Editors: The frequency of the audio tone is also reduced by the same value).

Due to the short duration of the signals, it is, of course, extremely important to know the frequency of both stations extremely accurately. One will not be successful if one has to firstly look for the signal. The frequency accuracy of transmitter and receiver must be better than 1 kHz.

In practice most MS-QSOs are arranged by post or on the 14.340 MHz MS-net frequency on Saturdays at 14.00 Z (GMT). One usually arranges to be on a certain frequency at a certain time. This is usually in the CW-portion of the band between 144.000 MHz and 144.150 MHz. A frequency of 144.100 MHz ± 4 kHz has been standardized for unscheduled MS-communications. A meteor scatter QSO is usually completed within one to two hours if a favourable time and shower have been selected and assuming that the partner station is really active on the correct frequency at the correct time. In the case of smaller showers and sporadic meteor scatter, the QSO's usually take longer. In spite of this, G 3 CCH has had far more than 50 QSO's with TF 3 EA via sporadic meteor scatter.

The majority of meteor-scatter communication is made on the 2 m band. Of course, 10 m would provide better signals, but such stations can often be worked normally when the band is open. The duration of the signals on 70 cm are extremely short which means that very high code speeds are required. Also, the signal strengths are much lower at 70 cm. However, a few successful MS-tests have been carried out on this band.

It is important to keep a careful log by listing each burst and ping. A separate line for each minute is very favourable. Example:

07.00	-	08.00	ping
07.01	burst ... r 2 ...	08.01	-
07.02	ping	08.02	burst ... bu 25 25 25 dj 5
07.03	-	08.03	ping ping
07.04	burst ... bu dj ...	08.04	-
07.05	transmit UR 2 BU DJ 5 DT	08.05	transmit rr 27 rr 27
etc.		etc.	

A copy of such a log should be sent to the QSO partner. A list of the approx. 130 active MS-stations is available from DC 7 AS: Alexander Schöning, D-1000 Berlin 28, Maximiliankorso 52. Please send an addressed envelope and a sufficient number of I.R.C.s. DC 7 AS would also like to hear of your own results using the meteor scatter mode.

10. REQUIRED EQUIPMENT

It is not necessary to have a super station with several hundred watts and large antenna array. From experience one can say that a 10 dB antenna gain and FET-converter are sufficient on the receive side, and 100 W output (e.g. from a QQE 06/40 (829 B)) into a 10 dB antenna are sufficient on the transmit side for satisfactory MS-operation. In order to obtain an idea of the strength and rates of bursts and pings it is advisable to set up the receiver on the frequency of an inaudible beacon on any normal day. One can expect to hear 2 to 4 pings each quarter hour from sporadic meteors, whereas this will increase to about 20 pings per quarter hour during the larger showers.

11. REFERENCES

(1) W. F. Bein: VHF Meteor Scatter Propagation
QST (1957), Pages 20-24

(2) K. R. R. Bowden: VHF Propagation by Meteoric Ionization
The Short Wave Magazine, Vol. 16 (1958), Pages 545-546

(3) B. Cooper: Meteor Showers: New Role in VHF Dxing ?
73 Magazine (1972), Pages 71-78

RECEIVE CONVERTER 145 MHz/9 MHz WITH SCHOTTKY DIODE RING MIXER

by J. Kestler, DK 1 OF

Wideband ring mixers using Schottky hot carrier diodes are available on the market at reasonable prices. These are complete ring modulators contained in a sealed metal case. They are provided with the required differential transformer and are thus suitable for a large number of applications. The operation and characteristics of Schottky diodes in such application were given in (1). It should be mentioned that these diodes possess hardly any charge storage effects due to the absence of minority carriers, and very fast switching times due to the very thin junction layer. It is these characteristics that are very favourable at high frequencies.

A receive converter is to be described which is suitable for use in a single superhet receiver. This converter uses such a wideband Schottky-diode mixer.

1. THE RING MODULATOR

The most important specifications of the inexpensive ring modulator (type SRA-1 of MCL) are given in Table 1. The given values are based on a signal frequency of 100 MHz and a local oscillator level of +13 dBm (= 20 mW).

Conversion loss:		typ. 6 dB; max. 8.5 dB
Max. input signal power for 1 dB compression:		typ. 6 dBm (= 4 mW)
Two-tone intermodulation rejection Signal level - 10 dBm (= 0.1 mW)		65 dB
Isolation between signal and oscillator input:		typ. 40 dB, min. 30 dB
Isolation between IF-output and oscillator input:		typ. 40 dB, min. 30 dB
Frequency range:	Signal input:	5-500 MHz
	Oscillator input:	5-500 MHz
	IF-output:	0-500 MHz
Nominal impedance of all connections:		50 Ω

2. DESIGN OF THE RECEIVE CONVERTER

Several considerations and calculations are to be given which led to the practical design of the input circuits. When designing the input circuits of a receiver, it is always necessary to find a rational compromise between sensitivity and large signal capabilities, since these two demands are often in conflict with another. It is especially in the two metre band where these two characteristics are so difficult to combine. It is not uncommon for level differences between a weak DX-signal and a local station to be greater than 120 dB. For this reason, it is advisable for the VHF gain before the mixer to be made variable so that it can be matched to the appropirate conditions.

For calculation of the required VHF gain factor, it is assumed that a crystal filter (single conversion superhet) is present at the input of the IF amplifier and that the first IF stage possesses a noise figure of 3 dB. The insertion loss of the first crystal filter including the matching losses is assumed to be 6 dB, and the passband width to be 2.5 kHz (SSB). It is sufficient for the noise level of the VHF stages to exceed the IF noise by 10 dB so that the latter does not noticeable add to the total noise figure.

At the given bandwidth B and IF noise figure (3 dB = F = 2), the inherent noise power of the IF amplifier referred to the input of the first IF stage will amount to:

$$P_{IFn} = F \times kT_o \times B = 2 \times 10^{-17} \text{ W}$$

It should be mentioned that the bandwidth of the filter can only be used as noise bandwidth when a second (identical) filter is present at the output of the IF amplifier. If this is not the case, the noise bandwidth is as wide as the bandwidth of the IF amplifier. This condition is to be mentioned later.

The noise level of the VHF stages amounts to approx. 2×10^{-16} W when referred to the input of the IF amplifier. Conversion loss and filter insertion loss result in a total of approx. 13 dB which means that the VHF noise level should possess a value of approximately 4×10^{-15} W at the input of the ring modulator. The dynamic range of the receiver is thus fixed, since Table 1 gives a maximum permissible signal level (for 1 dB compression) for the ring modulator of 4×10^{-3} W:

$$\text{Dynamic range: } \frac{4 \times 10^{-3} \text{ W}}{4 \times 10^{-15} \text{ W}} = 10^{12} = 120 \text{ dB}$$

If it is assumed that the input stage of the VHF amplifier also possesses a noise figure of 3 dB (F = 2), the VHF noise power referred to the antenna input will also amount to 2×10^{-17} W corresponding to 0.04 μV into 60 Ω. This means that a VHF gain factor (between antenna input and input of the mixer) is required which amounts to:

$$G_{VHF} = \frac{4 \times 10^{-15} \text{ W}}{2 \times 10^{-17} \text{ W}} = 2 \times 10^{2} = 23 \text{ dB}$$

If no further crystal filter is to be provided at the output of the IF amplifier, the noise bandwidth will amount to approximately 250 kHz (e.g. four resonant circuits at 9 MHz). The noise level of the IF amplifier is thus 100 times greater so that the VHF gain must amount to 43 dB instead of 23 dB. This is not possible since the dynamic range would be reduced considerably. If the receiver is only to be used for SSB operation, it would be possible for a steep AF filter to be provided after the product detector in order to reduce the noise bandwidth to twice the AF filter bandwidth (double-sideband demodulation). However, the generation of the control voltage (S-meter) still requires the whole IF bandwidth, assuming that the control voltage is not obtained from the AF voltage. This method cannot be used for AM and FM operation.

Another possibility is to provide an additional 20 dB of gain between the ring modulator and the crystal filter. The demands made on the large-signal capabilities of this stage are very high since the ring modulator provides a power level of nearly 1 mW with an input signal of 120 dB. The intermediate stage must therefore be able to handle 100 mW without distortion if it is to be able to amplify by the required level of 20 dB. Such an IF preamplifier is to be described at the end of this article.

The required VHF-preamplification of 23 dB is not made in a single stage, but split between two transistor stages. This also simplifies the provision of the resonant circuits required for ultimate and image attenuation. A low-noise junction FET is used in the input stage; the second stage is equipped with a MOSFET, mainly due to its control characteristics.

Figure 1 gives the block diagram and level plan of the receive converter. The local oscillator module should provide a level of approximately -7 dBm (corresponding to 0,2 mW or approx. 100 mV into 60 Ω). This means that the phase-locked oscillator (2) and the 80-channel synthesizer (3) from the same author can be directly connected.

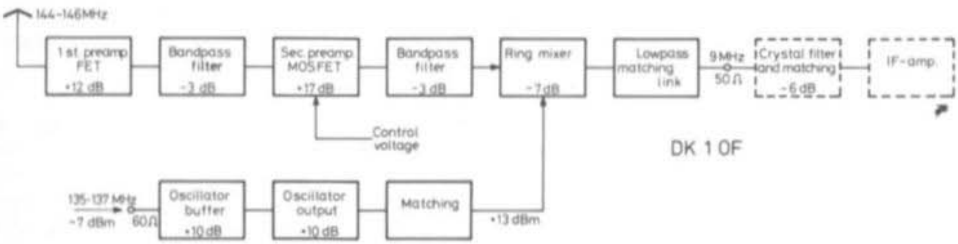


Fig. 1: Block diagram and level plan

3. CIRCUIT DESCRIPTION

The circuit diagram of the whole converter is given in Figure 2. The antenna is connected to Pt 161. The transformation link comprising C_A/L 160 matches the impedance of the antenna feeder to the first VHF preamplifier stage (T 161) which is operated in a common gate circuit. Since this type of circuit is somewhat unusual, it is to be described in more detail:

The required impedance transformation at the antenna input should be made with a minimum of signal loss since each dB of loss will deteriorate the noise figure of the whole receiver by the same value. For this reason, all components used for transformation should possess high static Q values, whereas the operating Q of the transformation link must be relatively low since the efficiency depends on the ratio of static to operating Q. The selectivity of the input circuit is therefore low. This means that it is not advisable for the usual parallel circuit with transformation link with a coil tap to be used. Tapped coils always exhibit a poor efficiency at higher frequencies since the stray inductivity between the parts of the windings is high due to the low number of turns.

This disadvantage is avoided in the high-pass filter link (C_A and L 160) shown in Figure 2. The high-pass filter coupling is very advantageous at this point since the image frequency is below the required frequency band and will therefore be better attenuated. The transformation ratio is fixed by the selection of C_A; L 160 is used to compensate the reactive values of the capacitor C_A and the transistor input.

Input circuits of amateur radio receivers usually use noise matching of the first transistor. This means that the input of the first transistor must be terminated with its optimum generator impedance $Z_{G\ opt.}$. It is the task of the transformation link to match the antenna impedance Z_A to this generator impedance

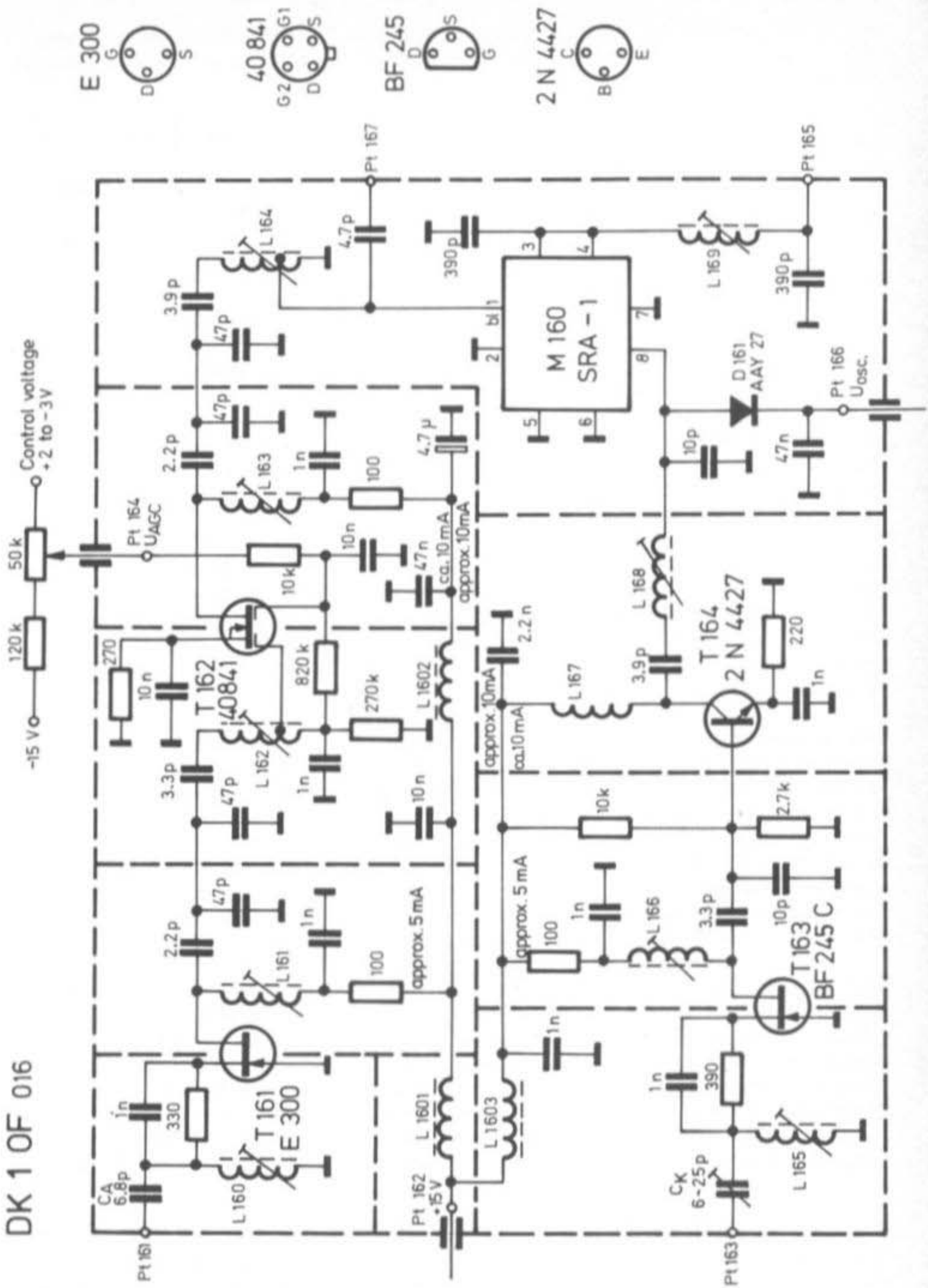


Fig. 2: Circuit diagram of the 144 MHz / 9 MHz receive converter with wideband ring mixer

and to compensate the reactive component C_{in} of the input transistor. Figure 3 gives further details. According to the rules of complex alternating current, the following are valid:

$$C_A = \frac{1}{\omega} \times \sqrt{\frac{1}{Z_A\,(Z_{G\,opt} - Z_A)}} \quad \text{and}$$

$$L = \frac{1}{\omega^2} \times \frac{1 + \omega^2 C_A{}^2 Z_A{}^2}{C_A + C_{in} + \omega^2 \times C_A{}^2 \times C_{in} \times Z_A{}^2}$$

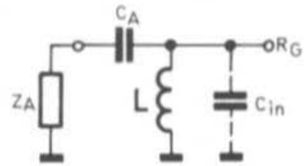


Fig. 3

The following is valid at $Z_A = 60\,\Omega$, $Z_{G\,opt} = 600\,\Omega$, $C_{in} = 4$ pF; $\omega = 9.1 \times 10^8$ (≙ 145 MHz): $C_A = 6.1$ pF, $L \approx 130$ nH.

In some cases it is advisable for power matching to be used instead of noise matching. This is especially the case when very long antenna feeders are necessary (4). In this case, it is not the optimum generator impedance $Z_{G\,opt}$ that must be used for the calculation, but the actual input impedance Z_{in} of the input transistor. This is equal to the inverse value of the transistor slope, e.g. approximately 200 Ω. In this case, the following values should be selected: $C_A = 12$ pF and $L \approx 100$ nH. L 160 should therefore be reduced to 5 turns.

The first VHF stage is equipped with the low-noise FET E 300 manufactured by Siliconix. The manufacturer gives a noise figure of 1.3 dB at a frequency of 100 MHz. The neutralization problems are avoided when the described common gate circuit is used. The amplifier stage operates extremely stabily and possesses a high output impedance that does not load the subsequent bandpass filter to any degree.

The subsequent two-link bandpass filter comprising inductances L 161 and L 162 is capacitively coupled. The advantage of this coupling over inductive coupling is that the degree of coupling is exactly reproduceable and need not be optimized with the aid of a VHF sweep generator that is usually not available. The filter is slightly overcoupled (passband dip approx. 1 dB), so that the 3 dB bandwidth of 2 MHz is obtained. Due to the base coupling used, favourable capacitance values (2 x 47 pF) are obtained. In order to reduce the effect of the connection inductances, the whole capacitance is divided over two ceramic disc capacitors with short connections.

The second VHF amplifier stage is loosely coupled to the bandpass filter in order not to dampen the secondary circuit. A dual-gate MOSFET type 40841 (RCA) is used. This is the follow-up type for the wellknown 40673 that is no longer in production. The gain control is active via both gates in order to ensure that the large-signal capabilities do not deteriorate to any extent when controlled. This stage is provided with a control voltage via connection Pt 164 from a 50 kΩ resistor that is not located on the PC-board. A variation of the control voltage from +2 V to -3 V provides a gain variation of 60 dB for this stage.

The input signal is then passed via a second, identical bandpass filter (L 163/ L 164) and fed to the VHF input of the ring modulator. This connection is specifically marked on the ring modulator, mainly with the aid of a coloured glass feedthrough for the associated connection pin. Part of the amplifier VHF signal can be tapped off at connection Pt 167 and fed, for instance, to a panoramic receiver. The output coupling (via 5 pF) is so loose that no reduction in gain is noticeable when this output is loaded.

The local oscillator frequency is fed to the receive converter via connection Pt 163. The input impedance can be aligned to 60 Ω (SWR = 1) with the aid of the transformation link comprising C_K/L 165. The buffer stage comprising transistor T 163 oscillates freely and efficiently in the common gate circuit. Since the output of the local oscillator (T 164) must provide a power of 20 mW, a low power transistor type 2 N 4427 has been used. Other types such as 2 N 3866, 2 N 3553, 2 N 5913 or similar can be also used. The oscillator signal is fed to the ring modulator via a matching circuit comprising inductance L 167, L 168 and capacitors of 4 pF or 10 pF. Diode D 161 is provided to assist alignment. A DC-voltage is available at connection Pt 166 whose amplitude is approx. equal to the peak voltage of the oscillator signal at pin 8 of the mixer.

The resulting intermediate frequency is tapped off at connections 3 and 4 of the ring modulator. A pi-filter comprising L 169 and two 390 pF capacitors suppress any residual VHF or local oscillator voltage on the intermediate frequency signal. The IF output Pt 166 should be loaded with approx. 50 Ω.

The operating voltage (Pt 162) should be approximately 15 V. With a slight reduction in gain, it is also possible to operate the converter from 12 V.

4. COMPONENTS

T 161: E 300 (Siliconix) or 2 N 5397
T 162: 40841 (RCA)
T 163: BF 245 C (TI) or W 245 C (Siliconix)
T 164: 2 N 4427 (RCA) or 2 N 3866, 2 N 3553, 2 N 5913 or similar

D 161: AAY 27 or AA 116 (Siemens) or similar point-contact diode

M 160: Wideband ring modulator SRA-1 (Mini-Circuits Laboratory) or similar

L 160: 7 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6 mm coilformer, 1 mm spacing between turns, VHF core (brown)
L 161: 9 turns, wire, coilformer and core as L 160
L 162: 9 turns, coil tap 1 turn from the cold end; otherwise as L 160
L 163: 9 turns, otherwise as L 160
L 164: 9 turns, coil tap 1 turn from the cold end; otherwise as L 160
L 165: 6 turns otherwise as L 160
L 166: 9 turns otherwise as L 160
L 167: 5 turns, as L 160 but self-supporting
L 168: 5 turns otherwise as L 160
L 169: 7.5 turns of 0.4 mm (26 AWG) enamelled copper wire in special coil set, approx. 0.8 μH
L 1601-L 1603: Wideband ferrite chokes with 6-hole cores (Philips)

C_K : Ceramic disc capacitor 6 - 25 pF, 10 mm dia. or plastic foil trimmer 2 - 22 pF (7 mm dia.).

With exception of the electrolytic (4.7 μF or more), all capacitors are ceramic disc types for 5 mm spacing.

A spacing of 12.5 mm is available for all resistors.

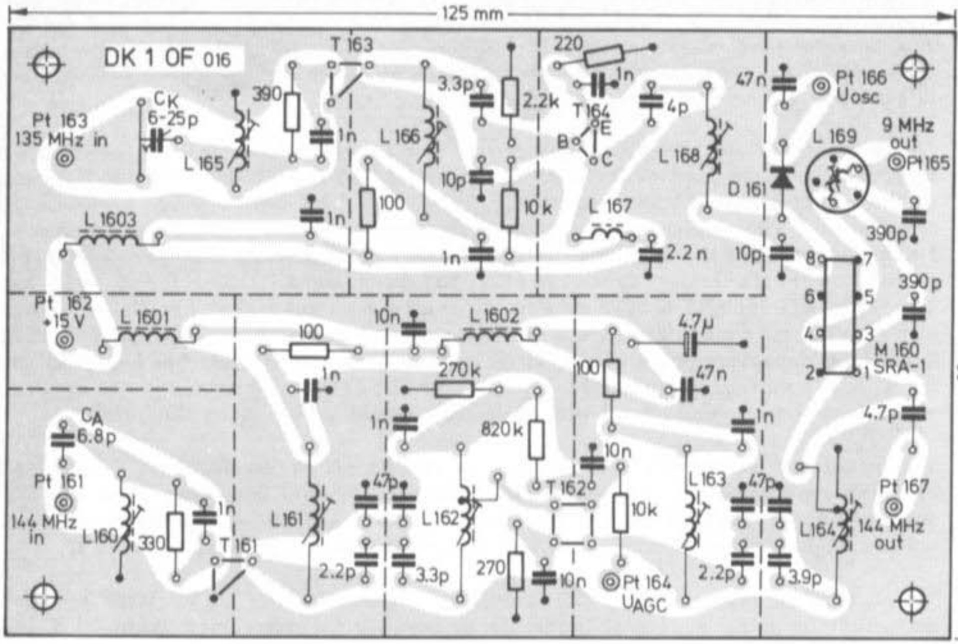


Fig. 4: Component locations on PC-board DK 1 OF 016

5. CONSTRUCTION

The single-coated PC-board DK 1 OF 016 has been developed for accommodation of the receive converter. The dimensions are 125 mm x 80 mm. As can be seen in the component location plan (Fig. 4), the converter is divided into a total of nine chambers with the aid of 30 mm high screening panels. The module is also provided with screening panels around the board (height also 30 mm) onto which the coilformers, feedthrough capacitors (Pt 162, 164, 166) and the RF cable connections (Pt 161, 163, 165, 167) are mounted. Figure 5 shows a photograph of the author's prototype.

6. ALIGNMENT

The alignment is commenced by aligning the amplifier of the local oscillator signal. This is achieved by connecting the output of the oscillator to Pt 163. A DC voltmeter (range 3 V, Z ≥ 100 kΩ) is connected between Pt 166 and ground. The oscillator should be tuned to the centre frequency (136 MHz) and the operating voltage connected to Pt 162. The core of inductance L 166 is now aligned for maximum reading on the meter. After this inductance L 168 is also aligned in a similar manner, completing the alignment with capacitor C_k. Due to the reaction of transistor T 164, inductances L 166 and L 168 will have some effect on another, so that it is necessary for these two inductances to be aligned alternately. A DC-voltage of approximately 1.5 V should be present at Pt 166 when the alignment has been completed correctly; a voltage of 1.2 V should still be available at the band limits. An oscillator voltage of approximately 0.2 mW is required at connection Pt 163 in order to achieve this. It is not possible to align the local oscillator amplifier circuits for maximum gain of the converter since the maximum gain will not continue to increase above a

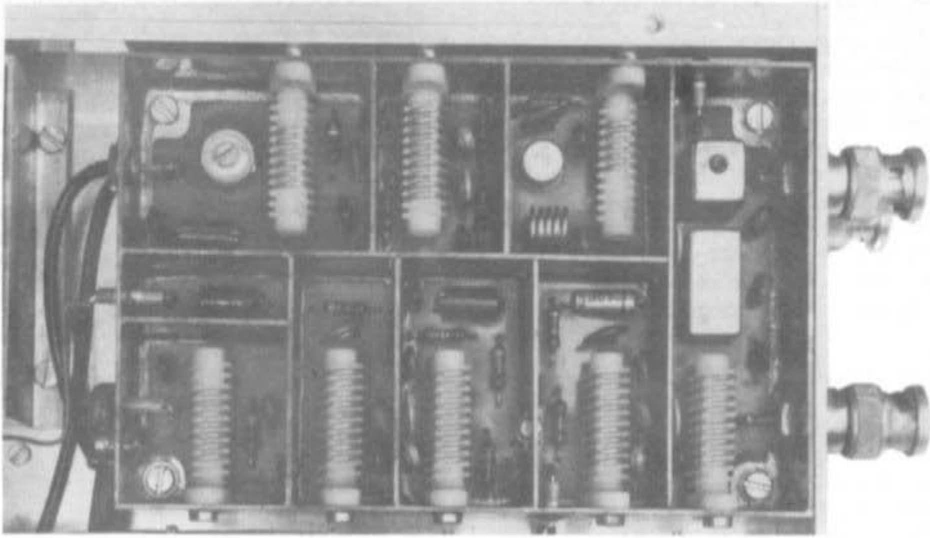


Fig. 5 : Author's prototype of the receive converter with Schottky diode mixer

certain oscillator level, (of approx. 5 mW at the ring modulator). However, the correct alignment of the oscillator chain is very important in order to obtain a favourable large-signal characteristic. For this reason test diode D 161 has been included on the board.

If a sensitive reflectometer is available (most HF standing wave bridges are suitable), input Pt 163 should be aligned for best SWR. This is achieved by observing the meter connected to Pt 166 and trimming C_k and L 165 alternately until minimum SWR and maximum voltage are indicated at the same time.

The VHF stages can be aligned very favourably with the aid of a sweep generator. The sweep generator is connected to Pt 161, and the probe of the oscilloscope to Pt 167. It is important that the local oscillator is in operation during the alignment since the input impedance of the ring modulator is very dependent on the local oscillator voltage. The IF output should also be terminated with 50 Ω.

If a sweep generator is not available, the passband curve should be measured point for point with the aid of a signal generator and corrected where required. If no measuring equipment is available, the converter should be aligned for constant noise over the whole band. However, it will hardly be possible to obtain an optimum passband curve in this manner. In this case, it will be better to align the four bandpass filter circuits for maximum with a receive signal of 145.5 MHz. This will allow the selectivity curve to be virtually symmetrical over the 2 m band. The input circuit comprising L 160 is aligned for best signal-to-noise ratio with a weak signal at the centre of the band.

The control voltage (Pt 164) at which the alignment is made is not critical. Due to the loose coupling of the MOSFET to the previous bandpass filter, the passband curve is practically independent of the control voltage.

7. PRACTICAL EXPERIENCE

The large-signal characteristics of the described converter should be considerably better than standard converters using FET or MOSFET mixers. It is, for instance, possible for the author to operate duplex only 30 kHz from his transmit frequency without desensitizing his receiver. In this case, a second antenna is used and the transmit power in SSB or FM amounts to 50 W. The decoupling between the two antennas amounts to approximately 60 dB, so that the author's own transmit signal is greater than 120 dB over noise at the converter input. Prerequisite for such tests are, of course, a very selective 9 MHz IF (3 crystal filters in the SSB mode) and oscillators having a very low sideband noise (for transmitter and receiver). The phase-locked oscillator as described in (2) satisfies these demands.

The characteristics of the receive converter are given in the following table:

Noise figure:	approx. 4 dB (F $\hat{=}$ 2.5)
Overall gain:	16 dB
3 dB bandwidth:	2 MHz
Passband ripple:	2 dB
Image rejection (127 MHz):	65 dB
Input matching:	50-60 Ω
IF output impedance:	50 Ω
Required oscillator level:	-7 dBm (= 0.2 mW)
Max. permissible interference signal	
at input: for 1 dB desensitization:	40 mV
for 6 dB desensitization:	100 mV
Dynamic range (1 dB compression):	120 dB

8. IF PREAMPLIFIER

As has already been mentioned in Section 2, an IF amplifier is to be provided between connection Pt 165 and the crystal filter if a second crystal filter is not to be provided at the output of the IF amplifier. In order to achieve a high output power having a good linearity, a low-noise high-current HF transistor has been chosen as the active element. A type 2 N 5109 (RCA) has been used, which was developed for applications in wideband amplifiers in TV-community antenna systems. Table 3 gives details as to the excellent characteristics of this transistor:

Limit values:

Collector-emitter voltage	40 V
Collector current	400 mA
Power dissipation	3.5 W

Dynamic values:

Transit frequency (U_{CE} = 15 V, I_C = 60 mA):	1.5 GHz
Noise figure (U_{CE} = 15 V, I_C = 10 mA, f = 200 MHz):	3 dB
Cross modulation rejection (at 5 mW output power):	57 dB
Reactive capacitance:	3.5 pF

8.1. CIRCUIT DETAILS

The circuit diagram of the IF amplifier is given in Figure 6. The two-stage capacitively-coupled bandpass filter at the input (Pt 171), provides the required ultimate selectivity. The subsequent transistor stage in a common-emitter circuit is built up in a conventional manner and offers no special features. The coupling link of filter F 3 is present in the collector circuit, and the tapped inductance of the resonant circuit is terminated with 50 Ω via Pt 173. The transformation ratio between this tap and the coupling winding amounts to 1 : 1 so that the output of the transistor is loaded with 50 Ω. For this reason no tendency to self-oscillation is present in spite of the relatively high reactive capacitance. A transformed impedance of 1200 Ω is present at the hot end of the resonant circuit, which means that the crystal filter XF-9 E can be directly connected. If the filter is to be connected via a coaxial cable to Pt 172, the 27 pF capacitor should be reduced to the value of the cable capacitance.

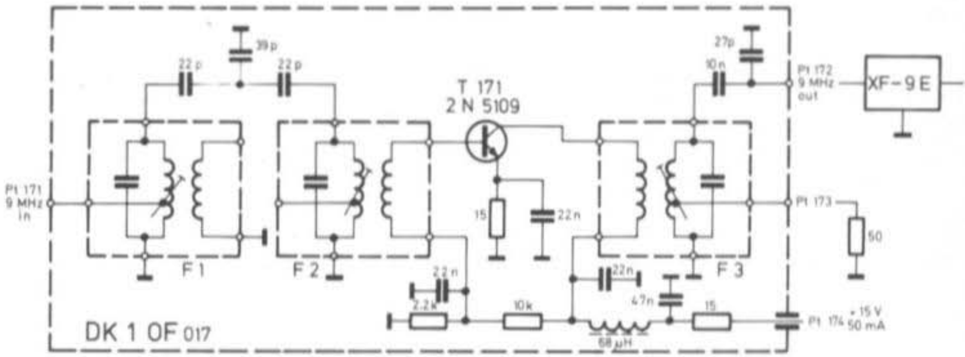


Fig. 6: Circuit diagram of the high-level IF amplifier

Since the transistor is to be operated with a collector current of 50 mA (it can be corrected by altering the value of the base voltage divider), it is important that it is provided with cooling fins. The operating voltage of 15 V (12 V is also suitable) is provided via Pt 174 and a filter link.

8.2. COMPONENTS AND CONSTRUCTION

T 171: 2 N 5109 (RCA) with cooling fins

F 1 – F 3: Screened 10.7 MHz miniature filters FM-FB

Choke: 68 µH miniature ferrite choke, value uncritical

All capacitors: Ceramic disc types for 5 mm spacing; tubular capacitors for 10 mm spacing can be used for the lower values.

The IF amplifier is accommodated on a single-coated PC-board whose dimensions are 125 x 30 mm. This PC-board has been designated DK 1 OF 017 and it is also provided with screening panels of 30 mm in height. Component locations and a photograph of the author's prototype are given in Figure 7 and Figure 8.

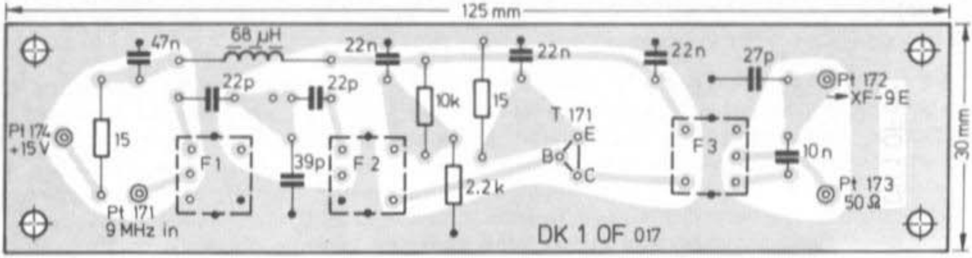


Fig. 7: Component locations on board DK 1 OF 17

8.3. ALIGNMENT OF THE IF AMPLIFIER

The alignment is relatively simple: After the module has been connected between the receive converter and the main 9 MHz IF amplifier, filters F 1 and F 2 are aligned alternately for maximum noise. Filter F 3 is responsible for the matching to the crystal filter and should be aligned for minimum ripple in the passband range of the XF-9E.

If the receiver is to be only designed for SSB operation, it is possible to replace the FM crystal filter with an SSB type XF-9B. Since such an SSB crystal filter is matched to 500 Ω, it is necessary to reduce the value of the terminating resistor at connection Pt 173 to 20 Ω. The reduction in gain due to this is not important in practice.

The described IF amplifier goes into saturation at approximately 500 mW. It operates completely linearly at an output voltage of 100 mW.

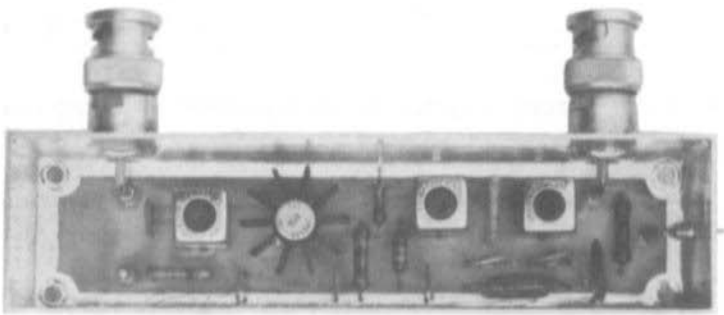


Fig. 8: Author's prototype of the IF amplifier

9. REFERENCES

(1) R. Lentz: A Wideband Ring Mixer with Schottky Diodes
VHF COMMUNICATIONS 4 (1972), Edition 2, Pages 121-124

(2) J. Kestler: A Phase-Locked Oscillator for 144 MHz
VHF COMMUNICATIONS 6 (1974), Edition 2, Pages 114-124

(3) J. Kestler: FM Transceiver with Multi-Channel Synthesizer
VHF COMMUNICATIONS 5 (1973), Edition 3, Pages 130-145

PRODUCT DETECTOR AND CRYSTAL OSCILLATORS FOR THE MODULAR RECEIVER

by D. E. Schmitzer, DJ 4 BG

1. MODIFICATIONS TO THE ORIGINAL SYSTEM

Whilst working on the SSB/CW IF-circuit of the modular receiver, it was found that the injected oscillator voltage was fed to the product detector not only directly but also via the IF amplifier if both circuits were accommodated in the same module. For this reason, the original concept (1) where a combined IF and product detector module were to be used with the triple crystal oscillator (DJ 4 BG 009) is no longer used. The product detector is now accommodated in a separate module from the IF circuit and combined with a simplified oscillator circuit with switchable crystals. This new module has been designated DJ 4 BG 014. A circuit for generating the required control voltage is now accommodated in the IF module in the space previously planned for the product detector.

2. CIRCUIT OF THE NEW PRODUCT DETECTOR AND CRYSTAL OSCILLATOR MODULE

The circuit diagram of this module (DJ 4 BG 014) is shown in Figure 1. The actual product detector is connected as push-pull mixer and uses the differential amplifier of the integrated circuit CA 3028. The IF signal (from Pt 5) is fed to the base of the constant current transistor (pin 2) where it drives the differential amplifier of the integrated circuit CA 3028. The IF signal (from Pt 5) is fed to the base of the constant-current transistor (pin 2) where it drives the differential amplifier mixer in push-push. Feedback is provided via the emitter of the constant-current transistor (pin 4) with the aid of resistor R 4 so that a very linear input characteristic results. The capacitor C 4 ensures that resistor R 4 is not able to alter the DC-voltage relationships in the integrated circuit.

The two base connections of the differential amplifier (pin 1 and pin 5) are connected to the same DC-voltage. One base is bypassed (C 2), and the other is fed with the oscillator signal. The cross coupling via the emitter of the integrated differential amplifier therefore provides a push-pull drive of the oscillator signal. The AF-voltage is taken from the collector resistor R 6 and passed via a low-pass filter comprising C 6, R 7, and C 7 in order to remove any residual RF voltages.This is followed by an emitter-follower comprising transistor T 1 which then provides a low-impedance AF voltage to the output connection Pt 10.

The crystal oscillator is coupled to the mixer via the dropper resistor R 3 which forms a voltage divider together with R 1. This ensures that no effect is made on the crystal oscillator even when the product detector is overdriven.

The three crystals provided in the oscillator are switched with the aid of diodes. Under normal conditions, the crystal frequencies used are those for the upper and lower sideband as well as for CW. The base bias voltages for the differential amplifier and the oscillator transistor are stabilized by the zener diode D 4. In addition to this, the stabilized voltage ensures that all switching diodes that are not used are blocked with a voltage of approximately 5 V so that it is not

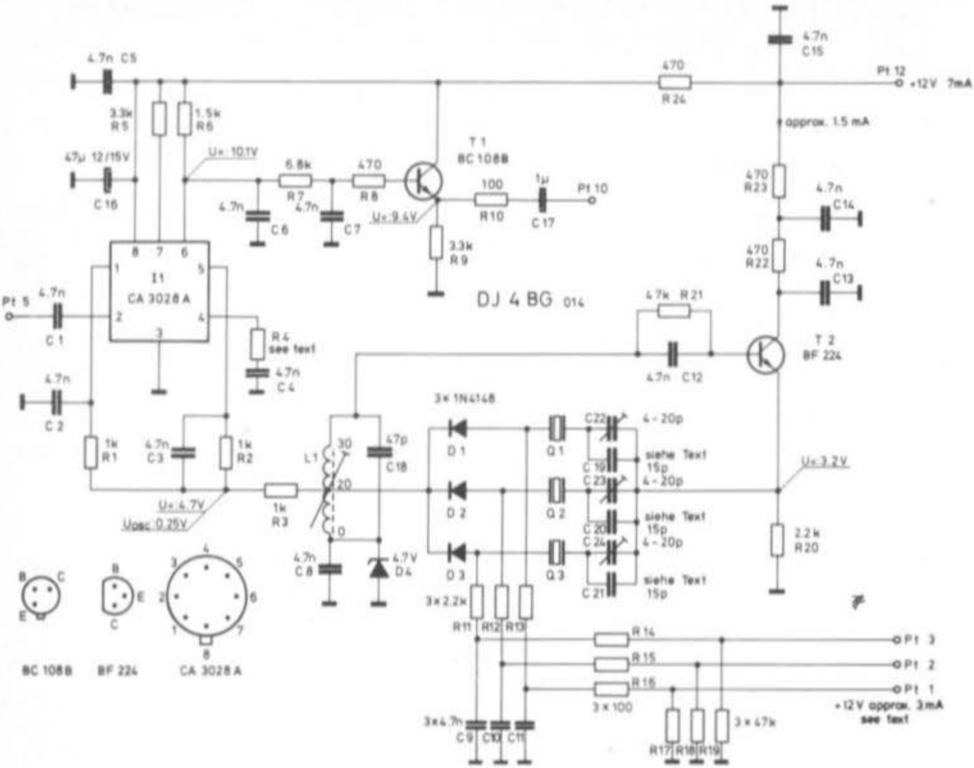


Fig. 1: Circuit diagram of the product detector and crystal oscillator module

necessary for an additional voltage to be used. This means that it is only necessary to connect the full operating voltage of 12 V to the appropriate connection (Pt 1 to Pt 3) in order to select one of the three crystals.

3. COMPONENTS

As with the other modules of the modular receiver system, a PC-board with the dimensions 65 mm x 90 mm is available. This board, which is designated DJ 4 BG 014, is shown in Figure 2 together with the component locations. It can be accommodated in a TEKO box 3 A and provided with a 13-pin connector.

T 1: BC 108 B, BC 183 B, BC 413 B, or similar silicon AF transistor with B min. 100

T 2: BF 224 (TI) or similar NPN RF transistor with low reactive capacitance. Attention should be paid to the connections when using other transistors.

I 1: CA 3028 A or CA 3053 (RCA)

D 1 - D 3: 1 N 914, 1 N 4148 or similar, low-capacitive silicon diodes.

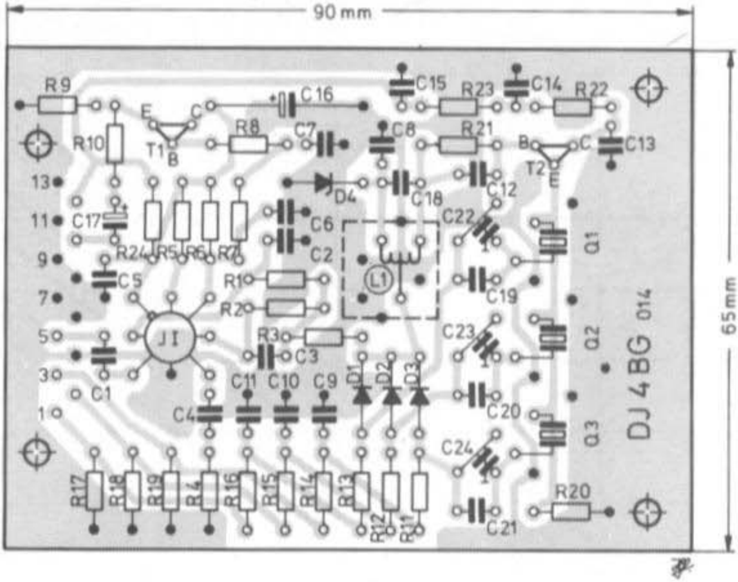


Fig. 2: PC-board DJ 4 BG 014 with component locations

C 22 - C 24: 4 - 20 pF ceramic disc trimmers 7 mm dia.
or plastic foil trimmers (green)
C 1 - C 15: 4.7 nF ceramic disc capacitors, spacing 5 mm
C 18: 47 pF ceramic capacitor
C 19 - C 21: 15 pF ceramic capacitor

L 1: 30 turns of 0.2 mm dia. (32 AWG) enamelled copper wire, coil tap 20 turns in special coil set.

4. ALIGNMENT

Trimmer capacitors C 22 to C 24 should firstly be brought to their central position and inductance L 1 aligned for maximum oscillator voltage. This condition is shown by a slight increase of the collector current of transistor T 2. This can be measured either as the voltage drop across resistor R 23 (approx. 0.7 V) or by connecting a mA-meter (approx. 1.5 mA) instead of R 23.

The inductance should not be altered after this adjustment since the inductivity also affects the oscillator frequency. The nominal frequency of each crystal can now be adjusted with the appropriate trimmer capacitor. The alignment range is generally sufficient, but can be extended if required by altering the values of the parallel capacitors C 19 to C 21. Although the crystal frequencies can be adjusted by ear, it is more favourable for them to be adjusted with the aid of a frequency counter. In this case, it should be connected at high impedance to the emitter of transistor T 2.

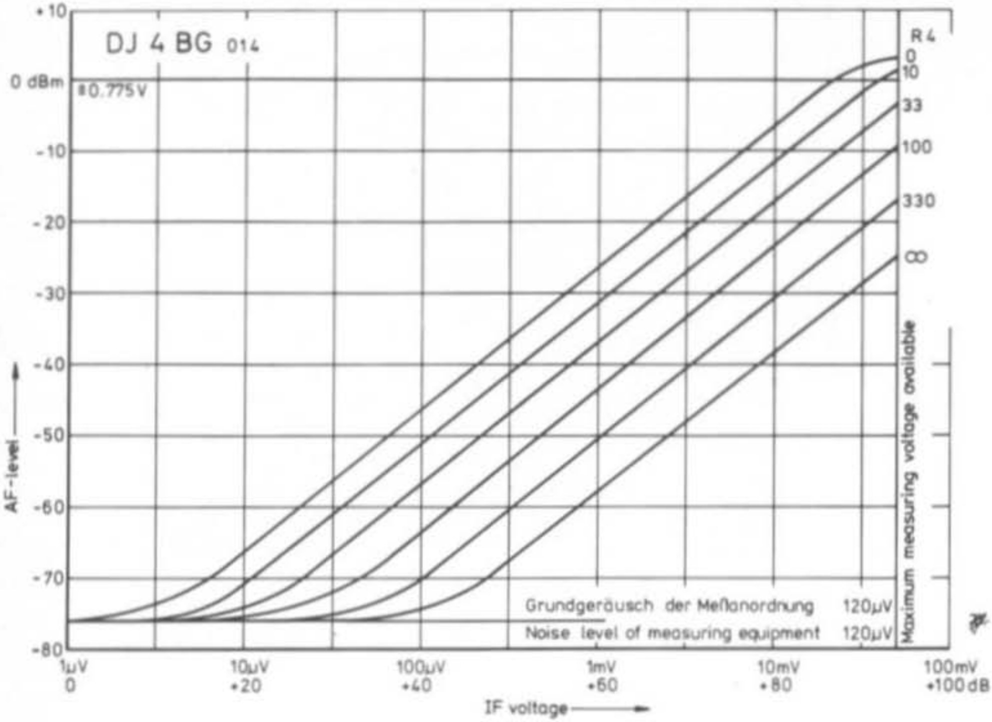


Fig. 3 : Characteristic curves of the product detector DJ 4 BG 014 at various degrees of feedback

The nominal frequencies of the crystals are:

XF 901: upper sideband 8998.5 kHz
XF 902: lower sideband 9001.5 kHz
XF 903: CW 8999.0 kHz

Since the active CW-filter for this receiver is designed for a somewhat lower frequency than 1 kHz, e.g. for approximately 800 Hz, it is advisable for the crystal XF 903 to be aligned to 8999.2 kHz.

5. MEASURED RESULTS

The relationships between the IF voltage and the demodulated AF voltage at various values of the feedback resistor R 4 is given in Figure 3. At maximum feedback, e.g. with R 4 = ∞, the gain is approximately equal to 1. This means that an AF output voltage of 10 mV will be present for an IF input voltage of 10 mV. Without feedback, e.g. with R 4 = 0, a total gain of approximately 32 dB will be provided. Even at this high gain, it will be seen that the characteristic curve is very linear up to IF voltages in excess of 10 mV. Distortion is observed only in excess of approximately 15 mV.

Of course, the highest possible value should be selected for the feedback resistor R 4 in order to provide the most linear range. The higher the IF voltage and/or the higher the sensitivity of the AF amplifier, the higher will be the value of R 4. On the other hand, the high sensitivity of the product detector at R 4 = 0 would allow a receiver to be designed that requires no IF amplifier. Further details regarding this will be given in the description of a single-band, shortwave receiver comprising the receiver modules which will be described later.

6. REFERENCES

(1) D. E. Schmitzer: A Modular Receive System
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 110-114

A SYSTEM BOARD FOR THE TEKO MODULES

by D. E. Schmitzer, DJ 4 BG

A system board has been developed that allows the various modules of the modular receiver system to be easily combined to form a versatile receiver for SSB/CW. This system board ensures that the only construction required is that the individual modules and for a case for the complete receiver.

The system board provides all electrical and mechanical connections for each of the modules. It is also equipped with a power supply to run them.

1. POSSIBLE VARIATIONS

The various possibilities and combinations are given in Figure 1 in the form of a flow diagram. A total of approximately 750 different receiver combinations are possible and far more if one or more VHF/UHF converters are also provided. These variations are based on the number of frequency bands that can be provided in conjunction with module DJ 4 BG 011, the bandwidth for the required modulation mode (SSB or CW), the possibility of using a variable or crystal-controlled local oscillator, as well as the various types of gain control and number of IF modules (with or without CW filter).

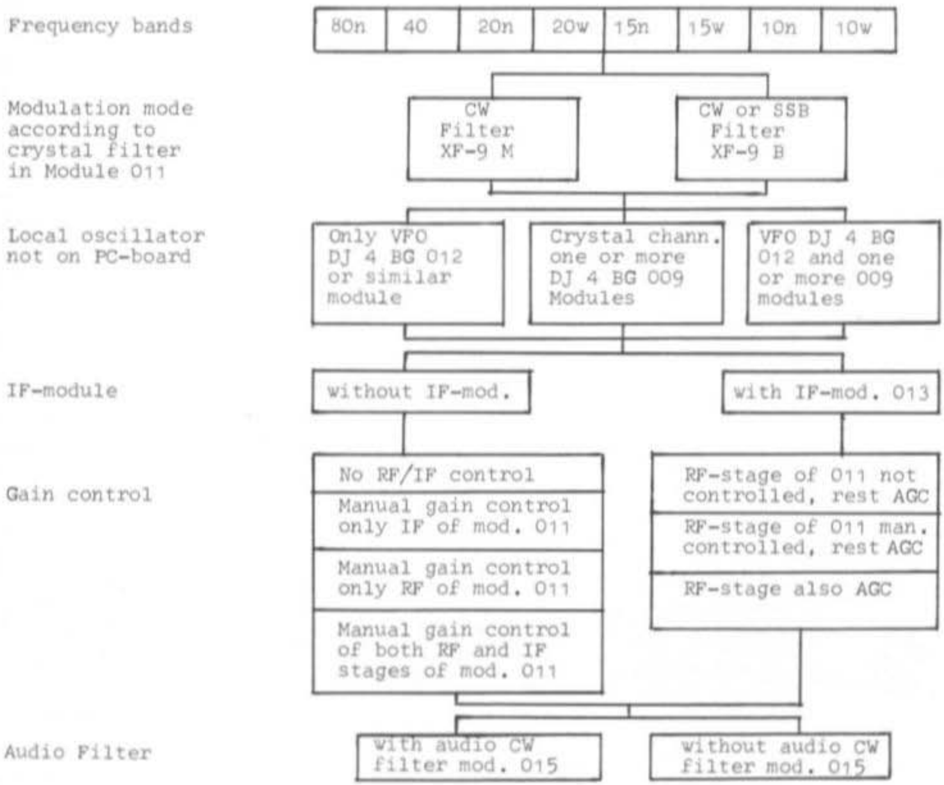


Fig. 1: Flow diagram of possible variations of the modular receiver

The maximum complement would include the following modules:

AF output and voltage stabilization DJ 4 BG 007
CW filter DJ 4 BG 015 in preparation
Product detector DJ 4 BG 014
IF amplifier with AGC, DJ 4 BG 013 in preparation
Shortwave receive converter DJ 4 BG 011

The system board also includes the power supply for the complete modular receiver comprising power transformer, rectifier and filtering.

The local oscillator for the mixer of the shortwave converter is not provided on the same board since it is advisable for it to be constructed separately due to the demands with respect to stability that are placed on it. Further details regarding this were given with the description of module DJ 4 BG 012. If only a few individual channels are to be received for certain applications (for example for fixed frequencies, emergency networks, calling frequencies etc.) it is possible for the triple crystal oscillator module DJ 4 BG 009 to be used.

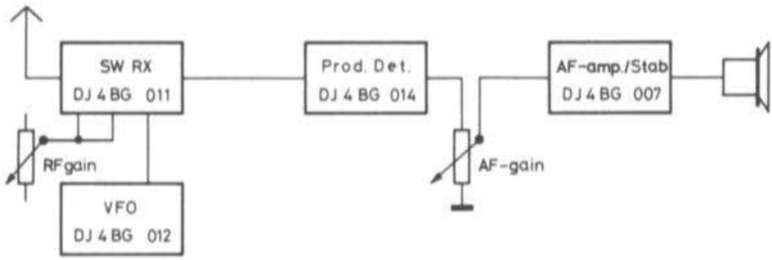


Fig. 2a: Simple SSB or CW receiver for one shortwave band

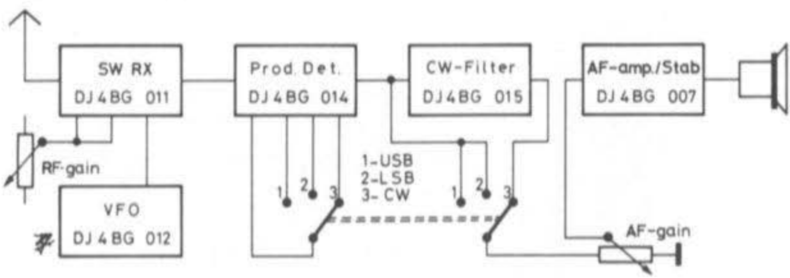


Fig. 2b: Receiver for SSB and CW

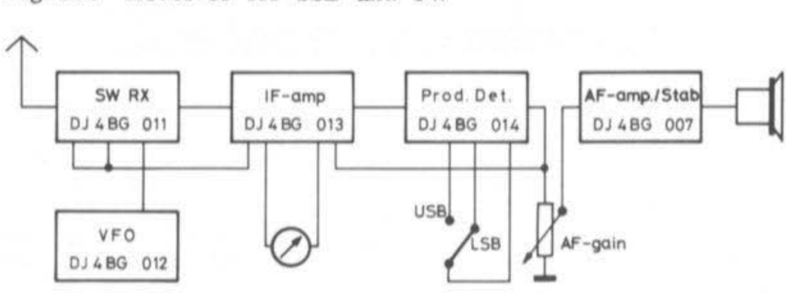


Fig. 2c: SSB receiver with automatic gain control and S-meter

A few examples from the large number of possibilities are given in Figure 2 in the form of block diagrams. Figure 2a shows the receiver combination mentioned in (7) for CW or SSB and allows such a receiver to be constructed very simply.

2. INDIVIDUAL MODULES

The following sections are to discuss details of the individual modules. It is advisable for the individual descriptions to be also available in order to avoid confusion. The module number is to be given together with all connection points and components. For instance, resistor R 4 of module DJ 4 BG 014 will be designated as R 4-014.

2.1. AF OUTPUT AND VOLTAGE STABILIZATION (DJ 4 BG 007)

Virtually no modifications are required to this module as long as the short ground connection from Pt 11-007 to Pt 12-007 is disconnected on PC-board DJ 4 BG 007. This is necessary because a stabilized voltage of + 12 V is required via the system board at Pt 12-007.

With the exception of the AF output stage, all modules mounted on the system board will be supplied from the +12 V stabilized voltage from module DJ 4 BG 007. This means that they place no further demands on the quality of the power supply filtering. Any residual hum voltages can only have an effect via the voltage divider R 2-007/R 1-007 of the AF output stage. This means that only a low degree of filtering is required at this point. It is even simpler for the voltage divider to be fed with the stabilized, and well-filtered +12 V supply. This is achieved on PC-board DJ 4 BG 007 by breaking the connection from R 2-007 to the unstabilized supply and making a small bridge on the lower side of the board to connection Pt 12-007 as shown in Figure 3.

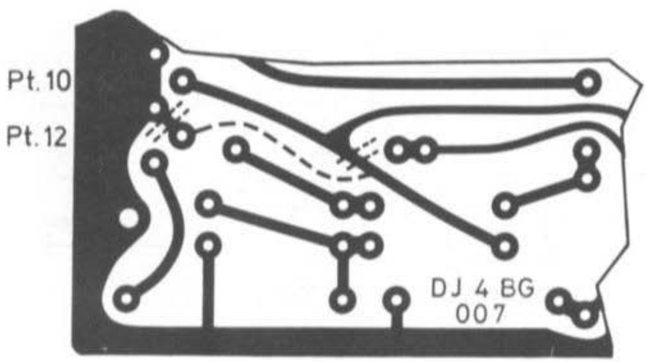


Fig. 3 : Part of PC-board DJ 4 BG 007 showing modifications

Resistor R 2-007 should be reduced to approximately 68 kΩ. The value should be selected so that approximately 8 V are present at pin 7 of the integrated circuit PA 237. This results in the most favourable drive range, since the operating voltage from the power supply (approx. 18 to 20 V) falls to approximately 16 V under full drive conditions. Further data regarding the output transformer Tr 2 are given in (2).

2.2. CW FILTER (DJ 4 BG 015)

This module, which is still in preparation, allows the SSB receiver (crystal filter XF-9B in module DJ 4 BG 011) to be modified for CW reception. If a CW receiver is to be constructed with a CW crystal filter (XF-9M in module DJ 4 BG 011), this filter will allow an even lower bandwidth to be selected (200 Hz) when the bandwidth of the crystal filter XF-9M of 500 Hz is still too wide. This module can be switched in and out of circuit via connections Pt 15/16/17 of the system board. If it is not to be used, Pt 15-016 should be connected to Pt 17-016 with the aid of a bridge.

2.3. PRODUCT DETECTOR (DJ 4 BG 014)

As has already been described in (7), the gain of this module can be matched to the required application by suitable selection of resistor R-4-014. If no IF amplifier module is to be used, full gain of the product detector will be required. In this case, R 4-014 should be replaced by a bridge. If, on the other hand, the IF module DJ 4 BG 013 is to be used, it is possible for the product detector to be provided with feedback so that the good linearity is even improved. In this case, a value of 100 Ω or more should be used for R 4-014; if the gain is sufficiently high, it is possible for this resistor to be deleted completely.

2.4. IF AMPLIFIER (DJ 4 BG 013)

This module is also in preparation. Full details regarding its application on the system board will be given when it is described.

2.5. SHORTWAVE CONVERTER (DJ 4 BG 011)

This module should be completed for an operating voltage of 12 V according to the original description.

No modifications are necessary to this module for all applications where only manual gain control is required. If this module, on the other hand, is to be partially or completely controlled by a AGC-circuit, which is only possible in conjunction with the IF module, it will be necessary for resistors R 3-011 and R 7-011 to be reduced to 4.7 kΩ. The following conditions will result:

2.5.1. RF STAGE NOT CONTROLLED

Connections Pt 6-011 should be connected to Pt 8-011 at the connector.

2.5.2. MANUAL GAIN CONTROL OF THE RF-STAGE, NO IF MODULE DJ 4 BG 013

A potentiometer of 10 kΩ should be connected between Pt 5-016 and Pt 14-016, with the wiper connected to Pt 8-016. The voltage at Pt 8-016 should then be variable in the range of +12 V (max. gain) and approximately -2 V (T 1-011 completely blocked).

2.5.3. MANUAL GAIN CONTROL OF THE RF-STAGE, WITH IF MODULE DJ 4 BG 013

A manual adjustment of the gain of the RF-stage can be made as previously described. In addition to this, it is possible for the potentiometer to be connected between connections Pt 5-016 and Pt 6-016 with the wiper once again at Pt 8-016. In this case, the voltage at Pt 8-016 is only variable between approx. +5 V and -2 V; the value of R 3-011 should be reduced to 4.7 kΩ in order to achieve maximum gain of the RF stage. With this complement, it is possible for the RF stage to be switched to manual or automatic gain control since the voltage levels now coincide. If this is required, Pt 8-016 can be switched between the wiper of the potentiometer in the manual mode and connection Pt 7-016 for automatic gain control.

2.5.4. AUTOMATIC GAIN CONTROL OF THE RF-STAGE

This mode has been described in the previous section; Pt 8-016 should be connected to Pt 7-016, and R 3-011 should be reduced to 4.7 kΩ.

2.5.5. FIRST IF STAGE UNCONTROLLED

This is only possible when no IF module is provided. In this case Pt 7-016 should be connected to Pt 14-016.

2.5.6. MANUAL GAIN CONTROL OF THE FIRST IF STAGE

This is only possible when no IF module is provided. In this case, a potentiometer of 10 kΩ should be connected as described in Section 2.5.2. between Pt 5-016 and Pt 14-016 with the wiper connected to Pt 7-016. The voltage at this connection should be adjustable in the range of +12 V and -2 V.

2.5.7. AUTOMATIC GAIN CONTROL OF THE FIRST IF STAGE

This is only possible when the IF module DJ 4 BG 013 is provided. The circuit required for this is already provided on the system board and it is only necessary for resistor R 17-011 to be reduced to 4.7 kΩ.

2.5.8. MANUAL GAIN CONTROL OF RF AND IF STAGES

This combination is only possible when no IF module is provided. A potentiometer of 10 kΩ should be connected between Pt 5-016 and Pt 14-016 with the wiper connected to Pt 7-016 and Pt 8-016.

2.5.9. AUTOMATIC GAIN CONTROL OF THE RF AND IF STAGES

This combination is only possible when an IF module DJ 4 BG 013 is provided. Connections Pt 7-016 should be connected to Pt 8-016 and resistors R 3-011 and R 17-011 should be reduced to 4.7 kΩ. The controlled IF stage provided in the IF-module DJ 4 BG 013 is always connected to the automatic gain control.

2.5.10 SELECTION OF THE REQUIRED COMBINATION

As can be seen in the previous sections, the system board offers a large range of possibilities for the gain control. This means that it is possible for the most favourable combination to be selected for each application. However, if one ignores the finer differences, the following combinations provide sufficient scope:

> If no IF module is provided, it will only be possible for manual gain control to be used. In order to achieve the greatest possible dynamic range, both the RF and IF stages of the receive converter DJ 4 BG 011 should be controlled as described in Section 2.5.8.

If an IF module is provided, it is simpler when the controllable stages of module DJ 4 BG 011 are also connected to the AGC voltage as was described in Section 2. 5. 9.

If a manual gain control of the RF stage is required in spite of the good cross-modulation characteristics of the receiver, this can be achieved as described in Section 2. 5. 3.; the IF stage should be connected to the AGC line as described in 2. 5. 7.

3. SYSTEM BOARD DJ 4 BG 016

Figure 4 gives the circuit diagram of the system board. The same power transformer is used as was described with module DK 1 PN 007. A negative voltage of approximately 2 V is obtained with the aid of diodes D 3 to D 5 (016). This voltage can vary in the range of 1.8 V to 2.5 V. It allows the controlled dual-gate MOSFETs of modules DJ 4 BG 011 and 013 to be well blocked so that the whole control range can be utilized. Figure 4 also gives the components that must be connected externally such as loudspeaker, potentiometers etc. As was described in Section 2. 2., the switch connected to connections Pt 15-016 to Pt 17-016 is only required when a CW filter DJ 4 BG 015 is used. The switch connected to Pt 11-016 to Pt 14-016 is provided in order to allow up to three different crystals to be switched into circuit in the product detector. This will allow CW, upper sideband and lower sideband to be selected. If the receiver is only to be utilized for CW transmissions, it is possible for the single crystal to be mounted in module DJ 4 BG 014. In this case, Pt 11-016 or Pt 12-016, or Pt 13-016 should then be connected to Pt 14-016.

As can be seen in Figure 4, the IF connection to the product detector is made with a piece of screened cable from Pt 22-016 (screen to Pt 23-016) or from Pt 25-016 (screen to Pt 26-016) according to whether an IF module is provided or not. For reasons of stability and neutralization of the IF module, this line has not been provided on the PC-board.

Figure 5 shows the conductor lanes and component locations on PC-board DJ 4 BG 016. The dimensions of this board are 170 mm x 120 mm. The on-off switch and fuses are not mounted on the system board since this would limit the possible applications. It will be seen that a protective ground ring is connected to Pt 1-016 that should be connected to the common ground of the unit in order to avoid any leakage currents under unfavourable conditions. Since the PC-board is only coated on one side for price reasons, this protective ring is not provided on the component side. It is, however, possible to provide such a protective ring using a conductive lacquer or by gluing a thin metal foil into place and connecting this to Pt 1-016. Of course, these safety measures are useless when the connection of the wiring between power-line connection, on-off switch, fuse and connections Pt 2-016 and Pt 3-016 are not made with the same care. Attention should be paid that quality, sufficiently rated components are used.

4. MOUNTING OF THE INDIVIDUAL MODULES

In order to accept the 13-pole connectors of the TEKO modules, matching connectors for solder mounting are used. However, since these stand approximately 4 to 5 mm higher than those designed for chassis mounting, it is necessary for

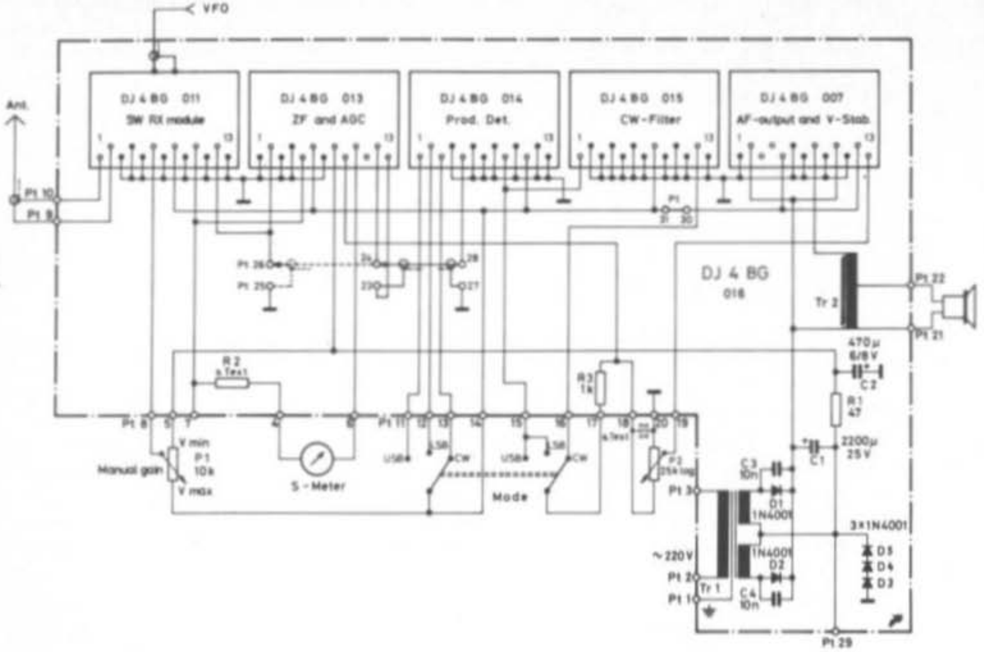


Fig. 4: Circuit of the system board DJ 4 BG 016

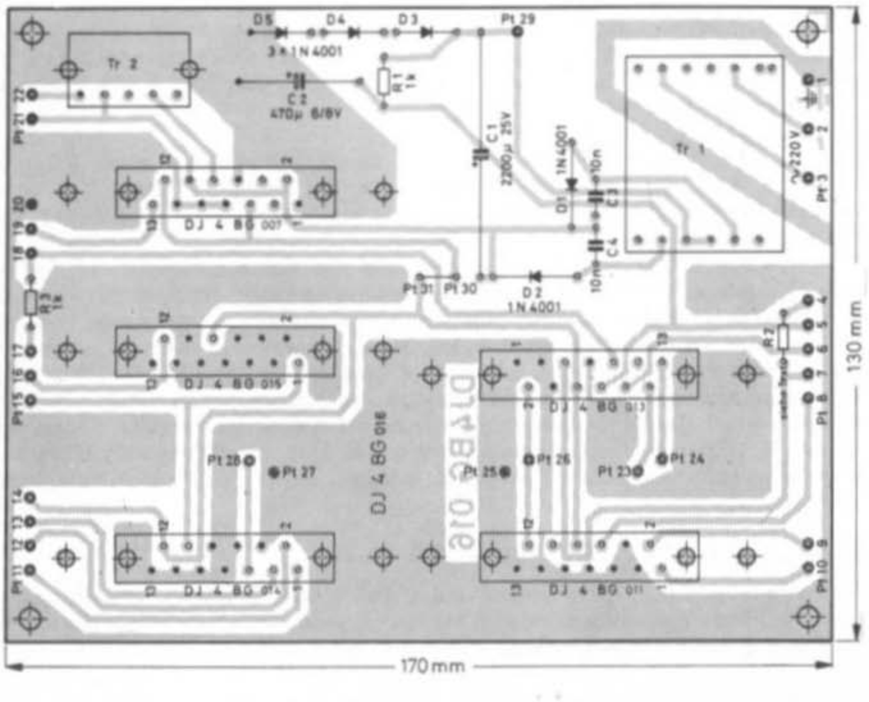


Fig. 5: PC-board DJ 4 BG 016

spacers of approximately 4 mm in length to be used between the system board and the modules. These can be in the form of brass bushings. If this is to be avoided, it is possible to place the individual boards approximately 3 to 4 mm higher in the TEKO boxes. This, of course, is possible when the modules are to be constructed at the same time as the system board. However, if completed modules are already available that were designed for chassis mounting, it will probably be advisable to provide the described spacers rather than to provide the TEKO boxes with additional holes. The connectors designed for normal wiring cannot be used since they possess a different connection pin spacing.

5. PREPARATIONS

The power transformer should be installed on the PC-board after which the voltages are checked; approximately +28 V across C 1-016 under non-load conditions at 220 V AC. Connect module DJ 4 BG 007. The voltage at connection Pt 30-016 should amount to 12 V. This value can be adjusted with the aid of resistor R 13-007. Since a little current is now flowing via diode D 3-016 to D 5-016, a negative voltage of approximately 1.8 V to 2.3 V to ground (Pt 19-016) should be measured at Pt 29-016. The stabilized voltage can be connected to the other modules by making a bridge between Pt 30-016 and Pt 31-016. If these modules have been aligned according to the individual descriptions and if the external components have been connected correctly to the system board, the receiver should be ready for operation as soon as an antenna has been connected. It may be necessary to correct the individual IF circuits carefully for maximum signal, as well as correcting the RF circuits at the centre of the band. If no frequency counter is available in order to establish the required frequency range of the oscillator, (e.g. DJ 4 BG 012), a calibration spectrum generator with switchable calibration lines (e.g. DJ 4 BG 004) can be used.

In this case, the generator is connected via a small capacitance of 1 to 2 pF and the receiver input terminated with the required impedance. The 1 MHz or 500 kHz marker should then be selected in order to find the lower frequency limit. After this has been achieved, the various other calibration lines can be selected in order to mark the upper band limit and to set the bandspread of the oscillator. The scale can then be calibrated using the finer calibration lines (10 kHz, 5 kHz).

6. S-METER

The control voltage is available at connection Pt 7-016 in the automatic gain control mode. This voltage can be indicated on a voltmeter. The meter should have a full-scale deflection of 6 V and the impedance should not be too low. Values of over 2 kΩ/V are sufficiently good. If a mA-meter with a full-scale deflection of less than 0.5 mA is available, it can be connected across resistor R 2-016 at Pt 4-016 on the system board. R 2-016 should be selected so that full-scale deflection is indicated on the meter when the plus connection is made to Pt 6-016 and minus to ground (Pt 19-016). The final connection is made by connecting the minus connection to Pt 4-016 and the plus connection to Pt 6-016. It is then possible for the meter to indicate zero when no signal is being received, and full-scale deflection with the strongest possible signal. Of course, the S-meter can be calibrated with the aid of a signal generator. If such a generator is not available, it is possible for a crystal oscillator to be constructed for the required frequency range that can be calibrated either on a calibrated receiver or wideband oscilloscope and brought to a defined output voltage of say, 100 mV. Relatively simple attenuators can then be used which allow the calibration of the S-meter to be made.

7. EXTENSION OF THE RECEIVER FOR RECEPTION OF FREQUENCY MODULATION

If a crystal filter having a sufficiently wide bandwidth (e. g. XF-9E) is used in module DJ 4 BG 011, it will be possible for frequency-modulated transmissions to be received. In this case, a short, screened RF cable is fed parallel to the IF line between the shortwave receiver and fed to a FM-module DJ 4 BG 008.

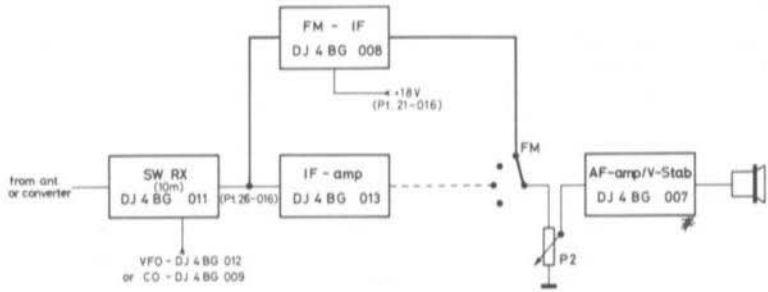


Fig. 6 : Extension for reception of FM transmissions

Since considerably lower bandwidths are advantageous for SSB and CW, a narrower filter can be used in the IF module DJ 4 BG 013. The corresponding filter for CW operation would be type XF-9M (bandwidth 0. 5 kHz), whereas the cheaper XF-9A will be suitable for SSB since the FM filter in the short-wave converter will provide sufficient ultimate selectivity. Figure 6 shows a suitable circuit in the form of a block diagram.

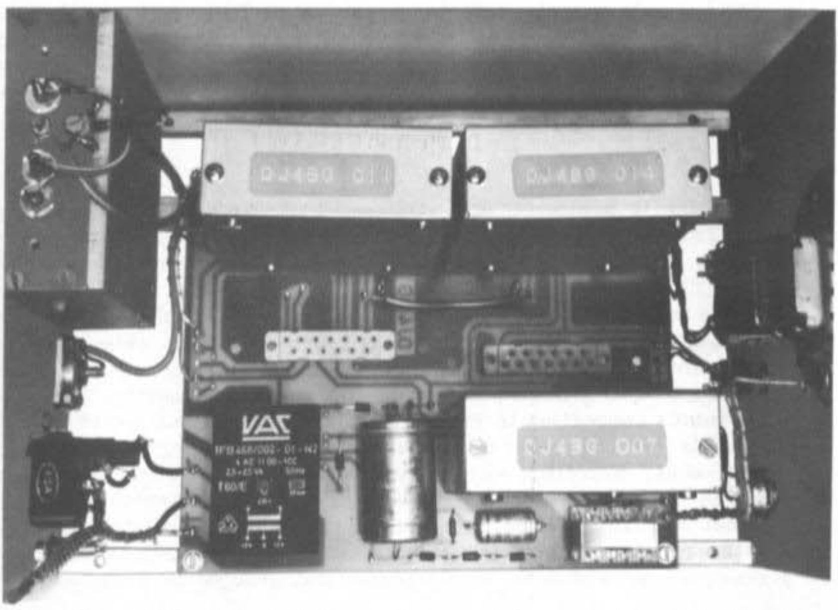


Fig. 7 : Author's prototype of a receiver as shown in Fig. 2a

8. NOTES

The author hopes that the described modules and the system board will form a basis of an efficient receive system. The receiver can be designed for any single shortwave band or together with suitable converters also for a number of VHF-UHF frequencies. The description of such modules together with the possible variations seems to be the most favourable compromise between the purchase of an expensive receiver, which is not usually able to satisfy all the demands placed upon it and the time, cost and experience required for a completely new development. If any of the components, such as integrated circuits, become no longer available, replacement modules will be developed, if possible pin-compatible with the modules that they are to replace.

9. REFERENCES

(1) D. E. Schmitzer: A Modular Receive System
VHF COMMUNICATIONS 3 (1972), Edition 2, Pages 110-114

(2) DJ 4 BG 007: Integrated AF Amplifier and Voltage Stabilizer
VHF COMMUNICATIONS 3 (1972), Edition 1, Pages 34-39

(3) DJ 4 BG 008: FM-IF Amplifier Module
VHF COMMUNICATIONS 4 (1972), Edition 1, Pages 40-45

(4) DJ 4 BG 009: Triple Crystal-Controlled Oscillator
VHF COMMUNICATIONS 4 (1972), Edition 3, Pages 175-179

(5) DJ 4 BG 011: SW Receiver
VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 24-32

(6) DJ 4 BG 012: Universal VFO Module
VHF COMMUNICATIONS 5 (1973), Edition 4, Pages 241-249

(7) DJ 4 BG 014: Product Detector Module with Crystal Oscillators
In this edition of VHF COMMUNICATIONS

(8) DJ 4 BG 004: Modular Calibration System
VHF COMMUNICATIONS 3 (1971), Edition 4, Pages 194-205.

A 500 MHz PRESCALER AND PREAMPLIFIER FOR FREQUENCY COUNTERS

by G. Bergmann, DJ 7 JX and M. Streibel, DJ 5 HD

1. INTRODUCTION

A large number of digital frequency counters are in use in amateur radio stations. The upper frequency limit of frequency counters using TTL-logic is in the order of 50 MHz. Higher input frequencies are possible using ECL-circuits.

Monolithic integrated frequency dividers for frequencies of up to 500 MHz are available from Motorola, Plessey (1 GHz), and Hewlett Packard (hp). A 4 : 1 divider for frequencies up to 1 GHz should also now be available from Fairchild (1). The most economical combination has now be determined that allows measurement of frequencies up to and in excess of the 70 cm band. For cost reasons, the described prescaler uses division ratios of 20 : 1 and 100 : 1.

2. PREAMPLIFIER

The task of the preamplifier is to increase the level of the frequency to be counted to a sufficiently high value. Due to the relationship $I_C \sim \exp U/u_T$ approximately $2.3\ u_T \approx 60$ mV input voltage is required to drive such ECL circuits at low frequencies (collector current ratio 10 : 1). The input voltage requirements increase at higher frequencies mainly due to the voltage drop across the base resistance. With the 2 : 1 divider used, approximately 100 mV (RMS) are required at 500 MHz.

Monolithic integrated circuits cannot be used as preamplifiers for the frequency range up to 500 MHz due to their high cost (more than 200 DM).

Relatively cheap transistors used in UHF antenna amplifiers are very suitable for use in this application.

2.1. POSSIBLE GAIN

If transformation networks are not used between the transistor stages, the following is valid for the gain per stage G_u:

$$G_u = \frac{-y_{21}}{y_{11} + y_{22}}$$

The following values for the four-pole coefficients at 500 MHz are taken from the data sheets of the transistor type 2 N 5179:

$$y_{11} = (13 + j\,12)\ \text{mS} \qquad y_{21} = (20 - j\,35)\ \text{mS}$$
$$y_{12} = (0 - j\,2.25)\ \text{mS} \qquad y_{22} = (2 + j\,5)\ \text{mS}$$

This means that the following is valid:

$$G_u = 1.77\ e^{j\,71.17^{\circ}}$$

The maximum power gain G_p including neutralizing network (unilateral gain) amounts to:

$$G_{p\,max} = \frac{(y_{21} - y_{12})^2}{4(R\{y_{11}\} \times R\{y_{22}\} - R\{y_{12}\} \times R\{y_{21}\})} = 14.16$$

When referred to equal impedances, the maximum gain per stage amounts to:

$$G_{u\,max} = 3.76$$

This shows that only twice the gain per stage would be possible if matching networks were used. The design of such neutralizing and matching networks for wide frequency range would, however, be somewhat complex.

3. ATTENUATOR

The normal circuit using two antiphase diodes at the input of the preamplifier is not sufficient for attenuating the signal. Even the use of Schottky diodes in a low inductance casing is not able to satisfactorily attenuate an input signal from a 50 Ω source.

The bridge-circuit given in Figure 1 has been found to be very successful:

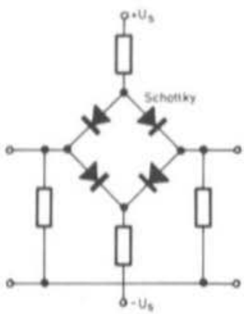


Fig. 1 : Attenuator using a bridge of Schottky diodes

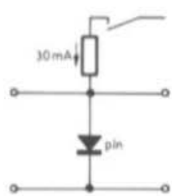


Fig. 2 : PIN-diode switch as attenuator

Due to the forward resistance of the diode, a certain amount of insertion loss is always present. PIN-diode switches have also been found to be successful (Fig. 2). A continuous attenuation as is often used in the input of TV-receivers is not possible in this application due to the large amplitudes that are present (snap-off effect).

The most suitable attenuator is the preamplifier itself. It is only important that the transistors are not able to trigger any parasitic resonant circuits when driven with negative voltages. Such resonant circuits are often used for correcting the frequency response (collector inductances). Figure 3 and Figure 4 clearly show the limiting characteristics as well as the effect of unfavourable collector inductances.

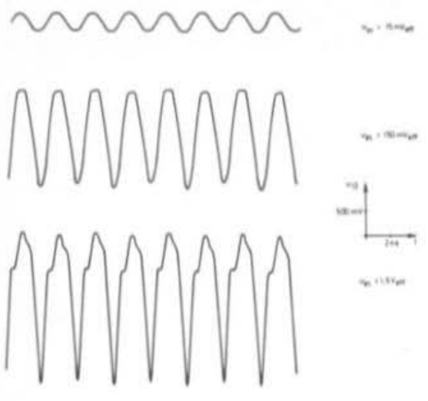


Fig. 3 : Input voltage of the 2 : 1 divider (Pin 13) for demonstration of the limiting of the wideband preamplifier

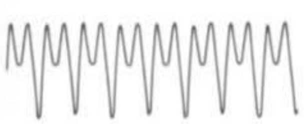


Fig. 4 : As Fig. 3 but using collector chokes (appr. 40 nH) for equalizing the frequency response (incorrect count)

4. CIRCUIT

The circuit diagram of the complete prescaler is given in Figure 5. The input signal is fed to the two-stage preamplifier comprising transistors T 1 and T 2. In order to avoid parasitic resonances, both transistors work into ferrite wideband chokes. Parallel and series feedback is used, which is widespread in the antenna amplifier field. The 2 : 1 frequency divider (I 1) is capacitively coupled to the 10 : 1 divider (I 2). Transistor T 3 is used as level converter from ECL to TTL-logic. According to the position of switch S, the TTL signal is fed via the gate direct to the output (total division ratio 20 : 1), or via integrated circuit I 3 which is connected as a 5 : 1 divider (total division ratio 100 : 1). Another flip-flop (50 MHz) is available at connections Pt 3 and Pt 4, which can be used for dividing the clock frequency by two. Increasing the gate time by two provides an effective division ratio of 10 : 1). Output Pt 7 should be directly connected with the gate of the counter. If it is not possible for the prescaler to be installed within the counter, the prescaler module should be connected via a coaxial cable to the preamplifier input of the counter.

4.1. COMPONENTS

I 1: 1820 - 0558 (hp) or 1820 - 0736 (hp)
I 2: 95 H 90 (Fairchild)
I 3: SN 74 196 N
I 4: SN 7400 N

T 1, T 2: 2 N 5179 (RCA)
T 3: BF 440, BF 414 (AEG-Tfk) or 2 N 5139, BF 272, BSX 29 (Fairchild) or BF 324, BF 450 (Siemens) or similar VHF PNP silicon transistor

L 1 - L 3: Ferrite wideband chokes (1.5 turns in a 6-hole core (Philips)
L 4, L 5, L 6: Ferrite wideband choke 2.5 turns in a 6-hole core (Philips)

Resistors approx. 0.25 W-types, spacing 10 mm

Capacitors: Ceramic disc types; with exception C 15: tantalium drop electrolytic

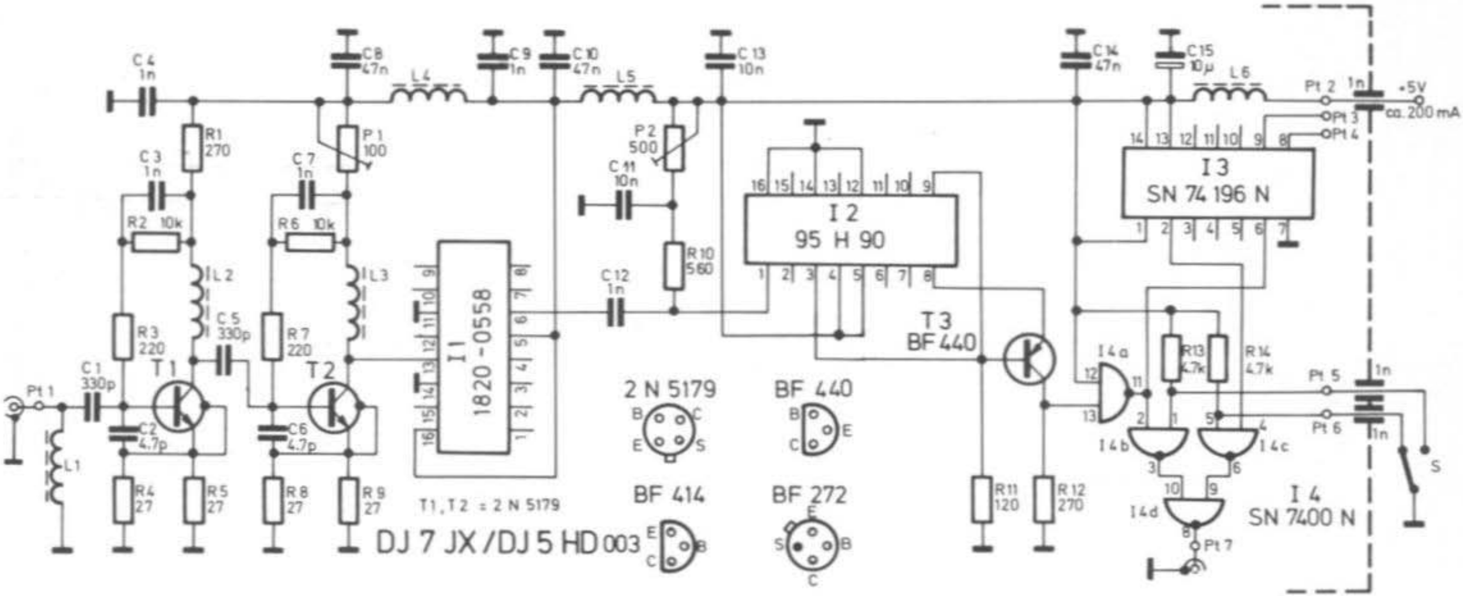


Fig. 5: Circuit diagram of the 500 MHz prescaler

5. MEASURED RESULTS

The input sensitivity of the counter is given in Figure 6. Figure 7 shows the input return loss $r_{in} = -20 \log (b_1/a_1)$ where:

a_1 = amplitude of the signal reflected from the input of the counter
b_1 = amplitude of the signal fed to the input of the counter.

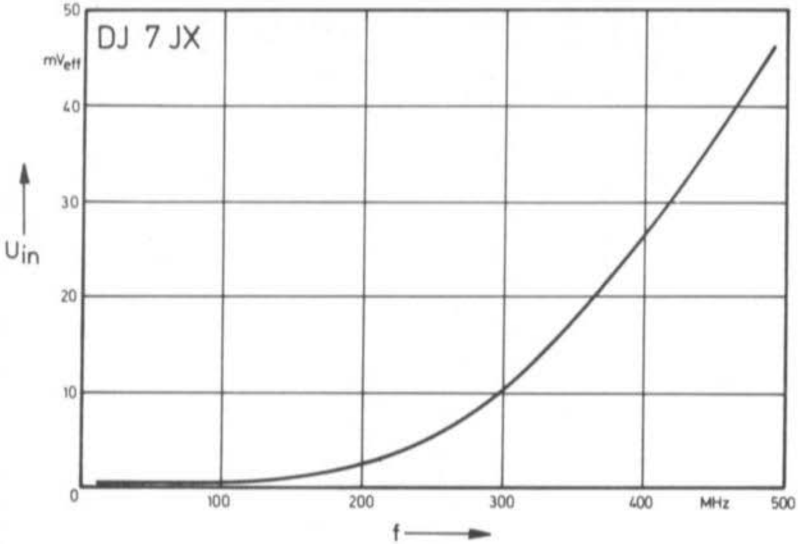


Fig. 6 : Input sensitivity of the prescaler as a function of frequency

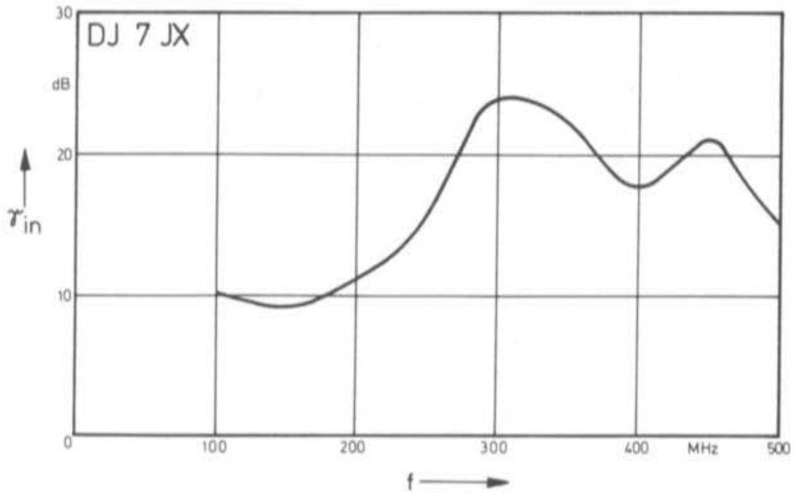


Fig. 7 : Input reflection of the prescaler as a function of frequency

The ripple is $s < 1.9$ for $r_{in} > 10$ dB. Due to the good input matching it is possible for the test signal to be fed to the counter via a relatively long coaxial cable. The input power should not exceed 150 mW which corresponds to a maximum input voltage of approximately 3 V. If frequencies are to be measured at higher power levels it is necessary to provide an attenuator with the required power handlung capability.

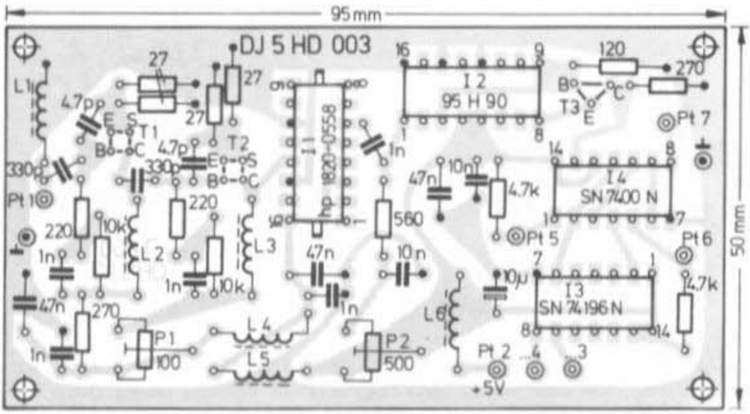


Fig. 8 : Component locations on PC-board DJ 5 HD 003

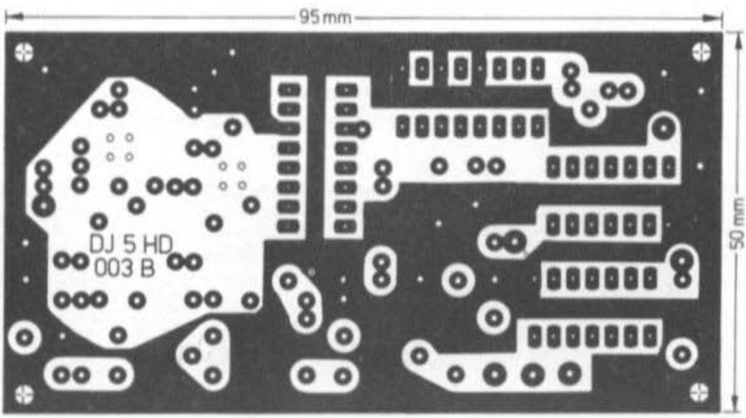


Fig. 9 : Upper side of board DJ 5 HD 003

6. CONSTRUCTION AND ALIGNMENT

The prescaler is accommodated on a double-coated PC-board having the dimensions 95 mm by 50 mm. The board has been designated DJ 5 HD 003. It is advisable for it to be enclosed in a metal box, but this is not absolutely necessary for correct operation. Ground loops must be avoided when installing the prescaler into a frequency counter.The conductor lanes and the component locations are given in Figure 8, and Figure 9 shows the conductor lanes on the upper side of the board. The printed board layout should not be altered; possibly with the exception of making a few wire bridges to interconnect the upper and lower ground-surfaces.

All components must be soldered into place with the shortest possible connections. This is especially important for the preamplifier transistors T 1 and T 2. The 0.8 mm diameter holes for the connection leads of the transistors are drilled as shown in Figure 10 directly adjacent to the ends of the conductor lanes so that the spacing between the leads is maintained as accurately as possible. The transistors are then soldered into place so that the case of the transistor is touching the top of the PC-board.

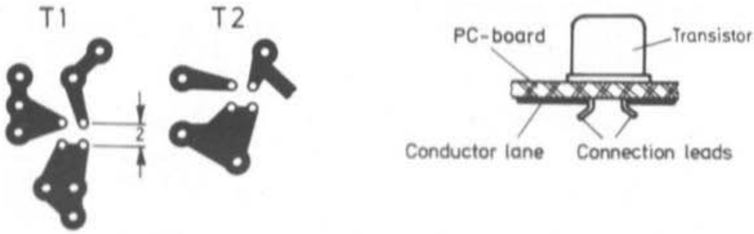


Fig. 10: Installation of transistors T 1 and T 2

Select a voltage of 4.75 V at pin 13 of I 1 with the aid of potentiometer P 1. If this is not possible, this will mean that the current gain of transistor T 2 differs greatly from the nominal values. The changing of the value of R 6 to the next standard value will most certainly be sufficient to achieve this aim.

Since the 2 : 1 divider (I 1) is usually able to count in excess of 600 MHz, it is the 10 : 1 divider (I 2) that determines the upper frequency limit of the module. The upper frequency limit is also effected by the operating temperature of the 95 H 90 which means that it is advisable for a small heat sink in the form of an aluminum bracket to be used that is glued to the top of this integrated circuit.

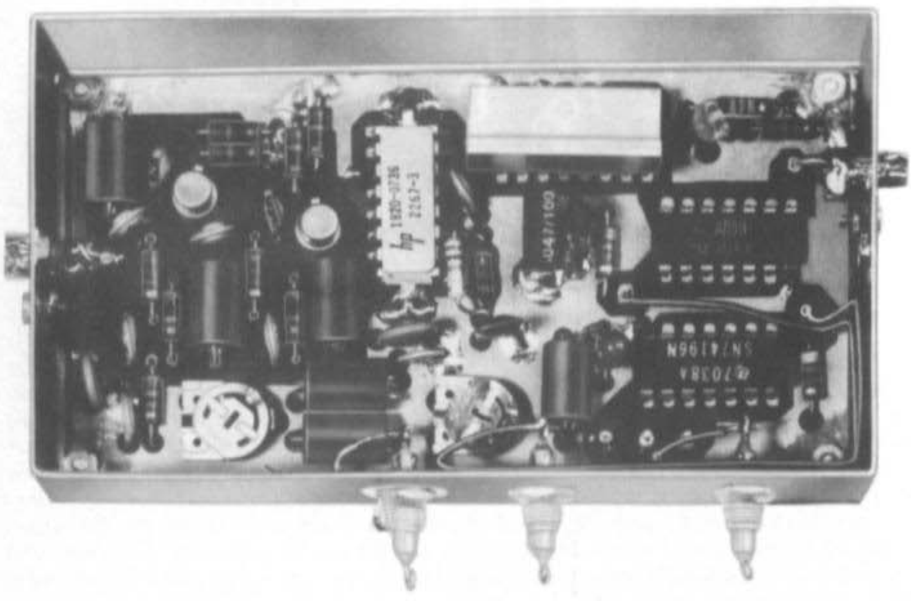


 Fig. 11: Author's prototype of the 500 MHz prescaler

Potentiometers P 1 and P 2 should be adjusted to obtain maximum sensitivity at the highest possible frequency. This is made after reaching the final operating temperature allowing a warm-up period of 2 to 3 minutes. Further alignment is not necessary. Figure 11 shows the author's prototype constructed by DJ 7 JX.

7. REFERENCES

(1) Electronics, February 1974

(2) W. R. Kritter: A 6-digit Frequency Counter for 1 Hz to Typically 100 MHz
VHF COMMUNICATIONS 5 (1973), Edition 2, Pages 95-103

(3) J. Grimm: A 10 : 1 Prescaler and Preamplifier for 250 MHz
VHF COMMUNICATIONS 5 (1973), Edition 3, Pages 154-159.

A STRIPLINE CONVERTER FOR THE 13 cm BAND

by Konrad Hupfer, DJ 1 EE

A converter module is to be described that comprises a preamplifier, push-pull Schottky mixer and intermediate frequency preamplifier. According to the oscillator frequency, the input frequency in the range of 2304 MHz (13 cm band) is converted either to the 2 m band or 10 m band. Two possibilities for provision of the local oscillator frequency are also given.

The receive converter is constructed in microstripline technology and is accommodated on a Teflon (PTFE) glass-fibre board whose dimensions are 95 mm x 65 mm. It is therefore suitable for mounting in a TEKO box 3A. The construction of this module should not cause any difficulties to those having a little experience in the UHF range. The converter provides a noise figure of approx. 5 dB, which according to AMSAT information should allow the 2304 MHz beacon of OSCAR 7 to be heard with a signal-to-noise ratio of 12 dB when used in conjunction with a 27 dB antenna (approx. 1.5 m diameter parabol) and at a receiver bandwidth of 500 Hz.

1. CIRCUIT DESCRIPTION

As can be seen in the circuit diagram given in Figure 1, the input stage is in the form of a common-emitter preamplifier feeding two Schottky-diode mixers via a ring configuration. The local oscillator voltage of 2160 MHz (or 2276 MHz for an IF of 28 to 30 MHz) is also fed via the ring to the diodes. The diodes are connected in parallel for the intermediate frequency with the aid of capacitor C 10. The mixer is followed by a low-noise preamplifier stage tuned to the intermediate frequency. The current flowing via the two diodes can be indicated on a meter.

1.1. THE 2.3 GHz PREAMPLIFIER

Network N 1 transforms the complex input impedance of the first transistor to a real input impedance of 50 Ω. The input signal is amplified in an unneutralized common-emitter circuit and is passed via network N 2 to the mixer. Point 1 of the ring also presents an impedance of 50 Ω. Inductance L 1 represents a short-circuited $\lambda/4$ line at 2.3 GHz, which shorts considerably lower frequencies to ground. Inductances L 2 and L 3 are also in the form of printed $\lambda/4$ striplines; they are used as chokes for the base and collector connections. The operating voltage supply has proved itself both under laboratory and operating conditions.

If the collector current is to be independent of large ambient temperature fluctuations (for example for portable operation), a certain amount of stabilization will be required. One possibility (1) is to connect the other end of potentiometer R 5 to ground and to use a 10 kΩ type instead of 5 kΩ. The resistance value of R 6 should then be increased to several kΩ and a resistor R_Z with a value of approximately 5 kΩ should be inserted. The voltage U_{CE} will then amount to approximately half the operating voltage at a collector current of between 2 and 5 mA. This corresponds to the data given by the manufacturer for low-noise operation.

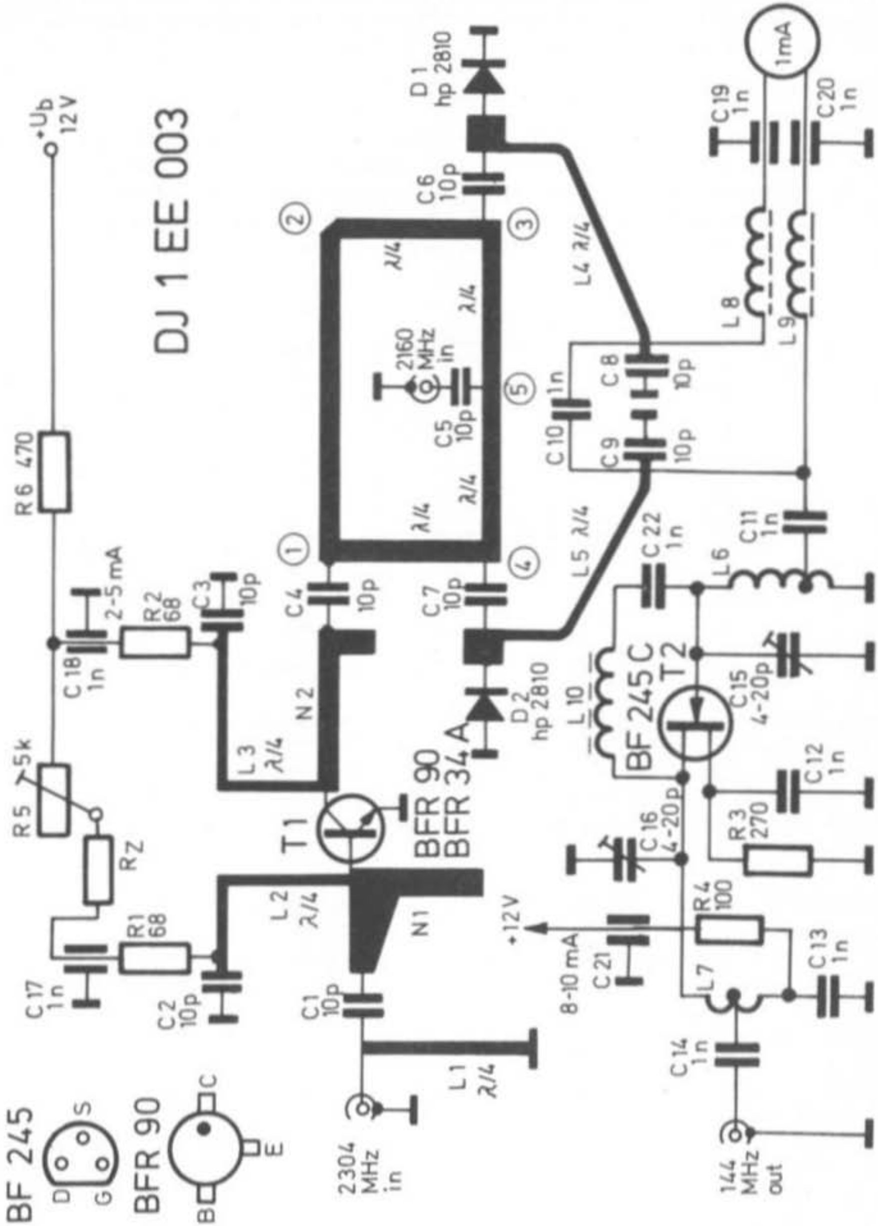


Fig. 1: Circuit diagram of the microstrip converter for 13 cm

The author has used transistor types BFR 90 (Philips) and BFR 34 A (Siemens) in the circuit with equal success. Preliminary measurements have shown the noise figure to be in the order of 5 to 5.5 dB.

1.2. RING CONFIGURATION

The operation of the mixer is to be described with the aid of the diagram given in Figure 2. The task of this ring configuration is to distribute the RF-energy from antenna and local oscillator equally to the two mixer diodes. At the same time, both inputs (1 and 5) must be decoupled from another so that the oscillator voltage is not present at the signal input of the ring and vice versa.

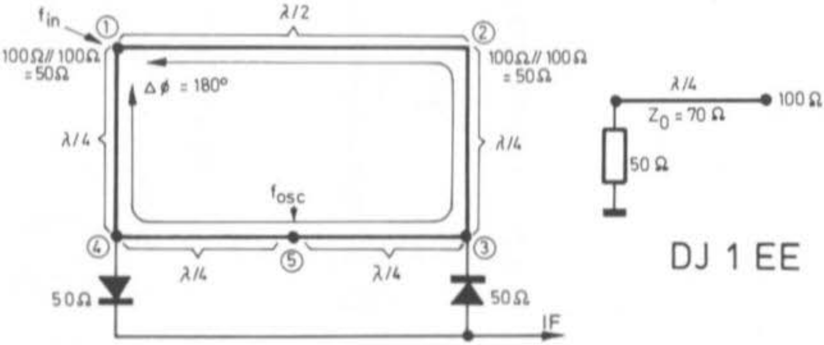


Fig. 2: Stripline length and impedance relationships on the ring

When commencing at the input for the received signal (phase = 0°), this signal is fed to diode D 1 via points 1-2-3 (3 x $\lambda/4 \triangleq 270^\circ$), and from point 1 direct to point 4 and to diode D 2 ($\lambda/4 \triangleq 90^\circ$). This means that the diodes will be provided with voltages that are phase-shifted by $270^\circ - 90^\circ = 180^\circ$. This means that it represents a push-pull circuit in which the diodes are connected at the same polarity to points 3 and 4.

The diodes are also decoupled from another: A voltage arriving at point 4 is fed to diode D 1 via the path 4-5-3 with a phase-shift of 180° and via the path 4-1-2-3 with a phase-shift of 360°. The difference is therefore 180° which means that they are cancelled out at diode D 1. The same is valid for a voltage at point 3 which will appear at point 4 and diode D 2 via two paths with a phase-shift of 180° and will be cancelled.

The decoupling between the UHF input 1 and the local oscillator input 5 is obtained in the same manner: The received signal is provided with a phase-shift of 360° via the path from point 1 via 2 and 3 to 5. It is also fed with a phase-shift of 180° via the path from 1 via 4 to 5. Both these voltages are therefore cancelled out at point 5 which means that the input signal is only fed to the two mixer diodes. The same is valid for the decoupling of point 5 from the signal input 1.

Since the ring can only be dimensioned exactly for one frequency, and since the signal delay cannot be maintained exactly due to manufacturing tolerances and component connections, the decoupling is not infinite. Practically, a decoupling of 20 to 25 dB can be achieved within a narrow band of 5 to 10% of the design frequency. This means that the ring is suitable both for the input frequency of 2300 MHz and the oscillator frequency of 2160 MHz.

1.3. IF-PREAMPLIFIER

The two $\lambda/4$ lines L 4 and L 5 ensure that no SHF energy is passed to the IF-amplifier. These lines have no effect at the intermediate frequency of 144 MHz or 28 MHz so that it is possible for the two IF voltage sources to be connected in parallel via capacitor C 10. Due to the low gain of the SHF preamplifier and the conversion loss of the diode mixer, the IF-preamplifier also has a contribution of the overall noise of the converter. For this reason, a very low-noise FET is used in a neutralized common-source circuit. Since the converter was to be used for mountain-top portable operation, the author's version was for an intermediate frequency of 144 MHz. The resonant circuits used for this intermediate frequency are in the form of air-spaced inductances (L 6 and L 7) providing a bandwidth of more than 2 MHz without additional damping. For an intermediate frequency in the 10 m band, inductances with core and possibly additional damping (as well as screening cans) should be used. The chokes L 8 and L 9 should block the intermediate frequency; this means that the inductance must amount to at least 5 μH at an IF of 28 MHz.

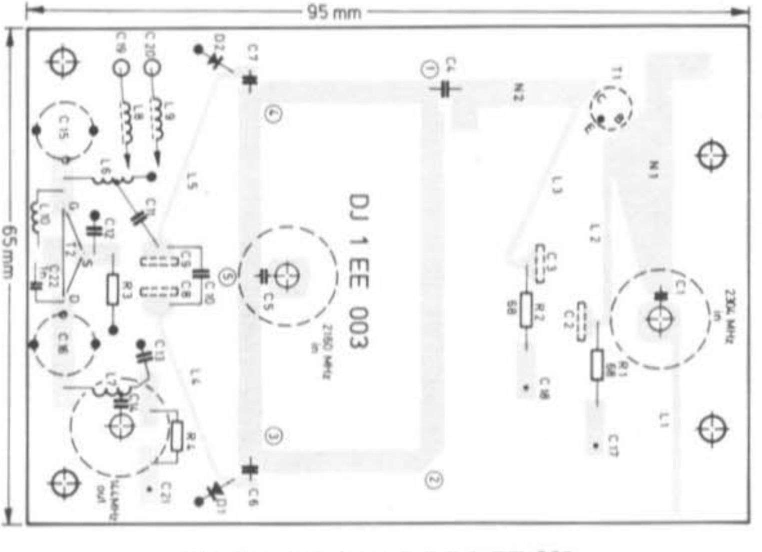


Fig. 3: PC-board DJ 1 EE 003

2. CONSTRUCTION

The SHF portion comprising preamplifier and mixer as well as the IF preamplifier are accommodated on a double-coated Teflon (PTFE) glass-fibre board. Figure 3 shows the PC-board which has the dimensions 95 mm x 65 mm and has been designated DJ 1 EE 003. This board offers the advantage of a trouble-free construction without tiresome mechanical work. The Q is lower than could be obtained with coaxial circuits with air dielectric, however, this is not important due to the short line length. The loss of sensitivity due to this probably will not exceed 5%.

The ground surface of the board is not etched; this means that striplines with a defined impedance can be made. The calculation is given in the appendix. The construction is now to be explained based on the finished PC-board.

2.1. DRILLING

The PC-board must be provided with holes for the inner conductors of the three coaxial sockets (BNC), for the ground connections of trimmers C 15 and C 16 as well as for the components: (emitter of) T 1, D 1, D 2, L 6, C 12, C 13 and R 3; furthermore, for the five feedthrough capacitors. The four mounting holes are only given as a recommendation; it would be better for the PC-board to be mounted in the case with the aid of the three connectors. A small slot is also required to ground one end of L 1 with the aid of a small metal strip. Finally, four other slots of 1 mm in width and 5 mm long are required for capacitors C 2, C 3, C 8 and C 9. No slots are necessary for the coupling capacitors since they are provided with small brackets on the PC-board in order not to break the ground surface.

2.2. SOLDER CONNECTIONS ON THE GROUND SURFACE

The only components that must be soldered to this surface are the feedthrough capacitors, the ground end of the components given in Section 2.1., one surface of the four chip capacitors C 2, C 3, C 8 and C 9, as well as the nuts of the three BNC connectors. The ground surface should be removed up to a distance of 1 mm around the holes for the inner conductor of the connectors so

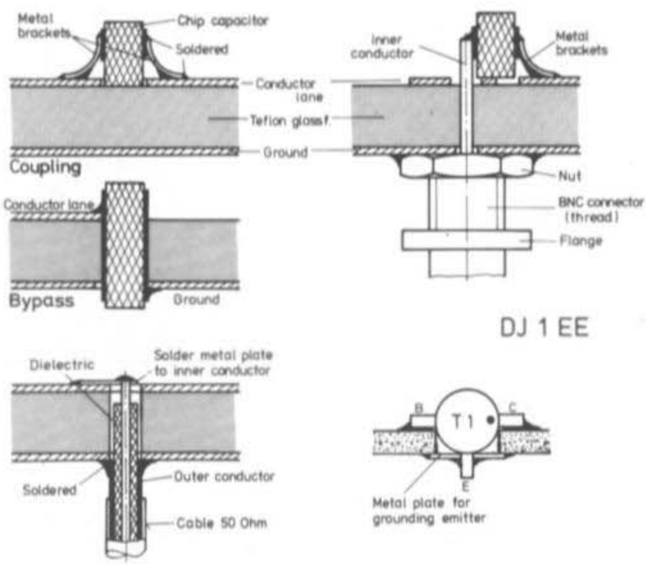


Fig. 4: Constructional details for stripline connections at UHF/SHF

that no shortcircuit can take place. The Teflon (PTFE) insulation and inner conductor of the connectors should be shortened as required. Of course, coaxial cable can be directly connected as shown in Figure 4 instead of the BNC-connectors.

2.3. CONNECTION ON THE CONDUCTOR SIDE

As has already been mentioned, the end of L 1 is grounded with the aid of a narrow, thin metal strip. The four bypass capacitors are placed into the slots provided, one surface soldered to the appropriate $\lambda/4$ line and the other to the ground surface. This is followed by inserting the BNC-connectors which are then screwed into place. The coupling capacitors C 1, C 4, C 5, C 6 and C 7 are soldered into place with the aid of small metal brackets of 3 to 4 mm in width and 2 to 3 mm in length. Figure 4 gives two drawings showing the connection of the coupling capacitors. Disc or drop-type capacitors can be used instead of the described chip capacitors if the protective lacquer and connection wires have been removed. The value of these capacitors should be in the order of 8 to 12 pF. The same is valid for the bypass capacitors.

Transistor T 1 and the two diodes are soldered into place with the shortest possible connections. The least possible amount of solder should be used for all UHF connections.

3. COMPONENTS

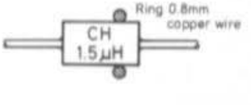
T 1: BFR 90 (Philips) or BFR 34 A (Siemens)
T 2: BF 245 B or C (TI), W 245 (Siliconix), MPF 102 (Motorola)

D 1, D 2: HP 2800 or 2804 (paired), HP 2810 or 2811 (Hewlett-Packard) or similar Schottky diodes (not very critical due to preamplifier stage); however, diodes HP 2817 or 2818 (paired) selected for SHF would be better.

C 1 - C 9: 8 to 12 pF chip capacitors or small disc (drop) types.
C 10 - C 14 and C 22: 680 pF to 1.2 nF ceramic disc capacitors.
C 15, C 16: Ceramic or plastic-foil trimmer 7 mm dia. approx. 20 pF max. cap.
C 17 - C 21: 500 pF to 2 nF feedthrough capacitors of max. 4 mm dia. for solder mounting.

L 6, L 7: 5.5 turns of 0.8 to 1 mm dia. (18-20 AWG) silver-plated copper wire wound on a 6 mm dia. former, self-supporting, coil length approx. 8 mm, coil tap 1 turn.
L 8, L 9: 1 μH (uncritical) miniature ferrite choke.
L 10: approx. 1.5 μH miniature ferrite choke.

Resistors for 1 mm spacing
3 BNC connectors for single-hole mounting.
1 TEKO box 3 A
1 1-ma meter.



4. ALIGNMENT

Figure 5 shows a prototype of the completed 13 cm converter. It is only necessary for the IF preamplifier to be aligned. The two trimmers should be aligned for maximum noise indication on the subsequent receiver when tuned to the centre frequency. If the amplifier stage should break into oscillation, it is necessary for it to be neutralized with the aid of a wire ring of approximately 0.8 mm dia. mounted on choke L 10 as shown above.

After testing the transistor currents, it is possible for a local oscillator to be connected that is able to provide a power of 1 to 2 mW. It is possible for the diodes to be checked individually by bridging them alternately with a 100 Ω resistor (capless resistor with short leads); the reduction of the indicated current should coincide to within approximately ± 10% in both cases.

A weak test signal should now be generated at 2304 MHz. This can be generated with the aid of a 57.6 MHz crystal (for 70 cm to 2 m converters). The frequency 57.6 MHz is multiplied firstly by 10 and finally by 4.

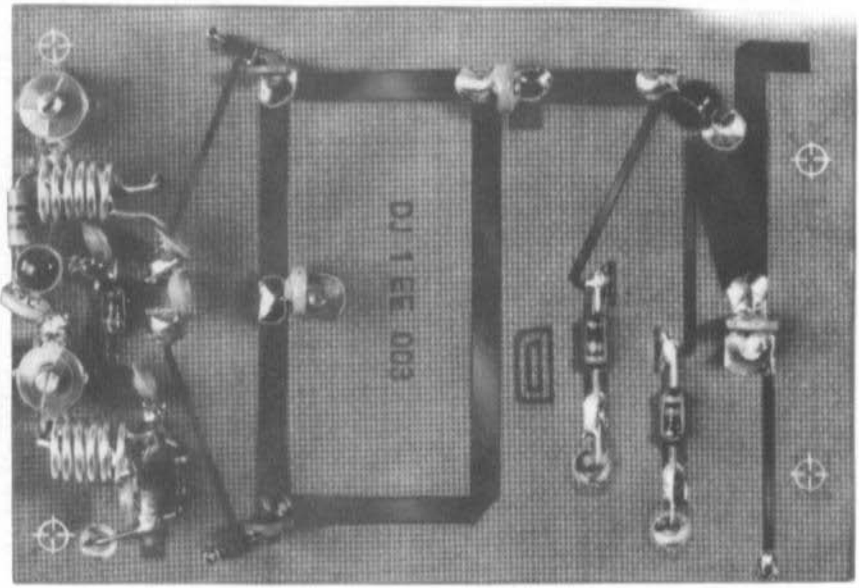


Fig. 5 : Author's prototype of the 13 cm converter DJ 1 EE 003

5. OSCILLATOR PORTION

There are two possibilities for obtaining the required local-oscillator frequency:

A 27 MHz oscillator is used in conjunction with a module DJ 4 LB 003 to multiply the frequency by five (T 3 with grounded emitter), doubling in T 302, doubling again using a BF 223 for T 303 and finally amplifying the resulting frequency of 540 MHz with the aid of a BFR 90 as T 304 to an output of approx. 50 mW. For this application, capacitors C 307 and C 310, as well as trimmers C 316 and C 321 are deleted; L 306 should be increased to 3 turns and R 310 should be decreased to a more favourable value. This signal can be fed to a Schottky-diode multiplier with subsequent interdigital filter as described in (2). The interdigital filter ensures that the required frequency of 2160 MHz is provided. The following article is to describe another suitable oscillator. The author is in the process of designing a matching local oscillator for this application.

6. APPENDIX

The geometrical dimensions of the ring are to be calculated so that it is possible for similar converters to be designed for other frequencies. The power matching described in Section 1.2. results in the ring consisting of a closed $\sqrt{2} \times 50\,\Omega$ line. This $70\,\Omega$ line has a length of $2 \times \lambda/2 + 2 \times \lambda/4 = 1.5\,\lambda$. The width of the printed stripline can be taken from the curves given in (4) for determining the impedance. With a substrate thickness of 1.6 mm and a Teflon (PTFE) glass-fibre dielectric of $\varepsilon_r = 2.2$, a ratio of width to thickness of $w/h = 1.9$ results. A stripline width of $1.9 \times 1.6 = 3.04$ mm results for the $70\,\Omega$ line.

All printed lines are shortened to the value of $\sqrt{\varepsilon_{r\,mean}}$. This effective dielectric constant $\varepsilon_{r\,mean}$ is less than ε_r. It is also dependent on w/h; this means that the velocity factor λ_o/λ_{st} is dependent on w/h. It is possible to establish from the given curves that λ_o/λ_{st} amounts to 1.35 for a ratio $w/h = 1.9$. The geometrical length of the $\lambda/4$ lines of the ring thus result as:

$$1\ \lambda/4 = \frac{300\ 000}{2304 \times 4 \times 1.35} = 24\ \text{mm}$$

7. REFERENCES

(1) K. Richter: Design DC Stability into your Transistor Circuits
Microwaves Vol. 12, No. 12 (Dec. 1973), Pages 40-46

(2) R. E. Fisher: Interdigital Converters for 1296 and 2304 MHz
QST Vol. 58, No. 1 (Jan. 74), Pages 11-15

(3) K. Hupfer: Striplines for VHF and UHF
VHF COMMUNICATIONS 3 (1971), Edition 4, Pages 207-216.

2160 MHz LOCAL OSCILLATOR FOR 13 cm CONVERTERS

by Konrad Hupfer, DJ 1 EE

A local oscillator module is to be described that is suitable for use with the converter described in (1). A matching local oscillator module for the converter is still under development and will be described later. However, this local oscillator module will allow the converter to be operational until the final, PC-board local oscillator module is described. The conception is similar to that given in (2) with respect to the frequency plan and multiplication with the aid of a Schottky diode.

The circuit diagram of the frequency multiplication up to 540 MHz is given in Figure 1. Figure 2 shows a drawing of the Schottky-diode multiplier using a shortened $\lambda/2$ coaxial resonator which multiplies the frequency by four. A lower frequency crystal could be used if a further stage was added, however, this would increase the danger of unwanted conversion products. The BFW 16 A (Philips) transistor has been found very suitable for use in the 540 MHz output stage; it provides a minimum output power of 50 mW at 540 MHz. The inductance L 3 consists of a piece of silver-plated copper tape, whose dimensions are not critical due to the large variation range of the trimmer (approximately 25 mm long, 4 mm wide).

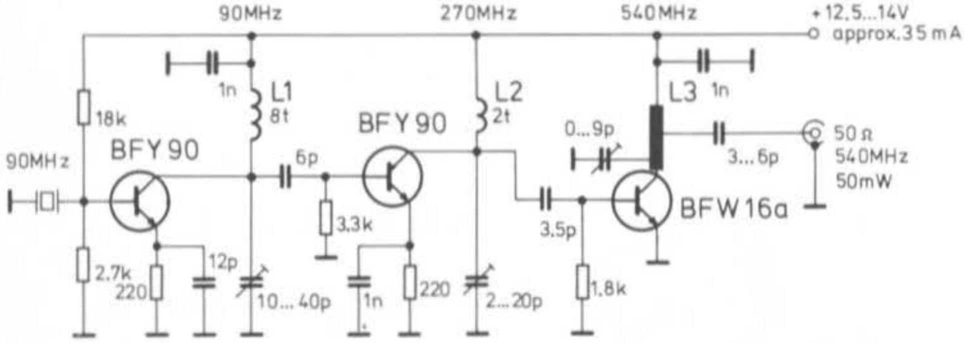


Fig. 1: Circuit diagram of a local oscillator module for 13 cm converters

The Schottky multiplier uses a type HP 2810 and is constructed as a separate module together with the matching circuit so that it is possible for each module to be aligned separately and to be modified as required. The construction of the Schottky multiplier and matching circuit is somewhat more critical, not with respect to the resonance at 2160 MHz which can easily be achieved with the given dimensions, but more with respect to the miniature components that must be used. The adjustment of the two trimmers of the matching circuit is somewhat critical since this part of the circuit is at high impedance. However, the spacing and design of the coupling link for the Schottky diode within the resonator is not very critical.

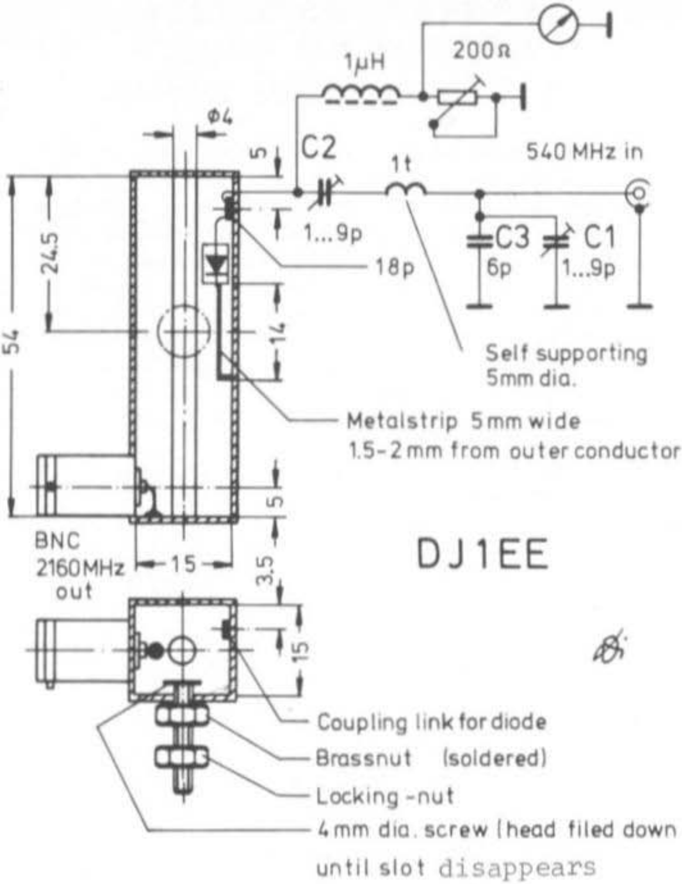


Fig. 2: Frequency multiplier and resonator for 2160 MHz

After the oscillator module has been aligned for 50 mW into 50 Ω, it is possible to connect it to the matching circuit of the multiplier. Of course, at 540 MHz, attention must be paid to keep the connections as short as possible. Trimmer capacitors C 1 and C 2 are aligned for maximum diode current which can be measured as the voltage drop across the bias resistor. The resonator is aligned to 2160 MHz with the aid of an absorption wavemeter (such as will be described in one of the next editions of VHF COMMUNICATIONS). The output power at this frequency amounts to approximately 1.5 to 2 mW, which should be sufficient to produce a diode current of 1 to 2 mA in the mixer of DJ 1 EE 003.

(1) K. Hupfer: A Stripline Converter for the 13 cm Band
In this edition of VHF COMMUNICATIONS

(2) R. E. Fisher: Interdigital Converters for 1296 MHz and 2304 MHz
QST Vol. 58, No. 1 (Jan. 74), Pages 11-15.

ANTENNA NOTEBOOK: ANTENNAS FOR OSCAR 7

by T. Bittan, G3JVQ/DJØBQ

Antennas are required for the 10 m, 2 m and 70 cm bands if both repeaters of OSCAR 7 are to be used. Suitable antennas are to be discussed for each of the repeaters. The major factor affecting the selection of a suitable antenna is the available output power from the transmitter. Operation is greatly simplified when high output power levels are available so that omnidirectional antennas can be used. However, this is not possible in many cases due to the non-availability of such linear amplifiers, or due to licence regulations. Since each of the repeaters requires a certain effective radiated power (ERP) for operation, the energy not provided by the transmitter must be compensated for by the gain of the antenna. If, for instance, an ERP of 100 W is required and only 10 W of output power is available, an antenna gain of 10 dB will be required.

Of course, an antenna having such a gain of 10 dB would have to track the satellite, since the beamwidths would be in the order of 45°, and no operation would be possible over the satellite once it has left the main lobe of the antenna. This means that some means of rotating and tilting the antenna must be provided. Such a simple, inexpensive tracking arrangement is to be described.

1. 145/29 MHz REPEATER

According to AMSAT information given in (1), an ERP of 80 W to 100 W is required for the two metre transmit signal. If this 80 W to 100 W is available from the transmitter, the most favourable antenna will be a crossed dipole (turnstyle) which not only provides a completely omnidirectional characteristic for linear polarisation, but also provides elliptical, and circular polarisation according to the position of the satellite with respect to the antenna: horizontal polarisation at the horizon and circular polarisation when directly overhead. The required sense of circular can be selected or switched as was described in (2) and (3).

If such an output power level is not available, a more directional antenna will be required as well as some form of antenna tracking. A switchable crossed Yagi antenna represents the most favourable antenna in this case.

The best antenna for reception of the 10 m output signal would also be a turnstyle antenna that can be switched to either circular or linear polarisation. A 28 MHz Yagi is at a disadvantage except when the satellite is at the horizon.

2. 432/145 MHz REPEATER

The required ERP for this repeater is given as being 400 W in (1). However, practical operation over the satellite has shown that the required ERP is in the order of 200 W at 432 MHz. Since relatively few stations have such an output power available on the 70 cm band, it will be necessary for a directional antenna and antenna tracking to be used. Once again, circular polarisation represents the best mode. The author recommends that the only switching that should be made with a 70 cm crossed Yagi is between clockwise and anticlockwise circular polarisation, since inhomgenities of the cables do not allow the switching of the four polarisation modes in the shack using two equal-length feeders.

Furthermore, the losses encountered using multiple coaxial relays for switching the polarisation at the antenna will be greater than the 2 to 3 dB loss due to the reception of linear polarisation on a circular-polarised antenna. The required switching between clockwise and anticlockwise circular polarisation can be made using a single coaxial relay and $\lambda/2$ delay line in one arm of the phasing cable as described in (3) and (4).

The same antenna can be used for reception of the 2 metre output band as was recommended for transmission using the 2 m/10 m repeater.

3. PRACTICAL ANTENNAS

Not all amateurs have sufficient space available for mounting several antennas. For this reason, methods of reducing the dimensions of such arrays are to be discussed. The 10 m crossed dipole is usually not tilted but is mounted stationary on the mast below the rotator. The author recommends the use of helical elements of say 1 to 1.5 cm in diameter wound on a glass-fibre or PVC-tube. It is possible to reduce the mechanical length of a resonant $\lambda/2$ element down to less than 3 metres in length. A total of approximately 350 turns are required for each dipole element. A suitable mechanical length of such a 10 m dipole would be in the order of 2.5 m.

It is possible to combine the 2 m and 70 cm antenna on a single boom. The most favourable construction would be for a 70 cm crossed Yagi to be mounted in front of a 2 m crossed Yagi. Separate feeders are used for each of the antennas and they will have little effect on another. Since no such antenna combinations are available on the market, it is necessary for the individual amateur to construct such mechanical mounting arrangements to join the antennas.

4. A SIMPLE SATELLITE TRACKING SYSTEM

The author has constructed a relatively simple tracking system for 70 cm/2 m that can be constructed simply, and which is also very suitable for normal terrestial communications. A similar method was described in (5). The actual antennas used are a Jaybeam 8XY/2 m crossed Yagi for 2 m (approx. 11 dB gain) and the new 12XY/70 cm crossed Yagi for 70 cm (13 dB). The 2 m crossed Yagi is switched as given in Figure 2 of (3) for four polarisation modes; the 70 cm crossed Yagi is only switchable between clockwise and anticlockwise circular polarisation. They are mounted side-by-side on a common glass-fibre

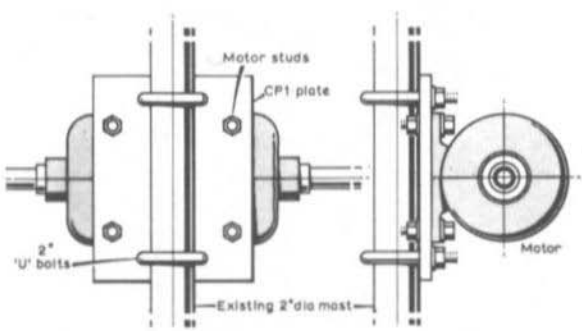


Fig. 1: Mounting the Stolle rotator horizontally on a mast

tube of 38 mm in diameter. An existing CDE-rotator is used for normal azimuth rotation. The vertical tilting is carried out by fixing the horizontal glass-fibre tube between the two crossed Yagis so that it passes through the Stolle automatic rotator that takes masts of upto 38 mm (1.5") diameter. The actual mounting of the Stolle rotator to the main mast is made as shown in Figure 1 using a standard CP 1 crossover plate offered in the Jaybeam range. It is only necessary for the holes in the crossover-plate to be filed slightly in order to mount the rotator. One modification is required on the Stolle rotator: The mast end stop of the rotator must be removed, or a slot cut in the cross over plate. The antennas have been mounted so that the indication "W" is the normal horizontal position, "N" is completely vertical, and "E" is once again horizontal to the rear. The control unit has not been modified but it would be possible to bridge the potentiometer of the control unit so that the whole range of "S" to "S" is used to cover the range horizontal-vertical-horizontal. However, this is not considered necessary.

The glass-fibre tube between the two antennas has not been supported except in the Stolle rotator. However, it is advisable to use a short "T" piece of metal tubing and use two Stolle alignment bearings RZ 100 to take some of the load from the rotator.

This antenna array has been in use for over three months and has proved to be very successful for normal terrestial communications where the vertical tilting sometimes offers over 3 dB of extra signal to be achieved by pointing the antenna slightly above or below the horizon. This means that such a tilting arrangement is extremely suitable for mountainous areas where the most favourable reflection or refraction of the signal can be selected.

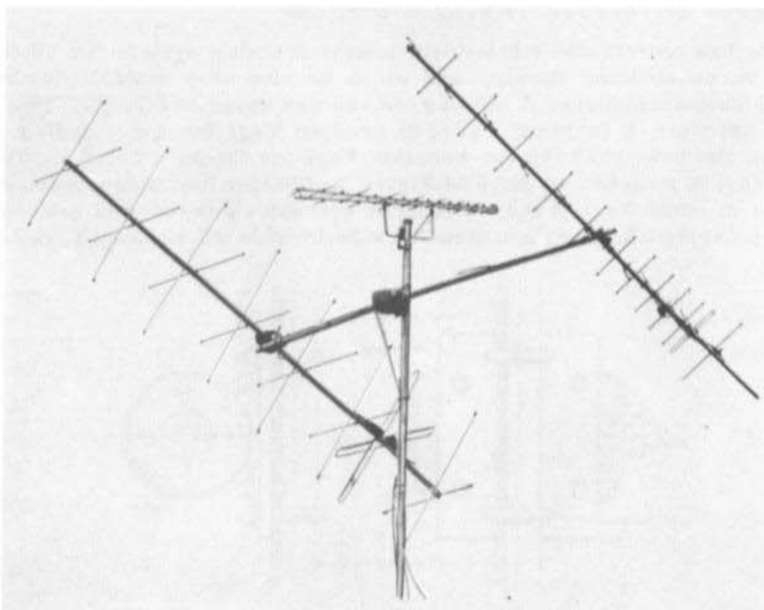


Fig. 2: Photograph of the author's antenna array
(uppermast antenna is a log-periodic for 400-1000 MHz)

Using the described array, it has been possible for the author to operate over the 432-145 MHz repeater with only an output power of approximately 6 W at the antenna. This corresponds to an ERP of only approximately 180 W, which means that the 70 cm crossed Yagi must carefully track the satellite. When the satellite is in the main beam it allows the authors own signal to be heard with 35 dB to 40 dB over noise (S 6 - S 7). All Europe as well as W 1, W 2, W 4, VE 2 and VE 3 have been heard at good signal strengths. Unfortunately, due to pressure of work the author has not been able to operate more than a few times over the satellite.

5. REFERENCES

(1) AMSAT: Amateur Radio Satellite OSCAR 7
VHF COMMUNICATIONS 6, Edition 1/1974, Pages 50-56

(2) T. Bittan: Antenna Notebook
VHF COMMUNICATIONS 5, Edition 4/1973, Pages 220-223

(3) T. Bittan: Antenna Notebook
VHF COMMUNICATIONS 6, Edition 1/1974, Pages 38-41

(4) T. Bittan: Circular Polarisation on 2 Metres
VHF COMMUNICATIONS 5, Edition 2/1973, Pages 104-109

(5) R. A. Ham: Tilting with the Stolle Rotator
Radio Communication, Vol. 49, No. 8, Page 539

NOTES AND MODIFICATIONS

1. 2 m/70 cm LINEAR TRANSVERTER DC 8 NR

Several enquiries have been received regarding obtaining too low an output power from the second transmit mixer: Since there is generally sufficient power available in the drain circuit comprising L 21, C 66 and C 67, the cause is probably that the coupling to the push-pull circuit (L 22/C 70) is too loose. Inductances L 21 and L 22 should nearly touch another. Since trimmer capacitor C 70 is virtually at minimum capacitance in the author's prototype, it is possible that the circuit will not be able to be brought to resonance under all conditions. In this case, the diameter of inductance L 22 should be reduced to 4.5 mm or 5 mm diameter. The number of turns should be maintained. It may be sufficient to increase the spacing between turns. The same is valid for inductances L 8 and L 9.

A loosely-coupled calibrated absorption wavemeter is more suitable for alignment than a low-impedance RF-probe and millivoltmeter, since the latter would load the circuits too greatly. If the sensitivity of the indication is not sufficient, it is possible for the tube-amplifier also to be used.

Resistors of 47 Ω should be placed in series with chokes Ch 6 and Ch 8 in the collector circuits of transistors T 12 and T 13. The potentiometer P 1 should be separated from R 3 and directly connected to the positive pole of the operating voltage. Since the given ceramic tubular trimmers are soon worn after long alignment processes, the author would like to suggest the Philips type 2222 802-20002 (6 pF) as a replacement.

The numbers of turns for inductance L 14 should be reduced from 2.5 to 1.5 turns (from PA ø GMS)

2. DUAL-INPUT PREAMPLIFIER AND 2 : 1 PRESCALER DL 8 TM 003

As has already been mentioned, the frequency counter often indicates a one instead of a zero. A rather radical method of suppressing has already been recommenced. This condition occurs when the output of the 2 : 1 divider 74 S 112 possesses logic-H level; it can be avoided when the divider is reset to logic L after each count. In order to achieve this, the reset inputs of the dual flip-flop must be fed with H-level during the count and receive a short L-level pulse at the end of the counting cycle. The reset pulse of the first decade counter (74196) is suitable for this.

Modifications: Disconnect pins 13 and 15 of I 302 from the operating voltage and connect them via a bridge to pin 13 of I 216 (74196). (from DC 8 UE)

3. MODULAR ATV TRANSMITTER DJ 4 LB 004

A group of ATV amateurs in Cologne, W.Germany, are using a Schottky diode ring mixer in module DJ 4 LB 004 instead of transistors T 401 and T 402. The output is connected via 100 pF to L 402 (0.5 turns from the cold end). An original oscillator module DJ 4 LB 003 is used and an output power of 25 mW is easily obtained when aligned on a sweep generator, inspite of the insertion loss of at least 6 dB. Advantages: uncritical alignment and operation as well as a continuously high carrier suppression. A ring modulator manufactured by TEKO was used that is not so high but wider and longer than the wellknown types SRA-1 and IE-500, which would be just as suitable.

4. PORTABLE SSB TRANSCEIVER FOR 144-146 MHz

Due to the very compact dimensions of this transceiver, a large number of recommended modifications have been received that are often in conflict with another and do not seem to be necessary in all cases. Since it is mainly the resonant circuits that are referred to, it seems that the resonant frequencies vary. Due to the compact dimensions, the differing ground connections and screening panels of the constructor seem to have a different effect in each case. The editors can only suggest that the constructor try the various modifications himself.

4.1. MODULE DC 6 HL 001

The gain of the stage (T 113) following the crystal filter is too great. This can be cured by providing an emitter feedback by removing C 173. It is still possible to tune the circuit with L 113 without reduction of the output power. The following measured values result after removing C 173 and removing the core of L 113 (measured on a Tektronix oscilloscope 485 with 350 MHz bandwidth): The AF-voltage at the input of the balanced modulator should not exceed 150 mV (peak-to-peak) otherwise transistor T 113 will limit. A peak-to-peak voltage of 3 V will be present at the collector of this transistor. If the level is increased, limiting will take place in one of the following stages. It is advisable to balance the balanced modulator (D 120-D 123) not only with respect to amplitude but also with respect to phase. The carrier suppression can be increased by 10 to 20 dB so that the long-term suppression is in the order of 50 dB to 60 dB. This is achieved by connecting a fixed capacitor of 8 pF to one end of L 111 and a 3-13 pF trimmer between the other end and ground (from DK 2 GU).

Diodes D 120 - D 123 should be selected so that they possess the same forward resistance. This improves the carrier suppression considerably.
(from PA Ø GMS)

The screening panels should not have any electrical contact to the case of the crystal filter, since this falsifies the filter grounding via the mounting screws and will cause deterioration of the filter curves. (from PA Ø GMS)

When the absorption circuit comprising L 121 and C 184 cannot be resonated, three turns should be removed from L 121 and the value of C 184 reduced to 3 pF. It is then possible to reduce the indication of a sensitive SWR-meter to zero (switch off 9 MHz oscillator). (from DK 2 GU)

The quiescent current of the output transistor T 118 (2 N 5641) is greatly dependent on the operating voltage in the given circuit. Since too low a quiescent current at a low operating voltage can lead to strong distortion, the following modification is very advisable: Resistors R 189 and R 192 should be removed and replaced by a 1 N 4148 diode (cathode to ground). Resistor R 190 should be reduced to 470 Ω and R 191 to 1 kΩ. The quiescent current should now be adjusted to a value of between 25 mA and 50 mA. (from DK 2 GU)

The emitter decoupling capacitor of T 101 was decreased in value to reduce IF-gain. The values of resistors R 138 and 139 were increased to 1.1 kΩ. Bias adjustment of T 118 proved critical since R 192 heated up and changed resistance. This means that one must wait 10 minutes before continuing adjustment.

An additional screen has been provided between collector and base circuit of T 118 to improve stability.

The transmitter is easy to align after matching diodes D 120-123.

Transistor types 2 N 918 has been used for T 113 - T 115 instead of the BF 224. (from G 8 DET)

4.2. OTHER COMPONENTS OF DC 6 HL

If transistor T 2 is to operate as constant-current source, it is necessary to use a BF 245 A and select the value of R 1. The high current of the BF 245 C is too high for this circuit which is the reason why it does not operate as a constant-current source. It is also possible to leave out T 2 completely. (from DK 2 GU)

4.3. MODULE DC 6 HL 003

If other constructors find the resonant frequencies of the circuits comprising L 310 and L 311 are too low, the number of turns on L 310 and the values of capacitors C 318 and C 319 should be reduced by half. If the core of the inductance is also aligned slightly out of the coil, it is possible to obtain 2.3 V_{pp} or approx. 0.7 V rms at Pt 306 with only 150 m V_{pp} at Pt 304. A higher VFO voltage will produce an even higher level at 135-137 MHz. A voltage of 100 mV at a frequency of 130 MHz is present at Pt 306 on switching off the VFO. This means that this signal is only suppressed by 26 dB. However, it is suppressed further by the absorption circuit in the linear amplifier. (from DK 2 GU)

If the short 4 mm dia. coilformers have been used for L 312, L 317 it will be necessary to use the brown VHF cores. If longer 4 mm dia. coilformers are available, it is possible to increase the spacing between turns and to use red HF-cores. This allows one to obtain a higher output voltage and thus higher output power (from PA Ø GMS)

The screening panels in the vicinity of L 308/L 309 to L 316/L 317 should be grounded at both sides of the long operating voltage line. Otherwise this amplifier can break into oscillation. (from PA Ø GMS)

The capacitively tapped coils have been replaced by inductive coupling and are now resonated with ceramic trimmers. This gives increased output when cores are not used.

Transistor type 2 N 918 have been used for T 302-T 305. The modulator diodes have been matched for minimum carrier. There is still a slight component of 130 or 131 MHz signal at Pt 306 but is well suppressed. The value of coupling capacitor C 306 was reduced to 3.3 pF in order to reduce the heating of T 302. (from G 8 DET.)

4.4. DC 6 HL 007/008

If higher demands are to be made on the very good local oscillator T 706 in the FM-module DC 6 HL 007, it is possible to replace inductance L 707 by a crystal and the fixed capacitor C 720 by a trimmer capacitor of 50 pF. A frequency of 9.455 MHz should be selected since 8.545 MHz produces a spurious signal in the band. Resistor R 720 should be reduced to approximately 200 Ω to 300 Ω in order to reduce the drive to the mixer. (from DK 2 GU)

The frequency stability of the free-running oscillator in DC 6 HL 008 is often unsatisfactory. Instead of making modifications at this point, it is simpler to frequency modulate the VFO by connecting a varactor diode via a 15 pF capacitor to the base of T 50. The carrier is then injected by unbalancing the ring modulator. This is achieved by placing a voltage of 9 V to 12 V to Pt 112; R 196 should be deleted, as should the whole module DC 6 HL 008.
(from DK 2 GU)

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 4/74 of VHF COMMUNICATIONS

DK 1 OF 016/017	144 MHz/9 MHz CONVERTER		Ed. 4/1974
PC-board	DK 1 OF 016	(with printed plan)	DM 14.--
PC-board	DK 1 OF 017	(with printed plan)	DM 9.--
Semiconductors	DK 1 OF 016	(4 transistors, 1 diode)	DM 28.--
Minikit	DK 1 OF 016	(8 coilformers with cores, 1 coilset, 3 ferrite chokes, 1 trimmer capacitor)	DM 9.60
Ring modulator	IE 500		DM 48.50
Kit	DK 1 OF 016/017	with above parts	DM 108.--
DJ 4 BG 014	PRODUCT DETECTOR with CRYSTAL OSCILLATORS		Ed. 4/1974
PC-board	DJ 4 BG 014	(with printed plan)	DM 10.--
Semiconductors	DJ 4 BG 014	(1 IC, 2 transistors, 4 diodes)	DM 18.--
Minikit 1	DJ 4 BG 014	(3 trimmer caps., 15 ceramic capacitors, 2 tantalum electrolytics, 1 coilset)	DM 22.--
Minikit 2	DJ 4 BG 014	(1 TEKO-Box 3 A, 2 HC-25/U crystal holders, 1 13-pin connector)	DM 20.--
Crystals	XF 901, XF 902 set		DM 40.--
Kit	DJ 4 BG 014	with above parts	DM 109.--
DJ 4 BG 016	SYSTEM BOARD for MODULAR RECEIVER		Ed. 4/1974
PC-board	DJ 4 BG 016	(with printed plan)	DM 26.--
Transformer	VAC encapsuled 220 V/2 x 12 V, 5 VA		DM 23.--
Kit	DJ 4 BG 016	with above parts	DM 49.--
Connectors	13-pin	male, each	DM 4.10
	13-pole	female for wiring, each	DM 5.40
	13-pole	female for PC-board, each	DM 4.90
KITS	FOR SYSTEM BOARD DJ 4 BG 016:		
DJ 4 BG 007	Edition 1/1972		DM 72.--
DJ 4 BG 011	Edition 1/1973		DM 73.--
DJ 4 BG 014	Edition 4/1974		DM 100.--
	DJ 4 BG 013 and 015 in preparation		
DJ 5 HD 003	500 MHz PRESCALER		Ed. 4/1974
PC-board	DJ 5 HD 003	(double-coated, no through contacts)	DM 18.50
Semiconductors	DJ 5 HD 003	(4 ICs, 3 transistors)	DM 365.--
Minikit	DJ 5 HD 003	(6 ferrite chokes, 12 resistors, 2 potentiometers, 3 feed-through caps., 1 tantalum electrolytic, 10 bypass caps., 4 ceramic caps.)	DM 20.20
Kit	DJ 5 HD 003	with above parts	DM 398.--
DJ 1 EE 003	13 cm CONVERTER		Ed. 4/1974
PC-board	DJ 1 EE 003	(Teflon/PTFE, double-coated)	DM 34.--
Semiconductors	DJ 1 EE 003	(2 transistors, 2 Schottky diodes)	DM 50.--
Minikit	DJ 1 EE 003	(10 chip caps., 5 feed-through caps., 6 ceramic caps., 2 trimmer caps., 3 ferrite chokes, 1 TEKO box 3A, 3 BNC-connectors)	DM 41.90
Kit	DJ 1 EE 003	with above parts	DM 125.--