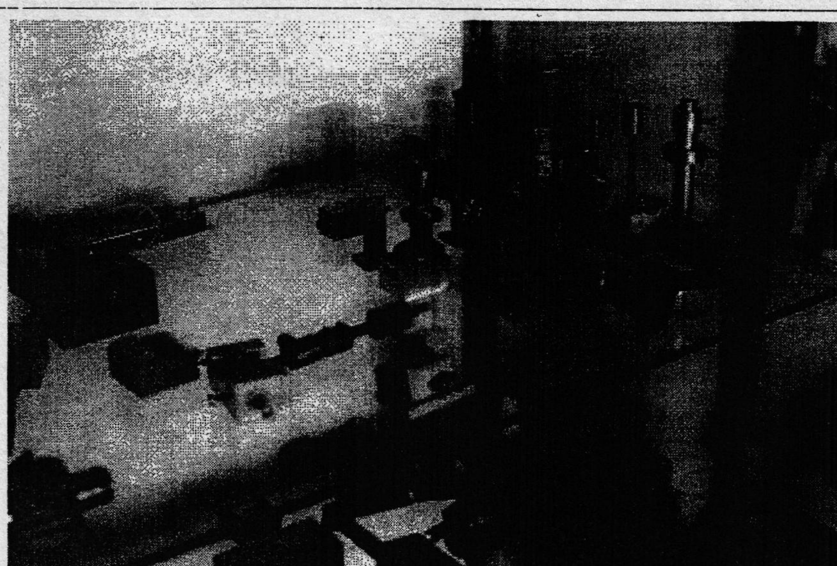
HYPER

BULLETIN D'INFORMATIONS
DES RADIOAMATEURS ACTIFS
EN HYPERFREQUENCES

HYPER

Numéro spécial

ondes millimétriques



Années 2001-2002

Les ondes millimétriques

47 Ghz et au-dessus ...

Selon l'encyclopédie Universalis: les « *hyperfréquences représentent des ondes électromagnétiques dont la longueur d'onde est de l'ordre du centimètre ... En fait les hyperfréquences débordent largement les longueurs d'onde centimétriques que du côté des ondes millimétriques. Les limites en fréquences sont donc très floues.* »

Cela va me permettre de vous dire que vous rencontrerez dans ce numéro spécial « millimétrique » de HYPER des descriptions de montages concernant le 47 Ghz jusqu'au 447 Thz !

Ce numéro spécial n'aurait jamais vu le jour sans l'énorme travail de collectage de Eric, F1GHB. Merci Eric.
Je remercie également les obcurs, les sans-grade qui ont oeuvré dans l'ombre pour ce numéro: F5AYE, F1CHF, F1AHO, F9HX, F4BAY, F6BVA, Jeanne (à la saisie du sommaire), ...
Merci également à Yves, F1AVY pour son article inédit sur le 447 Thz.

Pour les Oms intéressés par le 241 Ghz, je vous recommande le numéro spécial d'HYPER édité en 2001 qui décrivait un transverter 241 Ghz/144 Mhz conçu et réalisé par F1OPA et F5JWF.

Bonne lecture!
73
Alain, F5LWX, le pianiste d'HYPER

Ce numéro spécial « millimétrique » est en vente chez « ART-COMPO » 83 Avenue Louis Cordelet, 72000 LE MANS.

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Généralités

The bands above 24GHz

20.1 INTRODUCTION

Until 1979 the highest frequency microwave band available to amateurs was 24GHz. The World Administrative Radio Conference 1979 (WARC 79) allocated a number of new bands above 24GHz to the Amateur and Amateur Satellite Services. They are listed in Table 20.1 and in the text they are referred to as the 47, 76, 120, 142 and 241GHz bands.

Since that time there has been comparatively little work done on these new higher bands, largely because of the non-availability of suitable components to amateurs; components such as mixer, detector, Gunn and multiplier diodes, GaAs fets and the like have been available to professional engineers, but their price has precluded amateur use. As a result, the development of amateur techniques and practical designs have lagged behind developments on other microwave bands. Therefore much of what follows in this chapter is tentative and based on the limited experiences of a few amateurs. We look forward, however, to significant developments (at least at 47 and 76GHz) in the near future.

20.2 THE BANDS

The actual bands available to amateurs at present vary from country to country, but it is to be expected that at least the exclusive bands will be eventually allocated worldwide. Parts of these bands are harmonically related to presently active lower frequency microwave bands and the International Amateur Radio Union (IARU), part of the International Telecommunications Union (ITU), has recommended that initial operation in the new bands should use these harmonically related frequencies. It is thus possible to verify the frequency of operation by listening for harmonics from transmitters in the bands below and, eventually, when the necessary techniques have been developed, it will be possible to generate rf at the higher frequencies by frequency multiplication.

Fig 20.1 illustrates some of these harmonic relationships together with some other useful relationships *not* related to amateur frequency allocations. Fig 20.2 gives some simple relationships based upon harmonics from readily available Gunn sources in the 23 to 25GHz range. Some of the recent successful European work, described later, has been based upon multipliers from frequencies in this range. The awkward multiplication factor of ×7 from 144MHz is best avoided, for instance by generating 1,008MHz by some other route and providing amplification to a satisfactory power level at 1,008MHz, before further multiplying up into some of the millimetre bands.

It is also practical to start at higher frequencies in the tables; thus an oscillator at 8.064, 16.128GHz or in the range 23 to 25GHz would be a prolific source for the millimetre bands.

Table 20.1. Amateur millimetre bands

Band	Status
47.0 – 47.2GHz	Exclusive
75.5 – 76.0GHz	Eexclusive
76.0 – 81.0GHz	Secondary to radiolocation
119.98 – 120.02GHz	Footnote
142.0 – 144.0GHz	Exclusive
144.0 – 149.0GHz	Secondary to radiolocation
241.0 – 248.0GHz	Exclusive
248.0 – 250.0GHz	Secondary to radiolocation

These bands, with the exception of 120GHz, are allocated to the amateur and amateur satellite services.

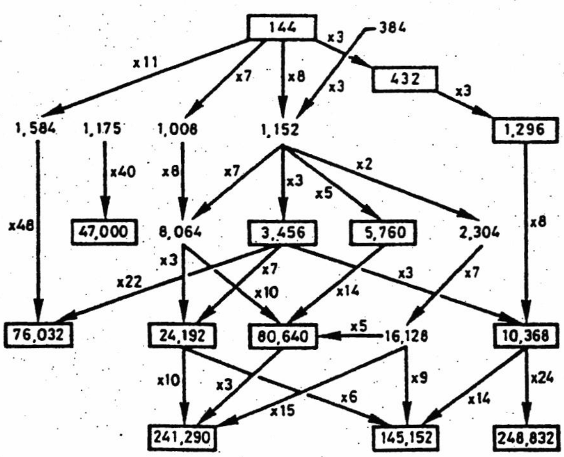


Fig 20.1. Harmonic relationships of the millimetre amateur bands. Frequencies in MHz. Frequencies in "boxes" are in amateur bands, others are not

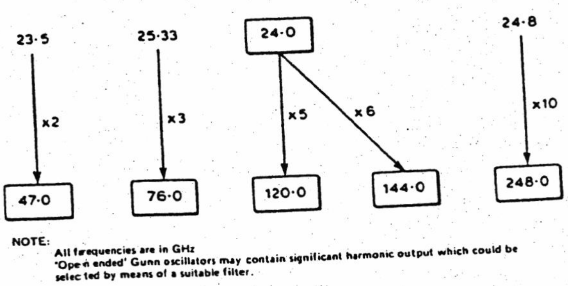


Fig 20.2. Simple harmonic relationships derived from tuneable Gunn oscillators. Frequencies in GHz

20.3 PROPAGATION

All of the propagation phenomena known on the lower microwave bands can be expected to manifest themselves in the millimetre bands, but the scale of the phenomena will be smaller. Effective ducts may be only a few metres thick. Smaller objects are effective reflectors and local weather may be of more consequence than is the case on the lower frequency bands.

One phenomenon appears on the millimetric bands which is of much less importance on bands below 20GHz – atmospheric absorption. We are used to referring to the atmosphere as transparent to light, but a moment's thought will remind us that we can see further on a "clear" day than a misty one. The same effect manifests itself with radio frequencies above about 20GHz. Depending on the humidity, there is a variable loss due to water vapour; there is a fixed loss due mainly to oxygen; and finally there is a considerable occasional loss due to rain (or other hydrometeors).

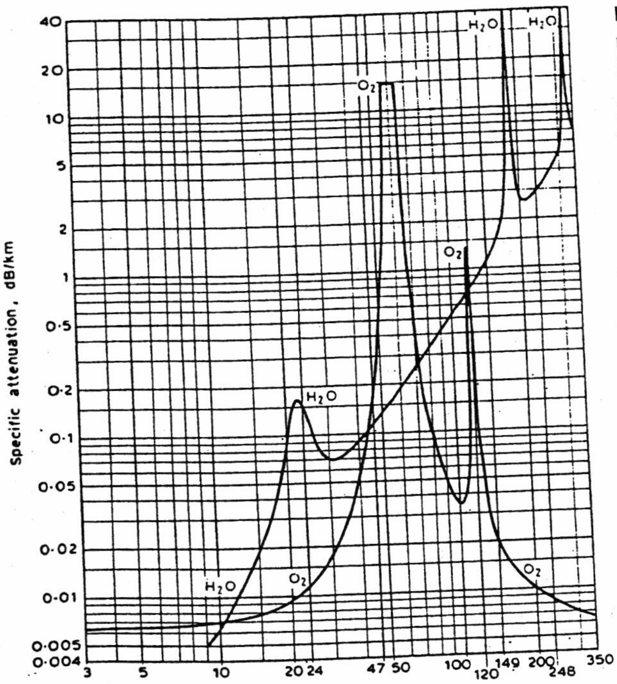


Fig 20.3. Attenuation due to oxygen and water vapour at pressure of 1 atmosphere, temperature 20°C, water vapour 7.5g/m^3 (source, CCIR)

Loss due to absorption follows a totally different law to that associated with radiation; the loss is a *linear* function of distance and is expressed in dB/km, ie. if there is a loss of, say 5dB in the first km, then there is the same loss in each subsequent km. This causes the loss to build up much more rapidly than is due to the inverse square law of radiation and, sooner or later, the loss due to absorption will determine the maximum range that can be worked.

Figure 20.3, taken from CCIR reports, shows the attenuation per km on a horizontal path due to oxygen and water vapour (at 7.5g/m^3), the two gases which cause the greater part of the losses. *This is in addition to the normal free-space loss.*

At a Microwave Round Table [1], G8AGN gave a valuable assessment paper which looked in more detail at the effects of atmospheric absorption and rain attenuation and considered the potential of the amateur millimetre bands. Much reference was made to the 24GHz band as a model, because it may be considered as a "longer mm waveband" and also lies close to the principal atmospheric water absorption band. The remainder of this section is based largely on that paper and it should become apparent that the mm bands above offer more potential than at first sight. A microcomputer prediction program was also given and is included later.

20.3.1 Free space loss

For line of sight paths, the free space path loss (in dB) may be calculated from the classical formula:

$$\text{Loss} = 92.45 + 20\log(f) + 20\log(d)$$

where f is the frequency in GHz and d is the path length in km. Fig 20.4 shows this relationship in graphical form for all the bands above 10GHz, with the exception of 120GHz.

20.3.2 Water vapour and oxygen attenuation

Fig 20.3 has already given a general overview of the additional path losses to be expected due to water vapour and oxygen absorption. The loss due to oxygen is fixed and there is nothing that can be done about it. In particular, at 120GHz there is an absorption band due to oxygen. The maximum loss occurs just below the amateur band at 118.75GHz but, even so, a loss of around 0.5dB/km is to be expected. This is enough to make long-distance contacts unlikely. The loss due to water vapour, however, is not fixed: it is dependent upon the total amount of water held in the air, the absolute humidity. This may be quite high in the summer but very low on a cold night in the winter.

A more useful picture is therefore given by the data derived from a Jet Propulsion Laboratory model which is shown in Fig 20.5. Here absorption predictions for several values of atmospheric water vapour content are shown.

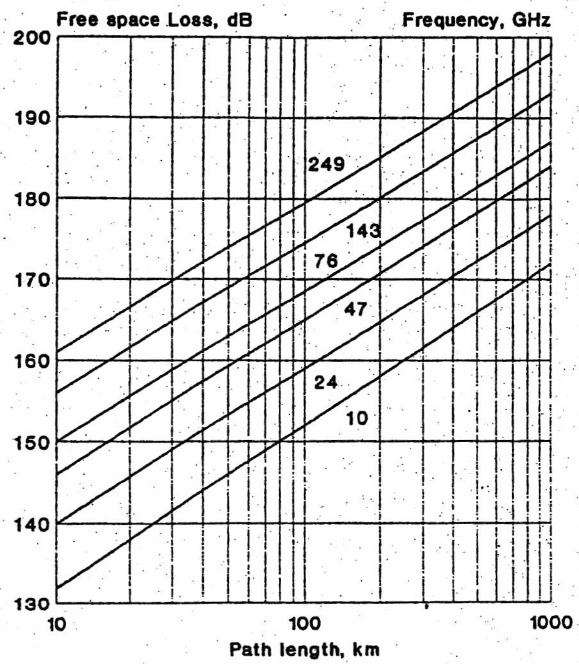


Fig 20.4. "Free space" path loss versus frequency for the bands from 10 to 249GHz (except 120GHz)

From these curves, a set of empirical relationships between absorption loss (in dB/km) and atmospheric water content (in gm per cubic metre) have been derived for the amateur bands above 10GHz and these are shown in Table 20.2. These relationships consist of two terms – a constant term, due largely to oxygen, and a term proportional directly to the water vapour content.

For these to be of use, a knowledge of the atmospheric water content over the path of interest is thus required but such information is not directly available and hence must be sought. Fortunately it is at hand in the form of the "relative humidity" (RH). This may be defined as:

$$\frac{\text{Actual water content of air in gm/m}^3\text{ at temperature T}}{\text{Water content in gm/m}^3\text{ if air saturated at temperature T}}$$

Note that RH is thus expressed as a percentage.

It is important to note that RH is mainly dependent on the temperature and, although not obvious from the formula given above, it is also slightly dependent on pressure, but this is ignored here. The weight of water, W, held in saturated air at different temperatures is given by the curve shown in Fig 20.6 and an approximate relationship between W and T has been derived from this. Hence, if the air temperature and RH can be measured, then the actual water content and, thus, the additional absorption term in the path loss can be determined.

The most satisfactory way for the radio amateur to measure RH is by using a pair of mercury thermometers, one of which has its bulb cooled by evaporation from a water-wetted wick

Table 20.2. Estimated atmospheric absorption loss

Band	Loss
10GHz	α = 0.0066 + 0.0011W
24GHz	α = 0.012 + 0.0185W
47GHz	α = 0.13 + 0.016W
76GHz	α = 0.2 + 0.034W
142GHz	α = 0.152W
241GHz	α = 0.417W

Loss (α) in dB/km; W in gm water/cubic metre

surrounding it. This records the so-called "wet bulb" temperature. Provided that the air flowing over the bulbs is moving at a velocity of at least 1m/sec, then the RH can be determined from simultaneous reading of both thermometers, as outlined in Table 20.3. Such measurements are facilitated by using an instrument known as a "whirling psychrometer" – two identical thermometers, one "dry" and one "wet", mounted side by side in what looks like an old-fashioned football supporters' rattle. This is whirled rapidly round and the wet and dry bulb temperatures read off the two thermometers without delay. Calculation is eliminated by use of a "psychrometric table" supplied with

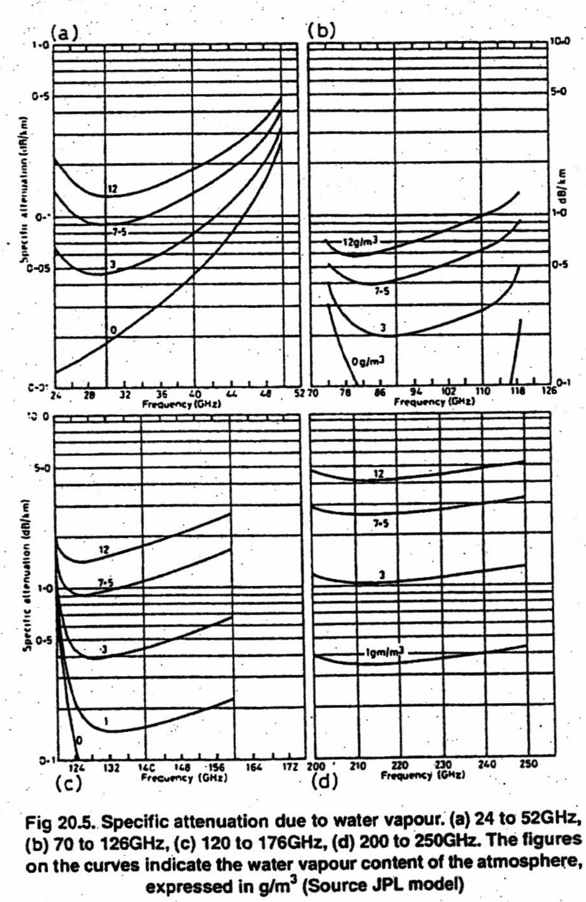


Fig 20.5. Specific attenuation due to water vapour. (a) 24 to 52GHz, (b) 70 to 126GHz, (c) 120 to 176GHz, (d) 200 to 250GHz. The figures on the curves indicate the water vapour content of the atmosphere, expressed in g/m³ (Source JPL model)

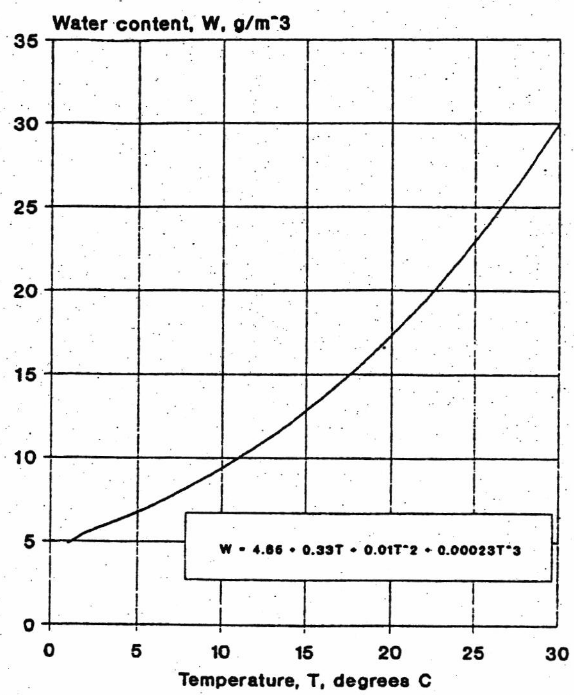


Fig 20.6. Graph of weight of water, W, in saturated air plotted as a function of temperature, T

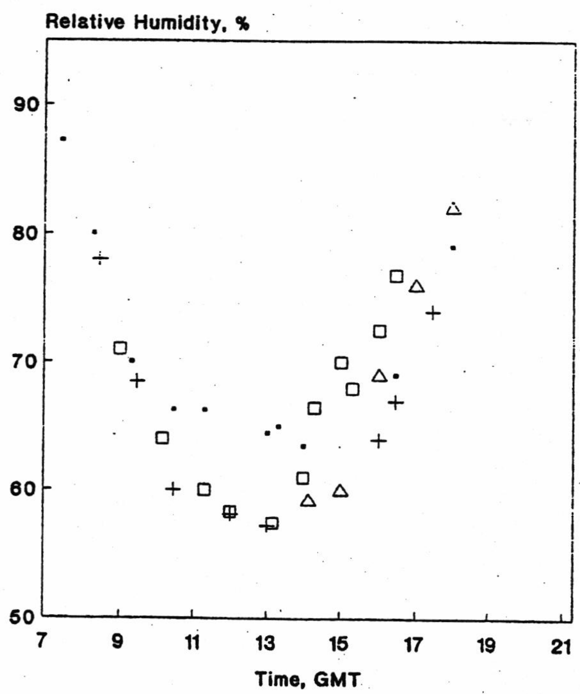


Fig 20.7. Typical diurnal variation of humidity in northern England as measured over several days by G8AGN

Table 20.3. Determination of relative humidity

$$RH = \frac{100[e_w - AP(T_d - T_w)]}{e_d}$$

P	=	atmospheric pressure in mB (milliBar)
T_d	=	dry bulb temperature in degrees C
T_w	=	wet bulb temperature in degrees C
e_d	=	Saturated water vapour pressure at T_d
e_w	=	Saturated water vapour pressure at T_w
A	=	A constant dependent on the velocity of the air flowing over the thermometer bulbs

For an air velocity of 1 to 1.5 m/sec and T_d 0 deg.C

$$A = 0.000799$$

e_d and e_w are related to T_d and T_w by:

$$e = t^{4\,9283} \cdot 10^{(23.5518 - (2937.4/t))}$$

where $t = T + 273$ (ie. degrees K)

Table 20.4. Effect of water vapour at various temperatures

Temperature (°C)	Maximum water content (gm/m³)	Attenuation (dB/km) 47GHz	76GHz	142GHz	241GHz
0	4.8	0.20	0.36	0.73	2.0
5	6.8	0.24	0.43	1.03	2.8
10	9.4	0.28	0.52	1.43	3.9
15	12.8	0.33	0.64	1.95	5.3
20	17.3	0.41	0.79	2.63	7.2
25	23.1	0.50	0.99	3.51	9.6
30	30.4	0.62	1.23	4.62	12.7
35	39.6	0.76	1.55	6.02	16.5
40	51.0	0.95	1.93	7.75	21.3

the instrument. It becomes merely a matter of reading off the dry bulb temperature against the difference "dry" minus "wet" (depression) to obtain RH directly.

Fig 20.7 shows the typical variation of RH throughout several days, as measured by G8AGN in the northern UK. At first sight it would appear that the most favourable time for operation on the mm bands would be in the early afternoon. This figure does not, however, tell the whole story since it is the *actual* water content of the air which matters and this is dependent on temperature. Thus, even though the RH may be quite low around the early afternoon, the air temperature will be high, as will be the amount of water that the air could hold if saturated. In practice, therefore, the optimum time for operation may be determined by measuring both RH and air temperature at both ends of the path and estimating the increased path losses due to water absorption.

Table 20.4 gives the maximum water content in the atmosphere at various temperatures and the corresponding attenuation for the four principal millimetre bands. Below 0°C, the absolute humidity falls rapidly so that, at least in the 47 and 76GHz bands, the loss due to water vapour can be ignored. The figures are maxima for the bands when the air is saturated with water, ie 100% relative humidity. The

Table 20.5. Estimated attenuation due to rain

Attenuation, alpha	= AR^B	–	(dB/km)
where	R (rainfall) is in mm/hour		
	A = cf^d (f = frequency in GHz)		
	B = ef^g (f = frequency in GHz)		
For	f < 2.9GHz	c = 6.39×10^{-5}	d = 2.03
	f > 2.9 and < 54GHz	c = 4.21×10^{-5}	d = 2.42
	f > 54 and < 180GHz	c = 4.09×10^{-2}	d = 0.699
	f > 180GHz	c = 3.38	d = –0.151
For	f < 8.5GHz	e = 0.851	g = 0.158
	f > 8.5 and < 25GHz	e = 1.41	g = –0.0779
	f > 25 and < 164GHz	e = 2.63	g = –0.272
	f > 164GHz	e = 0.616	g = 0.0126

actual loss can be taken as proportional to the relative humidity at any temperature. In temperate climates, 100% relative humidity over a large area is uncommon, but 90% occurs sufficiently frequently to limit the range of regular contacts. In arid regions the humidity may be less than 10% for much of the time. It follows that the best way of working very long ranges on the millimetre bands is either to operate in an arid climate or on a freezing winter's night!

20.3.3 Rain attenuation

The other main factor limiting long distance contacts is the increased path loss (α) due to rain attenuation. This may be estimated by using the relationship:

$$\alpha \text{ (dB/km)} = A \times R^B$$

where R is the rainfall rate in mm/hour and the constants A and B are given in Table 20.5. Actual values of rain attenuation may be estimated using the curves of Fig 20.8, given that typical rainfall rates are 0.25, 1, 4 and 16mm/hour for drizzle, light, moderate and heavy rain respectively. In practice, such rainfall rates do not apply uniformly along the whole path, since rain is often concentrated in localised "cells" which are typically 1 to 10km in extent.

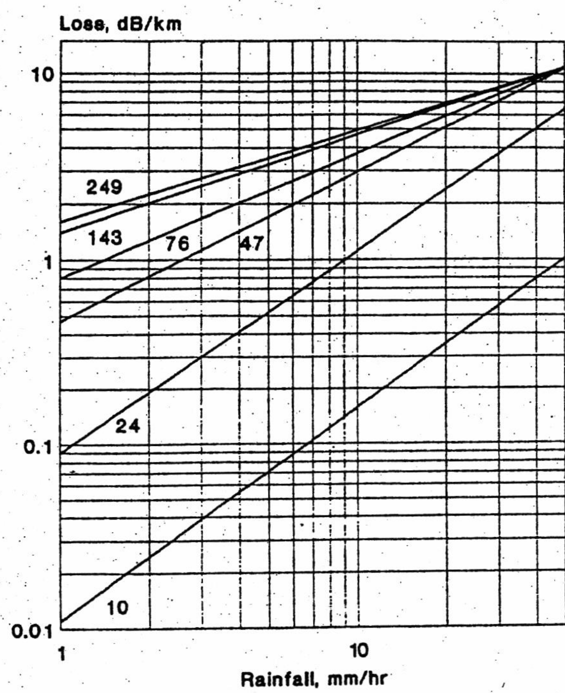


Fig 20.8. Attenuation in the amateur millimetre bands due to rain, plotted as a function of the rate of rainfall

20.4 BAND CAPABILITIES AND USES

20.4.1 The 47 and 76GHz bands

The 47 and 76GHz bands can be used for point-to-point operation in the same way as the 10 and 24GHz bands, with the proviso that, at least in Europe and the coastal regions of America, water vapour will limit ranges to a few tens of kilometres for most of the time. Fig 20.9 shows the expected path losses at 47GHz with the capability of simple wideband equipment and state-of-the-art low power

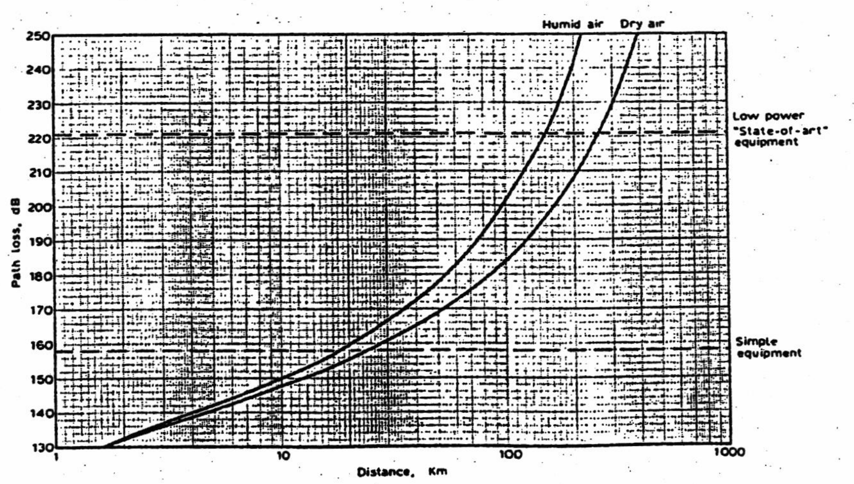


Fig 20.9. Path loss versus frequency for the 47GHz band, for both dry and humid conditions

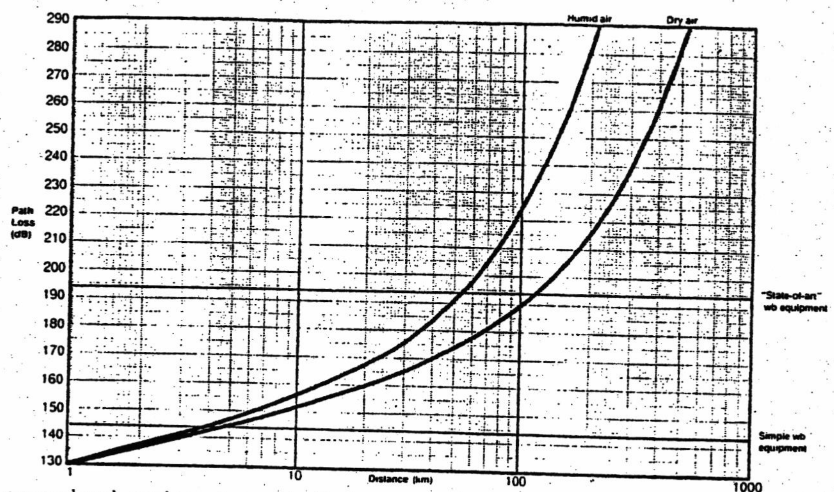


Fig 20.10. Path loss versus frequency for the 76GHz band, for both dry and humid conditions

narrowband equipment marked for reference. Fig 20.10 shows the expected path losses at 76GHz.

The additional path loss terms due to water vapour and oxygen absorption and rain attenuation have been incorporated into a special version of G8AGN's system analysis program given in chapter 2, "Operating techniques", and this is given in section 20.4.3. The modified program has been used to estimate the maximum potential of wideband equipment operating in the 47 and 76GHz bands under a range of climatic conditions. Fig 20.11 shows the path lengths possible at 47GHz for wideband equipment based, perhaps, on generating some second harmonic from a 23.5GHz oscillator and using a harmonic mixer. For comparison, the first world record for this band, a little over 50km, was made using just such wideband equipment (see later). Fig 20.12 shows that paths of 10 to 20km should be possible at 76GHz using very modest equipment, again based on harmonic transmitters and mixers.

It can be seen that, even with very modest equipment, the 47 and 76GHz bands offer some considerable scope for

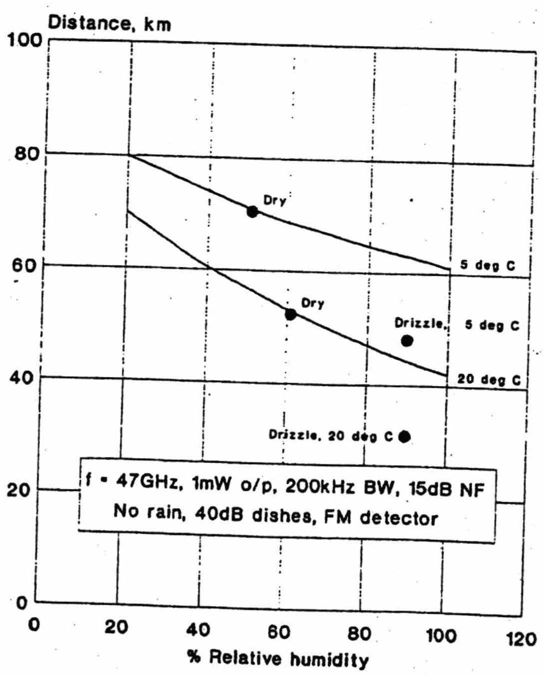


Fig 20.11. Predicted performance of simple wideband equipment at 47GHz

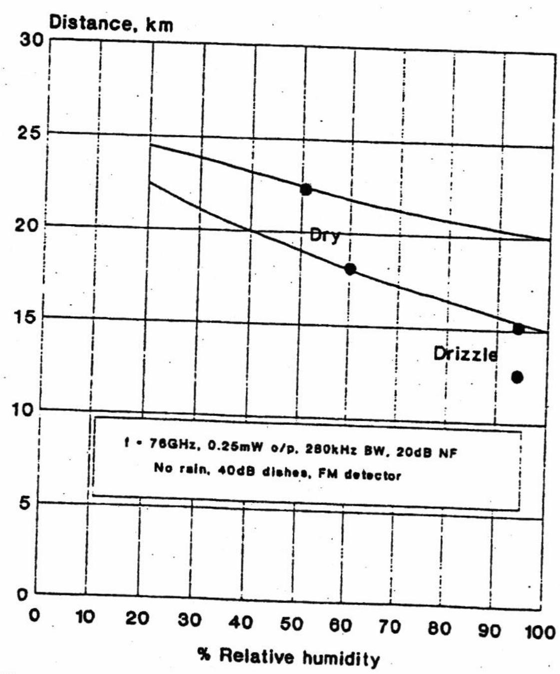


Fig 20.12. Predicted performance of simple wideband equipment at 76GHz

experimentation and dx working. As pointed out later, it is expected that mm-wave devices will be used more widely in both the professional and consumer areas and so will become available to amateurs on the "surplus" market. When this happens, then the prospect of narrowband operation at these frequencies will transform the situation and allow longer distance contacts that those predicted here – certainly on the lower mm wavebands. Indeed the recent establishment (August 1988), by USA amateurs, of a new world record of just over 105km for narrowband operation on the 47GHz band gives support to this thesis.

20.4.2 The higher mm bands

The 120GHz band is on the edge of an absorption band due to oxygen. The loss from this gas of about 0.4dB to 0.5dB per kilometre is more than that to be expected from water vapour and is such that quite potent equipment would be required for communication over any great distance. What this band does offer is privacy, because long distance transmission is practically impossible; so are interference and "eavesdropping".

Possible uses of the 142 and 241GHz bands are more problematical. Consolation can be found in the fact that, apart from the inevitable move to higher frequencies as the lower bands fill, no clear consensus has appeared among professional users as how to best exploit the special characteristics of the higher millimetre bands.

20.4.3 System performance analysis program

This program is a modified version of that given in chapter 2, section 2.7.6, and includes the losses due to water vapour, oxgyen and rain attenuation.

```
A MILLIMETRE - BAND RADIO SYSTEM PERFORMANCE PROGRAM

   1 REM  RADIO SYSTEM PERFORMANCE ANALYSIS
   2 REM
   3 REM  v 1.0 (C) 1985 B Chambers G8AGN
   4 REM
   5 REM  v 1.1 (C) May 1988, 24GHz water losses included
   6 REM
   7 REM  v 2.0 (C) Oct 1988, Atmospheric and water losses included
   8 REM               for the mm bands.
   9 REM
  10 REM  Program in BBC BASIC
  11 REM
  70 DATA 0,2.6,10,16,22: REM Detector thresholds
  75 DATA 3.41,3.66,3.93,4.22,4.52,4.83,5.19,5.56,6.36,6.80,7.26,7.75,8.27,
8.8,2,9.40
  76 DATA 10.01,10.66,11.35,12.07,12.83,13.63,14.84,15.37,16,21,17.30,18.34,
19.43,20.58
  77 DATA 21.78,23.05,24.38,25.78,27.24,28.78,30.38: REM Saturated air data
  78 @%=&20109: MODE 3: REM Output format F 9.1
  85 rfl=-1:afl=-1
  90 CLS:PRINT:PRINT"Microwave System Performance Analysis"
 100 PRINT:PRINT
 110 INPUT "Enter frequency in MHz   "F
 120 L=299.8/F
 130 INPUT "Enter Rx i.f bandwidth in KHz   "RB
 140 LRB=10*LOG(RB)+30
 150 INPUT "Enter Rx noise figure in dB   "NF
 160 INPUT "Enter Rx antenna gain in dB   "RG
 170 BR=SQR(27000/10^(0.1*RG))
 180 PRINT:PRINT "Nominal 3dB beamwidth =  ";BR;"  Degs."
 190 PRINT:PRINT "Enter Rx feeder loss in dB   "RF
 200 PRINT "Choose type of detector: "
 210 PRINT "  1.          SSB "
 220 PRINT "  2.          AM  "
 230 PRINT "  3.          FM  "
 240 PRINT "  4.          FM no limiter"
 250 PRINT "  5.          FM slope detectoin"
 260 PRINT: INPUT "Enter choice   "DT
 270 IF DT<1 OR DT>5 THEN 260
 280 RESTORE 70: FOR i= 1 TO DT: READ DTH: NEXT
 290 NT=290*(10^(0.1*(NF+RF))-1)
 300 PRINT:PRINT"Effective receiver noise temperature = ";NT;"Deg K"
 310 NT=10*LOG(NT)
 320 ERS=-228.6-LRB+NT+DTH-RG
 330 PRINT:PRINT"Effective receiver sensitivity =   ";ERS;"  dBW"
 340 PRINT:INPUT"Enter transmitter power in Watts   "TP
 350 LTP=10*LOG(TP)
 360 INPUT"Enter Tx antenna gain in dB   "TG
 370 BT=SQR(27000/10^(0.1*TG))
 380 PRINT:PRINT"Nominal 3dB beamwidth =   ";BT;"  degs"
 390 PRINT:INPUT"Enter Tx feeder loss in dB   "TF
 400 EI=LTP+TG-TF
 410 PRINT:PRINT"Transmitter eirp =   ";EI;" dBW"
 420 PLC=EI-ERS
 430 PRINT:PRINT"Path loss capability =   ";PLC;"  dB"
 440 PRINT:PRINT"  Options are: "
 450 PRINT"   1.    C/N over a given LOS path"
 460 PRINT"   2.    Maximum LOS range of the equipment"
 470 PRINT:INPUT"Enter choice   "C$
 480 IF C$<>"1" AND C$<>"2" THEN 440
 490 IF C$="2" THEN 600
 500 PRINT:INPUT"Enter path length in km   "PL
 510 LO=32.45+20*LOG(F)+20*LOG(PL)
 520 PRINT:PRINT"Free space loss =   ";LO;" dB"
 530 ATL=0:PROCatmos
 540 RL=0:PROCrain
 550 LO=LO+ATL+RL
 580 PRINT:PRINT"Nominal path loss =   ";LO;" dB"
 590 PRINT:PRINT"Estimated LOS CARRIER TO NOISE =   ";PLC-LO;" dB"
 600 PROCpl
 605 PRINT:PRINT"Atmospheric absorption loss =  ";ATL;" dB"
 610 IF RL>0.0 THEN @%=&020205:PRINT:PRINT"Rain loss =   ";RL;" dB": @%=&20109
 620 PRINT:PRINT"Max. LOS path length = ";PL;" km":@%=10:END
 670 DEFPROCpl
 680 DL=1:PL=1
 700 ATL=0:PROCatmos:PROCrain
 710 X=PLC-20*LOG(4000*PI*PL/L)-ATL-RL
 720 IF X>0 THEN PL=PL+DL:GOTO 700
 730 IF X=0 THEN 770
 740 IF X<0 AND DL>0.01 THEN PL=PL-DL:DL=0.1*DL:PL=PL+DL:GOTO700 ELSE PL=PL-
0.05:GOTO 770
 760 PRINT:PRINT"No solution found for path length"
 770 ENDPROC
2000 DEFPROCrain
2004 IF rfl=0 THEN ENDPROC
2005 IF rfl=1 THEN 2120
2010 PRINT:INPUT"Is there rain along the path ?   "RP$
2020 rfl=INSTR(RP$,"Y"): IF rfl=0 THEN ENDPROC
2030 PRINT:PRINT"Enter type of rain: ":PRINT: PRINT"1.   Drizzle over whole
path"
2040 PRINT"2.   Light rain over a 10km cell"
2050 PRINT"3.   Moderate"
2060 PRINT"4.   Heavy"
2070 PRINT:INPUT"Enter choice   "CH
2080 IF CH<1 OR CH>4 THEN 2070
2090 IF CH=1 THEN R=0.25
2100 IF CH=2 THEN R=1
2110 IF CH=3 THEN R=4
2112 IF CH=4 THEN R=16
2120 f=0.001*F
2130 IF f<2.9 THEN c=6.39E-5:d=2.03
2140 IF f<54 AND f>=2.9 THEN c=4.21E-5:d=2.42
2150 IF f<180 AND f>=54 THEN c=4.09E-2:d=0.699
2160 IF f>=180 THEN c=3.38:d=-0.151
2170 A=c*f^d
2180 IF f<8.5 THEN e=0.851:g=0.158
2190 IF f<25 AND f>8.5 THEN e=1.41:g=-0.0779
2200 IF f<164 AND f>=25 THEN e=2.63:g=-0.272
2210 IF f>=164 THEN e=0.616:g=0.0126
2220 B=e*f^g
2230 alpha=A*R^B
2240 IF CH=1 THEN RL=alpha*PL ELSE RL=alpha*10
2245 IF rfl=0 THEN @%=&020205:PRINT:PRINT"Rain loss = ";RL;" dB":@%=&20109
2250 ENDPROC
3000 DEFPROCatmos
3005 IF afl=1 THEN 3100
3010 REM Determine atmospheric water content
3020 RESTORE 75
3030 PRINT:INPUT"Enter air temperature in deg C   "T
3040 PRINT:INPUT"Enter relative humidity in %RH   "RH:RH=0.01*RH
3050 FOR IT=-5 TO T
3060 READ W
3070 NEXT
3075 w=W*RH
3080 PRINT:PRINT"Atmospheric water content =   ";w;" gm/m^3"
3100 f=0.001*F
3110 IF f>=10 AND f<=10.5 THEN al=0.0066*0.0011*w
3120 IF f>=24 AND f<=24.25 THEN al=0.012+0.0185*w
3130 IF f>=47 AND f<=47.2 THEN al=0.13+0.016*w
3140 IF f>=75.5 AND f<=76.0 THEN al=0.2+0.024*w
3150 IF f>=142 AND f<=144 THEN al=0.152*w
3160 IF f>=248 AND f<=250 THEN al=0.417*w
3170 ATL=al*PL
3175 IF afl=0 THEN PRINT:PRINT"Atmospheric absorption loss = ";ATL;" dB"
3180 IF ATL=0 THEN PRINT:PRINT"Atmospheric data not available"
3185 afl=1
3190 ENDPROC
```

20.5 TRANSMISSION LINES

20.5.1 Waveguide and flanges

Table 20.6 lists the standard waveguide sizes suitable for the amateur millimetre bands together with a range of rectangular brass tubing available from model shops and which can be used for the 47 and 76GHz bands. The quite high losses to be expected from standard rectangular waveguide should be noted. The values given in Table 20.6 are for copper; brass waveguide will give figures some 30% worse and silver about the same amount better.

Two flange types are available for millimetre waveguide: square flanges with four bolt holes and round flanges with

Table 20.6. Waveguide for the millimetre bands

Type no.	Outside dimensions (mm)	Inside dimensions (mm)	Cut-off (GHz)	Band (GHz)	λ_o (mm)	λ_g (mm)	λ_g / λ_o	Loss dB/m
WR42 WG20	12.7/6.35	10.67/4.32	14.047	24.0	12.49	15.41	1.234	0.45
K&S 268	9.53/4.76	8.74/4.00	17.1	24.0	12.49	18.18	1.454	0.60
WR28 WG22	9.14/5.59	7.11/3.56	21.07	24.0	12.49	26.11	2.090	0.75
K&S 266	7.93/3.97	7.14/3.18	21.0	24.0	2.49	25.72	2.060	0.75
WR22 WG24	7.72/4.88	4.78/2.39	26.34	47.0	6.38	7.70	1.208	0.78
K&S 264	6.35/3.18	5.55/2.38	26.98	47.0	6.38	7.79	1.221	0.90
K&S 262	4.76/2.38	3.96/1.60	37.8	47.0	6.38	18.06	2.830	1.8
WR12 WG26	5.13/3.58	3.10/1.55	48.35	76.0 81.0	3.94 3.70	5.11 4.61	1.296 1.247	2.5
WR8 WG28	3.96 dia	2.03/1.02	73.84	120.0	2.50	3.17	1.268	5.0
WR7 WG29	3.96 dia	1.65/0.825	90.84	142.0 149.0	2.11 2.01	2.75 2.54	1.301 1.261	6.0
WR5 WG30	3.96 dia	1.30/0.648	115.75	142.0 149.0	2.11 2.01	3.64 3.19	1.725 1.587	9.0
WR4 WG31	3.96 dia	1.09/0.546	137.52	241.0 250.0	1.24 1.20	1.51 1.44	1.215 1.196	12.0

removable, screwed rings. The latter are more popular among professional users but are more expensive. Fig 20.13 gives the dimensions of a general purpose waveguide flange much used professionally for test gear. The same-sized flange is used with a range of waveguide sizes and it can therefore be used as a standard connector between non-standard waveguides.

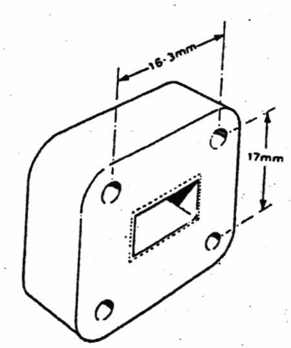


Fig 20.13. General purpose waveguide flange. The holes are 0.144in for either 4BA or M3 bolts. The waveguide is central with respect to bolt holes. Overall dimensions are not critical

As an alternative to the use of flanges, it is possible to adapt popular coaxial connectors for use with waveguide. Fig 20.14 shows the use of a BNC or TNC connector in this way. Since these two connectors have the same basic dimensions and differ only in the use of a bayonet or screwed clamp, it is possible to use the same arrangement with BNC fittings for easy connection and disconnection in the workshop and TNC fittings for a more reliable connection in the field. The internal coaxial components of either type of plug are discarded and for the larger waveguides the connector will need to be drilled through to allow the waveguide to fit.

20.5.2 Overmoded waveguide

As can be seen from Table 20.6, the loss of conventional waveguide is prohibitive, except for very short runs inside the equipment. If we wish to operate with an elevated antenna which involves a long transmission line, then the losses associated with a waveguide run are too high and other techniques are required. If a waveguide is larger than necessary to propagate the fundamental waveguide mode then a variety of other modes are possible in the guide. If not controlled, this "overmoding" is undesirable but, if controlled, particular modes may propagate with much lower loss than the fundamental TE10 mode used in conventional rectangular waveguide.

The most useful low-loss modes occur in a circular tube, and are the TE01 and TM01 modes. They are symmetrical; the TM01 mode has a radial electric field and a circumferential magnetic field, while the TE01 mode has a radial magnetic field and a circumferential electric field. These are shown in Fig 20.15. In each case the fields are low at the circumference of the metal guide; this means that the currents flowing in the waveguide wall are minimal, so the losses are correspondingly small.

The TE01 mode shows the lowest loss; a 15mm diameter copper tube operating at 47 or 76GHz can be expected to have a loss of not more than 0.01dB/m. The mode can be launched from a rectangular guide by the arrangement shown in Fig 20.16. The transition is quite badly mismatched and matching screws are required in the rectangular guide as shown. The TM01 mode has a higher loss; a 15mm copper tube, as above, would have a loss of 0.1dB/m, but the mode is much easier to launch by the arrangement shown in Fig 20.17 and, for the short feeder runs likely to be used by amateur stations, the additional loss is not important. Since these waveguide modes are radially symmetrical, the connections at the two ends need not be in the same plane, and if the circular guide is run up a mast, the antenna can be arranged to rotate about the centre line of the circular guide and a simple choke joint is all that is required to provide for rotation.

It is important that the overmoded waveguide runs are kept straight; the effect of bends is to couple between

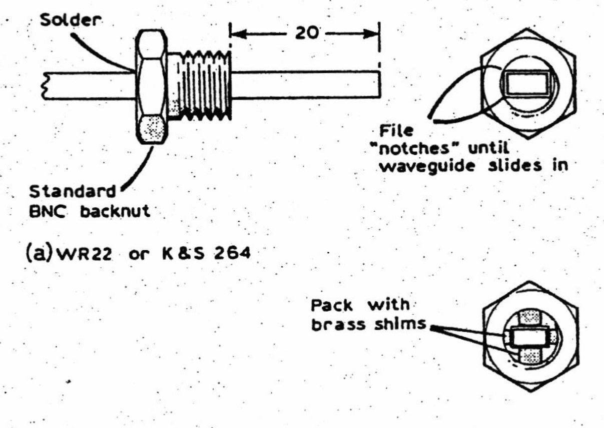


Fig 20.14. Use of BNC or TNC connector instead of a flange

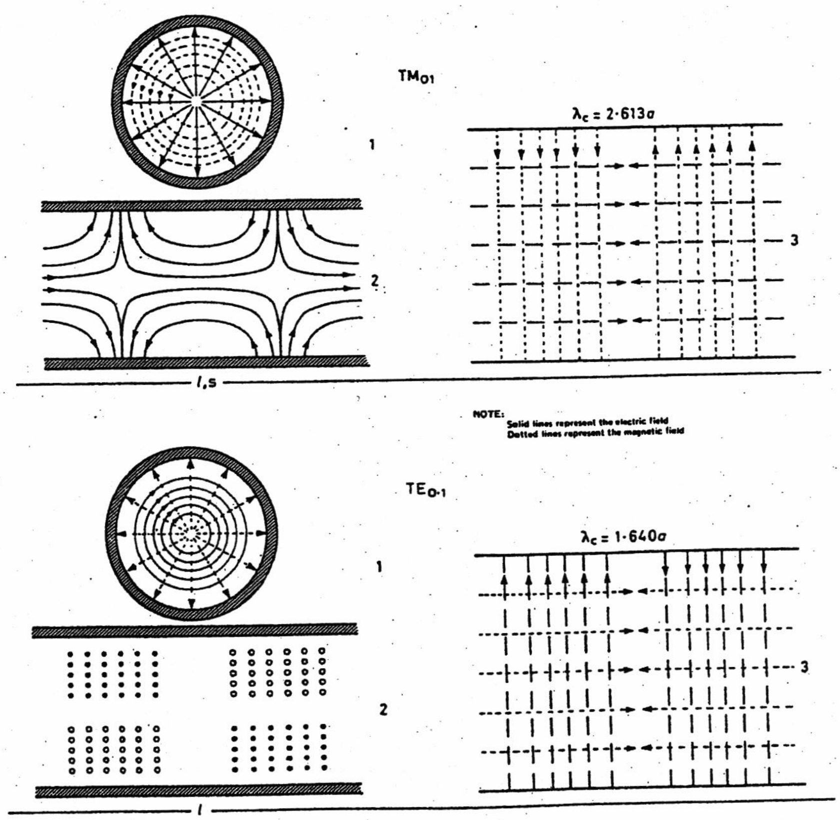
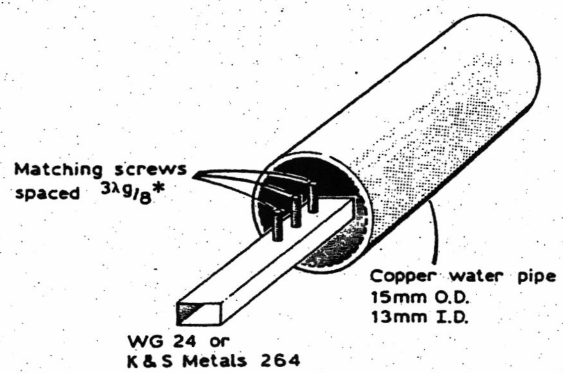


Fig 20.15. TM and TE modes in circular waveguide. Solid lines represent E-field and dotted lines the H-field. © Artech House, Inc., from Theodore S. Saad, *Microwave Engineers Handbook*, Vol 2, 1971. Reprinted by permission



*$3\lambda g/8$ (WG24) = 2·89mm

$3\lambda g/8$ (K & S 264) = 2·92mm

Fig 20.16. Mode transition for rectangular to circular waveguide

modes, and energy is transferred into other higher loss modes. These modes are also mismatched at the terminations, so that very high ratio standing waves are set up and a variety of unacceptable resonances will be produced in the guide, leading to further losses.

20.5.3 Other transmission lines

Whilst amateurs are most likely to use waveguide, it has several disadvantages for the professional user, including: close tolerances are needed in building components, it is difficult to use mass production techniques and waveguide does not lend itself to integration techniques. There are alternative transmission lines can be used that overcome these problems; one is microstrip, which is popular at centimetre wavelengths, but at millimetre wavelengths the losses start to become excessive. An alternative is dielectric waveguide.

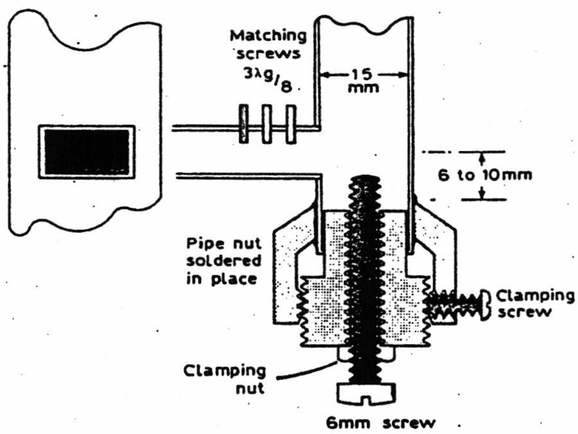


Fig 20.17. Launching the TM01 mode: rectangular to circular guide transition

20.5.4 Dielectric waveguide

This consists of a rectangular strip of dielectric and is normally used spaced from a ground-plane by a separate, thick dielectric layer, as shown in Fig 20.18. The electromagnetic field is confined in the rectangular dielectric due to its refractive properties, much as optical fibres behave with light. It can provide fairly good performance, much better than that of microstrip, as shown in Table 20.7. Most of the familiar waveguide components can be made from dielectric waveguide – directional couplers, attenuators, phase-shifters, filters and isolators are all used professionally.

20.5.5 Coaxial lines

Coaxial cables are almost impractical, in the 47GHz band and above, because of overmoding which occurs when the circumference of the cable approaches one wavelength in the dielectric. For example, at 47GHz the largest coaxial line with polythene dielectric (dielectric constant of 2) that can be used would have a circumference of:

$$\frac{300}{47 \times \sqrt{2}} = 4.5\text{mm}$$

giving a diameter of 1.4mm. At 76GHz, the maximum safe diameter is 0.9mm. Cables of such diameter are available (small diameter semi-rigid cable), but are generally too lossy to be useful.

20.5.6 "Optical transmission lines"

In the higher frequency bands it is possible to dispense with the metal walls of the waveguide by focusing radiation into a sufficiently "tight" beam and "catching" all of

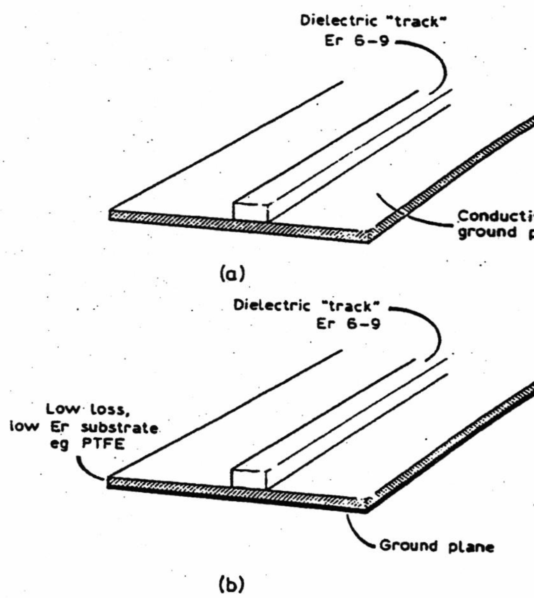


Fig 20.18. Dielectric waveguide

Table 20.7. Comparison of transmission line losses

Transmission line	Loss at 47GHz (dB/cm)	Loss at 76GHz (dB/cm)
Rectangular guide (copper)	0.011	0.022
50Ω microstrip (on quartz)	0.112	0.225
Dielectric guide (alumina)	0.039	0.078

the radiated power in a similar collector. By using a high-gain antenna which produces an almost parallel beam for a short distance and collecting the power in another antenna of similar size, it is possible to transmit power over tens of metres with relatively low loss. The losses due to side-lobes or other defects of the antennas are likely to be less than those from the same length of conventional wave-guide. The antennas are necessarily located within each others' near fields and the best results will be obtained with identical antennas at the two ends of the link.

20.6 SOLID STATE DEVICES

Devices for the millimetre bands are not yet commonplace. There are only a few manufacturers of both oscillator and detector/mixer diodes up to about 100GHz, notably Alpha Industries and M/A-COM (Microwave Associates). Devices at these frequencies are often listed in a catalogue by band (as in X-band for 10GHz), but confusion can arise since there are several letter designation systems and some frequency bands are different in each system. Common designations are:

- Ka-band, covering 18 to 26.5GHz. This stands for "K-above", since K-band is 12.4 to 26.5GHz, and Ku (for "K-under") is 12.4 to 18GHz.
- Q-band, covering 40 to 75GHz. However, the range 40 to 60GHz is sometimes referred to as U-band (eg by Alpha Industries).
- W-band, covering 75 to 110GHz.

See also chapter 13, "Data". All in all, it is considered better to avoid using these lettering systems and to use frequency ranges instead. GaAs fets will soon be available for the millimetre bands, but their cost is likely to be too high for amateur use in the near future.

20.6.1 Oscillators

Klystrons may be found and semiconductor oscillators, Gunn devices and impatt diodes are becoming available for the lower frequency millimetre bands. Low power Gunns will give around 10mW of output power, while high power devices giving some 100mW are available but are very expensive. They are usually operated with 4.5 to 5.5V bias, similar to 24GHz devices. A typical package for a device operating in the 40 to 75GHz range is shown in Fig 20.19.

20.6.2 Mixers and detectors

Probably most diodes in the pill package (SOD31, SOD45, SOD46 and similar outlines) will function as useful rectifiers even if they are not too good as mixers. In general, the chips used in modern microwave diodes are usable up to at least 100GHz; it is the mount which sets the limit to the frequency of use and the problem is getting the rf into the chip. Techniques to overcome this are described later.

Quite low noise figure diodes are available using Schottky barrier construction. Noise figures as low as 8dB at 60GHz are attainable. Like high power Gunn diodes, these better performance devices are very expensive.

20.6.3 Multipliers

At millimetre-wave frequencies, multiplication usually means high-order multiplication and step recovery diodes are available, suitable for output frequencies of up to 100GHz. They, too, are expensive and the amateur is probably better off trying to use lower frequency devices. These will function with reduced efficiency, but should give enough output power for a receiver local oscillator or a harmonic generator/calibrator.

20.7 TECHNIQUES

20.7.1 Mounting components

Most of the readily available pre-packaged devices are physically large compared with a wavelength and it is quite impracticable to arrange "connections" to "terminals" in the sense appropriate to lower frequencies. Usually the actual semiconductor chip is physically small and, in devices designed to operate efficiently at frequencies of 20 or 30GHz, it is usual for the chip to perform quite well at several times that frequency. The problem is, then, to get the rf into or out of the chip.

Fig 20.20 illustrates the internal construction of a typical mixer diode in a pill (SOD31) package, showing the device

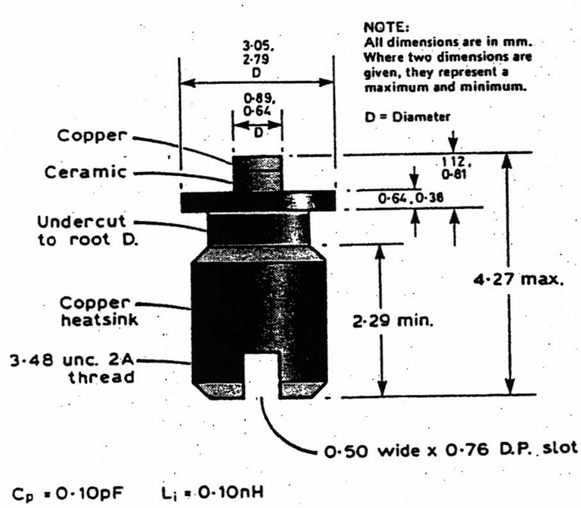


Fig 20.19. Typical Gunn diode package

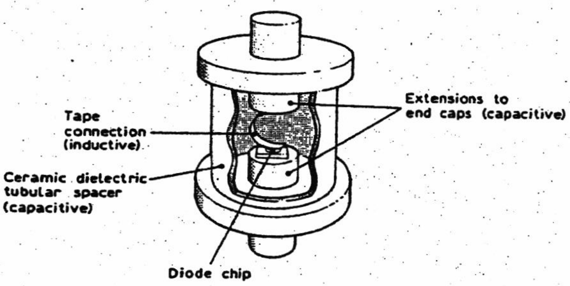


Fig 20.20. Internal construction of a 'pill' mixer diode

chip mounted on one cap of the package and tape connection made from the other cap. Gunn devices, varactors and pin diodes usually have similar arrangements. Since the physical size of the device package is an appreciable part of a wavelength even at 47GHz, we need to find some way to bypass the terminals. The technique used in waveguide mixers is, therefore, to immerse the diode package in the rf field and to compensate as best we can for the reactance introduced by the packaging.

The ceramic package is equivalent to bulk capacitive susceptance across the waveguide and is tuned out by a sliding short behind the diode. A series of tuning screws then match the resistive term and tune out the internal inductance of the mount. Since we are interested only in relatively narrow bandwidths – the whole 47GHz band is only 0.43% wide – the restricted bandwidth introduced by these matching methods is of no consequence. The method is illustrated in the description of a mixer mount for 47GHz which follows later.

20.7.2 Self-oscillating mixers

A receiver can be built using a Gunn diode as both the local oscillator and a mixer – it is referred to as a "self-oscillating mixer". This technique means that a complete receiver (and transmitter, of course) can be built which requires only one expensive semiconductor device. Construction is also greatly simplified – only one diode mount need be built and, of course, there is no need for a cross-coupler.

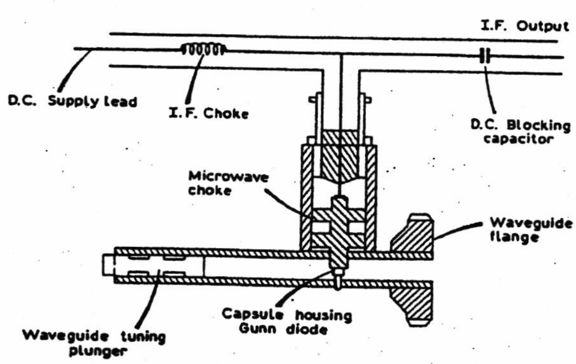


Fig 20.21. Self-oscillating mixer assembly

Fig 20.21 shows the cross section of a 47GHz self-oscillating mixer assembly, with separate dc and i.f feeds. The oscillator uses a plunger tuning arrangement. One problem to watch out for is the occurrence of low frequency parasitic oscillations which are normally suppressed in Gunn oscillators by a shunt capacitor across the choke. A capacitor cannot be used here because it would shunt the i.f signal to ground; in the example given, the Gunn was operated at the highest permissible voltage and this cured the problem. At these frequencies the receive performance can approach that of a separate oscillator and mixer and a system built in this way is an easy way of getting started on the band.

20.8 ANTENNAS

20.8.1 Horns

The simplest antenna to make and use is a horn; quite large gains, eg 30dB, can easily be achieved in a compact size. For higher gains, a parabolic dish can be used. There are other alternatives, eg lenses and microstrip antennas, but these are really only of practical interest to professional users.

Small horn antennas are easy to design, using the design methods described in chapter 4, "Microwave antennas". Being so small, they are not particularly easy to make but are, at least, not as difficult as dish antennas of the same gain!

Methods described for 10 and 24GHz horns can be adapted to make horns for these higher frequencies. A satisfactory method is to make a former from hardwood and build up the horn soldering the four sides on a cut-and-try basis, and finishing off with a wrapping of copper wire or some reinforcing plates at the joint of the horn with the waveguide. The same former can be used for several horns, producing transmit and receive antennas for two or more experimental stations. Some examples of calculated horn dimensions for 47 and 76GHz are given in Fig 20.22.

20.8.2 Dishes

Small dishes are useful as directional antennas but large dishes are hopelessly impracticable. For instance, a gain of 46dB implies a beamwidth of less than one degree in both azimuth and elevation, an accurate knowledge of the location of the station to be worked and a very rigid mounting to enable the antenna to be aligned and kept pointed correctly. Exceptionally precise setting-up would be required to point the dish at the other station, probably involving optical sighting tubes or telescopes. Fig 20.23 extends the range of dish size versus gain to the millimetre bands. Note that the diameter is now given in centimetres, not feet!

Parabolic antennas for the millimetre bands can be cut from solid on a lathe. If a block of metal is used, then the reflector can be used immediately. If a block of hardwood is used then it should be covered in kitchen aluminium foil. The foil should be laid in parallel strips with an overlap of at least λ/4 at the lowest frequency of operation and parallel to the electric plane of polarisation. Make strips wide enough so that nowhere are there more than two layers and do not attempt to lay the strips radially as it is too difficult to avoid a "muddle" at the centre of the dish.

The required accuracy is not very high, λ/10, which is 0.6mm at 47GHz, falling to 0.1mm at 240GHz. This should present no difficulty on a metal-working lathe. If a front feed is used then an f/D ratio of 0.5 is recommended; this is the ratio correctly fed with a sectoral horn that tapers in the narrow face only. Initially, a satisfactory feed can be made for experimental work by just radiating from an open guide. Such dishes for the millimetre bands are so small that it is quite good practice to make them oversized and then "underfeed" them. This produces a highly efficient design in the sense that no rf is "wasted" and the best gain is obtained for a given beamwidth. For example, for an f/D ratio of 0.5, the complete dish is turned and the feed is in the plane of the aperture where it is very easy to support.

20.8.3 Lenses

At millimetre wavelengths, techniques similar to optics become practical. Components are much larger than the wavelength and dielectrics exhibit properties which refract the waves in the same way as light is refracted ("bent") in a lens. Thus, not only will a parabolic dish focus microwaves as a concave mirror does light, so a dielectric "lens" will focus millimetre waves. It might be worthwhile for the experimenter to produce a plano-convex lens by turning a disc of a dielectric such as polythene or perspex and trying this as an alternative to a dish or horn!

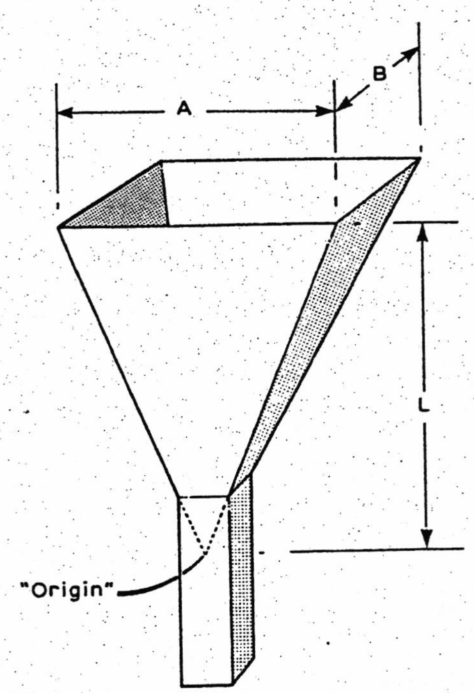


Fig 20.22. Typical horn dimensions for 47 and 76GHz. For 47GHz WG24 is used. 20dB horn: A=28.5, B=22.3, L=33.2; 30dB horn: A=88.6, B=71.5, L=385.7; For 76GHz, WG26 is used. 20dB horn: A=17.6, B=13.8, L=20.4; 30dB horn: A=54.8, B=44.2, L=238. All dimensions in mm. Optimum gain and sectoral horns can be easily designed using the program in Chapter 4, "Microwave antennas"

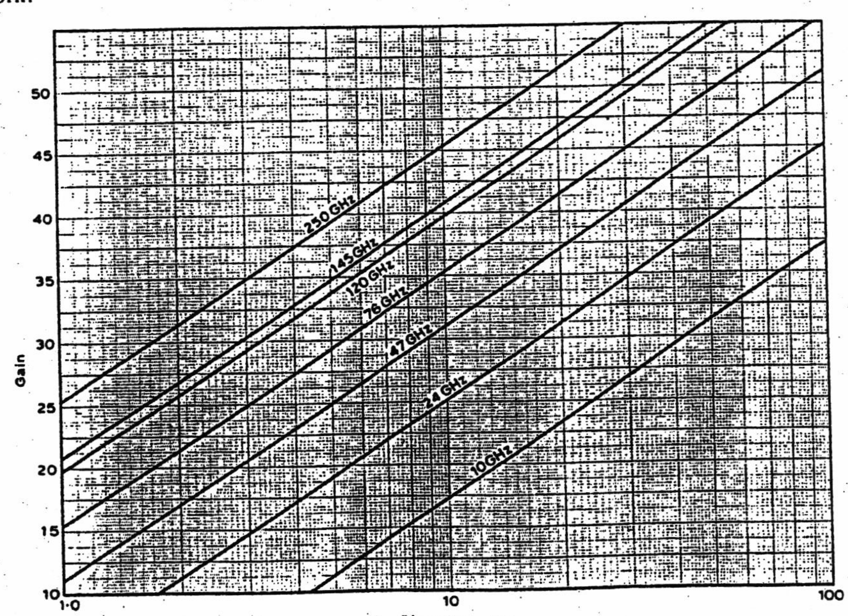
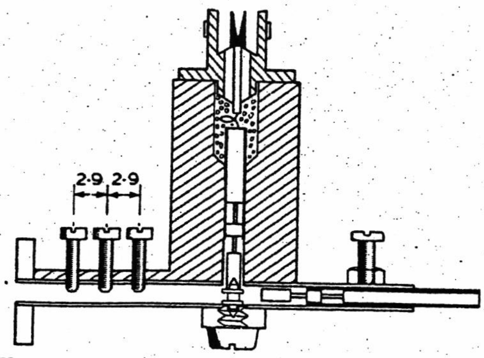
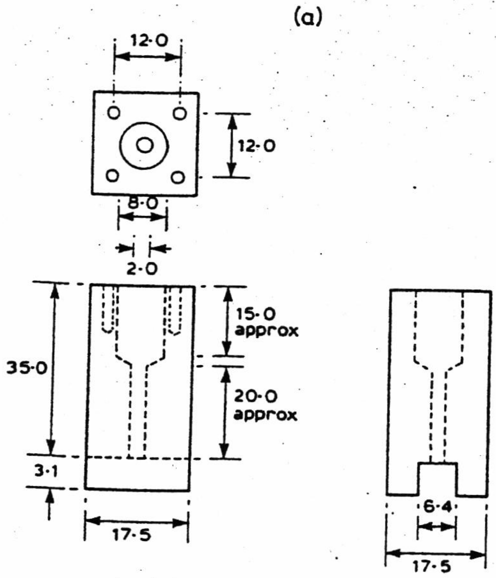


Fig 20.23. Gain of various sized dish antennas



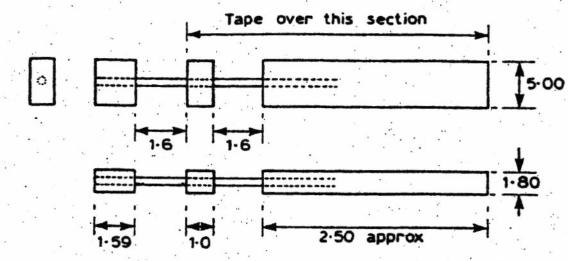
NOTE:
The space between the output choke and the co-axial connector is filled with foam rubber to provide a spring for the diode mount.

(a)

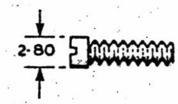


(b)

Fig 20.24. (a) and (b): A mixer/detector for 47GHz



(a) Sliding short tuning plunger for waveguide mixer



(b) Matching screw with turned down head

Fig 20.25. (a) and (b): Detail of tuning plunger for the 47GHz mixer/detector

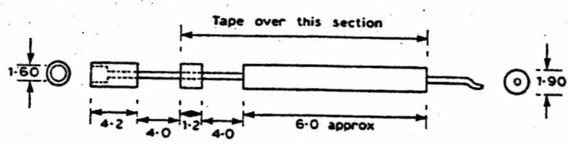


Fig 20.26. Choke for the 47GHz mixer/detector

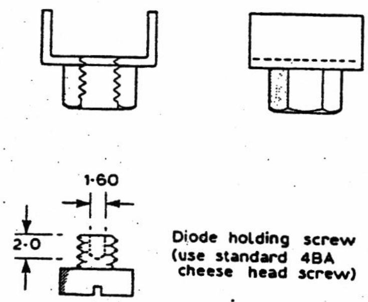


Fig 20.27. Screw to hold the diode for the 47GHz mixer/detector

20.9 EXAMPLES OF EXPERIMENTAL EQUIPMENT FOR 47GHz

20.9.1 Availability of components

The bands immediately below the four principal amateur bands are assigned to the fixed and mobile services while the upper sections of the bands are shared with radiolocation. It is thus to be expected that components will be developed that can be adapted to conventional station-to-station QSOs in the lower parts of the bands and that it will be worth watching the "surplus" market for components suitable for wideband operation in the higher parts of the bands.

20.9.2 A mixer/detector for 47GHz.

This design is based on old designs for 10GHz, but using a "pill" diode instead of a 1N23/1N415 type diode. Fig 20.24 shows the general assembly and the loose components are shown in Figs 20.25, 20.26 and 20.27. The waveguide is soldered to the end of a squared-up brass block. The drawing shows the waveguide set into a slot cut into the block, but this is not really necessary. The size of the block is not important; the sizes shown are suitable for a BNC connector and can be adjusted to suit a different connector. The minimum height is 25mm to allow for the coaxial choke in the output lead and the spill on the output connector. A piece of brass sheet is also soldered to the top of the waveguide at the input to provide thickness for the threads of the matching screws. A flange, or other connector, should be soldered in place at the same time.

The sliding short comprises two sections of high/low transformer, the first of which has an air dielectric and the second plastic. It is backed up by a larger piece of metal for convenience of handling. The prototype was made of brass. The block is first filed up in the solid to be about 0.1mm (say 5 thou inch) smaller than the inner dimension of the waveguide you are using. The dimensions given are for K & S Metals size 264 tubing. The changes for "professional" waveguide are obvious. Do not be too concerned if these dimensions are not precise; what is important is that the sides of the block are parallel.

Next, the block is drilled lengthways with a number 74 drill (for 22swg wire) or 0.5mm drill (for 0.5mm wire). First select your wire and then drill to suit. It does not matter if the hole is off-centre (in fact the block is easier to assemble if the hole is not exactly on the centre line) but do try to make the hole parallel with the long axis. Cut off a small piece of the block and file it down to 1.59mm long. It is worth taking time and care over this dimension, making sure that the block is not tapered. This is one of the reasons the original block was made too long; you can afford to throw away your failures! Make two of this length. Cut off another small piece and file down to 1mm. This component is less important electrically, its main job is for mechanical support.

Now re-assemble on a piece of wire as shown in Fig 20.25. The best method is to apply a touch of solder paste to the parts to be joined and to apply a hot iron to the remote end of the big block and allow the heat to work its way through. Do not underestimate the time it takes for the block to heat and cool. Clean up the block, eg with a nylon burnishing pencil available from model shops.

Next, bend and twist the end sections gently until they look straight and wind on a few layers of thin, self-adhesive plastic tape. The tape should preferably be ptfe, but polyester or even Sellotape will do. Here, accuracy of the end section tells; the better it is, the less important is the dielectric used in the back section. Don't use double sided sticky tape. Now peel off the tape, one face of the block at a time, until the main block is a nice, smooth fit in the waveguide and make final adjustments to the alignment of the end sections so the whole block moves smoothly. The point here is to test the size using the large block and "tweak" the smaller blocks to line up. Do not cut extra tape from the small blocks to make them slide. Now cut away the tape from the end block so that this block is left with a 0.05mm air gap between it and the waveguide.

Next, we tackle the output choke. Because a coaxial transmission line of too large diameter will permit a waveguide mode so that rf bypasses the filter, we are limited to a maximum diameter of 2.0mm. We first drill through the whole block and the waveguide with a 2mm drill. Next open out the top of the block with a larger drill to clear the connector you are using and to provide for the compression pad. The choke inner is turned from brass to a diameter of 1.9mm and while the piece is still in the lathe, a hole (drill 71 or 0.5mm as before) is drilled the full length. Next, pieces are cut off and faced off to lengths of 4.18 and 1.0mm. The three pieces are then assembled on to a straight piece of wire, which at this stage can protrude at both ends, and soldered up. The wire extended at the connector end becomes the output terminal and is cut to length. The piece is then returned to the lathe and the copper wire faced off flush with the brass end and touched with a centre drill. Now drill a 1.6mm (1/16in) hole, 2mm deep, into the end of the piece to receive the end of the diode. As before, wrap tape around the piece until it is a smooth fit into the 2mm hole and then cut away the tape as shown in Fig 20.26 to form the complete choke section. By now you will have noticed that the 2mm hole is too small to permit assembling this mixer by dropping the diode down the output tube, as we do with a 1N23/1N415 diode mixer on 10GHz. Thus a detachable diode mount is needed. This is a 4BA screw fitted to a nut soldered to the face of the waveguide. First measure a few nuts and pick half a dozen or so with the same thickness; these will be our stock of such mounts. Cut off a short (10mm) piece of waveguide, drill a 3.1mm hole (size 32) in one face and cut away the other face, taking care not to distort the piece. Tap the hole 4BA and using a chrome plated or rusty steel screw to align the nut, solder this to the outer face of the waveguide. Ease out the end of the nut with a 4.0mm drill (size 26) so that a 4BA screw will screw down hard and now you will have the jig shown in Fig 20.27(a). This will enable you to file down a 4BA screw to the required length. Do this and put the screw in the lathe and drill the 1.6mm hole, 2mm deep, as shown in Fig 20.27(b). Drill and tap 10BA for the matching screws and fit the nut for the clamping screw for the sliding short, once again using an untinnable screw to locate it.

Finally make some matching screws. The heads of the screws must be turned down so that they can be spaced 2.9mm ($3\lambda/8$) apart without fouling each other. 10BA screws are preferred, but 8BA are just possible if the head is turned down to the same diameter as the thread.

Testing and setting up

The instructions for setting up assume you have no source of 47GHz, but do have a source at 24GHz. There is no need, yet, to re-tune it to 23.5GHz. Set the matching screws to be flush on the inside of the guide. Connect the most sensitive meter you have to the output of the diode unit and offer it up to the open end of the source waveguide. Adjust the sliding short for maximum reading; if you are lucky you will find a series of settings about 4mm apart corresponding to half wavelengths in the guide. Choose the setting nearest to the diode and clamp the short in place. Finally set the matching screws for maximum meter reading.

If your source has a high harmonic output, you may be confused by the presence of 72GHz, and the setting of the short may not be too obvious. Move the mount as far from the source as possible while still getting a reasonable meter reading, and draw up a graph of the meter reading against position of the short. You should then see a pattern and be able to select the best working position.

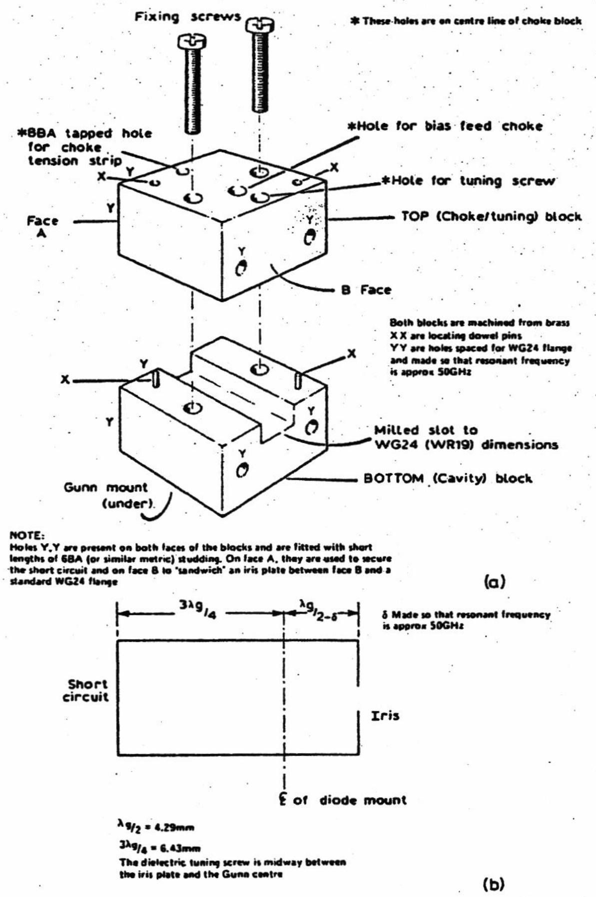


Fig 20.28. (a) and (b): 47GHz Gunn oscillator construction. General view of oscillator block

Finally, retune the oscillator to 23.5GHz and repeat all the optimisations described above for the new frequency, which should now lie in the 47GHz band.

20.9.3 An experimental Gunn oscillator for 47GHz

The following design ideas (due to G3WDG and G8AGN) are mainly unproven in practical operation. The description is given to indicate the type of approach which may be feasible for the more advanced microwave constructor but the design cannot, at the moment, be considered as a really practical one. First, the cost of the Gunn devices is very high, their availability low and their robustness suspect. Second, the dimensions and form of construction are critical, as was indicated by the findings on a prototype oscillator of somewhat different construction. The first prototype used a milled cavity with internal dimensions similar to WG22, but with rounded corners. It was found to be off frequency, despite careful construction, and unstable.

The tests carried out suggested the dimensioning of both the choke feed and the profile of the cavity to be very critical and, while some rf power was generated in the vicinity of 42GHz, there was considerably more power being developed over a range of undefinable (parasitic) frequencies. Without spectrum analysis available, it was assumed that the parasitic output was similar to that occasionally seen in 10GHz oscillators of early design, where at some bias and tuning settings the device output breaks up into what is best described as rf "white noise" centred roughly around the cavity resonant frequency but extending for perhaps several hundred MHz in a 10GHz oscillator.

The main features of the second design is shown in Fig 20.28(a) and (b). A completely different method of construction has been adopted to ensure *exact* cavity dimensions, with the cavity reduced in size to that of WG24. The choke bias feed arrangement of the first prototype was retained and the general format of the oscillator is similar to that commonly used in the recent iris coupled designs for 10GHz Gunn oscillators described in chapter 18.

It must be stressed that at the time of writing the second prototype oscillator has not been tested (another difficulty in millimetre band construction!), so the design is not offered as a practical solution, but solely to indicate methods of construction which might produce the right results. An accurate milling machine, a model maker's lathe, accurate measuring instruments and considerable machining skills are *essential* before tackling a job of this nature.

First, a piece of suitable brass stock is machined into a right-angled block with dimensions 10.43mm by 25.4mm by, say, 29mm long, with the first two dimensions accurate and flat, as these are the finished dimensions of the oscillator block. Next, four holes, D and E, are drilled to the sizes in Fig 20.29, to a depth of about 20mm. Initially, holes D should be 6BA tapping (or a comparable metric size, not critical in size or placing) and holes E of a size to suit any dowel pins available. The centre line of the long dimension is now scribed around the block and small punch marks made on the block, one each side of the scribe line. These will identify which way round the blocks will fit after the next operation. Now carefully saw the block into two smaller, equal sized blocks. The newly cut faces of each block are milled, flat and square, to the final dimensions ie so that the height of each block is 12.7mm. The two blocks should now measure 25.4mm by 12.7mm by 10.43mm. The two blocks should fit closely together if the job has been done correctly. The block which has holes right through it is the choke block (A) and the other, with blind-ended holes will be the cavity block (B). Holes D in the choke block should now be opened up to clearance size for the bolts to be used and then three more through-holes of the sizes and centres shown in Fig 20.29 drilled in this block. Note that the sizes and placing of the holes A and B are critical, hole C is not.

The cavity block (B) now needs more work. First, holes D are tapped to the appropriate size, taking care to avoid the "bell mouth" effect which was described in chapter 7, "Constructional techniques". The block is returned to the milling machine and a channel 4.775mm wide and 2.387mm deep (critical) is milled centrally across the face which mates with the choke block, as shown in Fig 20.30. Finally, the diode mount collet hole, detailed in Fig

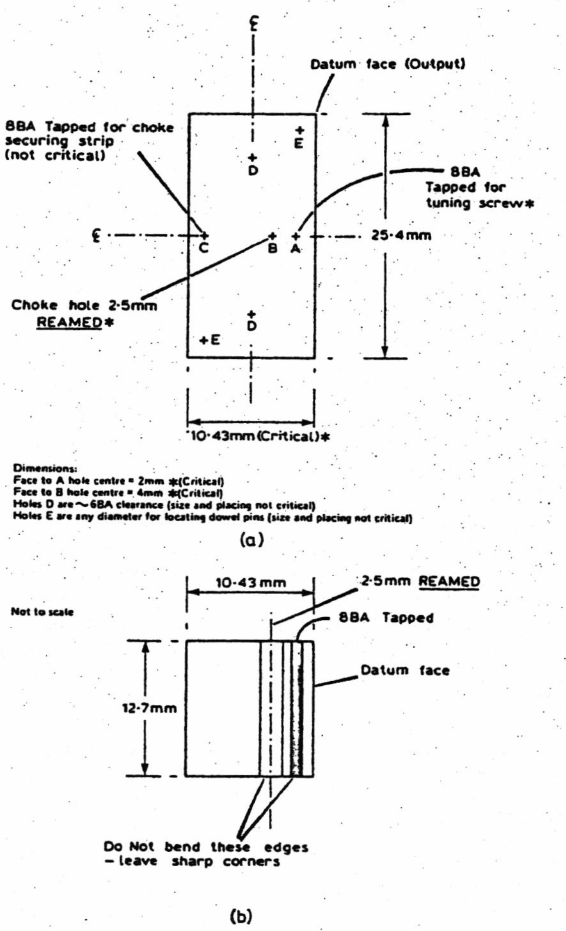


Fig 20.29. (a) and (b): Machining of brass block for the 47GHz Gunn oscillator - choke block

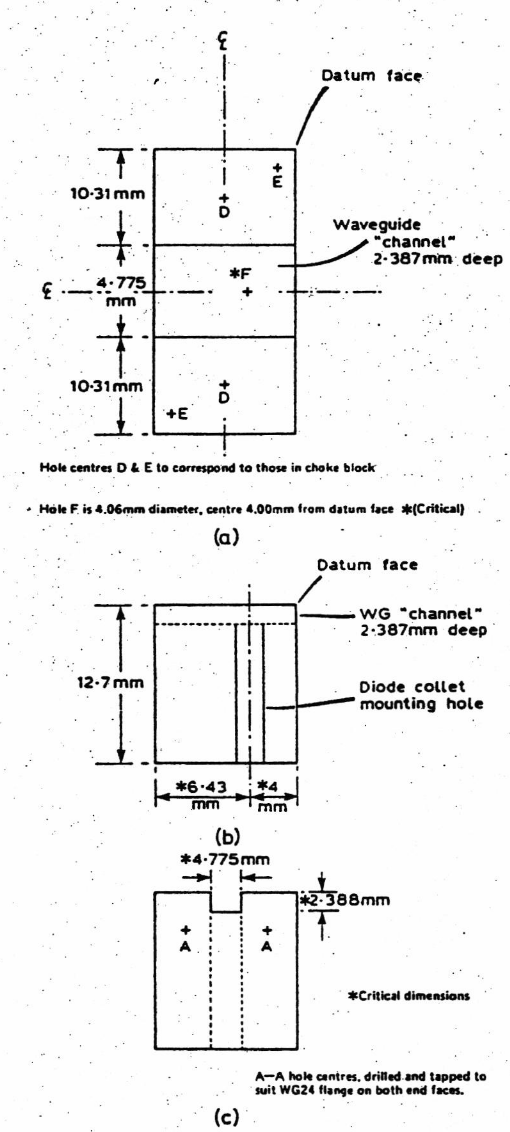


Fig 20.30. (a), (b), and (c): Detail of the cavity block

20.31(b) is drilled and tapped in the position shown in Fig 20.29(b). During milling of the channel it is *essential* that sharp, right-angled corners are achieved.

The dowel pins are gently driven into one of the blocks; the corresponding holes in the other block should be enlarged very slightly to avoid the pins binding when the two blocks are fitted together. Now try the blocks for fit and temporarily bolt them together: they should be a very neat fit! Whilst fitted together, mark the centres for the waveguide flange bolts on each of the two outer faces of the waveguide channel in the assembled block. The holes should be drilled 6BA tapping (or similar metric size) to a depth sufficient to take a short length of studding, which can now be cut, fitted and carefully soldered in place.

A cavity back-plate or short circuit is made from thick (eg 4mm) brass or copper plate, drilled with 6BA clearance holes to centres to correspond to the studs. In the final assembly it is held in place with nuts and shake-proof washers. A diode mounting adapter collet is made from *copper* (heat conduction is important) to the dimensions of Fig 20.31(a) and, on completion, is checked for diode fit and fit into the cavity block mounting hole. At each stage of all the above operations care must be taken to ensure clean, flat, unblemished surfaces on the blocks and to removal of swarf and burrs from surfaces and holes.

The next operation is to fabricate several choke "pins" to the dimensions of Fig 20.30(c). It is suggested that a series of pins be made with the diameter of the diode contacting pin varying between 2.6 and 1.2mm in 0.1mm steps, because it is believed that this dimension is critical and best selected on test. The length of the contacting pin should be

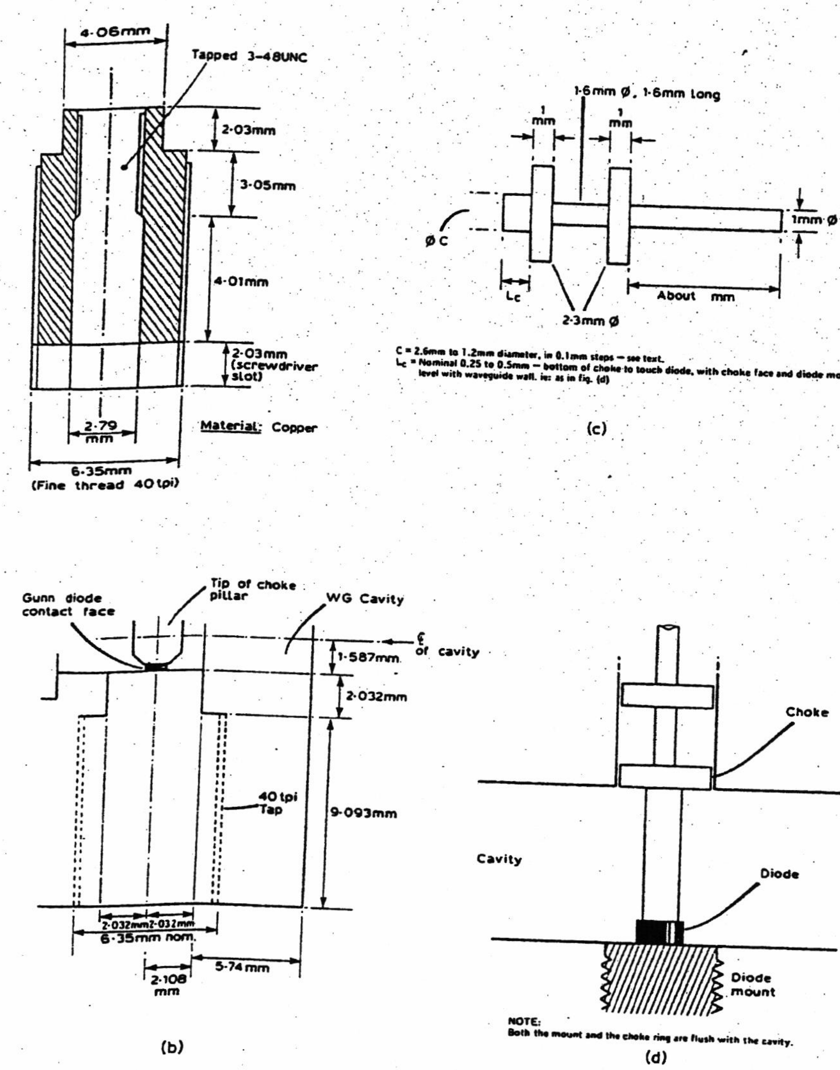


Fig 20.31. (a) Detail of the Gunn diode mounting screw, (b) detail of the diode mounting hole in the oscillator block, (c) detail of the choke, (d) Gunn diode and choke mounting positions

adjusted so that the face of the diode is contacted with the diode and choke in the positions shown in Fig 20.30(d).

Insulation of the choke was provided by means of 0.1mm thick shrink sleeving fitted over the choke after completion. The overall diameter should be a sliding fit into the choke block hole and the choke is held in place by a thin plastic strip attached to a screw fitted into the rear tapped hole in the choke block; contact pressure must be small otherwise the diode may be damaged.

An 8BA tuning screw is fabricated either from ptfe or from a brass screw with a ptfe insert fitted into a small centre-drilled hole in the end of an 8BA screw of appropriate length. The iris plate, which completes the assembly, is made from thin brass shim the same size as the standard flange, drilled accordingly and sandwiched between the output end of the cavity and the output flange. The size of the iris hole might start at 0.5mm diameter and be enlarged, on test, to provide the best compromise between power output and stability, as with the corresponding 10GHz designs. A photograph of the completed prototype module is given in Fig 20.32. This should give some idea of the size and general form of construction of the oscillator.

20.9.4 Alternative approaches to 47GHz

The following is an abstract of the ideas used by HB9MIN and HB9AMH [2] which led to the establishment of the first world record for amateur communication at 47GHz, of slightly more than 50km, which stood from 1984 until August 1988. One of the main problems for the amateur (and, it seems, to many professionals too!) is access to suitable test equipment; some home-made test equipment is described in outline in this section.

Three possibilities for the generation of 47GHz signals were considered and examined:

Fig 20.32. Photograph of a prototype 47GHz Gunn oscillator (G8AGN)

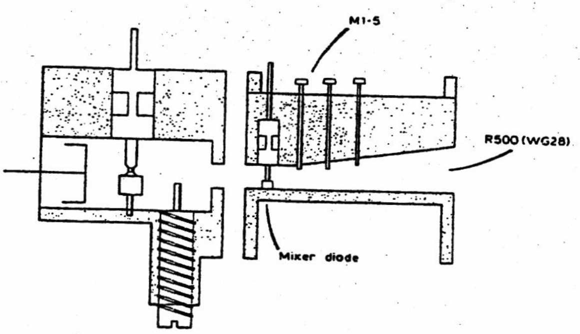


Fig 20.33. Use of 23.5GHz oscillator followed by a doubler (HB9MIN)

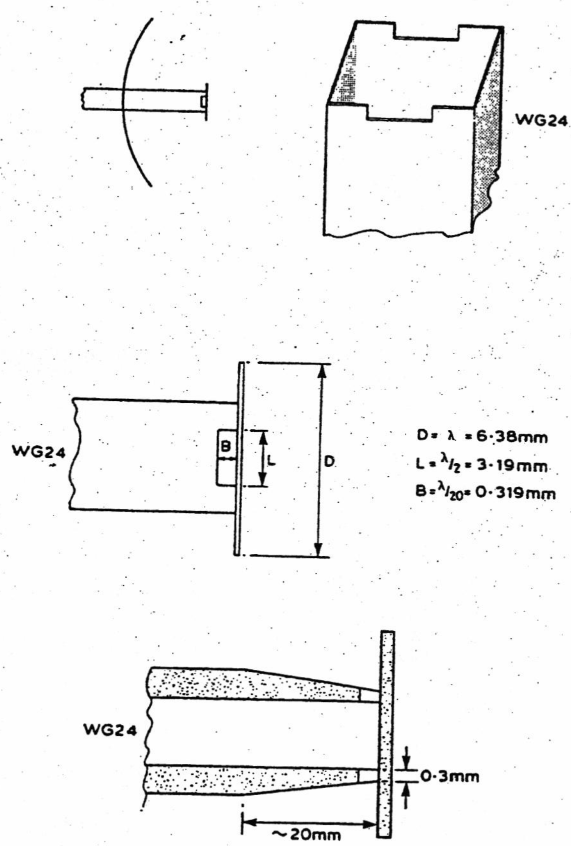


Fig 20.34. Dish feed for 47GHz (HB9MIN)

1. Use of the second harmonic from a 23.5GHz Gunn oscillator, with the fundamental suppressed by filtering.

2. The use of a 23.5GHz Gunn oscillator, followed by a Schottky diode doubler.

3. The use of a 47GHz fundamental Gunn oscillator.

The third alternative was immediately dismissed because of the very high cost of Gunn diodes and professional oscillator assemblies. Such oscillators are very temperature sensitive (about 3MHz/°C) and are pulled by at least 100MHz by slight load variation. Exact design frequency is difficult for the amateur to attain because of the unpredictable nature of the effect of the package parasitic elements on the cavity frequency.

Alternative 1 offers perhaps 1 or 2mW output from a commercial 24GHz oscillator; the temperature effect is typically 1.5MHz/°C whilst frequency pulling will be around 10MHz, both figures being much better than with a 47GHz oscillator.

Although alternative 2 is possible, the increase in power output did not seem to be worth the effort of mounting an expensive diode in a special mount which tapers from WG20 to a reduced height diode section and then back to WG23, especially when the 23.5GHz oscillator could be used effectively as either a sub-harmonic self-oscillating mixer or with a low-barrier or zero-bias Schottky diode in an in-line configuration. Ideally the mixer device should be GaAs because silicon diodes are near their limit, although still just usable.

The equipment used consisted of an M/A-COM 24GHz oscillator retuned for 23.5GHz and a low barrier Schottky diode mounted in WG23 waveguide and used in an in-line configuration as shown in Fig 20.33. The antenna used was a 70cm dish (lamp-shade) with an f/D ratio of 0.3, measured gain of 40dB and fed with a scaled-down version of the "penny feed", shown in Fig 20.34. Great accuracy is needed in the construction of the feed and it may be more practical to use a large horn. The pointing accuracy needed for such a dish is better than 0.5°!

To test and align such transceivers, some form of measuring equipment is essential: professional equipment at these frequencies is not common! Several pieces of home-made equipment are described briefly. The first is an up/down converter based on the self-oscillating mixer principle, the i.f output of which is routed to a counter or spectrum analyser at lower

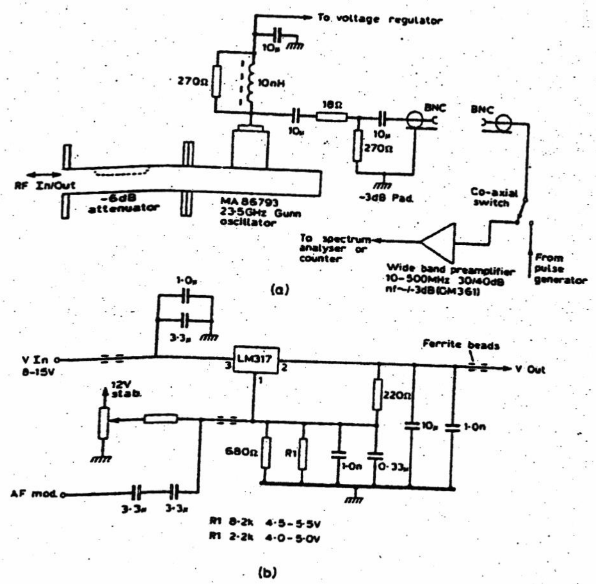


Fig 20.35. (a) Schematic for an up-down converter, and (b) stabilised power supply for the Gunn oscillator. (HB9MIN)

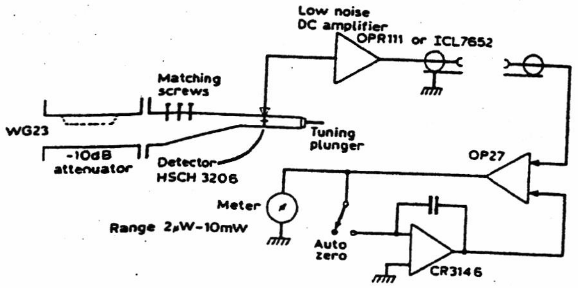


Fig 20.36. Schematic for a power meter (HB9MIN)

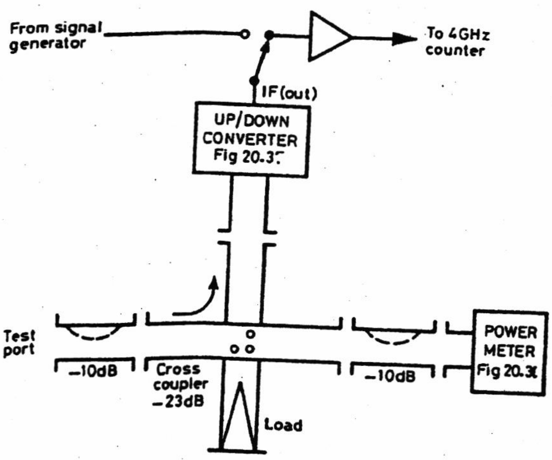


Fig 20.37. Test set up using the power meter (HB9MIN)

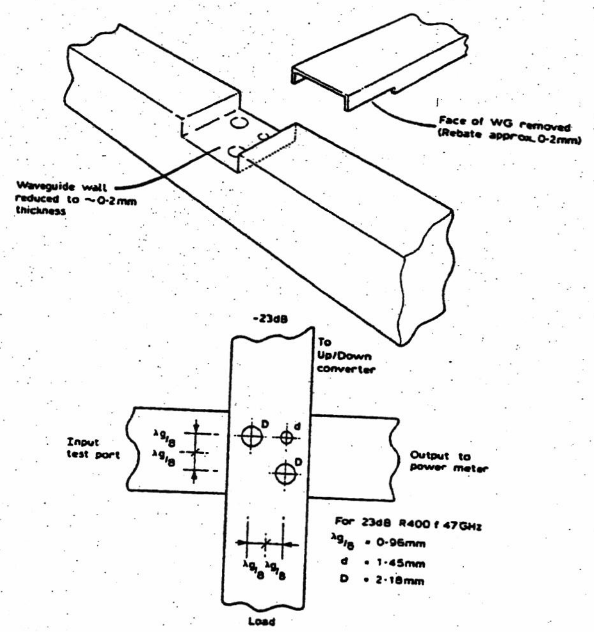


Fig 20.38. Cross coupler dimensions for 47GHz (HB9MIN)

frequency. The block diagram of the converter is given in Fig 20.35. The same converter, when fed from a pulse generator at i.f, will generate test signals in the band.

A power meter was also constructed using a zero bias Schottky diode type HSCH3206, mounted in a reduced height guide, taper matched into WG23 as shown in Fig 20.36. This is normally used with the test set-up shown in Fig 20.37. Fig 20.38 gives some detail of the -23dB cross coupler. This, too, is difficult to make and great accuracy is needed in its construction. It must be pointed out that these ideas, whilst practical, do need a great deal of care and attention in their construction and the degree of inaccuracy which might be tolerable at, say 10GHz, certainly is *not* at 47GHz!

20.10 ACKNOWLEDGEMENTS

The authors would like to acknowledge the assistance of E Zimmermann, HB9MIN, in preparing this chapter.

20.11 REFERENCES

[1] "Propagation above 10GHz – the potential of the amateur mm-wave bands", B Chambers, G8AGN. *Proceedings of the Martlesham Microwave Round Table,* November 1988, published by the Martlesham Radio Society, with acknowledgements to the British Telecom (BT) Research Laboratory, at whose establisment the meeting was held.
[2] "47GHz Praktischer Teil", E Zimmermann, HB9MIN.

Divers

WAVEGUIDES FOR MILLIMETER WAVES

By John Anderson WD4MUO/0

DATA

Attached are mechanical data and drawings for waveguides and flanges used at frequencies between 18.0 and 325.0-GHz [1]. I find this information valuable in building adapters and transitions to connect the odd pieces of guide and devices one collects and must integrate into a system.

ADAPTERS

Much of the mm-waveguide available from surplus sources has round flanges. This was the industry standard in the early days and probably still is at the higher frequencies. Newer equipment tends to use square flanges, particularly at frequencies below 40-GHz. To some extent this has dropped the value of round flange items on the commercial surplus market to where Hams can afford pretty good equipment. However, this presents a significant problem when we are faced with a variety of flange configurations and hole patterns on the individual equipment items that we want to integrate into a system.

The guide pin holes in round flanges are tantalizingly close to the holes in square flanges. This has lead some Hams to simply bore out the guide holes and mate the flanges using through bolts and nuts. It works, but is not recommended as the two sets of holes have a small, but significant offset (see flange diagrams) and it is very difficult if not impossible to drill a hole in the correct position in the flange with the existing hole off-center. Thus some filing and elliptical or oversized holes are necessary, misalignment of the joined waveguides normally results. The situation is more serious when mating choke and O-ring sealed flanges, very little lines up and there will probably be a gap somewhere. The correct solution is to use a flange adapter. Unfortunately these are very rare at hamfests and cost a fortune new (>$100) or surplus (I have found some used adapters that are more expensive than new ones!)

So I build my own. I buy cheap pieces of copper, brass or bronze waveguide or equipment with assemblies of waveguide just to get the flanges and salvage short scraps of waveguide. The more dings, holes, crimps, mud, etc. the better - gives me room to negotiate the price. When I need an adapter I usually select a round flange with about 0.5 to 1.0 inch of good waveguide attached. I square up the open end and silver solder a salvaged square flange to it. If I do not have a commercial square flange at hand I make one from a piece of brass plate 0.125 to 0.25 inch thick. Forget the fancy grooves and milled sockets for the waveguide, just cut a flat flange with a jewelers' saw to fit over the waveguide, solder it and then lap/file the mating surface flat. The flange holes can be threaded or drilled as clearance holes as required (you may want both types of holes on the same flange to mate to a particular device). Give some thought to the length of the waveguide between the flanges. If the bolts will be screwed into the adapter then the flanges can be quite close together. But if the bolts must go through the adapter flange and then into "blind" holes [2] in the mating device, you must have a long enough piece of waveguide to insert the bolt and the wrench/screwdriver to tighten the assembly. One inch between flanges is about right for allen wrenches.

If a very short adapter is necessary you can build a block type by soldering two flanges back to back or fabricate it from 0.25 inch plate. I usually saw a square with sides as long as the diameter of the round flange, saw and file the center hole to the interior dimensions of the waveguide, and then bore and thread the necessary holes from each side of the plate. When necessary it may be possible to countersink or counterbore some holes so all bolts come from the same direction; be careful of mating choke and O-ring groves. Since the true block adapter does not need to be soldered, aluminum is a good choice for the plate.

TRANSITIONS

Now that everyone(!) wants to get on a mm-wave band (24-GHz), WR-42 waveguide and devices have become very expensive - a relay costs the national debt. Several of us [3] have experimented with using the next lower or higher size of available waveguide. WR-62 appears to work at 24-GHz without over-moding (test it, there maybe a mode close to our frequency) and WR-28 certainly works with its cut-off frequency at 21.07-GHz; loss is a little higher than WR-42. WR-28 might work at 47-GHz, I have not tested it. I have been given or paid pennies for several power meter heads (the old, uncompensated type) that appear to be WR-50. These may be a pain to use (now - they were state-of-the-art in 1960!), but a surplus WR-42 head costs about $400. A length of WR-35(?) flex guide was also free, several WR-62 relays were $5 to $20 each. The problem here is getting a transition from one waveguide series to another [4]. Guess what: these are rarer than flange adapters and cost more - try $400. A transition is only a little harder to make than a flange adapter. Take a length of the larger size waveguide, saw four, thin triangles in the broad walls, squeeze a taper (ideally three guide wavelengths long) and put the necessary flanges on each end, then silver solder [Fig. 1]. Takes about an hour to build one and costs a buck or two (depending on how well you negotiate for scrap waveguide). Devices like the power heads may not be flat across the band, but they will be usable. The transitions are virtually lossless and have very low VSWR.

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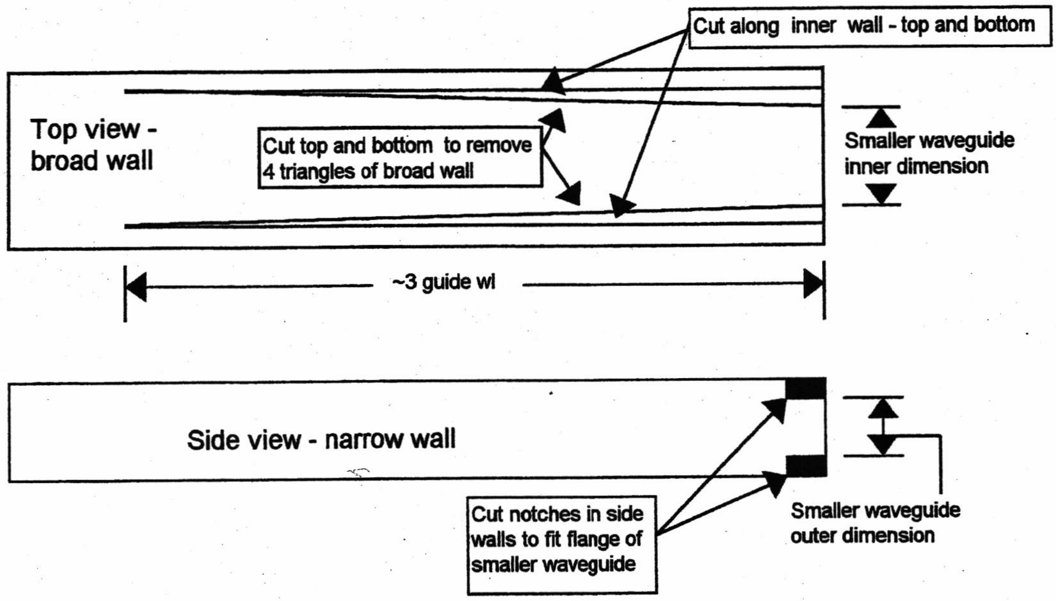
1. I thank Al Hislop of Pacific Millimeter Products for providing the Aerowave inc. catalogue and Ted Kozul of Aerowave for authorizing reproduction of the data pages.
2. Many mm-wave devices have flanges so close to the chassis that it is impossible to get a bolt into the back side of the flange.
3. WA7CJO, AA0BR and WD4MUO.
4. WA5VJB drills matching holes in the flanges and simply bolts the different size waveguides together. It works, but I find transitions somewhat more pleasing, aesthetically if not electrically.

WAVEGUIDE TRANSITIONS



Following preparation of the waveguide as shown above, spring the sides to the proper dimensions for the flange of the smaller series and place the appropriate flanges on each end. Bind the taper with iron wire at several locations to minimize the gap and tie down the flanges so they do not shift when the assembly is heated. Silver solder the flanges and slits with caution, use minimal amounts of solder as excess solder builds up along the inner portion of a slit and is difficult to file out. It is probably easier to close a hole or gap by re-soldering a small portion of the joint than to file out excess solder.

Figure 1

Following preparation of the waveguide as shown above, spring the sides to the proper dimensions for the flange of the smaller series and place the appropriate flanges on each end. Bind the taper with iron wire at several locations to minimize the gap and tie down the flanges so they do not shift when the assembly is heated. Silver solder the flanges and slits with caution, use minimal amounts of solder as excess solder builds up along the inner portion of a slit and is difficult to file out. It is probably easier to close a hole or gap by re-soldering a small portion of the joint than to file out excess solder.

mm- wave Waveguide and Flange data WR 42 to WR 03 (18 to 325 GHz)
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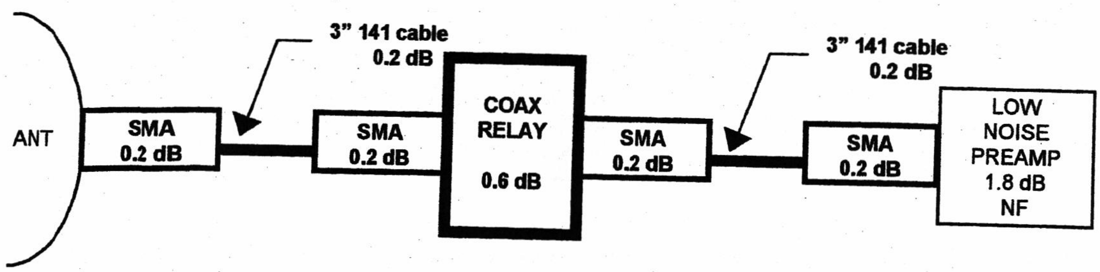
PCB to WAVEGUIDE TRANSITIONS for MILLIMETER WAVE BANDS

By John Anderson WD4MUO/0

One of the items I find puzzling and ironic in Amateur construction of mm-wave (and some SHF) receivers is our obsessive quest for minimum noise figure in our preamplifiers and then placing these state-of-the-art devices after two lengths of high loss coaxial cable, four connectors that are not designed for mm-wave bands and a coaxial relay of questionable lineage. On the VHF/UHF bands we continually preach 'no loss before the preamp' and go to great heights to mount the preamp *at* the antenna, but somehow forget this principle at the higher frequencies. At mm-wave frequencies, *close* to the antenna is not good enough!

Insertion loss for SMA relays at 18 to 26.5-GHz is specified by several manufactures at 0.6-0.7 dB. Rule of thumb for loss in a SMA connection at 10-GHz is 0.1 dB, it is probably at least twice this at 24-GHz. At 18-GHz the loss in 0.141 semi-rigid cable is 0.68 dB/ft, for 0.085 cable 1.1 dB/ft, it will be higher at 24-GHz [1]. Examining a typical front end at 24-GHz we find the following estimated losses:



Thus, the total loss between the antenna and the very 'hot' preamp is estimated at 1.8 dB which adds directly to the noise figure. The **system noise figure** now starts at 3.6 dB, to which one must factor in the contribution of following stages. It would probably be generous to place the system noise figure at 4.0 dB - it could be much worse.

For WR-42 waveguide the loss is 0.2 dB/ft. The loss in a waveguide relay is similarly low. Several Amateurs (DB6NT, JE1AAH and W0EOM, that I know of) have used direct preamp circuit board to waveguide transitions in an attempt to overcome these losses. However, a standard design does not appear to exist in Amateur literature.

With this in mind I approached Al Hislop of Pacific Millimeter Products who most kindly provide for Amateur use the following design information on circuit board to waveguide transitions that he uses in his commercial products.

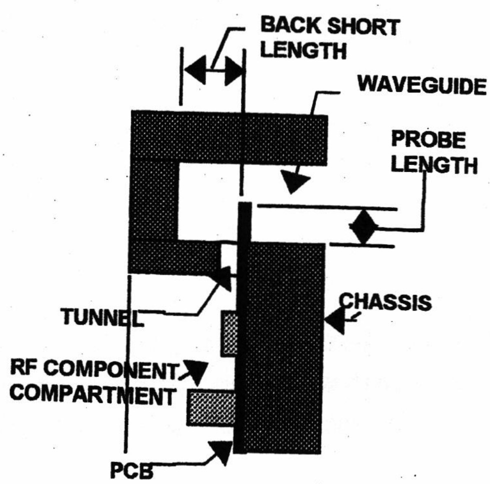
Circuit Board to Waveguide Transitions:

For waveguide with a 2:1 aspect ratio use a probe (strip line) Zo of 94 ohms.

Waveguide	Probe Length	Back Short
WR-28	0.080 in.	0.105 in.
WR-22	0.064	0.080
WR-19	0.056	0.064
WR-15	0.042	0.053
WR-12	0.035	0.047
WR-10	0.027	0.038
WR-8	0.022	0.030

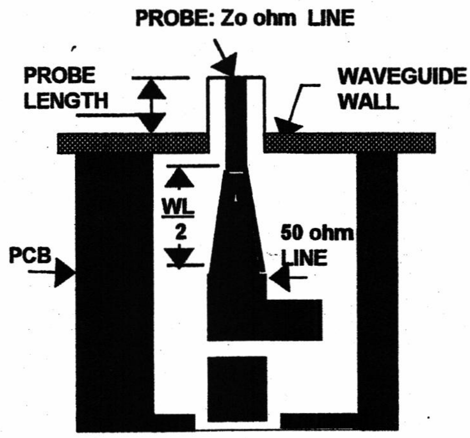
For waveguide with the aspect ration of WR-42 use a probe Zo=85 ohms.

WR-42	0.118	0.140



Transition Cross Section

Pacific Millimeter designs use suspended substrate stripline, a circuit board without ground plane backing that has a milled channel in the chassis above and below the circuit trace. From previous research, Mr. Hislop believes that Amateurs should be able to produce transitions in the more familiar microstrip using the above dimensions. The probe trace width for the specified Zo is calculated from the dielectric constant and thickness of the circuit board being used. The line should taper from 50 to probe Zo ohms over as long a section as possible, one half wavelength is the minimum. The ground plane of the circuit board should end at the waveguide inner wall with no projection into the waveguide. A tunnel (outer to inner wall) approximately three times the pcb thickness (3x h) high should be over the stripline at the entry point. The tunnel width is not critical, several times the trace width is suggested. Either side of the probe PCB can face the waveguide short as long as the back short length is measured to the metallic surface of the probe. In fact the probe width can be perpendicular to the back short. However, this introduces ambiguity in the measurement of the back short length and the data given above should be modified by experiment.



Typical PCB

Mr. Hislop did not have data for larger waveguide, but indicated that designs could be scaled from WR-28 using Zo=94 ohms for WR-75 and WR-62 as they have a 2:1 aspect ratio.

This design data is based on achieving a best match to 50 ohm line over the frequency range covered by the waveguide. Tuning may improve the match at a specific frequency. It is possible to calculate data to match lines with impedances other than 50 ohms.

It is probable that the chassis for an integrated preamplifier/waveguide transition can be fabricated using the sandwich technique described elsewhere in these proceedings [2]. Splitting the chassis at the plane of the PCB is suggested.

My sincere appreciation to Mr. Hislop for making the above data available to the Amateur community and for reviewing this paper.

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[1] Please indulge my estimates. I was not able to find data for cables and connectors at 24-GHz and my test equipment is not calibrated at this frequency.
[2] Anderson, John, "The Brass Sandwich: An Alternative to Milled Box Chassis," Proceedings of Microwave Update 1998.

Les O.L.
Les O.C.X.O.

Voici l'OCXO 106,5 ou 108 MHz de F5AYE avec tous les éléments pour le réaliser, schémas, nomenclature, dessins des circuits imprimés, implantation des composants, dessin de la partie mécanique et photographies.

OCXO 106,5Mhz et 108Mhz par F5AYE

Voilà un OCXO de plus ! Pourquoi ne pas avoir utilisé les descriptions déjà parues : elles nécessitent toutes un boitier cuivre pas facile à réaliser, celui-ci est réalisé avec du matériel de plombier. Résultats après 2 mois de vieillissement le passage de 23° à 3° (dans le réfrigérateur) donne une dérive de 2 Hertz. Soit 0,02PPM. D'autre mesures seront faites prochainement.

Boîtier et régulateur de température :

J'ai utilisé comme boîtier un tube de cuivre de 26mm intérieur sur lequel un fil résistant est collé entre 2 couches de ruban en capton.

Le système de régulation en température comporte un capteur LM 335 (fixé au tube de l'OCXO)qui génère une tension proportionnelle à la température (10mV/degré). Un modulateur d'impulsions (1/2 LM392) compare la tension du LM335 à la tension divisée de la référence de tension LM336.5. Les impulsions sont amplifiées jusqu'à un BD139 (régime saturé) qui alimente la résistance de 25 ohms enroulée autour du tube.

Avantages : une répartition uniforme de la zone chauffée, une régulation "nerveuse", une inertie thermique entre le tube régulé en température et le print de l'oscillateur ce qui peut amortir les éventuelles variations de température du tube.

La référence de tension LM336.5 ne doit pas être soudée directement sur le CI, mais reliée via des fils souples de façon à être disposée contre le tube de l'OCXO. Ceci permet d'améliorer la stabilité de température, la réf. de tension étant un peu sensible aux variations thermiques.

Température:

Le montage décrit ici fonctionne à 40 degrés et est utilisé avec un quartz taillé pour 20 degrés. Si l'on désir l'utiliser avec un quartz taillé pour 60 degrés, il faut jouer sur le rapport R4 1K et R5 3,3K de façon à "monter" la tension de 200mV.

Oscillateur :

L'oscillateur est inspiré d'un L.O. de DD9DU, une varicap a été ajoutée pour retuner la fréquence après vieillissement du quartz (10 tours du potentiomètre donne 250 Hz à 106Mhz).

Suit un buffer avec un BF199 puis un atténuateur en T pour isoler l' oscillateur des variations de charge. Il est réalisé en résistances SMD (Les valeurs des résistances sont à calculer en fonction du niveau mesuré et du niveau désiré).

Un filtre bouchon suit pour atténuer l'harmonique 2, les harmoniques sont toutes inférieures à 50dB.

Montage :

Boîtier en tube :

Constitué d'un tube cuivre de 26mm intérieur et de bouchons (cap chez les plombiers) qui sont raccourcis à 9mm. Sur le bouchon avant, souder une SMA et un bypass de 1nF, sur le bouchon arrière percer un trou en face de l'axe du potentiomètre multitours. Un adhésif capton de 8mm (1)est enroulé en spirale sur 5 tours avec un espace de 1 mm, aux 2 extrémités fixer une borne isolée.

Enrouler sur le capton 5 spires de fil résistant 100 Ohms/m (2) et souder les extrémités aux bornes. Recouvrir le fil d'une nouvelle couche de capton (une goutte de colle à 2 composants peut protéger l'extrémité du fil entre la borne et le capton).

Au milieu des 5 spires percer un petit trou pour une vis paker de 2,2 qui fixera un cavalier pour maintenir le capteur LM335. Dégager le capton sous le capteur et mettre de la graisse thermique.

Le CI est maintenu par 2 colonnettes de 10mm. La liaison entre le CI et la prise SMA doit être faite par 2 courts fils souples de façon à ne pas créer de liaison mécaniques entre le bouchon, la première version où le print était soudé au bouchon était moins stable.

Isolation thermique :
Le tube est placé dans une mousse isolante (Style polystyrène expansé) entouré d'une feuille découpée dans une "couverture de survie".
Le tube ainsi que la platine de régulation sont placés dans une boite en aluminium moulée (chez différents fournisseurs de boîtiers). L'alimentation DC est faite au travers d'un by-pass et la sortie HF au travers d'une SMA de châssis.

Oscillateur :
Le circuit réalisé est un double face avec la face supérieure comme plan de masse. Les passages des pattes composants seront chanfreinés.
Le boîtier du pot Neosid sera soudé au niveau de ses pattes à la face supérieure et côté inférieur une patte doit être soudée pour la continuité de la masse pour le potentiomètre.
Le boîtier du quartz doit être soudé en un point au plan de masse du circuit.
Dans un premier temps ne pas monter l'atténuateur en SMD mais faire un pont.

Réglage:
Mettre le potentiomètre en position médiane, régler le noyau pour amener l'oscillation à la fréquence désirée, régler le circuit du BF199 pour un maximum de niveau. Avec un analyseur de spectre régler le circuit "bouchon" de façon à minimiser l'harmonique 2, puis reprendre le réglage du circuit du BF199 cela plusieurs fois de suite.
Mesurer le niveau, en fonction de la puissance désirée, calculer la valeur de l'atténuateur puis supprimer le pont et souder les résistances SMD de l'atténuateur.
Eventuellement retoucher la fréquence avec le potentiomètre.

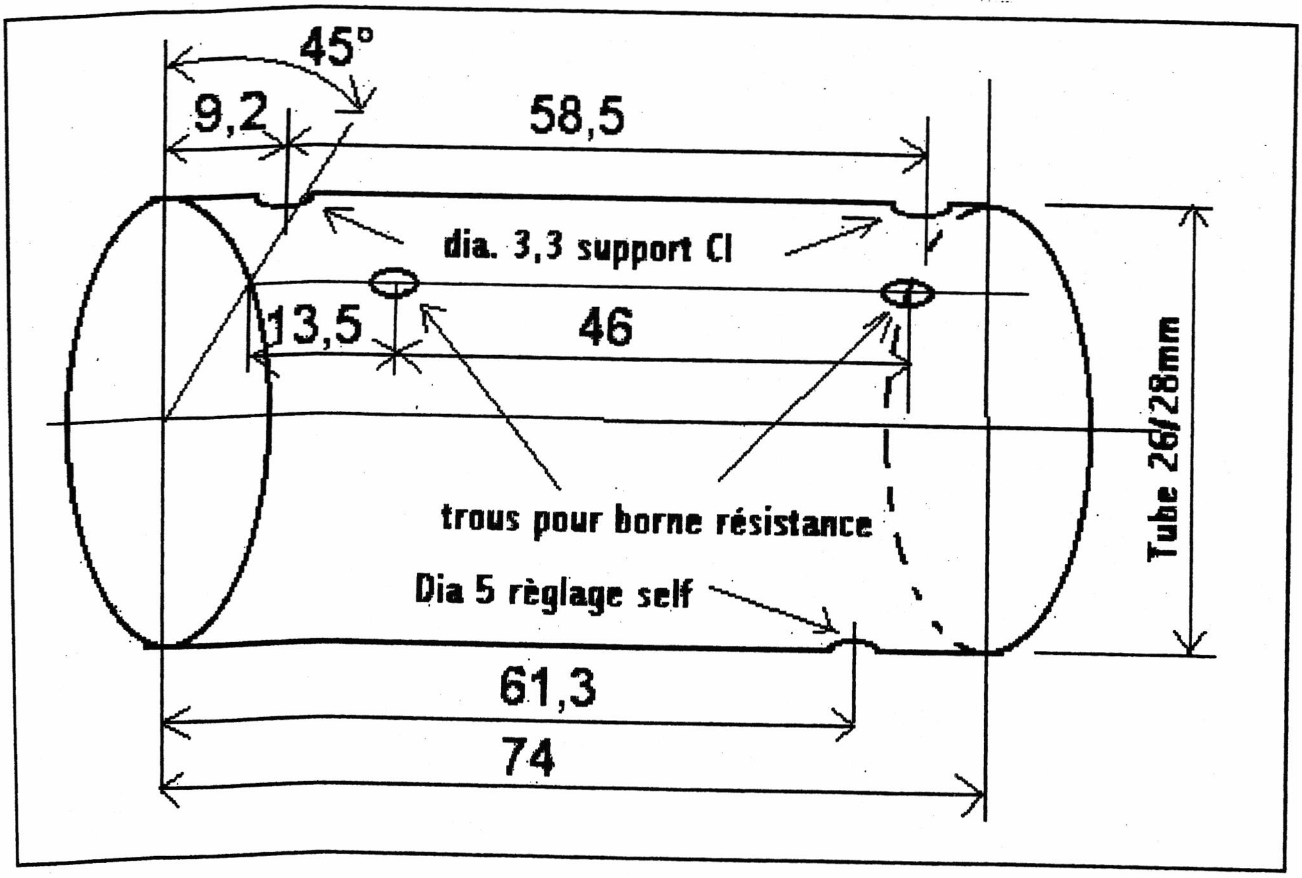
Notes:
1: Je peux donner du ruban capton auto-collant.
2: Le fil résistant et tous les composants du chauffage sont dans le catalogue "Conrad".
Photos sur mon site : **http://perso.wanadoo.fr/f5aye/**

Avertissement: ce montage a été conçu par un "amateur" dans tous les sens du terme et n'est sûrement pas fait suivant les règles pro. Seule garantie : 3 pièces fonctionnent bien.

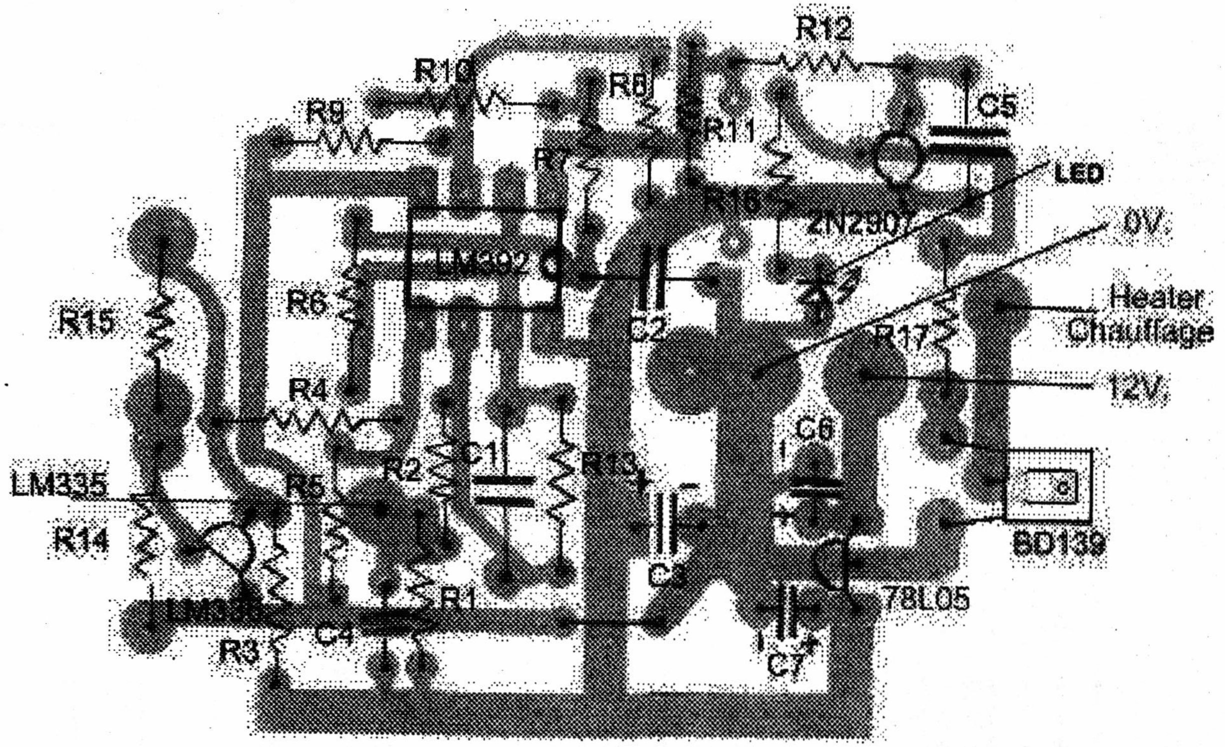
Bonne réalisation et finis les RDV à + ou – X kHz
73 Jean-Paul F5AYE

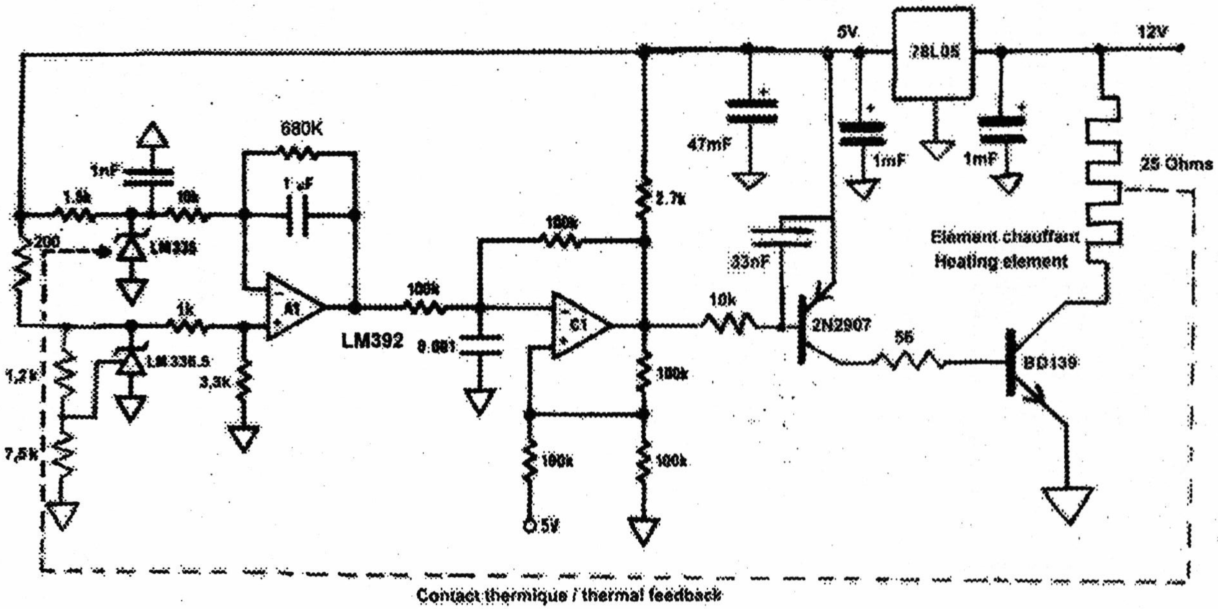
Liste composants OCXO		
Référence	Composant	Remarques
P1	5K	Pot. multi axe horizontal
C1	1microF	Tantal
C10	15pF	Coef. Temp. -750
C11	82pF	Coef. Temp. -150
C12	2-22pF	Plastique bleu
C13	1.6-15pF	Plastique vert
C2	1microF	Tantal
C3	1nF	
C4	1nF	
C5	1nF	
C6	1nF	
C7	10nF	
C8	1pF	
C9	1pF	
CI1	78LO9	
D1	BB505	
L1	Neocid 5061	
L3	10microH	
Q1	Quartz	106.5 ou 108Mhz
R1	220	
R2	100	
R3	4.7K	
R4	47K	
R5	220	
R6	A calculer atténuateur	SMD
R7	A calculer atténuateur	SMD
R8	A calculer atténuateur	SMD
TR1	J310	
TR2	BF199	
	L1	0.8mm 8 TOURS DIAM.5 prise à 2 tours coté C7
	L2	0.8mm 6 TOURS DIAM.6

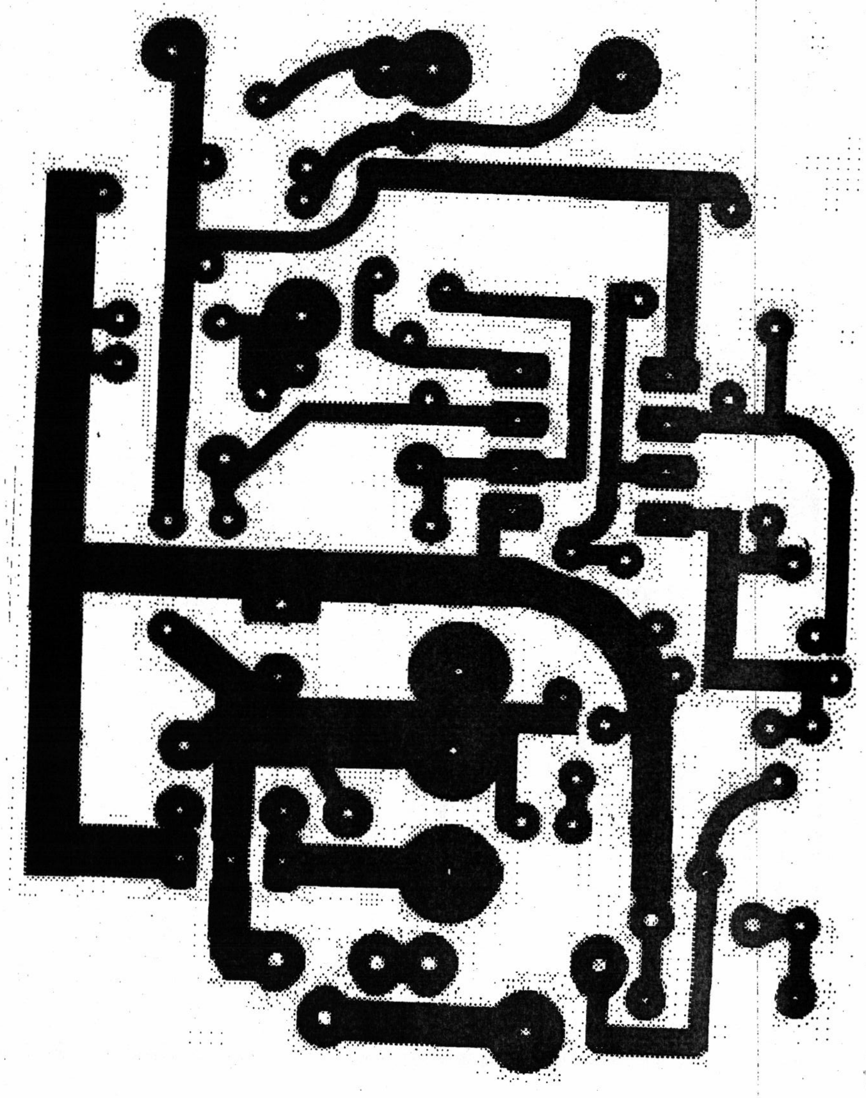
Liste composants chauffage thermostaté		
Référence	Composant	Remarques
R1	1.5K	
R2	10K	
R3	220	
R4	1K	
R5	3.3K	
R6	100K	
R7	100K	
R8	100K	
R9	100K	
R10	100K	
R11	2.7K	
R12	10K	
R13	680K	
R14	7.5K	
R15	1.2K	
R16	560	
R17	56	
C1	1microF	non polarisé.
C2	1nF	
C3	47microF	
C4	1nF	
C5	33nF	
C6	1.5microf	Tantal
C7	1.5microf	Tantal
IC1	LM335	TO92
IC2	LM336.5	
IC3	LM392	
IC4	78L05	
TR1	2N2907	
TR2	BD139	



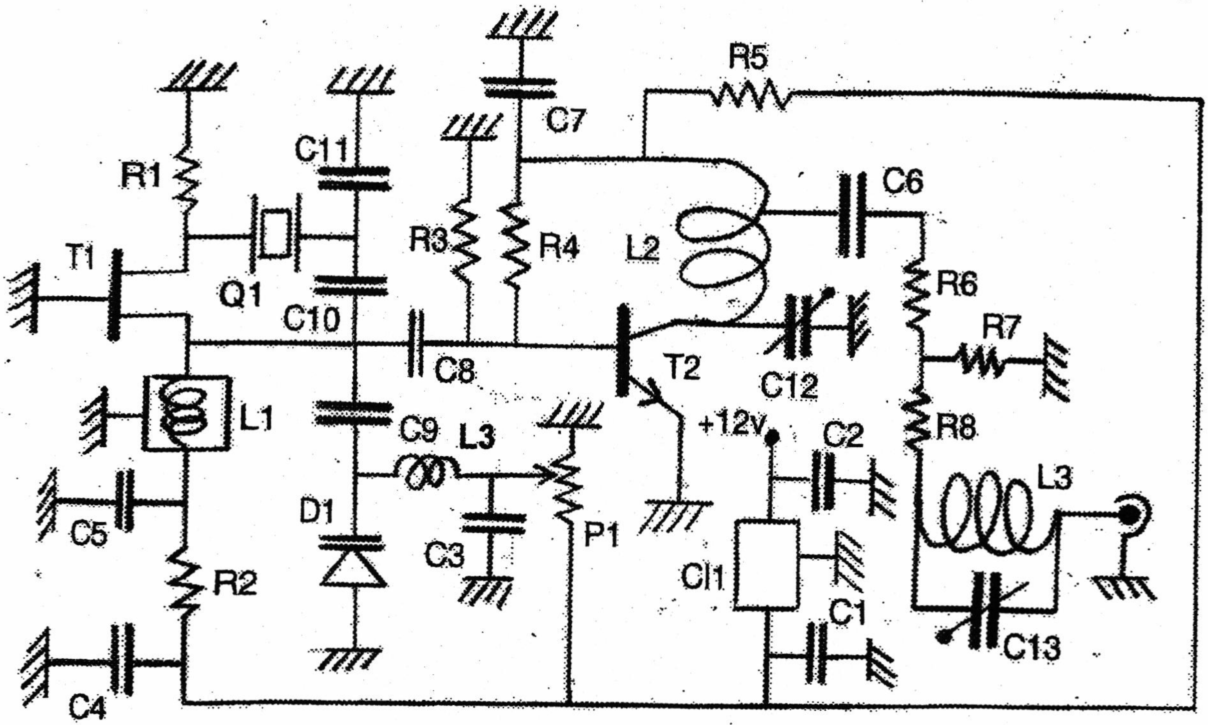
REGULATEUR DE TEMPERATURE

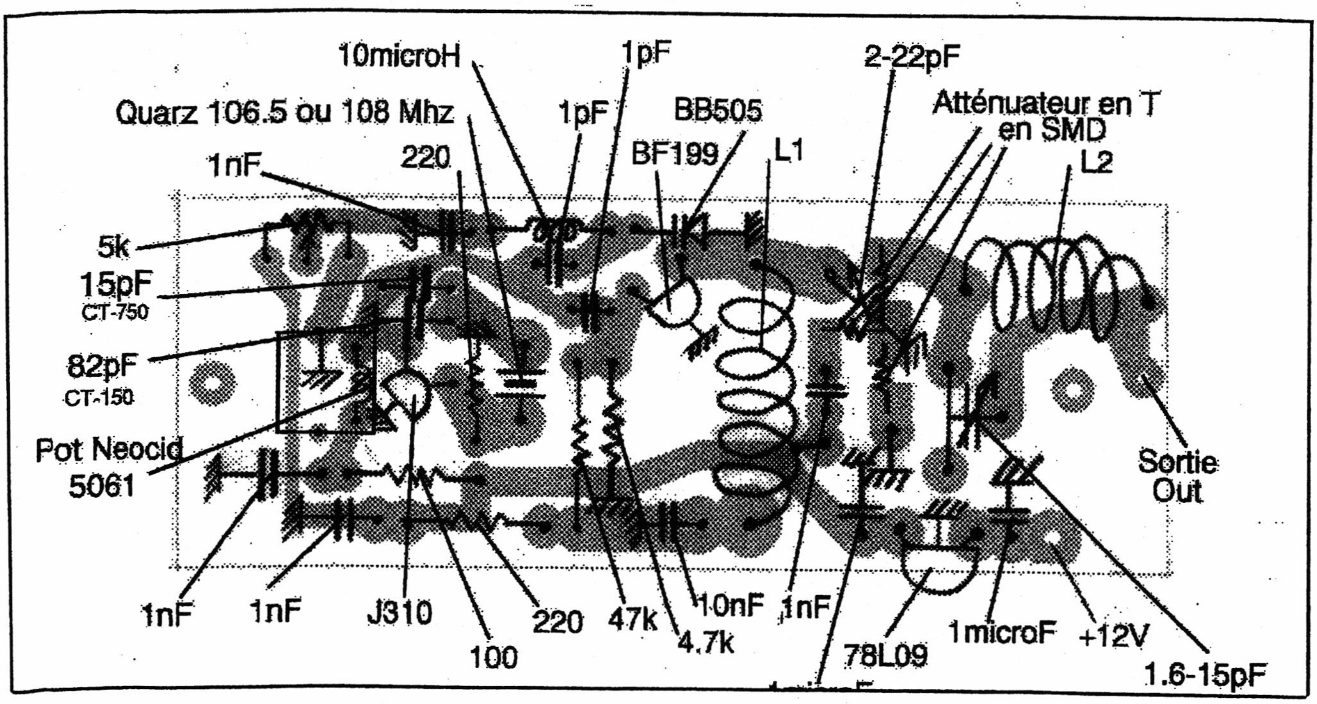


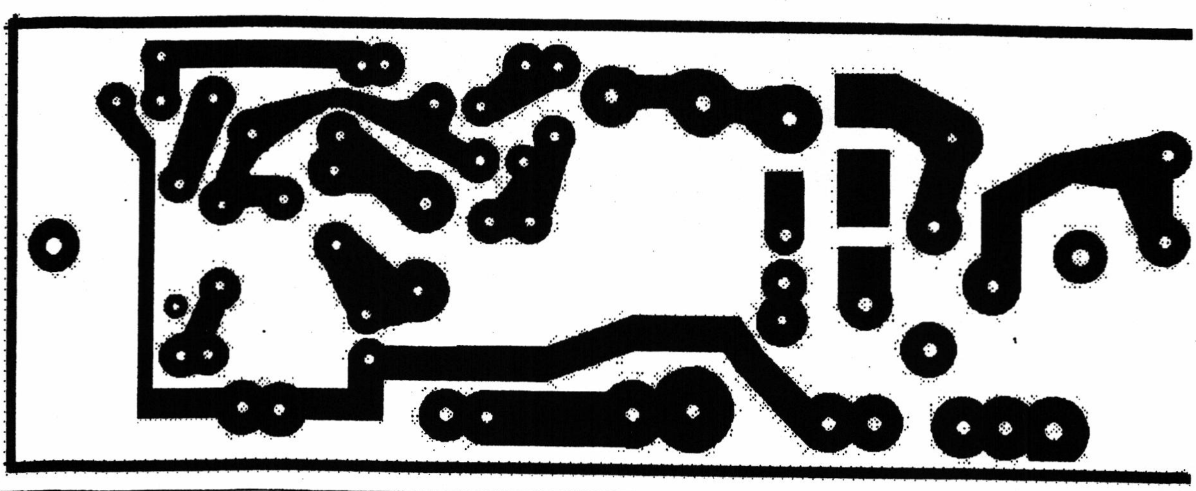


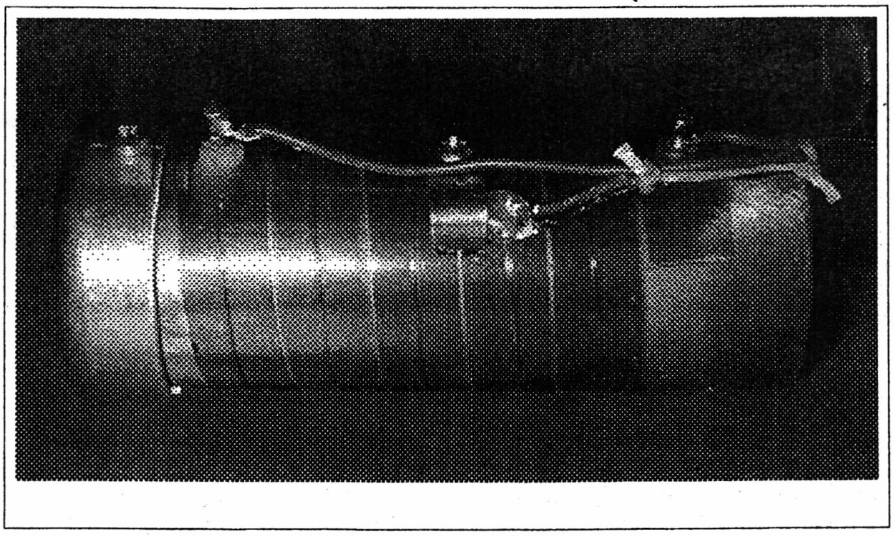


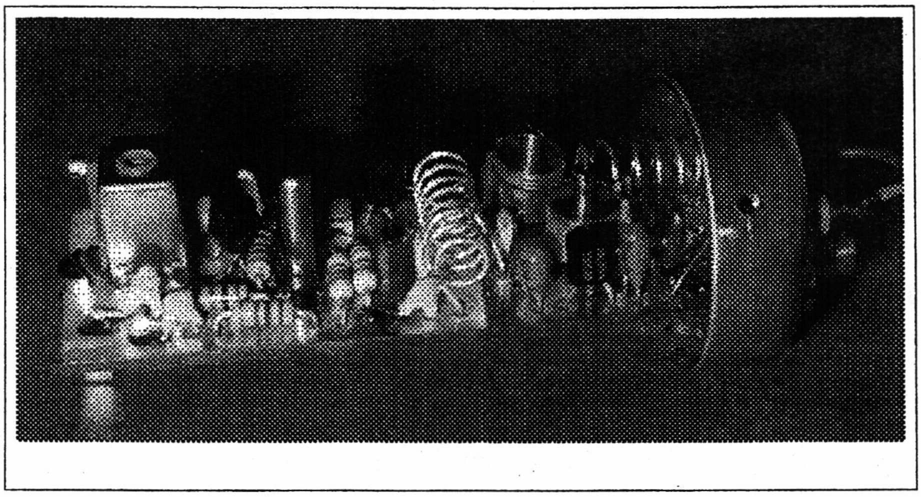
OSCILLATEUR

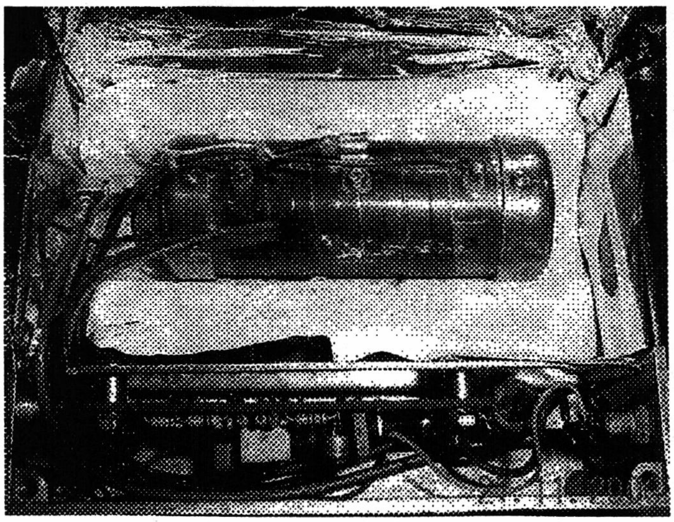












Cet OCXO est largement utilisé comme base d'oscillateur local par les OM trafiquant sur 10, 24 et 47 GHz, en copie conforme ou en apportant des modifications. Il est aussi employé pour les fréquences supérieures, mais sa stabilité est un peu limite pour cet usage. (voir à la suite de l'article des indications sur les performances qu'il permet d'obtenir.)

Oven-stabilized XO for VHF

Uwe Nitschke, DF9LN

Kurzbeschreibung: Ein temperaturstabilisierter Quarzoszillator ist das Herzstück einer jeden Frequenzaufbereitung für Mikrowellen-Transceiver. Die Frequenzen liegt um 100 MHz. Eine Butler-Schaltung sorgt für niedriges Phasenrauschen.

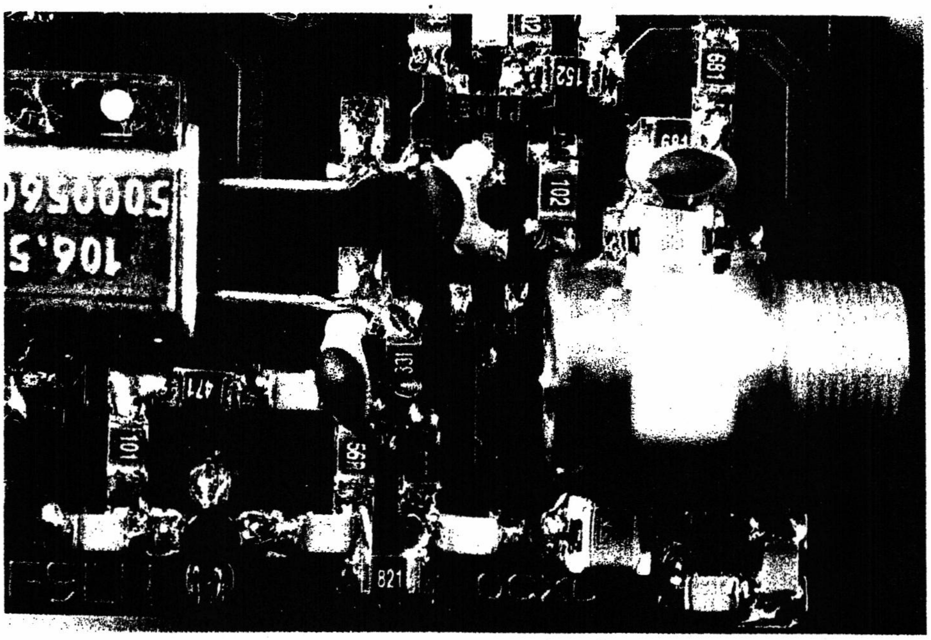
Abstract: A Butler type oven stabilized crystal oscillator on VHF (around 100 MHz) can be used for any serious microwave work. Low phase noise and high stability are features of this circuit.

Beschreibung

Der hier vorgestellte rauscharme Oszillator wurde auf extreme Stabilität, extreme Rückwirkungsfreiheit und sehr gutes Seitenbandrauschen gezüchtet. Damit eignet er sich vorzüglich zur Vervielfachung für 10GHz und auch höher.

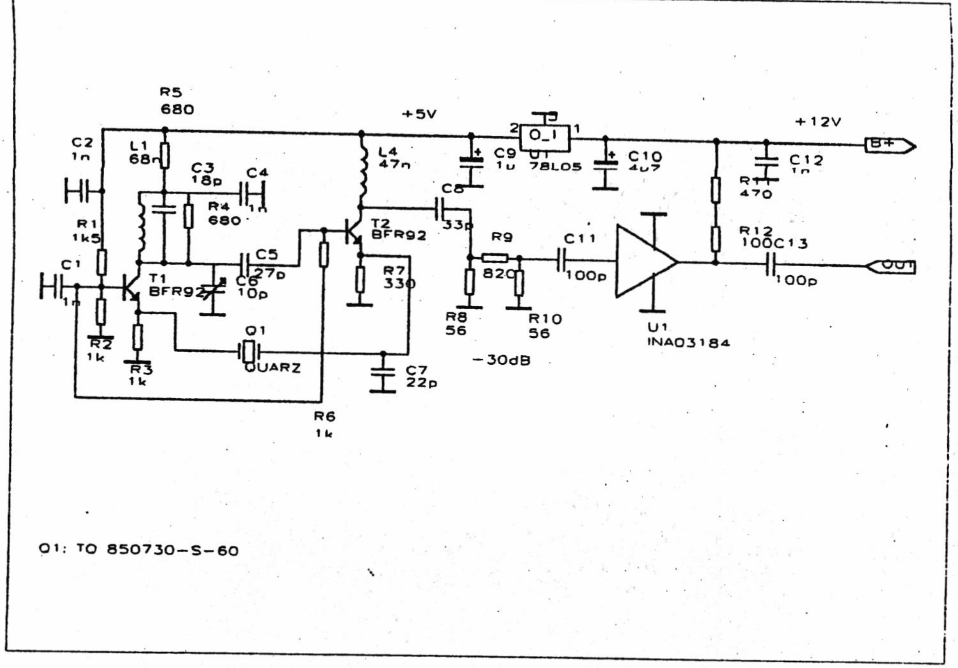
Die Transistoren T1 und T2 bilden eine Butler-Schaltung, in welcher der Quarz niederohmig eingebettet ist, d.h. der Quarz wird von einem Emitterfolger mit niedrigem Innenwiderstand angesteuert und er 'sieht' den niedrigen Innenwiderstand der Basisstufe. Das sorgt für niedriges Phasenrauschen.

Nach einem Dämpfungsglied mit 30 dB Dämpfung folgt ein Puffer mit einem MMIC INA03184



OCXO for 106.5MHz

Fig. 1: Circuit Diagram XO



und verhindert wirkungsvoll Rückwirkungen durch Lastschwankunegn.

Technische Daten:

1. Frequenzbereich: 106,500MHz je nach Quarz auch andere Frequenz 90...110 MHz
2. Ausgangsleistung: 0dBm
3. Versorgungsspannung: 12,6V= (8V bis 24V)
4. Stromaufnahme: 40mA, sofort nach dem Einschalten 700mA

Description

A Butler circuit with T1 and T2 (Fig. 1) is a well proven oscillator design for low phase noise and high stability crystal oscillators in the VHF-range. A buffer with the MMIC INA03184 isolates the circuit against any variations in load impedance.

Technical Data

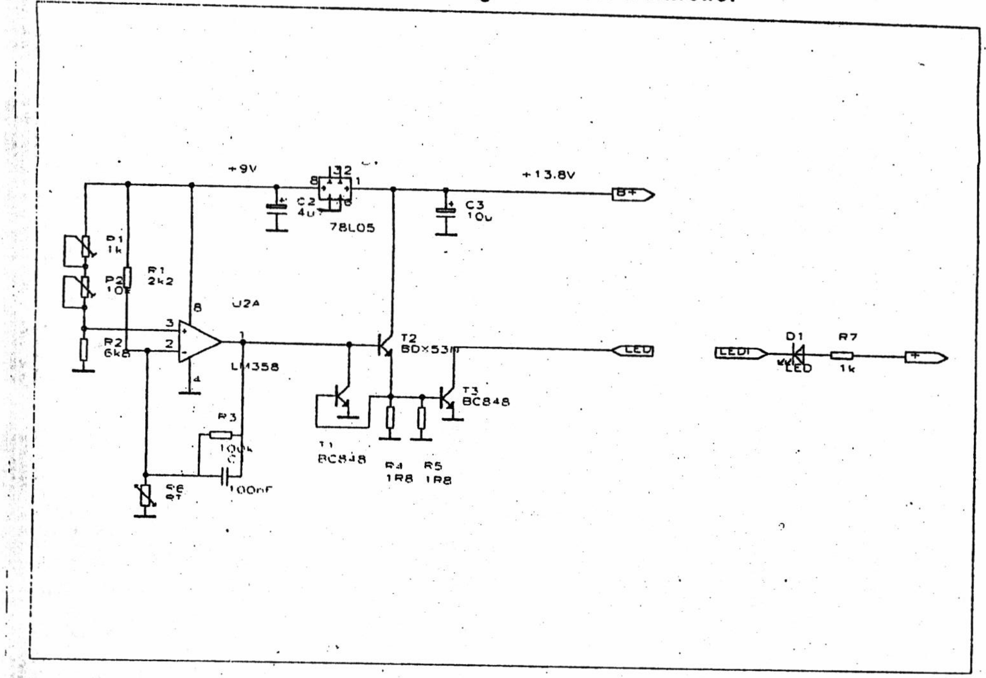
1. Frequency Range: 90..110 MHz, 106.5MHz typical
2. Output power: 0dBm
3. Supply voltage: 12.6V (8...24V)
4. Supply Current: After Startup. max. 700 mA., after warm-up: 40 mA

Konstruktion

Die Oszillatorplatine wird wird nach dem Bestücken auf den Aluheizkörper mit Silberleitkleber aufgeklebt und bei ca. 100°C ausgehärtet. Bei Verwendung des Kupferheizkörpers wird die unbestückte Platine nach vorheriger Verzinnung des Kupferheizkörpers mittels einer Heizplatte aufgelötet. Die Temperatur des Kupferheizkörpers sollte nur so hoch sein, daß das Lötzinn gerade fließt. Bei zu hoher Temperatur verbrennt die Leiterkarte. In den Heizkörper muß vor der weiteren Bearbeitung ein M 2,5mm Gewindeloch geschnitten werden. Zum Anzeichnen verwendet man die 3 mm Bohrung in der Heizplatine. Hier wird später die Heizplatine mit dem von unten montierten BDX53 aufgeschraubt. Beim Bestücken der Heizplatine ist darauf zu achten, daß die beiden Metallschichtwiderstände (1,8 Ω), der Temperaturfühler (KTY10-6), das 20-Gang Potentiometer und der BDX53 von Unten bestückt werden. Jetzt wird die vorher

Uwe Nitschke, DF9LN: Oven-stabilized XO for VHF

Fig. 3: Circuit Diagram Oven Controller



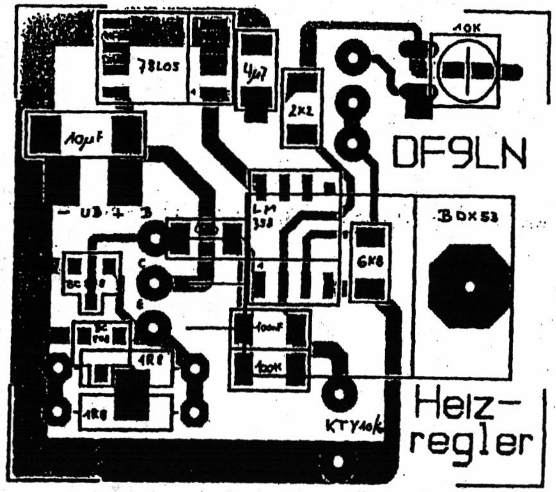


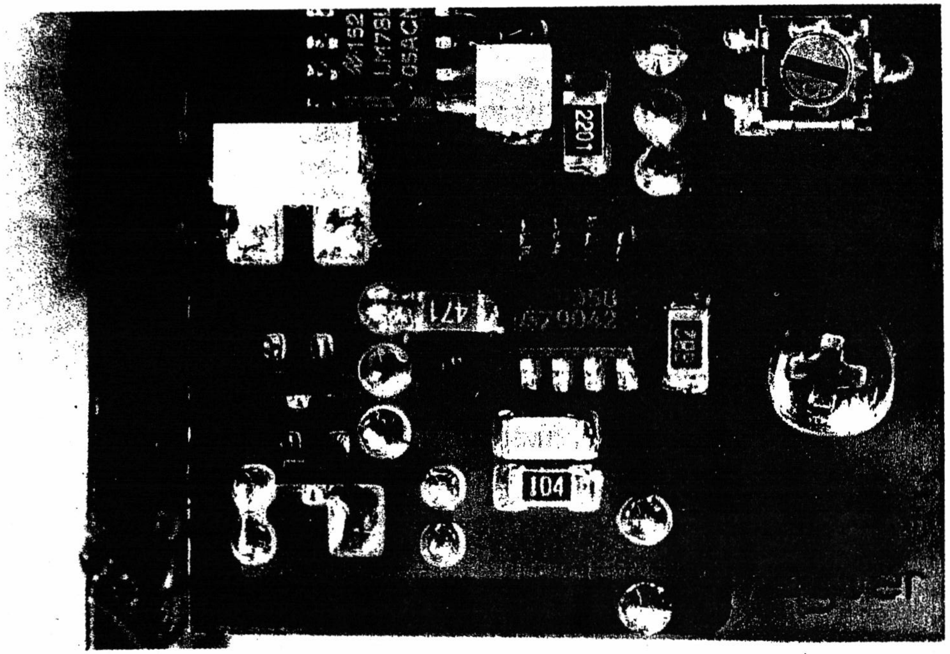
Fig. 2: PCB oven controller

bestückte und geprüfte Leiterkarte "Heizung" an die Aluheizplatte (oder Kupferheizplatte) angeschraubt. Dabei sollte die Glimmerscheibe unter dem Transistor BDX53 nicht vergessen werden. Der Wärmefühler KTY10-6 muß fest gegen die Heizplatte gedrückt werden. Dazu ist eine kleine Gummischeibe zwischen Platine und Wärmefühler zu schieben. Die Heizung wird jetzt in Betrieb genommen. Mit dem 10k Potentiometer wird die Temperatur des des Alu-/Kupferheizkörpers auf ca. 100°C eingestellt. Nun kann die ebenfalls auf 100°C aufgeheizte Oszillatorplatine sehr einfach bestückt werden.

Construction

After mounting all parts the XO board is mounted to the heatsink by means of silver expoxy cement.

OCXO for 106.5MHz



The BDX53 is fastened by a M2.5 screw. The complementary thread has to be cut into the heatsink.

When mounting the parts of the controller circuit, take care that the 1.8 Ω resistors, the BDX53 and the temperature sensor KTY have to be mounted onto the bottom side.

When mouting the controller board don't forget for the mica insulator for the BDX53 and to establish a solid contact of the KTY to the heatsink for a small temperature gradient.

Abgleich

Bei dem ersten Abgleich des Oszillators auf die Sollfrequenz wird wie folgt verfahren:

1. Quarz durch Drahtbrücke kurzschließen.

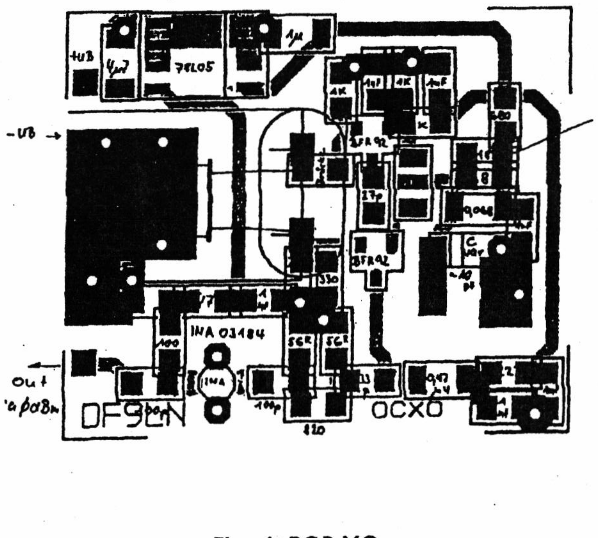


Fig. 4: PCB XO

Table 1: Parts List of OCXO Oven Controller

Stück/ Count	Art/Sort	Wert/Value	Hersteller/ Manufacturer	Bauform/ Size
1	SMD-R	470 Ω	Sie	1206
1	SMD-R	2200 Ω	Sie	1206
3	SMD-R	6800 Ω	Sie	1206
3	SMD-R	100 kΩ	Sie	1206
2	R	1,8 Ω	Vitrohm 490	
1	Poti,SMD, CERMET	1K	Beckmann	64W
1	Poti,SMD, 10K	Beckmann	23B	
1	SMD-C	100 nF	Sie	1206
1	SMD-Tantal	10 uF/16V	Sie	C
1	SMD-Tantal	4.7 uF/10V	Sie	B
1	Regulator, SMD	78L05		SO8
1	Opamp, SMD	LM358		SO8
2	SMD-NPN	BC848	Sie	SOT-23
1		BDX53		TO220
1	Temperaturfühler	KTY10-6	Sie	
1	Leiterkarte	FR-4 "OCXO-Heizung"		
1	Heizkörper	ALU oder Kupfer		30,5mm x 25,5mm Stärke min 4mm

2. Mit Frequenzzähler oder Spectrumanalyzer am Ausgang die Frequenz messen.
3. Mit dem Trimmkondensator die gewünschte Frequenz einstellen. Falls dies nicht gelingt, ist der Parallelkondensator (18pF NP0 bei 106,5MHz) zu verändern. Er ist zu verkleinern, falls die gemessene Frequenz zu tief ist. Er ist zu vergrößern, falls die gemessene Frequenz zu hoch ist.
4. Wird die Sollfrequenz erreicht, kann die Drahtbrücke über den Quarz entfernt werden. Die genaue Frequenz sollte jetzt mit dem Trimmer eingestellt werden.
5. Der komplette Oszillator sollte jetzt im eingeschalteten Zustand 2 - 3 Stunden im Backofen bei ca. 100°C gealtert werden. Nachdem die Schaltung in ein wärmeisolierendes Gehäuse eingebaut wurde, kann jetzt der Endabgleich (genaue Frequenz und Temperatur) erfolgen. Zum Temperaturabgleich könnte ein Sekundenthermometer mit dünnen Drahtfühlern verwendet werden (z.B. Greisinger GTH1160 mit Fühler GTF300). Der Oszillator sollte vor der endgültigen Verwendung ca. 1 - 3Wochen im Dauerlauf betrieben werden. Wegen der in dieser Zeit auftretenden Alterung der Bauelemente muß nach Ablauf der 3Wochen die Frequenz noch einmal überprüft werden.

Alignment

Proceed in the following manner:

1. Short crystal by a small piece of wire.
2. The oscillator now is in a free running mode. Count the frequency at the output
3. Adjust the parallel trimmer to the wanted frequency. If this does not succeed vary the parallel cap which has a nominal value of 18 pF accordingly. Reduce the value if the frequency is too small and vice versa.

Table 2: Parts List of OCXO

Stück/ Count	Art/Sort	Wert/Value	Hersteller/ Manufacturer	Bauform/ Size
1	SMD-R	22 Ω	Sie	0805
2	SMD-R	56 Ω	Sie	0805
1	SMD-R	100 Ω	Sie	0805
1	SMD-R	330 Ω	Sie	0805
1	SMD-R	470 Ω	Sie	0805
2	SMD-R	680 Ω	Sie	0805
1	SMD-R	820 Ω	Sie	0805
3	SMD-R	1000 Ω	Sie	0805
1	SMD-R	1500 Ω	Sie	0805
1	SMD-C	33 pF NP0	Sie	0805
2	SMD-C	100 pF NP0	Sie	0805
5	SMD-C	1 nF NP0	Sie	0805
1	C	22 pF NPO	Sie	EGPU
1	C	27 pF NPO	Sie	EGPU
1	C	18 pF N470	Sie	EGPU
1	SMD-Tantal	1 uF/16V	Sie	A
1	SMD-Tantal	4.7 uF/10V	Sie	B
1	Var.-C	1.1...11pF	Tronser 60-0715-100011	
1	SMD-L	68 nH	Sie	SIMID01
1	SMD-L	470 nH	Sie	SIMID01
2	SMD-NPN	BFR92	Sie	SOT-23
1	Regulator	78L05	Siliconix	SO8
1	MMIC	INA03184	HP	86
1	Quarz	106.5MHz, TQ85 07 30-S-60	Telequarz	
1	PCB	FR-4, OCXO	Eisch	

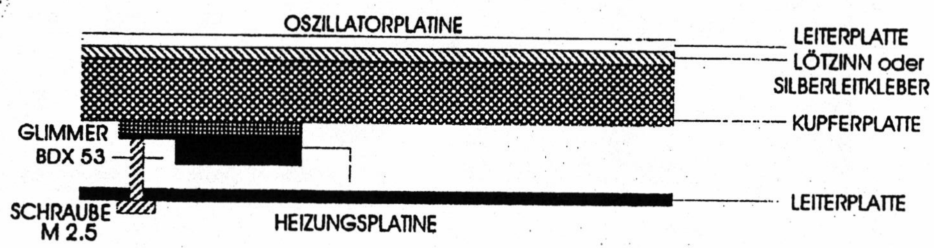


Fig. 5: Mounting details

4. If the nominal frequency of the crystal can be achieved remove the short circuit and adjust to the exact frequency of the crystal
5. It's not a bad idea to age the whole circuit at a temperature of 100 °C in a kitchen stove.

Mount the whole circuit into a isolating box and adjust the temperature controller to a oven temperature of 60 °C. After 3 weeks of operating the OCXO can be tuned again for the exact frequency.

Teile/Parts

Parts, PCBs and kits are available from:

Annemarie Eisch-Kafka, Abt-Ulrich-Str.16, 89079 Ulm-Gögglingen, Tel.: 07305/23208, FAX: 07305/23306

Articles à consulter pour connaître les performances que l'on peut obtenir avec l'OCXO de DF9LN :
- mesures effectuées sur un OCXO DF9LN, HYPER N° 28 octobre 1998
-le manque de répétabilité des OCXO, HYPER N°32 février 1999

(24 et) 47 GHz

Ampli 'Plug and play' 24Ghz sortie 500mW, gain 37dB, coût: 375 US Dollars

Entrée 0.1 mw , sortie 500 mw en 24 Ghz dans 9 cm^3

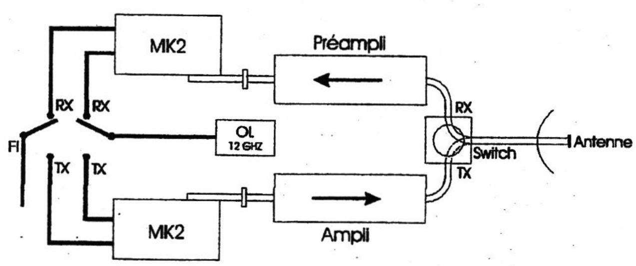
Voilà quelques mois en surfant un peu, j'ai découvert un produit intéressant, un ampli de 27 dBM sur 24 Ghz avec un gain typique de 37 dB, voici les caractéristiques principales :

-Fréquence d'utilisation : 24.00 – 24.25 Ghz
-Puissance de sortie : 27dBM avec –13 à-15 dBM d'entrée, (-10 dBM max)
-Mode classe AB
-Alimentation : +6 Volts @ 1.2 A
-5 Volts @ 13 ma
-Entrée SMA, sortie Guide
-Dimensions 50X23X8 mm

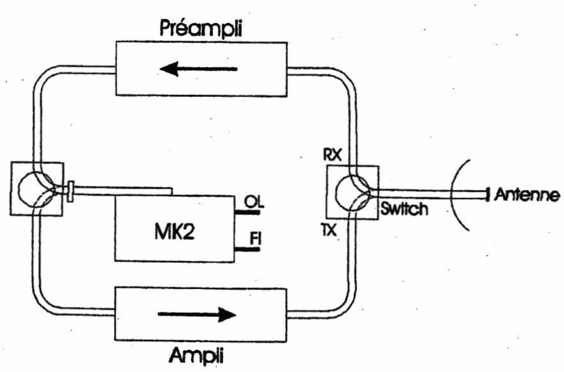
Deux pièces ont été commandées pour Marc F6DWG et moi-même. Après modifications des transverters existant, des essais ont été effectués, le premier entre JN18CV et JN19AJ soit 57 km report des deux cotés 58 à 59+. Deux jours plus tard, second essai entre JN18CV (92) et JN09VT (76) soit 106 km, QSO validé avec des reports de 52 de part et d'autres. Bien sur rien d'extraordinaire, mais compte tenu des reliefs de la région parisienne ces premiers résultats sont très encourageants et nous permettent d'espérer de bonnes liaisons lorsque « dame propag. » sera là !

Deux types de montages ont été réalisés selon la disponibilité des composants.

-Première solution F6DWG : 1 relais guide, 1 relais coax 10 GHz, 1 relais 144, préampli, ampli, 2 transverters et oscillateur.



-Seconde solution F1PYR : 2 commutateurs guide , préampli, ampli, 1 transverter et 1 oscillateur.



Ces amplis sont disponibles auprès de Carl chez Jeremy Communication, email : km1h@juno.com . Le prix unitaire est de US$ 375.00 plus port. Pour de plus amples informations n'hésitez pas à me contacter, André F1PYR.

Module doubleur, sortie 47 Ghz 12 mW max. Le doubleur en utilisant 2 diodes beamlead HSCH-5312 sort 5mW et avec un varactor Russe A92220-5 , 12 mW. Niveau d'entrée nécessaire en 23,5 Ghz : 100 mW.

Anmerkungen:

1. Die Leiterplatte sollte mit Silberleitkleber und vier M2 Schrauben in das Gehäuse montiert werden.
2. Aufbauhinweise in DUBUS 2/92, S. 2 bis 14 beachten

Abgleich

Mit angeschlossenem Leistungsmesser wird nach Anlegen der Steuerleistung auf 23,5GHz die Ausgangsleistung durch Veränderung von R_A sowie durch Anbringen von Abstimmfähnchen optimiert. Mit dem Kurzschlußschieber (M5 Schraube) kann die Anpassung auf den Hohlleiter eingestellt werden.

4. Doubler 23,5->47GHz

Figures 11, 12 and 13.

The doubler (Fig. 8) uses beamlead-diodes. The circuit on a PCB is mounted into a aluminum box (Fig. 9).

With 100mW drive power on 23.5GHz you can get more than 5mW on 47GHz with two times HSCH-5312 (Fig. 10) diodes in parallel ($R_A \sim 100\Omega$).

With a Russian varactor A92220-5 from ELECS Ltd. more than 12mW can be achieved ($R_A \sim 5k$).

Remarks:

1. Mount PCB with silver adhesive and four M2 screws into box
2. Look for hints in DUBUS 2/92

Adjustment:

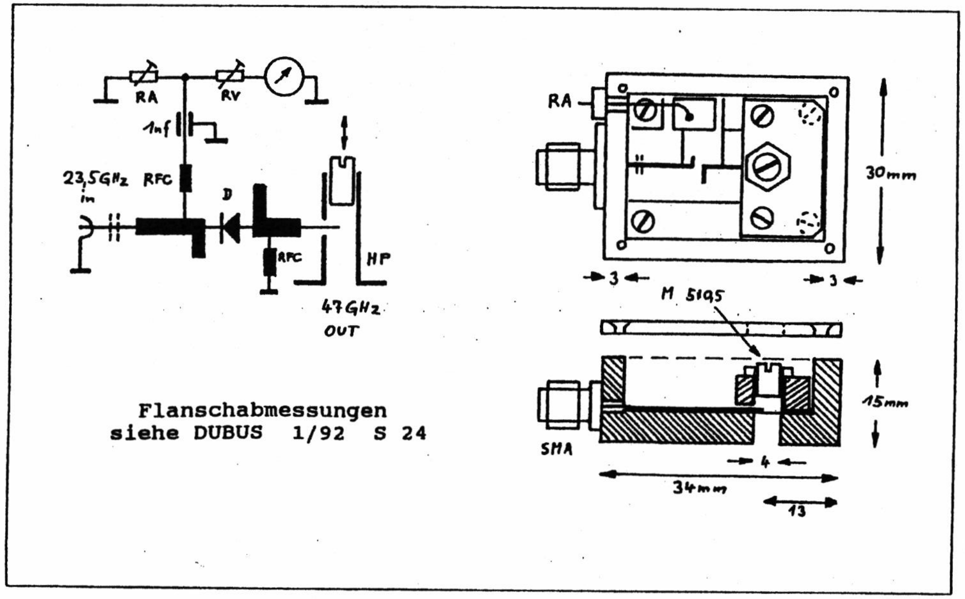
Connect powermeter to 47GHz output and tune for maximum power by adjusting R_A and by means of tuning stubs. The shorting glider can be adjusted for best match.

5. Teile:

Leiterplatten für Koax- oder Hohlleiteraufbau sind bei DB6NT Michael Kuhne, Birkenweg 15, D-95119 NAILA erhältlich. Alugehäuse sowie Hohlleiterkomponenten sind bei DG1KBF, Hubert Krause, Berghagen 60, D-53773 HENNEF erhältlich.

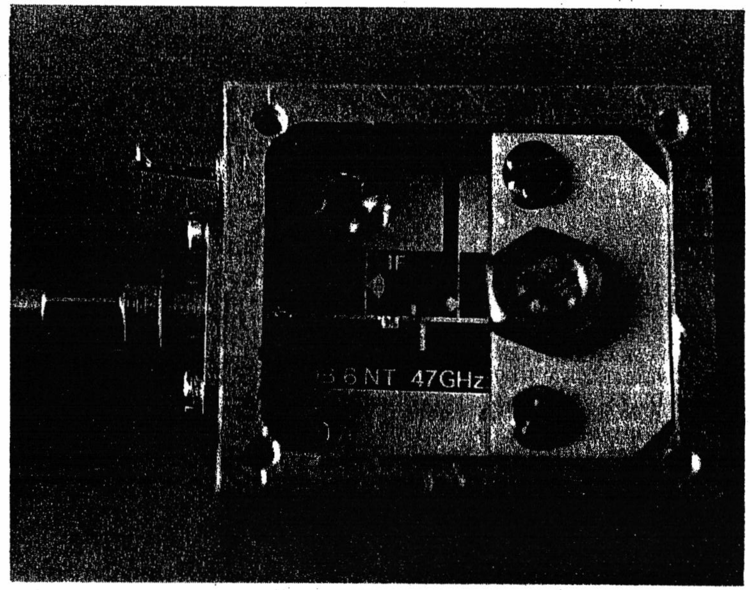
5. Parts:

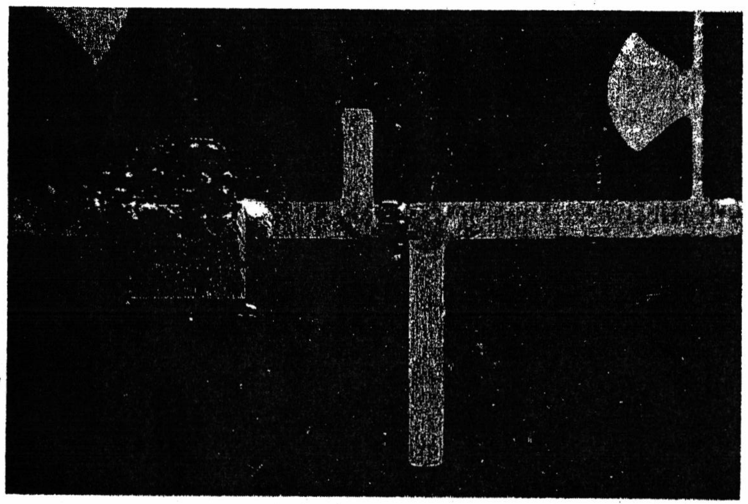
You can get PCBs at DB 6 NT Michael Kuhne, Birkenweg 15, D-95119 NAILA. Machined boxes and waveguide components you can get at DG1KBF, Hubert Krause, Berghagen 60, D-53773 HENNEF.



Bild/Figure 11: Frequenzverdoppler 23,5 -> 47GHz - Diagram

Bild/Figure 12: Frequenzverdoppler 23,5 -> 47GHz





Bild/Figure 13: Einbau der HSCH-5312 Dioden

Module doubleur, sortie 46 Ghz 1,5 dBm max. Utilisation de 2 HEMT FHX05LG, le 1er en ampli 23Ghz puis 2eme en doubleur sur un PCB RT/Duroid 588 0,5mm. Le module a une perte de conversion de -4dB à –1dBm de sortie et de –8dB à 1,5 dBm de sortie.

47GHz Active Doubler

Toshihiko Takamizawa, JE1AAH

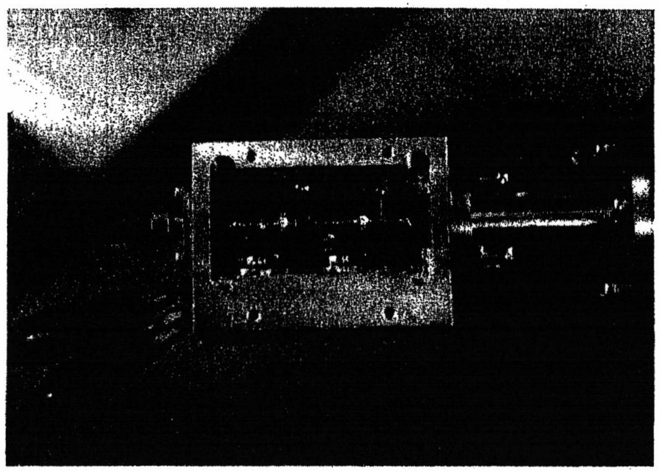
Abstract: A 46 GHz active frequency doubler has been developed using conventional packaged low noise HEMT and soft-board technology instead of MMIC technology. The doubler utilizes low noise HEMT FHX05LG from Fujitsu for 23 GHz buffer amplifier and times 2 multiplier. The board used is RT/Duroid 5880, 5 mil thickness with 18 µm copper clad soft board.

Kurzfassung: Ein aktiver 47GHz Verdoppler benutzt gehäuste HEMT Transistoren FHX05LG von Fujitsu auf einem 5mil RT-5880 Teflon-Substrat. Ein 23 GHz Verstärker steuert einen Verdoppler auf 47 GHz an.

Introduction

By recent performance improvement of packeged HEMT, this project has been addressed and targeted to realize millimeter wave component using consumer level products which means to utilize packaged device and soft-board. To date, there have been many papers reported regarding this frequency region's LNA, HPA and frequency doubler, etc. However all the work has been done in MMIC technology.

Years ago DBS receiving systems had started using MIC or MMIC technology, but now there is no product using this technology and conventional packaged device has been widely used instead. Currently, a future 23 GHz DBS system is being considered to employ MMIC technology but there are several papers reported to extend 12 GHz DBS circuit technology to 23 GHz systems.



47 GHz Active Doubler

Based on this history, I decided to develop 46 GHz components using current low cost production technology.

Figure 3 shows the block diagram of the doubler. Since there are no S-parameters published above 20 GHz by manufacturers, tunig has been done by tweeking a small piece of copper foil along with micro stripline track. Stripline to waveguide transition is done by quarter wavelengh probe.

Figure 2 shows wide band output power performance at +10 dBm input level and figure 2 shows input/output conversion loss. As shown in the figures, conversion loss is less than 4 dB at low input level and less than 10 dB at saturated and high power output level. -3 dB bandwidth is more than 1 GHz at output frequency region.

Since gain of input buffer amplifier is estimated to about 7 to 8 dB, actual doubler stage conversion loss in the linear operating region is about 7 to 8 dB. This means total conversion loss is about 0 dB. In fact, the doubler could be optimised to

0 dB of conversion loss at 0 dBm output power level to achieve better conversion loss. For fixed frequency application, this good conversion loss setting could be used.

For wider frequency range application, naturally packaged device is not good for this purpose due to excessive parasitic elements. Packaged device frequency response varies more than that of a chip device. However, by operating buffer amplifier and multiplier at saturated power region, it is possible to widen its operating frequency range.

47 GHz Doubler: Top View

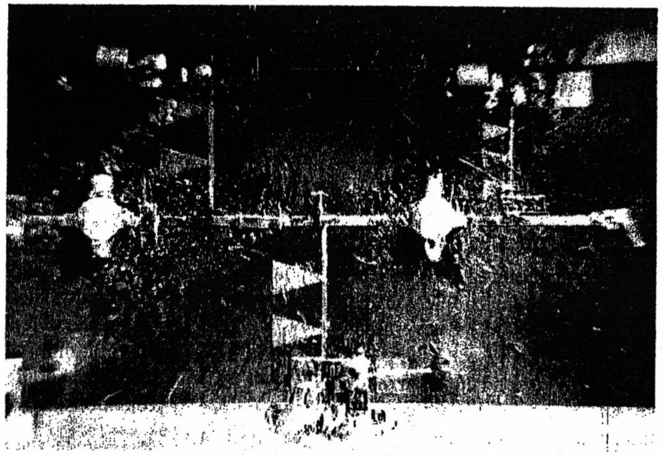


Figure 2 could be obtained using this technique.

Spurious response of the doubler for fundamental frequency it is less than -50dBc and for times 3 frequency it is less than -40dBc with wide band response setting.

To conclude a paper, the project was very successful and good performance has been obtainable to use packaged dvice on 46GHz. Next target is to develop LNA and HPA for this frequency region.

Deutsche Version

Das Ziel dieses Projekts war, Millimeter-Komponenten zu entwickeln, die normale gehäuste FET-Transistoren und Teflon-Substrat verwenden. Bisherige Komponenten benutzten MMIC Technologie, die für Amateure unerreichbar ist. Beispielsweise kann man sehen, daß heute alle 11 GHz DBS Systeme in konventioneller Technik aufgebaut sind, obwohl die ersten Labormuster in MMIC Technik aufgebaut waren.

Abb. 3 zeigt das Schaltbild des aktiven 47 GHz Verdopplers. Da keine S-Paramater oberhalb von 20 GHZ zur Verfügung stehen, müssen die FET's in Striplines mit kleinen Abstimmfähnchen eingebaut werden. Der Hohlleiter-Übergang wird durch einen Viertelwellen-Monopol gebildet.

Abb. 1 zeigt die Breitbandausgangsleistung bei einem Eingangspegel von +10dBm. Aus Abb. 2 ist die Verstärkung bei 47GHz zu sehen. Die Verstärkung (Verlust) ist ca. -4dB bei kleiner Ausgangsleistung und ca. -10dB bei hoher Ausgangsleistung. Die 3dB Bandbreite ist ca 1GHz.

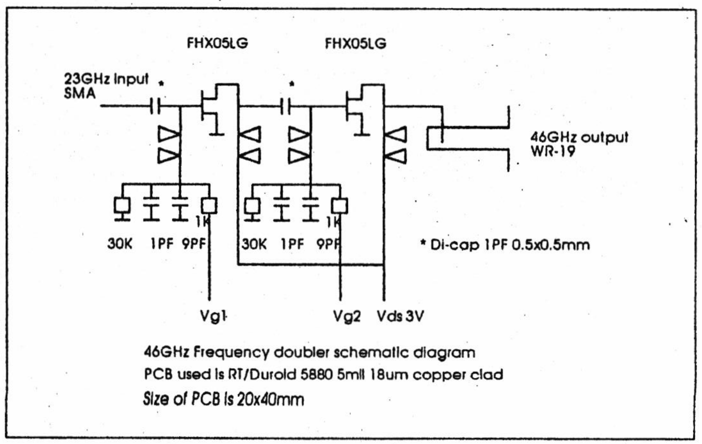
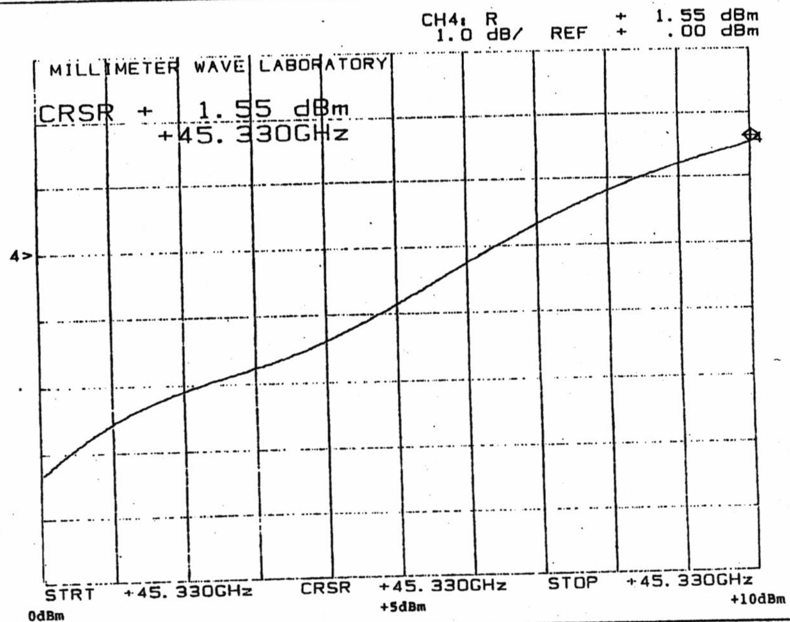


Abb./Figure 3: Circuit Diagram

Abb./Figure 1: Frequency Response



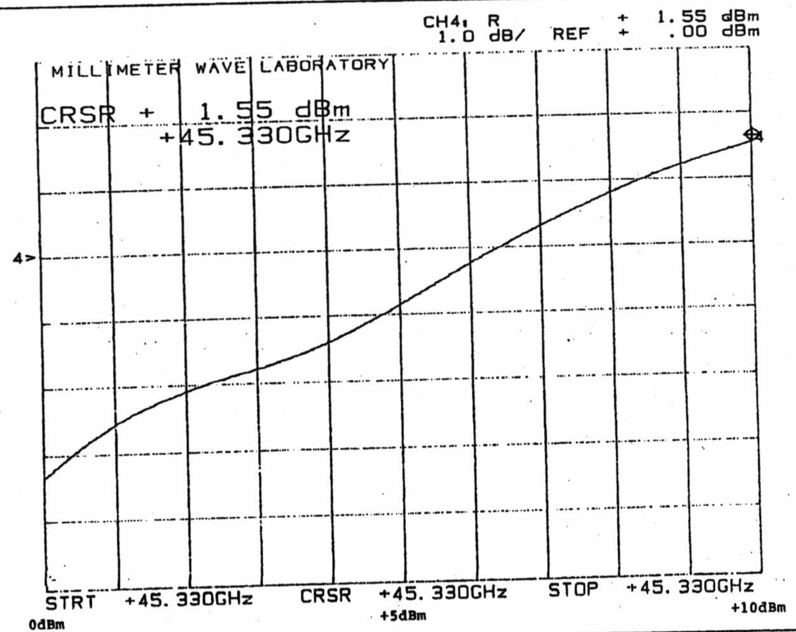


Abb./Figure 2: Pin @ Pout

Méthode et calcul du subréflecteur hyperbolique dans un système d'antenne Cassegrain. L'auteur ayant acquit une parabole avec un F/D de 0,38 et désirant utiliser un feed W2IMU adapté pour un F/D de 0,8 , le système Cassegrain rend compatibles parabole et source.

Large Dish Cassegrain Development Using CAD & Spreadsheet For Millimetric Bands & Practical Implementation.

Martin Farmer G7MRF
Email: g7mrf@compuserve.com

August 14, 2000

Introduction

The scope of this paper is to make available the design calculations and the spreadsheet program for which now carries out the long-winded and repetitive process to design the hyperbolic sub reflector for manufacturing, the W2IMU feed horn used and how the implementation of quick change bands on my portable set up.

After using flat plate type reflectors on both portable systems on 24GHz 47GHz band to good effect the decision was made to incorporate both of these bands into one box during a major rebuild over the winter months. The decision also included using one dish with interchangeable bands contained within a quick-change system eventually ending up with 5.7, 10, 24 & 47GHz as a possible combination on a portable expedition.

The decision to move away from the flat plate type reflectors was made by the small improvement in system gain but more importantly I wanted to use a W2IMU dual mode feedhorn using circular waveguide to feed the new dish which has a F/D ratio of 0.38.
Looking at the reference books the optimum F/D for the dual mode feedhorn is 0.8 but can be used on parabolic dishes with F/D ratios as low as 0.5. The use of a cassegrain system and in particular the ability to slightly change values in the calculation you can end up modifying the F/D ratio of the virtual parabolic dish to suit what F/D you require for your feedhorn as in my particular case from 0.38 if I used a flat plate type reflector the virtual dish would appear to be 0.38 and not suit the feedhorn type that I wanted to use.
For details of the 24GHz and the 47GHz W2IMU dual mode feed horn see figure 8.

The original flat plate reflector used on 24GHz is shown in figure 1 it was constructed from double-sided PCB material and brazing rods for the supports from the WG20 dish feed.

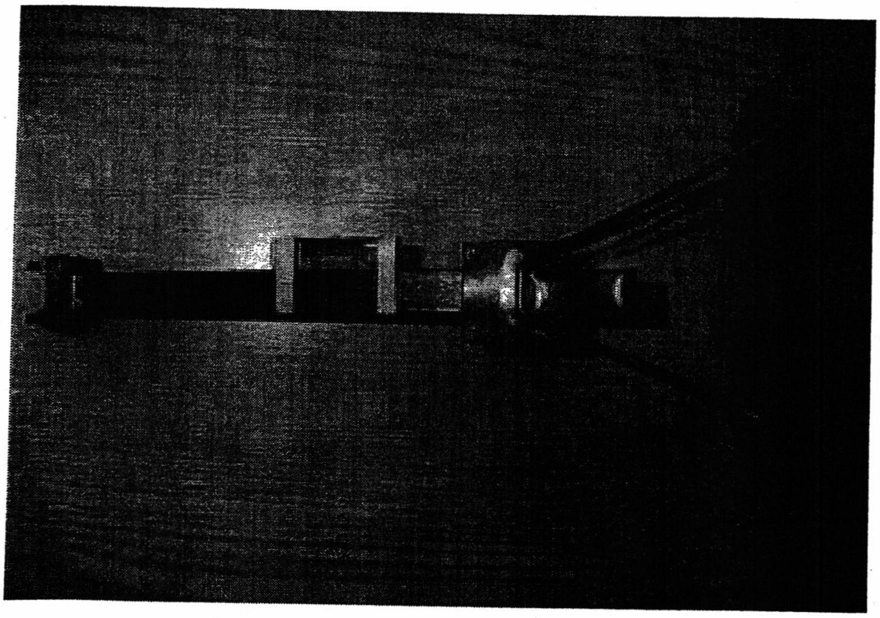


Figure 1

A similar approach was used on my original 47GHz transverter using a 60mm circular disc reflector and brazing rod supports but with a circular collar on the round 4mm tube see figure 2

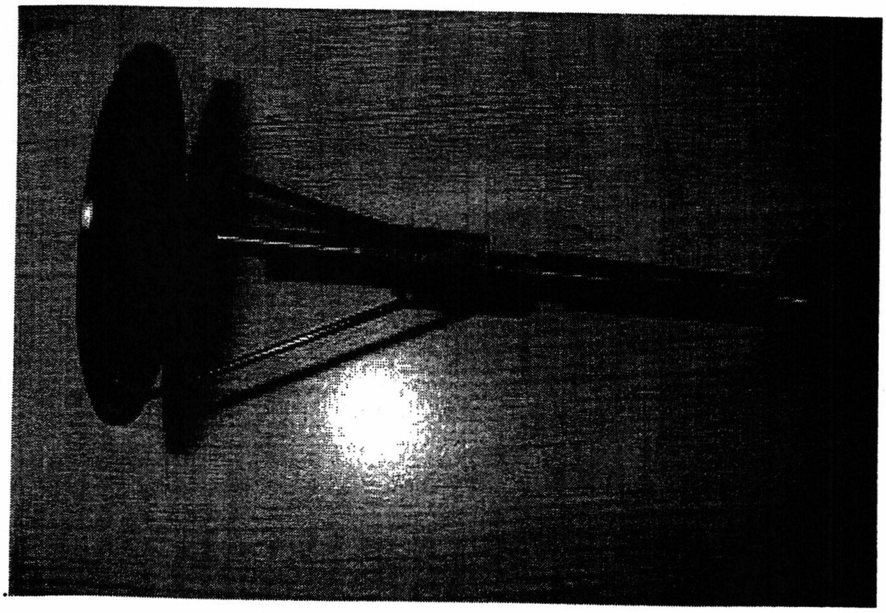


Figure 2

Discussions with Dr Dennis Hawkins of Qpar Angus [1] at the end of one of the UK Microwave Round Tables during 1999 resulted in the exchange of correspondence on the subject of cassegrain feeds. Armed with this information and calculations from a paper [2] regarding Microwave antennas derived from the cassegrain telescope we started to lay out in AutoCAD the profile from the main dish that was going to be used and by using geometry we arrived at a shape for the reflector. This profile from the CAD system was sent to Dr Dennis Hawkins who compared with his commercial antenna software what we had laid out and confirmed this to be ok. It was decided to go ahead with the manufacture. I found a machine shop that was willing to make the component for me using CNC technology and they had the ability to take the CAD profile of the sub reflector in to his computer system to generate his NC code automatically. The end result is shown in figure 3.

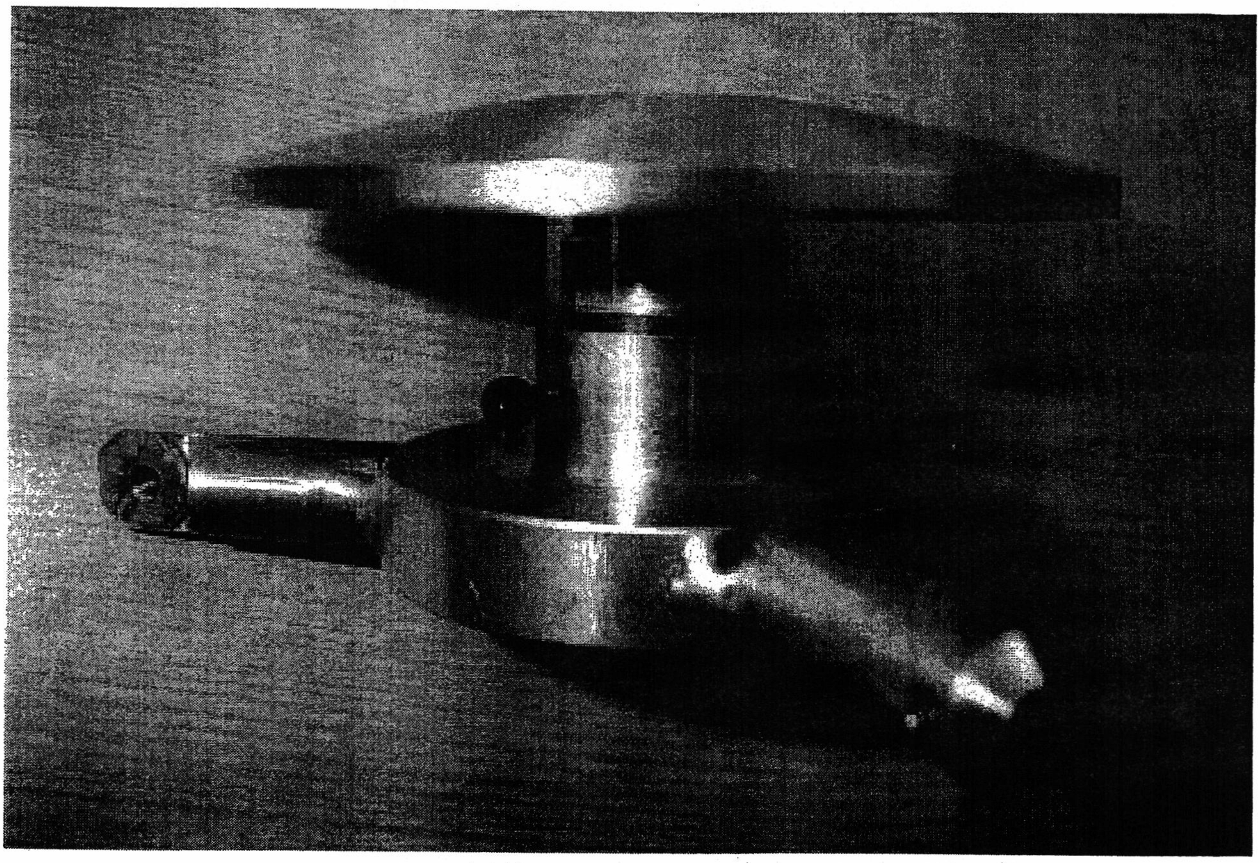


Figure 3

Above figure shows the 100mm-diameter sub reflector and support boss that is used on my new portable system.

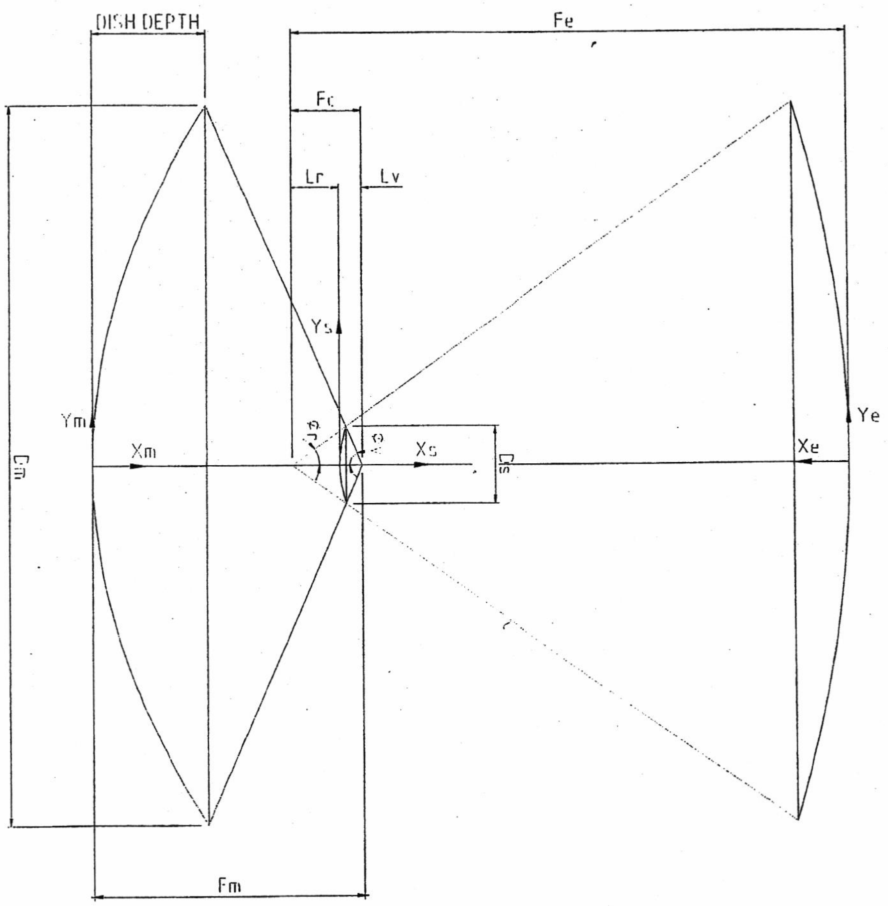


Figure 4

Dm = EFFECTIVE DIAMETER OF CIRCULAR MAIN DISH (TO EDGE RAYS).
Ds = EFFECTIVE DIAMETER OF CIRCULAR SUB DISH (TO EDGE RAYS).
Fm = FOCAL LENGTH OF MAIN DISH.
Fc = DISTANCE BETWEEN FOCI OF SUB DISH.
Fe = EQUIVALENT FOCAL LENGTH OF CASSEGRAIN SYSTEM.
Lv = DISTANCE FROM VIRTUAL FOCUS (OR MAIN DISH FOCUS) TO SUB DISH.
Lr = DISTANCE FROM REAL FOCUS (OR FEED) TO SUB DISH.
ϕv = ANGLE BETWEEN AXIS AND EDGE RAY, AT VIRTUAL FOCUS.
ϕr = ANGLE BETWEEN AXIS AND EDGE RAY, AT REAL FOCUS.
E = ECCENTRICITY OF CONIC SECTION.
A = TRANSVERSE HALF-AXIS OF CONIC SECTION.
B = CONJUGATE HALF-AXIS OF CONIC SECTION.
Xm = AXIAL CO-ORDINATES OF MAIN DISH.
Ym = RADIAL CO-ORDINATES OF MAIN DISH.
Xs = AXIAL CO-ORDINATES OF SUB DISH.
Ym = RADIAL CO-ORDINATES OF SUB DISH.
Xe = AXIAL CO-ORDINATES OF VIRTUAL DISH.
Ye = RADIAL CO-ORDINATES OF VIRTUAL DISH.

Manual Calculation Method

Below is a worked example based on the following data:

Dm = 935 mm
MAIN DISH DEPTH = 152 mm
Ds = 100 mm
Fc = 94 mm

CALCULATING F/D RATIO

F/D RATIO = Dm/(16 x DISH DEPTH)
= 935/(16 x 152)
= 935/2432
= 0.3845

CALCULATING MAIN DISH FOCAL LENGTH

Fm = Dm x F/D RATIO
= 935 x 0.3845
= 359.5 mm

CALCULATING ANGLE ϕv

ϕv = INV TAN [(Dm ÷ 2) ÷ (Fm – DISH DEPTH)]
= INV TAN [(935 ÷ 2) ÷ (359.5 –152)]
= INV TAN [(467.5) ÷ (207.5)]
= 66.0659°

CALCULATING ANGLE ϕr

(1 ÷ TAN ϕv) + (1 ÷ TAN ϕr) = 2 (Fc ÷ Ds)
(1 ÷ TAN 66.0659°) + (1 ÷ TAN ϕr) = 2 (94 ÷ 100)
0.4439 + (1 ÷ TAN ϕr) = 2 (0.94)
(1 ÷ TAN ϕr) = 1.88 – 0.4439
ϕr = INV TAN (1 ÷ 1.4361)
ϕr = 34.85°

CALCULATING DISTANCE Lv

1 - [SIN ½ (ϕv - ϕr) ÷ SIN ½ (ϕv + ϕr)] = 2 (Lv ÷ Fc)
1 - [SIN ½ (66.0659 – 34.85) ÷ SIN ½ (66.0659 + 34.85)] = 2 (Lv ÷ 94)
1 – (0.269 ÷ 0.771) = 2 (Lv ÷ 94)
0.651 = 2 (Lv ÷ 94)
Lv = (0.651 ÷ 2) x 94
Lv = 30.597 mm

CALCULATING ECCENTRICITY 'E'

$E = SIN\ ½\ (\phi v + \phi r) \div SIN\ ½\ (\phi v - \phi r)$
$E = SIN\ ½\ (66.0659 + 34.85) \div SIN\ ½\ (66.0659 - 34.85)$
$E = SIN\ ½\ (100.9159) \div SIN\ ½\ (31.2159)$
$E = 0.771 \div 0.269$
$E = 2.866$

CALCULATING VALUE 'A'

$A = Fc \div 2E$
$A = 94 \div 2 \times 2.866$
$A = 16.4$

CALCULATING VALUE 'B'

$B = A \sqrt{E^2 - 1}$
$B = 16.4 \sqrt{(2.866)^2 - 1}$
$B = 16.4 \sqrt{7.213956}$
$B = 16.4 \times 2.6859$
$B = 44.05$

CALCULATING CONTOUR OF MAIN DISH (PARABOLA)

$Xm = Ym^2 \div 4\ (Fm)$
FOR $Ym = 50$
$Xm = (50)^2 \div 4\ (359.5)$
$Xm = (50)^2 \div 1438$
$Xm = 2500 \div 1438$
$Xm = 0.017$

REPEAT FOR VARYING VALUES OF Ym TO CALCULATE CORRESPONDING VALUE OF Xm.

EG.	FOR Ym = 50	Xm = 1.739	
	Ym = 100	Xm = 6.954	
	Ym = 150	Xm = 15.647	Etcetera

CALCULATING CONTOUR OF SUB DISH (HYPERBOLA)

$Xs = A\ [\sqrt{1 + (Ys \div B)^2} - 1]$
FOR $Ys = 5$

$Xs = 16.4\ [\sqrt{1 + (5 \div 44.05)^2} - 1]$
$Xs = 16.4\ [\sqrt{1 + 0.01288} - 1]$
$Xs = 16.4\ (0.006419)$
$Xs = 0.105$

REPEAT FOR VARYING VALUES OF Ys TO CALCULATE CORRESPONDING VALUES Xs.

EG.	FOR Ys = 5	Xs = 0.105	
	Ys = 10	Xs = 0.417	
	Ys = 15	Xs = 0.925	Etcetera

CALCULATING EQUIVALENT FOCAL LENGTH Fe

$Fe \div Fm = (E + 1) \div (E - 1)$
$Fe \div 359.5 = (2.866 + 1) \div (2.866 - 1)$
$Fe \div 359.5 = 3.866 \div 1.866$
$Fe \div 359.5 = 2.072$
$Fe = 2.072 \times 359.5$
$Fe = 744.88$

CALCULATING CONTOUR OF VIRTUAL DISH (PARABOLA)

$Xe = Ye^2 \div 4Fe$
FOR $Ye = 50$
$Xe = (50)^2 \div 4\,(744.88)$
$Xe = 2500 \div 2979.52$
$Xe = 0.839$

REPEAT FOR VARYING VALUES Ye TO CALCULATE CORRESPONDING VALUES OF Xe.

EG.	FOR Ye = 50	Xe = 0.839	
	Ye = 100	Xe = 3.356	
	Ye = 150	Xe = 7.552	Etcetera

After going through the above set of calculations a series of X, Y points are obtained, with this the shape of the sub reflector could be laid out onto paper to produce a template. This could then be used to check profile accuracy if manually turning the reflector in a lathe.

After searching the Internet for any programs to calculate the reflectors profile nothing could be found and also not everybody has access to a CAD workstation on their desk so the decision was made to make the calculations available to other amateurs by writing a simple Microsoft Excel spreadsheet to do this task.

Figure 5 shows the layout of the Excel screen and user data input is on the left hand side moving to the right the data tables for the sub reflector, actual dish parabola and virtual dish parabola profiles are given. The dish and virtual dish profiles tables are given to allow the user to draw out the whole scheme if required. I am currently writing an AutoCAD Lisp program to take this data and draw a 1:1 layout within the CAD system.

The Excel program is available to download from [3] but we are getting close to the limits of what we as a group are able to do within Excel. Please feel free to play with the routine and comment.

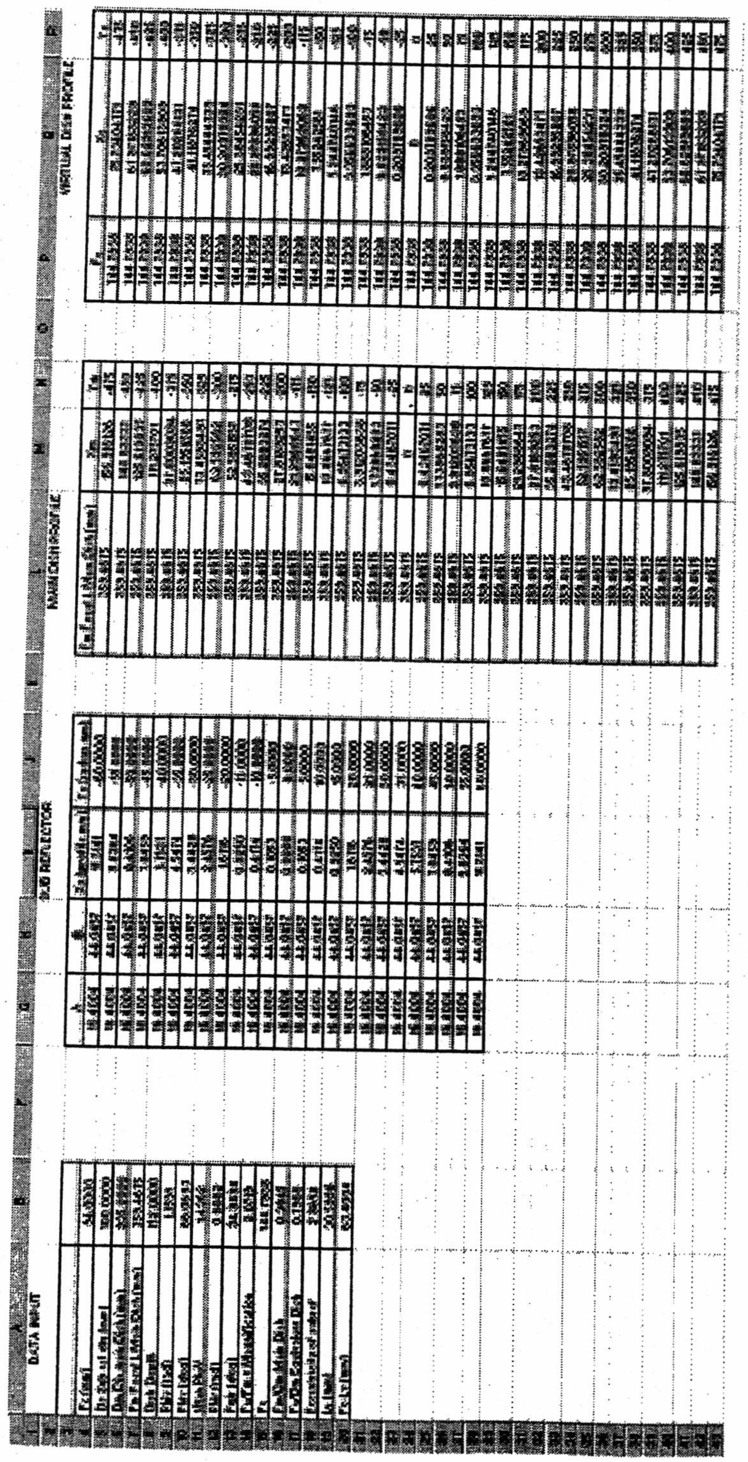

DATA INPUT
SUB REFLECTOR
MAIN DISH PROFILE
VIRTUAL DISH PROFILE

Figure 5

Figure 6 shows the completed portable transverter in use. With using such a large dish initial alignment when arriving onto the portable site is done by fitting into the sub reflector boss on the front of the dish a rifle site that looks back though a hole where the feed fits and by optically aligning the dish onto a visible landmark (180 degrees out) then working out the beam heading to this point and altering the compass on the neck of the tripod to suit. After an initial contact refinement to the compass can be made.

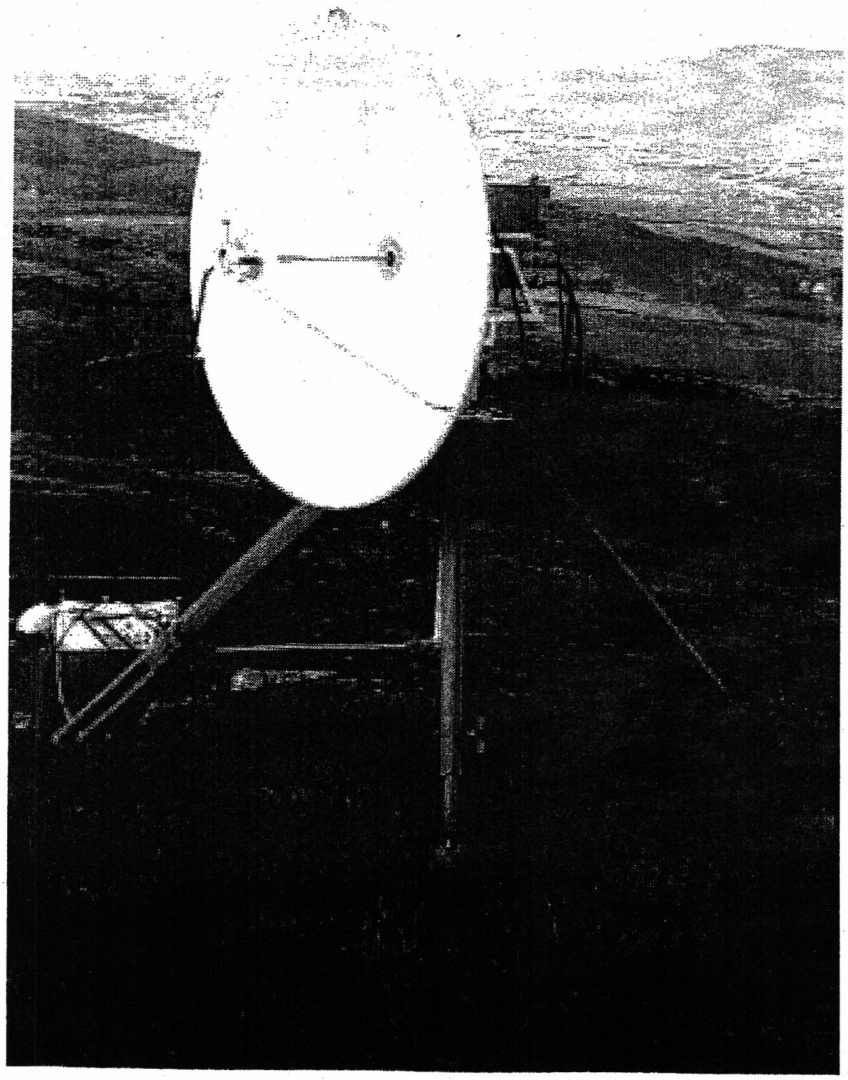


Figure 6

After working somebody with the 24GHz transverter the head unit can be replaced with the 47GHz transverter whilst still maintaining the correct beam heading by the following method

I start with standard type die-cast boxes for the equipment and a three-point female conical mounting on the lid of every box that is to be used. This is done before any equipment is installed & the position of the waveguide feed is marked using a laser pen innards mounted inside a machined cylinder that is slid into position where the sub reflector is located. The laser light beam projects through a very small hole in the cylinder and marking of the box entry are quite straightforward. Mounted on the metal framework that supports the dish is three male conical points that allow the boxes to be repeatedly positioned with good accuracy.

Figure 7 shows the 24/47GHz transverters mounted on the metal frame that holds the dish onto the tripod.

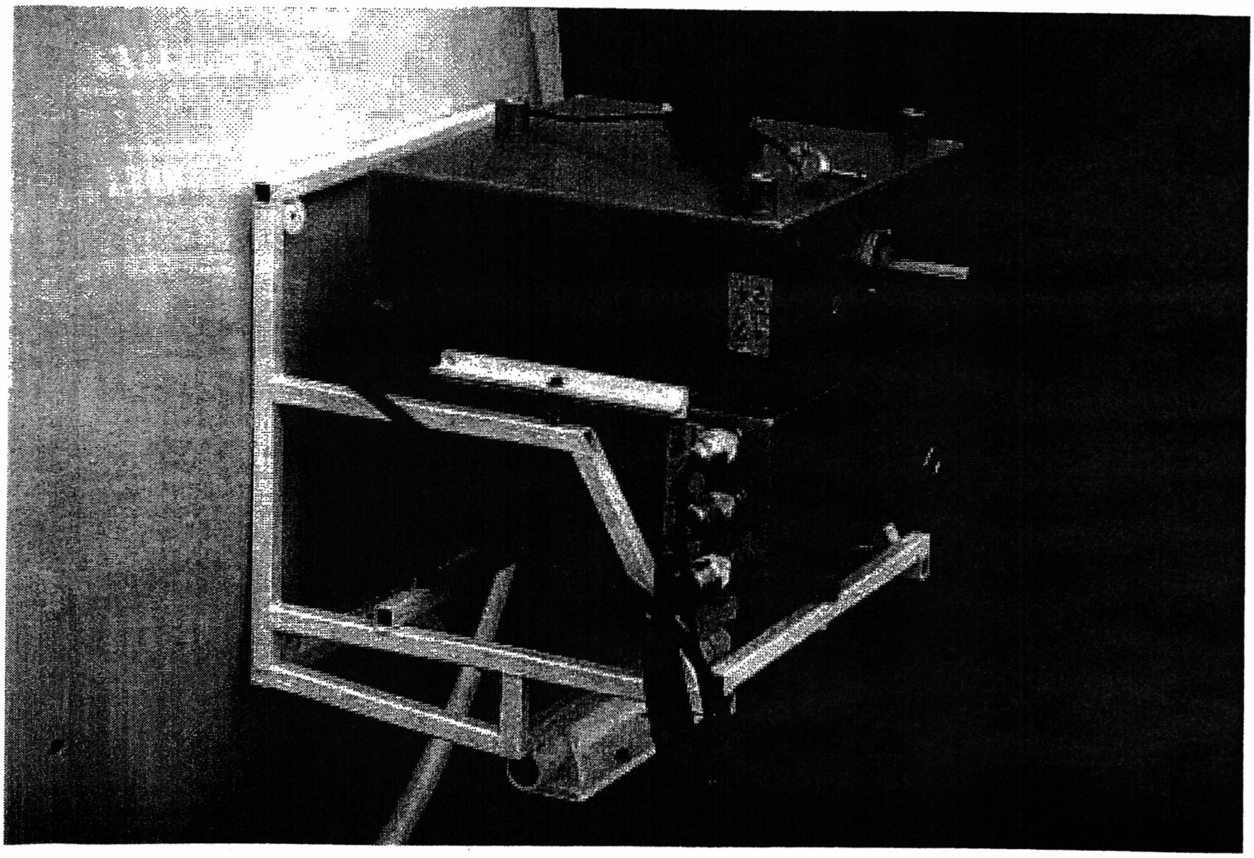
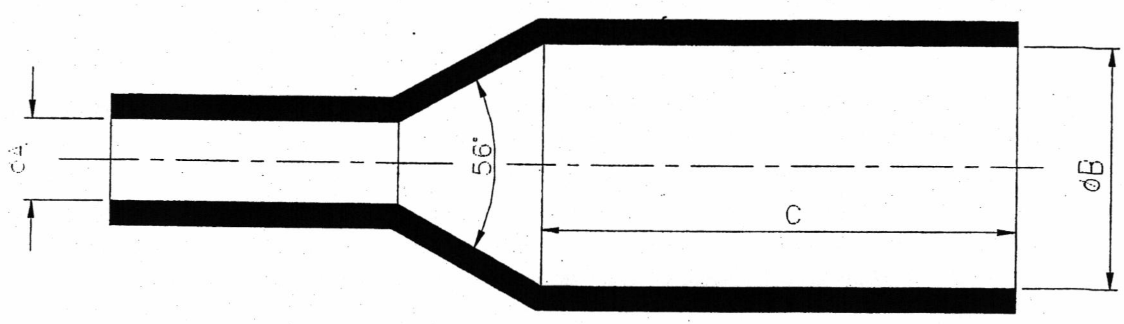


Figure 7



BAND	DIMENSION A	DIMENSION B	DIMENSION C
24GHz	8.73	23.06	46.50
47GHz	4.0	11.81	23.81

W2IMU Dual Mode feedhorns for 24GHz and 47GHz

Dimension A in the table above shows the inside diameter of K&S modelmakers brass tubing that is being used for circular waveguide. For 24GHz K&S Number 135 and for 47GHz K&S Number 129. Figure 8 shows the completed W2IMU dual mode feed horns.

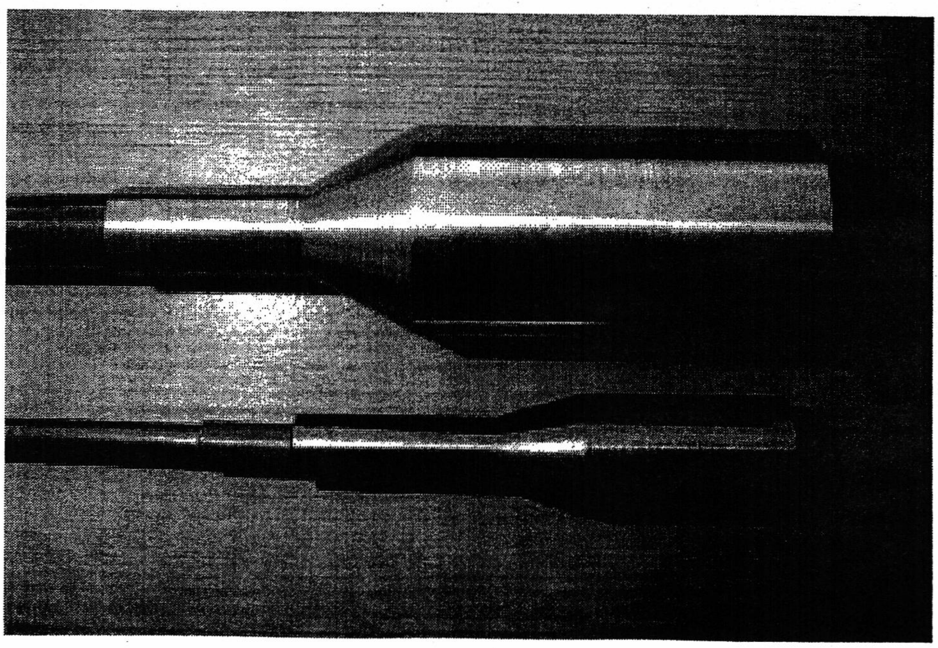


Figure 8

Conclusion

After initial outings with the system I feel it was worth the effort to go to the trouble of researching and wading through the calculations. The ability to change the parabolic dish F/D ratio to suit a particular type of feed horn in my circumstance will only be proven in time but at the moment the 24GHz system is not yet near its path potential and the 47GHz has worked 140Km without much problem.

As seen the calculations for the sub reflector are long but with the use of the spreadsheet it was easy to nudge figures one way then the other to see the effects of what if.
The spreadsheet is available to anybody to use and I welcome comments and improvements to it

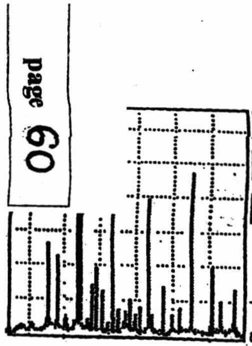
Acknowledgments

I would like to thank Dr Dennis Hawkins of [1] for assistance at the early stages of this project and confirming our profile results David Woodward G0IVA for his assistance in locating [2] and for his assistance with the calculations. Simon John Pritchard for help with the understanding of the manual calculations and work done on the spreadsheet [3] alongside John McNeil who co-wrote the routine. It was a team effort on this project.

References

1. Qpar Angus Limited, Barons Cross Lodge, Leominster, Herefordshire, HR6 8RS
2. Microwave Antennas Derived from the Cassegrain Telescope by Peter W Hannan
3. http://www.qsl.net/g3pho/cassdesign.xls

Nélangeur externe 24 Ghz et plus pour analyseur HP8555A, réalisé à l'aide d'un guide WG20, d'une diode de mélange de LNB, d'une SMA et d'un peu de mécanique.



An external mixer for the HP8555A Spectrum Analyser Plug-in

~ by Chris, G8BKE

This article describes an extension of this useful instrument to 24GHz and above, by the addition of a simple add on mixer.

In its standard form, the 8555A plug-in for the HP141T Spectrum Analyser will accept inputs in the range 10MHz to 18GHz via the input N type connector. There is a modification available to allow access to at least 24GHz via this connector but 24GHz via an N type is not really recommended. Indeed, I believe the modification precludes one from using the external mixer capability.

The 8555A will allow the use of an external HP mixer via a front panel BNC socket and this will extend the range of the instrument to 43GHz, just short of the amateur band! To obtain displays at this sort of frequency the 8555A LO is automatically multiplied in the external mixer and mixed with the test signal. In conversation with Brian, GM8BJF it turned out that he was using a "home-brew" external mixer to look at a 24GHz signal quite successfully and its was decided to try this approach to look at 47GHz. Although the amplitude calibration using this method is not really absolute, adjustments for maximum output, sideband selection and signal cleanliness can nevertheless be examined.

With no genuine surplus HP external mixers available, a small unit would have to be made as GM8BJF had done. However, without access to a 1N26E diode that Brian had used, experimentation with more readily available diodes was needed.

Since the mixer was going to be used at 24GHz and above, it was decided to make the unit in a very short length of WG20 which was to hand.

Figure 1 shows the outcome of using this approach. The SMA socket was mounted on a small 4.5mm thick brass block soldered to the guide broad face, with a ~7.5mm diameter. hole in it to accommodate the mounting of the diode. A short circuiting brass block was soldered in to the end of the guide at the position shown. The probe from the SMA protrudes into the guide via a 2mm hole in the guide wall. The probe itself can be a wire extension of the SMA pin, or better still the SMA pin itself.

A number of diodes were tried including those out of LNBs. All worked well, but, in order to obtain a readily available source, the commonly available HP HSMS 8101 diodes which are available from Farnell (Stock No. 994-649) were used. Only really applicable for use in a surface mount applications their packaging is not ideal, but since maximum efficiency is not required, they seem quite satisfactory here. Care should be exercised in soldering in the diode as static can damage the device. Correct orientation of the diode should also be observed, since a positive bias voltage is applied to it from the 8555A. There are three leads on the diode package, but only two are used so it should be mounted with the lettering as shown, ensuring that it is correctly orientated. It is quite possible that other diodes such as the DDC4561 or even a 1N21, could also be used, but these have not been tried.

The external mixer arrangement on the HP 8555A is clever in that the DC bias, IF and LO all share the same cable to the mixer. Thus if a short length of good SHF cable is used, suitable for use at 2-4GHz (e.g SUCCOFLEX) no difficulty should be experienced in obtaining results. Adjustment of the "Ext. mixer bias" pot on the 8555A optimises the mixer /multiplication process for best signal into the analyser.

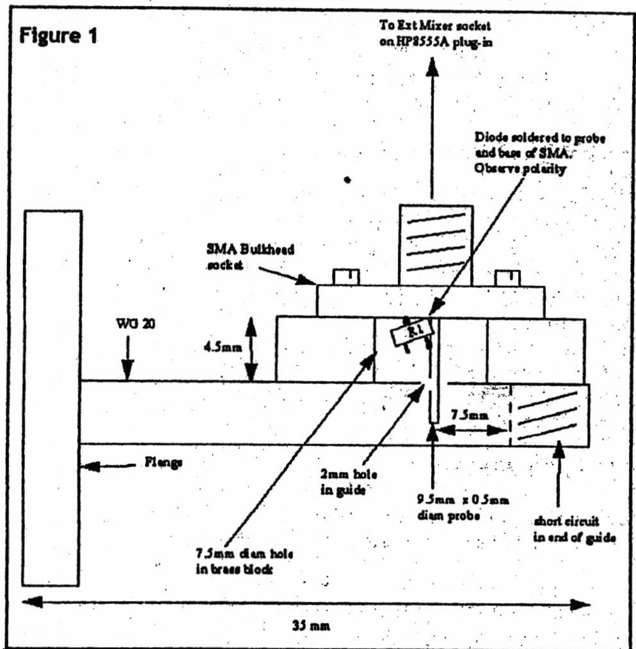
Since the frequency dial on the 8555A runs out at 43GHz, some other means has to be employed to determine the frequency of the wanted signal. Luckily HP also put the LO frequency on the top of the scales. Thus one knows what LO frequency the mixer is seeing!

The IF centre frequency on the highest ranges of the 8555A is 2.05GHz. Knowing this, one

can, by some arithmetic, work out what the LO frequency should be.

As an example, to display a 47.088GHz signal, 47.088 minus the 2.05GHz IF will require an LO of 45.038GHz. This is obtained by a x12 multiplication of 3.75GHz LO in the mixer. Thus the frequency set pointer should be set as close as possible to the LO frequency of 3.75GHz on the top scale of the instrument. Although I have not tried this, due to lack of a good 3GHz counter, it should be possible to connect a counter to the ext. mixer socket (watch the DC bias!) or the first LO output socket and with the 8555A set to "manual sweep" set the LO to precisely this frequency and then reset to "Int. sweep mode" again with the span set to say 1MHz. However, having said this, it appears that the "Signal Identifier" on the 8555A still operates in this non-standard mode thus one can check by the normal means if the signal being displayed is the correct one by the "usual two divisions to the left" offset.

Figure 1



Note that for the external mixer to function and for the LO to be directed out of the instrument correctly the 8555A *must* be operated with the band selector set to one of the frequency ranges <u>above</u> 18GHz. Also note that in this mode the "Input Attenuator" of the 8555A is not functional, although the "Log Ref. Level" control is.

If the power of the signal has been measured previously, on a power meter, then a crude calibration of the vertical scale can be made.

The signal being checked into the mixer should be kept to 2mW or less as driving it harder only produces more mixer products and tends to confuse the measurements. Keeping the input to this sort of level also ensures the longevity of the diode!

It is hoped than when I'm operational on 76GHz that the same method can be used to examine the signal at that frequency.

A similar diode mounting arrangement is being examined with a view to producing a useful waveguide noise source for noise figure measurements above 24GHz. If successful, this will form the basis of a future article.

Editor's comment... many thanks are due to Chris for providing us with yet another interesting and useful article. Other readers who have developed useful items of test gear are invited to forward details to the Newsletter.