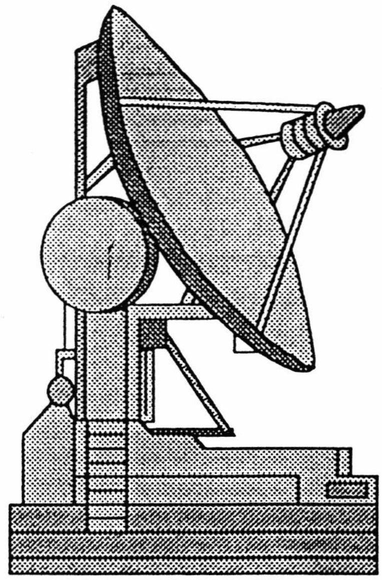
# HYPER



BULLETIN D'INFORMATIONS
DES RADIOAMATEURS ACTIFS
EN HYPERFREQUENCES

# NUMERO SPECIAL

# 5,7 GHZ

ERIC MOUTET
28 , Rue de KERBABU
SERVEL
22300 LANNION
02-96-47-22-91

**Pour " s'abonner " à HYPER :**

**Envoyer des enveloppes self-adressées et timbrées à 4,20 FF .**

## PRESENTATION

Une demande a été faite , lors de CJ 97 , à propos des 5,7 Ghz :

Pourquoi si peu de stations Françaises sur 6 cm ? Et la réponse a été par manque de documentation !

Nous avons donc préparé cette compilation de tous les articles que nous avons pu trouver à ce sujet .

Ces documents sont extraits des ouvrages ou revues suivantes :

- Feedpoint
- Hurk Infos
- UHF VHF Manuel
- VHF Communications
- Microwaves Newsletter
- Proceedings Weinheim
- Proceedings CJ
- Radio REF
- DUBUS
- CQ DL
- QEX

ou , ont été rédigés par des Oms pour le bulletin HYPER . Saluons à cette occasion le beau travail de Vincent , F1OPA et les descriptions de Jean-Luc F1BJD .

D'autres OMs ont participé à la redaction de ce numéro : F5EFD , F8UM et F9HX remercions - les des infos transmises .

# SOMMAIRE

# GENERALITES

# *LE 5,7 GHZ EN FRANCE*

La bande des 6 cm ( 5760 Mhz ) est fort délaissée en france et c'est bien dommage ! Pourtant le matériel est plus facilement récupérable ( Bande pro à 6 Ghz ) , les montages sont disponibles comme sur 3 cm ( merci à DB6NT ) , les réglages plus souples , le réglage d'un ampli est du vrai " gâteau " et les performances bien meilleures l'atténuation en espace libre étant plus faible . Alors pourquoi aussi peu de volontaires ??

PERFORMANCES :

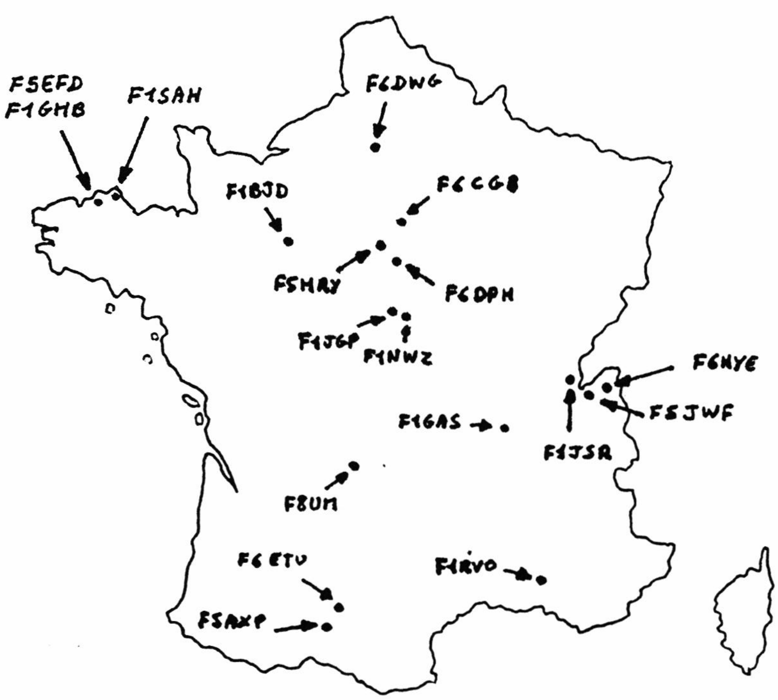
Aujourd'hui , on peut atteindre facilement le Watt sur 6 cm en émission et , pour un QSJ raisonnable ( ≅ 700 FF ) on peut même avoir un transistor de 5 W . Côté reception , un NF de 1 à 2 dB est facilement accessible et pour l'antenne , les schémas de source ne manquent pas . Le record de France en SSB est aujourd'hui de 507 km entre F1BJD/P 72 et F5JWF/P dans le 01 .

## LES STATIONS FRANCAISES

STATIONS FRANCAISES ACTIVES EN 5,7GHz

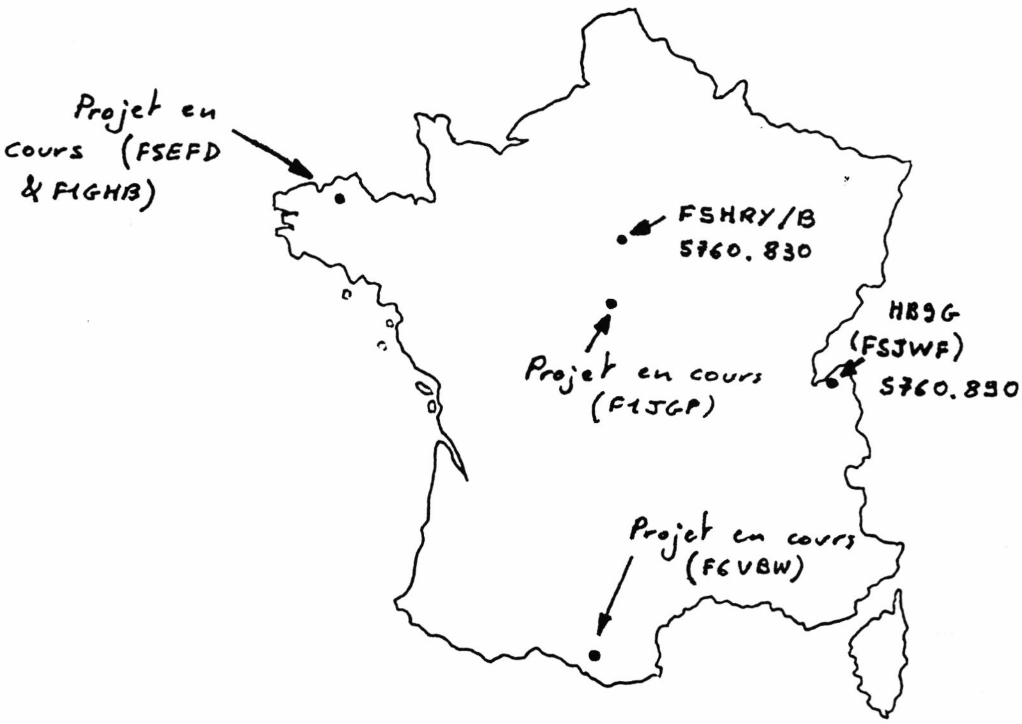
| INDICATIF | BANDE | LOCATOR | PWR | ANT | NF | PRENOM | TELPH. | REMARQUES |
|---|---|---|---|---|---|---|---|---|
| F1BJD/P | C | IN98WE | 15 | 0,9 | ? | JEAN LUC | 02-43-81-81-04 | |
| F1GAS | C | ? | ? | ? | ? | BERNARD | | |
| F1GHB/P | C | IN88IN | 10 | 0,9 | 2 | ERIC | 02-96-47-22-91 | |
| F1JGP | C | JN17CX | 17 | 0,9 | 1,7 | PATRICK | 02-38-65-51-96 | |
| F1JSR/P | C | JN36FG | 0,1 | 1,6 | ? | SERGE | 04-50-72-00-52 | ATV |
| F1NWZ | C | JN17 | 0,25 | 0,9 | 2,7 | PIERRE | 02-38-57-20-79 | |
| F1RVO/P | C | JN05 | 0,18 | 0,8 | ? | MICHEL | 04-90-85-98-39 | GJ6WDK |
| F1SAH/P | C | IN88MS | 0,2 | ? | ? | ERIC | | |
| F5AXP/P | C | JN03RQ | ? | ? | ? | DOMINIQUE | 05-61-70-45-14 | |
| F5EFD/P | C | IN88GT | 0,2 | 0,9 | ? | MAURICE | 02-96-91-04-37 | |
| F5HRY | C | JN18EQ | 2 | 0,7 | 2 | HERVE | 01-69-96-68-79 | |
| F5JWF/P | C | JN35BT | 4 | 0,9 | 0,9 | PHILIPPE | 04-50-56-72-03 | |
| F6CGB | C | JN18FW | 1 | 0,7 | ? | RENE | 01-48-30-71-04 | |
| F6DPH/P | C | JN18 | 12 | 1 | ? | PHILIPPE | 01-60-59-13-96 | |
| F6DWG/P | C | JN19DL | 0,2 | 0,6 | ? | MARC | 03-44-84-73-84 | |
| F6ETU/P | C | JN13FK | 0,25 | ? | ? | JEAN-MARIE | 05-61-20-73-90 | |
| F6HYE/P | C | JN36BI | ? | ? | ? | PATRICK | 04-50-94-19-14 | |
| F8UM/P | C | JN05 | 3.2 | 0,9 | ? | RENE | 05-55-27-90-32 | |

Note : 5,7Ghz = C, PWR en Watts, ANT en Mètres, NF en dB

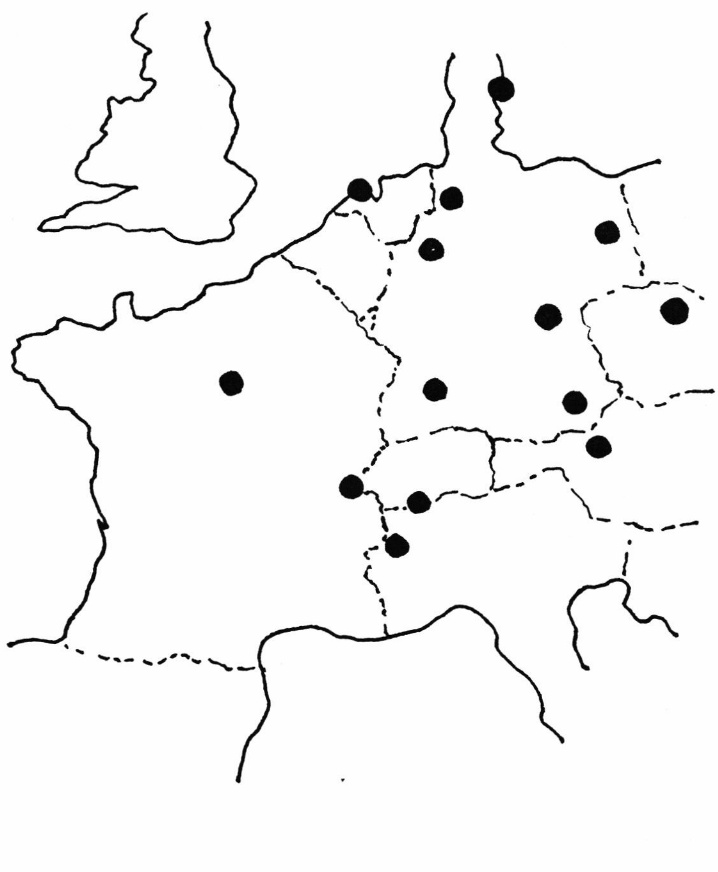


LES BALISES 6 CM

LES BALISES EN FRANCE




LES BALISES EN EUROPE



# Getting Started on the Microwave Bands

*By Rick Campbell, KK7B*

(from *QST*, February 1992)

Several pockets around the country contain much of the US amateur microwave activity. These pockets come into being when one amateur in a geographic location builds a station for the "next higher band," and then serves as a mentor to a small group of interested enthusiasts—often contesters who want to improve their scores. Recent advances in microwave technology have made access to the microwave bands easier than ever before. Building a complete microwave station in the 1990s is about as challenging as assembling a mobile rig in the 1950s—and just as much fun!

I intend this tutorial to help you get your feet wet in engineering, building and operating Amateur Radio stations on the microwave bands, with a focus on the 5.7-GHz band (which I'll sometimes refer to as "5760"). Although the focus of this article is weak-signal contest and "grid-expedition" style communication, the information provided here is useful to anyone interested in getting on any of the microwave bands. After a little introductory theory, I'll describe a basic 5760 station that you can use as-is for line-of-sight hilltopping, or that you can enhance in stages all the way to moonbounce (also known as earth-moon-earth, or EME) capability.

### Introduction

Commercial equipment is now available for narrowband, weak-signal work on all amateur bands from 1.8 MHz through 10 GHz. Until recently, the only gap in that coverage was the 5760-MHz band, a void that Down East Microwave filled with a no tune 5760-MHz transverter.[1] Amateurs who have operated on 2.3, 3.4 or 10 GHz will find no surprises at 5.7 GHz; those who haven't done so will find 5760 to be an interesting and surprisingly far-reaching band.

It's easy to find other hams interested in getting on 5760—their calls are published in the VHF and UHF contest scores in *QST*. Look for a nearby call with lots of letters after it, but no *h*.

A few of the developments responsible for the dramatic increase in amateur microwave activity are:

• Unconditionally stable MMIC amplifiers are now available at far less than the cost of the surplus amplifiers used by amateurs a few years ago. Destroying a component is now a three-dollar mistake instead of a total catastrophe!

• Printed-circuit no-tune bandpass filters, widely used in industry, have been designed for the amateur bands. These filters include the manufacturing and circuit-board tolerances in the design, and use many low-Q resonators in a flat-response bandpass filter, instead of a few high-Q, critically tuned resonators. It's now possible to obtain better performance with *no tuning at all* than was previously available with multiple interacting microwave tuning adjustments made by carefully tweaking a device while watching its output with a spectrum analyzer!

• You can build a complete microwave transverter system, using all new components, at a cost competitive with radio equipment for the lower frequencies. This makes the use of surplus components optional and removes the greatest barrier to amateur microwave operation: the lack of reasonable priced equipment.

• A pair of basic transverters can be used to test receive and transmit amplifiers, filters and antennas, *with essentially no other test equipment.*

## Propagation

Amateurs commonly use four propagation modes at 5.7 GHz: line-of-sight, obstructed paths, over-the-horizon tropospheric, and EME. Line-of-sight paths provide the strongest signals, and an excellent way for microwave experimenters to get their feet wet. The other modes are discussed in more detail in my article in the *Proceedings of the 1991 Central States VHF Conference.*[2]

## Free-Space Path Loss

The Friis path-loss formula, often quoted but seldom understood by communications-systems engineers (amateur or professional), states that: *The loss between a pair of isotropic antennas increases as the square of the distance and the square of the frequency.* The statement, "Path loss increases with frequency," taken out of the context of the Friis path-loss formula, has been repeated so many times in the literature that many amateurs and professionals believe it is a physical law. *It's not!*

Friis actually worked with *three* "path loss" formulas, the most basic of which states that the loss between an isotropic transmit antenna and a one-square-meter receiving antenna goes up as the square of the distance—independent of the frequency. This is called "geometric spreading," and it *is* a physical law.[3] The second formula is for the loss between a pair of isotropic antennas, and includes a correction factor for the effective size of an "isotropic receiver." The third formula states that the loss between two antennas of a given size goes up as the square of the distance, *and down as the square of the frequency.*[4,5,6]

Because antenna gain can be converted to effective area and vice versa, any of the three formulas can be used to obtain the correct path loss for any transmit and receive antennas. The only difference between the three formulas is in the correction factors needed for the different antenna combinations.

Most of us would rather remember a few simple rules of thumb than a bunch of formulas in a textbook. Let's convert the various "path loss" formulas into English:

1) For systems with antennas like dipoles and Yagis, where we specify antenna *gain*, the "path loss" goes up as the frequency goes up. A dipole is a fine receive antenna for 80 meters, but a 5760 dipole isn't very useful—it doesn't intercept much signal, and it's too small for a tie clip!

2) For systems with dishes and lenses, where we specify antenna *size*, "path loss" goes *down* as frequency goes up. If you use the same dish for 2304, 3456 and 5760, *and run the same power and noise figure on all three bands*, signals will be 8 dB stronger on 5760 than 2304.

*Reference Data for Radio Engineers* concludes its section on free-space path loss with this statement: "As frequency is increased, the transmit power or the antenna sizes can be reduced; it is clearly better to use the highest frequency for which generators and receivers are available."[7]

Remember: *Loss on a free-space path is independent of frequency*; the antenna gain-size correction factors are frequency dependent.

On line-of-sight paths with no ground reflections, path loss increases as 20 times the log of the path distance. In very useful terms, that works out to 6 dB every time you double the distance, or 20 dB if you multiply the distance by 10.

Let's put all this into practice. Suppose you set up two 5760-MHz systems 100 meters apart. Using a calibrated variable attenuator in the transmit feed line, you determine that the signal received on the 100-meter path is 46 dB above the noise. How far apart can these two stations be and still have a signal above the noise level? First, double the distance to 200 meters—this reduces the signal by 6 dB, to 40 dB above the noise. Next, multiply the distance by 10, to 2 km—this reduces the signal by 20 dB, to 20 dB above the noise. Multiplying the distance by 10 again, to 20 km, reduces the signal another 20 dB, down to the noise level.

Setting up portable (rover) systems on a known path and measuring the signal-to-noise ratio is an excellent way to both determine the maximum range of the system on line-of-sight paths and to verify system performance.

## Obstructed Paths, Trees and Tropospheric Scatter

When you're attempting to work over non-line-of-sight paths, many other variables become important. A few of these are frequency dependent, such as antenna height above ground. Most loss mechanisms are not very frequency dependent. In my experience, all of the microwave bands have very similar propagation on obstructed paths, *after all the bugs are worked out at both ends of the system.*

Over much of the country, even the best microwave paths have a few trees. How much attenuation do a few small trees have? If they are close to one end of an otherwise line-of-sight path, it's about 10 dB. If a few small trees occlude each end of an otherwise line-of-sight path, the tree attenuation will be about 20 dB. In other words, if the microwave system is capable of operating over a 100 mile line-of-sight path with no trees, it should be able to work over a 10-mile path with a few trees at each end.[8,9]

What about dense or big trees? Once the direct signal is attenuated by 20 dB or so, the signals bounced (or "scattered") from the treetops will be stronger than the signals coming directly through the trees. These scattered signals can be used to communicate when line-of-sight signals are attenuated by more than 20 dB. If one end of the path is in the clear and the other is inside a dense forest, expect 20 or 30 dB of additional path loss. If both ends of the path are forested, but with a clear line-of-sight path between the treetops near the receiver and transmitter, 40 to 60 dB of additional path loss may occur. The path from the forest floor to the nearby treetops near the receiver, and then down to the receiver is called a "scattered lateral wave."[10] A microwave system that can communicate 100 miles in free space may be able to communicate 1 mile in dense forest.

These numbers are speculative, but they represent a cur-

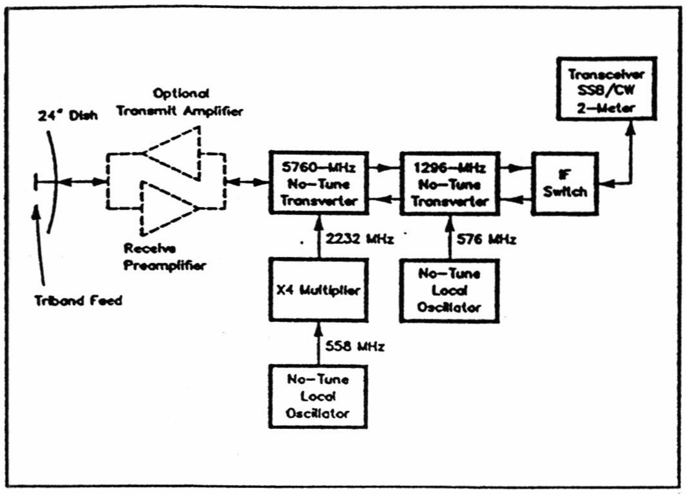



**Fig 1—Block diagram of a basic 5760-MHz transmit/receive system.**

**Approximate Gain, Size and Beamwidth of Dish Antennas**

$$\text{gain} = \eta \frac{4\pi A}{\lambda^2} \cong 5D^2$$

$$\text{Gain (dB)} \cong 7 + 20 \log (D)$$

where

$\eta$ = efficiency, typically about 0.5
A = area of dish = $\pi r^2$
D = dish diameter in wavelengths

$$\text{3-dB beamwidth} \cong \frac{90}{D} \text{ degrees}$$

$A_e$ (effective receive aperture) = $\eta$ (area of dish) $\cong$ 0.4 (diam)$^2$

These formulas are generally accurate to within 10%, or less than 1 dB.—*KK7B*

rent consensus of experimental and anecdotal data by professional and Amateur Radio scientists. In any case, they are the best information currently available. This is an area in which amateurs can make a meaningful contribution to radio science.

*Tropospheric scatter* (also known as *troposcatter*) is a practical propagation mode on all of the amateur microwave bands. The US Army has operated long-range portable troposcatter links above 3 GHz for many years. Amateur experience indicates that a 5.7-GHz station equipped with a small TVRO dish, a GaAsFET preamp and a 10-watt amplifier is approximately equal to a 150-watt, single-Yagi 2-meter station—reliable for SSB communications out to about 250 miles and occasionally supporting CW work out to about 400 miles. During enhancements, much greater distances are possible.

## Antennas

Selecting an antenna for 5760 MHz is easy: Nothing matches the performance of a small dish. Approximate formulas relating the gain, diameter and beamwidth for dish antennas are given in the sidebar "Approximate Gain, Size and Beamwidth of Dish Antennas." For example, a 24-inch dish at 5.8 GHz has the following gain, beamwidth and effective aperture, according to the formulas given in the sidebar:

Dish diameter: 11.7 wavelengths
Gain: 28 dB
3-dB beamwidth: 7.7 degrees
Effective aperture = 0.15 $m^2$

The gain can be improved by up to about 2 dB by using an optimized dish feed. Several excellent articles have been published on such feeds. The recent one by Barry Malowanchuk, VE4MA, is particularly useful.[11]

Unlike the lower-frequency bands, it's fairly easy to develop *too much* antenna gain at 5.7 GHz. Recall that gain and beamwidth are inversely related. My 19-inch dish on a small camera tripod has a nice 10-degree beamwidth. It is ideal for loaning to high-school students running a grid expedition to a

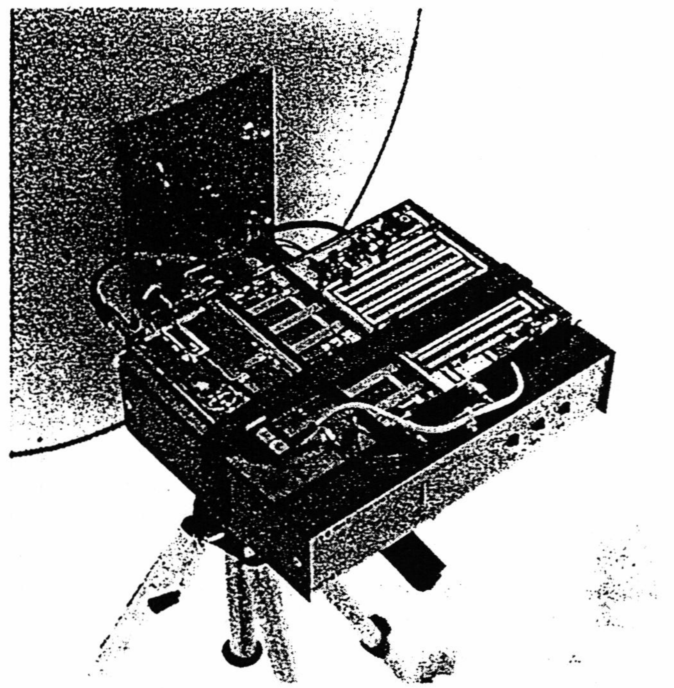


**Fig 2—A complete 5760-MHz station and 24-inch dish antenna, less IF transceiver and 12-V power supply, mounts on an inexpensive camera tripod and is ready to go anywhere!**

windy hilltop for the first taste of microwave Amateur Radio. Even a 10-degree beamwidth is awful narrow when you can't see the other station and aren't sure where to point the antenna. In addition to antenna aiming uncertainty, there's also frequency uncertainty. For the contact to occur, the antenna aiming and radio tuning must be correct at the same time. For beginning operators, or experienced operators on cold, wet hilltops with state troopers slowing down to ask questions, aiming a narrow-beamwidth antenna can be frustrating.

## Building a Basic Station for 5760 MHz

This section presents a basic 5.7-GHz station. You can improve this station by adding amplifiers, TR switching and a bigger dish. This is in contrast to another common microwave system concept, in which an operator builds a "simple" duplex station to make a few contacts with one other station, and then has to build a whole new station if he or she becomes interested in working someone else on the weak-signal calling frequency (5760.1 MHz).

The entry-level station I'll describe here is simple, almost foolproof, can be built at low cost using only new components, and can serve as the core of a moonbounce or troposcatter system. Of equal importance is that the entry-level station serves as test equipment for all of the antennas, feed lines and amplifiers as they're added to the system.

There are two attractive approaches to entry-level 5760 stations that can be easily enhanced up through the EME level (and beyond). The first approach, used by many hams already on the band, is to convert a Frequency West-type surplus phase-locked oscillator to 5760 plus or minus 144 MHz, use a surplus microwave mixer, and retune a surplus filter to pass 5760 mHz. The surplus route has some advantages—if you have access to the parts, expertise and test equipment needed to get it up and running:

- It can be very inexpensive
- Converting surplus is a long and honorable Amateur Radio tradition.
- There may be a psychological advantage—it seems easier to start with something that already works and modify it rather than to make something new from a little Teflon board and a teaspoonful of parts.
- Converting surplus is fun—and is the perfect excuse for buying and using a spectrum analyzer—which is even more fun, but negates the first advantage.

The second approach to an expandable entry-level 5760 station involves assembling a station from no-tune boards, as discussed in October 1990 *QST*.[12] This approach has numerous advantages, mostly because microwave technology has come a long way since those Frequency West local oscillators were built a quarter century ago. Some of the advantages of the no-tune approach are:

- You don't need microwave engineering talent to get the basic station on the air.
- It uses all new, inexpensive parts, and you can duplicate or repair it at low cost.
- You don't need any RF test equipment.
- It is very easy to assemble and very reliable.
- Most importantly, a station can be put on the air by anyone with basic soldering and construction skills—not just mystical guru spectrum-analyzer engineers.

## System Block Diagram

The block diagram of a 5760-MHz transverter system using no-tune boards is shown in Fig 1. (The block diagram of the entry-level system using surplus components is not shown—it depends on the available components, and if you're capable of assembling a microwave station from surplus components, you don't need my block diagram!) Note that a 1296-MHz IF is needed with the no-tune 5760 transverter board. Three options for a 1296-MHz IF are (1) a 1296-MHz multimode transceiver (expensive but increasingly popular); (2) an existing 1296 transverter with a multimode IF transceiver; and (3) a no-tune 1296 board with a multimode 2-meter IF radio.

### Where to Get the Pieces

In addition to flea markets and VHF conferences, here are a few places where you can get the parts you'll need to get on any microwave ham band.

Down East Microwave, RR 1, Box 2310, Troy, ME 04987, tel 207-948-3741, fax 207-948-5157: no-tune transverters, antennas, triband dish feeds, components, etc; catalog available.

Microwave Components of Michigan, PO Box 1697, Taylor, MI 48180, evening tel 313-753-4581: components; price list available.

The Antenna Center, 505 Oak St, Calumet, MI 49913, tel 906-337-5062: spun-aluminum dishes; price list available.

Steve Kostro, N2CEI, RD 1 Box 341A, Frenchtown, NJ 08825, tel 201-834-1304 (9AM to 7 PM) and 201-996-3584 (after 9 PM): components; price list available.

### *2-Meter IF Transceiver*

Any 2-meter multimode radio can be used as a microwave IF. Older rigs are simple, inexpensive, and have poor sensitivity, which is fine, because the system noise figure is set by the 1296-MHz IF or 5760-MHz preamp. My favorite is the ICOM IC-202, a nearly ideal microwave IF rig in the $100 to $150 price range. These radios, unfortunately, are becoming quite hard to find. The original Yaesu FT-290R is my second choice, as it is small, has a few useful bells and whistles and low battery drain. The typical price range for these rigs is about $200 to $300, depending on condition and accessories. For those who prefer a "real" radio, the analog-tuned Kenwood and Yaesu radios from the TS-700 era are widely available in the $250 to $350 price class. I personally prefer digging for weak CW signals with an analog knob, rather than a 100-Hz click-step synthesizer knob.

### *Antennas*

As mentioned earlier, for an entry-level antenna, it's hard to beat a 24-inch dish (UPS shippable and currently about $70 from The Antenna Center [see the sidebar "Where to Get the Pieces"]). Use the triband feed designed by Tom Hill, WA3RMX[13] (available for $15 when this book was being prepared from Down East Microwave), mounted on an inexpensive tripod from K-Mart or another variety store. Feed efficiency can be improved later, but the 24-inch dish is just about optimum for both hilltopping and long-haul tropo work. For what other band can you buy the same antenna the big guns are using, brand new, for $70?

## Communication with Entry-Level Stations

As Table 1 shows, a pair of entry-level stations are capable of surprising DX from hilltop to hilltop—roughly the same as a pair of 10-GHz Gunnplexer stations with small dishes. A pair of operators willing to locate the appropriate set of grid squares and hilltops can easily work VUCC. Entry-

**Table 1**

**Basic 5760-MHz Station Characteristics**

Power output: –10 dBm (100 µW, or 0.1 mW)
Noise figure (NF): 10 dB
Bandwidth: 100 Hz or 2.5 kHz*
Antenna gain ($G_a$): 25 dB†
Effective aperture ($A_e$): 0.1 m²†
Maximum communication distance: 8000 km (5000 miles)**
Routine SSB communication distance: 160 km (100 miles)††
SN100:*** 84 dB

*100 Hz is an approximation for the ear-brain bandwidth when copying weak CW through an SSB filter.

†$G_a$ and $A_e$ are for a 24-inch dish with 25% efficiency (includes illumination and feed-line losses).

**Represents a 0-dB signal-to-noise CW signal at the human hearing threshold.

††Represents "armchair copy" SSB (20 dB above the noise in a 2.5-kHz bandwidth).

***Expected signal-to-noise ratio in a 2.5-kHz bandwidth of two identical 5760-MHz systems aimed at each other over an unobstructed 100-mile path.

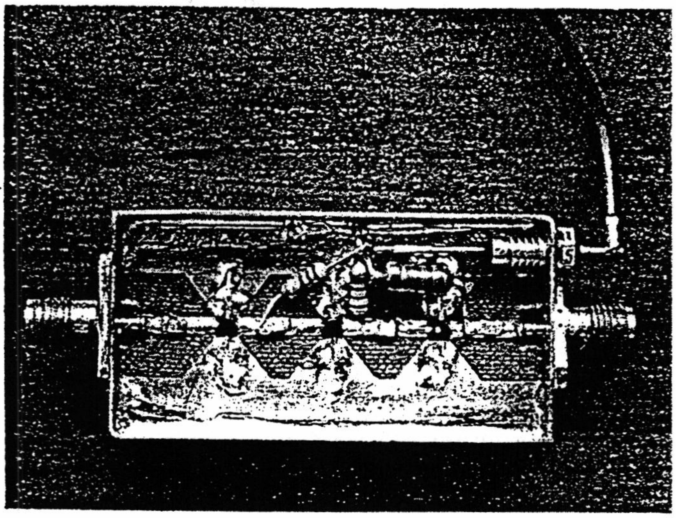


Fig 3—This simple utility amplifier, suitable for 903, 1296, 2304, 3456 and 5760-MHz transmitting and receiving use, consists of three MMICs, three ordinary resistors and five chip capacitors.

level stations are nearly ideal for contesting because they use no amplifiers that could oscillate or fail, no expensive microwave relays with their associated connectors and switching-voltage complications, and they can be mounted right at the dish to eliminate feed-line loss and all the other complications of using feed lines. The entry-level station is all many operators will ever need, and smart contesters experiment with entry-level stations before adding the complexity of amplifiers and relays. Fig 2 shows a complete entry-level system built entirely from Down East Microwave no-tune boards.

### Entry-Level Stations as Test Equipment

The nature of microwave hamming is such that you either own two complete systems for each microwave band or are within driving distance of another station. The ideal piece of test equipment for a 5760-MHz station is a second 5760 station. If each is an entry-level station made from no-tune boards, then the approximate power output, noise figure, antenna gain, and so forth, are known and, more importantly, stable.

One rule needs to be followed: The local oscillators (LOs) in the two transverters need to be on slightly different frequencies. Otherwise, it's impossible to separate the up- and down-converted microwave signals from IF-signal leakage (sometimes called *breakthrough*). Normally the natural offsets of fifth-overtone crystals are more than enough to provide the needed LO frequency offset, but if you're not sure whether the signal you're hearing is at 5760 MHz or IF breakthrough, wave your hand near the antenna and listen for Doppler fluctuations, or breathe on the LO crystal to change its temperature, which should provide enough frequency shift for you to hear easily.

A number of useful tests are possible by setting up two stations as described in Table 1 facing each other over a path of about 100 m.

• Place a 40-dB pad in the 1296-MHz IF line of each system. SSB signals should be just barely readable both ways. With pads in the IF lines and signals at the "just-barely readable" level, you can compare the effects of different dish feeds, feed lines and filters. By using CW, disabling receiver AGC and using an audio-output meter calibrated in decibels, you can make quantitative comparisons. If a standard-gain antenna is available, you can also make antenna-gain measurements.

• You can determine frequency offsets and study the effects of time and temperature on stability. I write the frequency offset (ie, "5760.100 MHz = 144.126 MHz") on each transverter with an indelible marker.

• On the bench, an entry-level system can serve as a low-level source for aligning filters and measuring amplifier gain, filter loss and transmission-line loss. In receive, the entry-level system provides a way to listen and compare signal levels in transmitters.

### Enhancing the Entry-Level Station

The entry-level transverter is the heart of a more elaborate station for working over obstructed paths, and via troposcatter and moonbounce. Moving up to the next level requires adding 5760-MHz gain to the transmit and receive signal paths. You need three things to do this: (1) a way to separate the transmit and receive signal paths; (2) a receive preamplifier; and (3) a transmit power amplifier.

The easiest way to split the transmit and receive paths is with a pair of microwave relays. These are expensive, but are often available surplus at hamfests and flea markets. Once the transmit and receive paths are separated, you can add amplifiers. A particularly useful amplifier for low-level transmit and receive use is shown in Figs 3 and 4. At 5760 MHz, the three-MMIC circuit has a gain of about 15 dB, a noise figure (NF) of about 7 dB and a few milliwatts output. Station enhancements are discussed in more detail in my article in the *Proceedings of the 1991 Central States VHF Conference.*[14]

### Grid-Hopping Operating Hints

Here are four suggestions that have been useful in the last

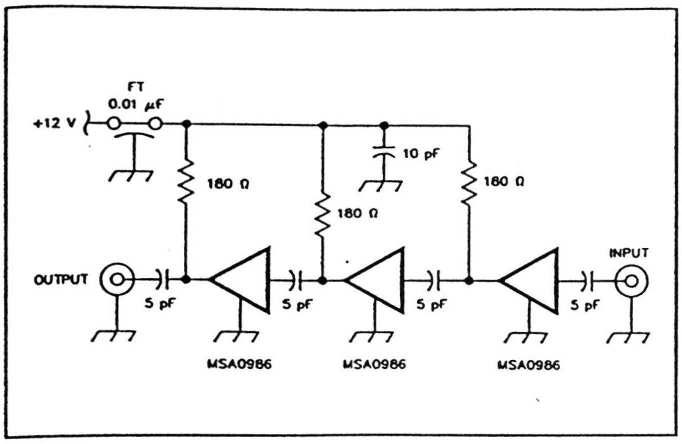



**Fig 4—Schematic of the utility amplifier. Chip capacitors and MMICs are available in small quantities from Down East Microwave, Microwave Components of Michigan, and other sources.**

ten years of microwave grid hopping at KK7B:

• Set up and test everything (both stations) the week before the contest or grid expedition. Fix any loose connectors, charge batteries and make sure everything is ready to go.

• Pick a reasonable liaison frequency. Two-meter FM handhelds are useless for all but the shortest microwave paths. Two-meter SSB radios with ⅝-wavelength whips are okay for medium-length paths, but remember that even the simplest pair of 5760 stations with 24-inch dishes is capable of operating SSB over a 100-mile path. With amplifiers, the 5760-MHz system will outperform 2-meter SSB stations with 150-watt "bricks" and Yagis! During the last few KK7B/8 multiband grid expeditions, I've used 5760 MHz as the liaison frequency for the lower bands!

• Work your own grid square first by setting up both stations a couple of hundred feet apart. This confirms that everything still works, permits a final check on operating frequencies, and gives you at least one contact on the band, in case the weather turns foul or something fails.

• Most "line-of-sight" microwave paths have a ground-reflected component. Ground-reflected signals may add or subtract with the direct signal. Sometimes aiming a dish a few degrees above the horizon improves signal strength by reducing the ground-reflected component.

## Conclusions

The 5760 stations described in this article are portable, lightweight, reliable and can be battery powered. They can be operated from home stations, but the *real* fun begins when you take the station to a hilltop or beach. You can earn a 5760-MHz VHF-UHF Century Club (VUCC) award by working just five grid squares on 5760. Obtaining VUCC from a portable location is not only practical, but is generally *easier* than working five grids from a home station. This is one of the few areas in ham radio where you can be a "big gun" while living in an apartment, dormitory, barracks or "controlled community"! Two-foot dishes fit nicely in the trunk or back seat of a car, and a backpack station is entirely feasible.

An operator who concentrates on 5760 and coordinates with other 5760 operators can stay happily busy for an entire contest weekend. And this is without even considering adding a couple more bands to your setup, which you can do easily using no-tune transverters and a multiband dish feed.[15] You don't even have to get permission in advance to set up a camera tripod and small dish at a scenic overlook.

Perhaps best of all, you get to drive around and enjoy the hilltops and beaches between contacts, instead of gritting your teeth in front of an overloaded 2-meter rig while your neighbor works the rare grids with his legal-limit amplifier!

## Notes


[1]R. Campbell, "A Single-Board Bilateral 5760-MHz Transverter," *QST*, Oct 1990, pp 27-31 (reprinted in Chapter 3 of this book).

[2]R. Campbell, "A 5760 MHz Primer," *Proceedings of the 1991 Central States VHF Conference* (Newington: ARRL, 1991), pp 7-24.

[3]A. Ishimaru, *Electromagnetic Wave Propagation, Radiation and Scattering* (Englewood Cliffs, NJ: Prentice-Hall, 1991), pp 123-127.

[4]*Reference Data for Radio Engineers*, 6th Ed. (Indianapolis: Howard W. Sams, 1975), pp 33-3, 33-7.

[5]R. Campbell, "Does Path Loss Increase with Frequency?" Technical Correspondence, *QST*, Jan 1991, p 38.

[6]A. Hambley, *An Introduction to Communications Systems* (New York: W. H. Freeman and Co, 1990) pp 168-169.

[7]See note 4.

[8]D. Hilliard and B. McCaa, "Preliminary Results of Foliage Attenuation Measurements at 902 MHz and 3456 MHz in the Boulder, Colorado Area," *Proceedings of Microwave Update '89* (Newington: ARRL, 1989), pp 133-147.

[9]W. Vogel and J. Goldhirsch, "Roadside Tree Attenuation Measurements at UHF for Land Mobile Satellite Systems," *IEEE Transactions on Antennas and Propagation.*

[10]R. Campbell and H. Wang, "A Theoretical and Experimental Study of Scattered Lateral Wave Propagation in Forest," *Digest of URSI North American Radio Science Meeting*, London, Ontario, Jun 1991.

[11]B. Malowanchuck, "Use of Small TVRO Dishes for EME," *Proceedings of the 21st Conference of the Central States VHF Society* (Newington: ARRL, 1987), pp 66-77.

[12]See note 1.

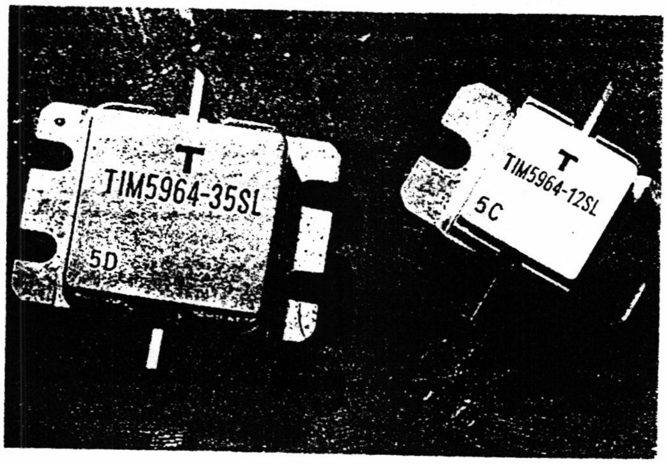
[13]T. Hill, "A Triband Microwave Dish Feed," *QST*, Aug 1990, pp 22-27 (reprinted in the Antennas chapter of this book).

[14]See note 2.

[15]See note 13.

# *REVONS UN PEU !!!*

Ces derniers mois , nos " fournisseurs " préférés ont sorti une nouvelle génération de GaAsFets de puissance dans la bande C ( aux environs des 6 Ghz ) , peut être nos futures " récupérations " . Pour se faire saliver , un petit aperçu :




**TOSHIBA**

**The C-band GaAs FETs at a glance**

| Model | Frequency range (GHz) | Output power (dBm) | Gain (dB) | Power-added efficiency (percent) | Typical third-order intermodulation distortion (dBc) |
|---|---|---|---|---|---|
| TIM3742-4SL | 3.7 to 4.2 | +36.5 | 10.5 | 37 | –45 |
| TIM3742-8SL | 3.7 to 4.2 | +39.5 | 10.0 | 36 | –45 |
| TIM3742-16SL | 3.7 to 4.2 | +42.5 | 9.5 | 36 | –45 |
| TIM3742-30SL | 3.7 to 4.2 | +45.0 | 10.0 | 41 | –45 |
| TIM5964-4SL | 5.9 to 6.4 | +36.5 | 9.0 | 35 | –45 |
| TIM5964-8SL | 5.9 to 6.4 | +39.5 | 8.5 | 35 | –45 |
| TIM5964-16SL | 5.9 to 6.4 | +42.5 | 8.0 | 34 | –45 |
| TIM5964-30SL | 5.9 to 6.4 | +45.0 | 8.0 | 38 | –45 |
| TIM6472-4SL | 6.4 to 7.2 | +36.5 | 8.0 | 34 | –45 |
| TIM6472-8SL | 6.4 to 7.2 | +39.5 | 7.5 | 33 | –45 |
| TIM6472-16SL | 6.4 to 7.2 | +42.5 | 7.0 | 32 | –45 |
| TIM7785-4SL | 7.7 to 8.5 | +36.5 | 6.5 | 32 | –45 |
| TIM7785-8SL | 7.7 to 8.5 | +39.5 | 6.0 | 30 | –45 |
| TIM7785-16SL | 7.7 to 8.5 | +42.5 | 5.5 | 29 | –45 |
| TIM7785-30SL | 7.7 to 8.5 | +45.0 | 6.0 | 34 | –45 |

**Typical C-Band Performance**

| | | | |
|---|---|---|---|
| FLC091WF | 6.0 | 28.8 | 8.5 |
| FLC161WF | 6.0 | 31.8 | 7.5 |
| FLC253MH-6 | 6.4 | 34.0 | 9.0 |
| FLM5964-4D | 5.9-6.4 | 36.0 | 9.0 |
| FLM5964-6D | 5.9-6.4 | 38.0 | 8.0 |
| FLM5964-8D | 5.9-6.4 | 39.0 | 8.0 |
| FLM5964-12DA | 5.9-6.4 | 41.0 | 9.5 |
| FLM5964-18DA | 5.9-6.4 | 42.5 | 8.5 |
| FLM5964-25DA | 5.9-6.4 | 44.0 | 8.5 |

FUJITSU

# TRANSVERTERS 6 cm

# TRANSVERTER 6CM

**GRIGIS Vincent F1OPA**
**16 rue Eugène Delacroix**
**38000 GRENOBLE**
**Tél : 04-76-15-33-64**

Suite à CJ et à l'apparition sur le marché de nouveaux MMIC, l'idée m'est venue de réaliser un transverter 6cm ayant des performances correctes, une mise au point simple et un prix de revient réduit.

## DESCRIPTION (Fig. 1)

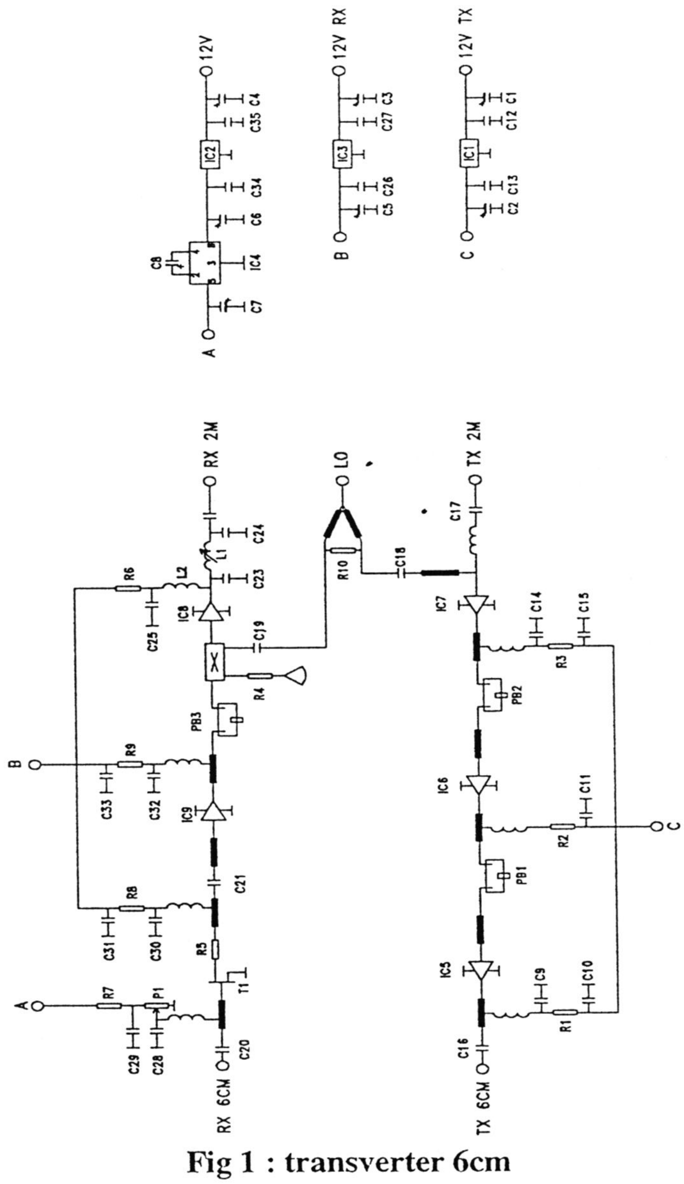



**Fig 1 : transverter 6cm**

| Désignation | Référence |
|---|---|
| C1,C2,C3,C4,C5 | Condensateur 2.2uF |
| C6,C7,C8 | Condensateur 10uF |
| C9,C14,C23,C24,C28,C30,C32 | Condensateur CMS 10pF, boitier 805 |
| C10,C11,C15,C17,C22,C25,C29,C31,C33, | Condensateur CMS 100pF, boitier 805 |
| C12,C13,C26,C27,C34,C35 | Condensateur CMS 10nF, boitier 1206 |
| C16,C18,C19 | Condensateur CMS 3.3pF, boitier 805 |
| C20,C21 | Condensateur CMS 1.5pF, ATC |
| R1 | Résistance CMS 91 Ohms, boitier 1206 |
| R2,R3 | Résistance CMS 120 Ohms, boitier 1206 |
| R4 | Résistance CMS 47 Ohms, boitier 805 |
| R5 | Résistance CMS 12 Ohms, boitier 1206 |
| R6,R10 | Résistance CMS 100 Ohms, boitier 1206 |
| R7 | Résistance CMS 22 KOhms, boitier 1206 |
| R8 | Résistance CMS 270 Ohms, boitier 1206 |
| R9 | Résistance CMS 470 Ohms, boitier 1206 |
| L1 | NEOSID BV5061 |
| L2 | Inductance CMS 2.2uH |
| IC1 | Régulateur 7808 |
| IC2,IC3 | Régulateur 78L05 |
| IC4 | Convertisseur 7660 |
| IC5,IC8 | MMIC ERA2 |
| IC6,IC7,IC9 | MMIC ERA3 |
| T1 | HEMT ATF36077 |
| PB1,PB2,PB3 | Voir description |
| | 5 connecteurs SMA |
| | Boîtier 111*74*30 |

**Nomenclature des composants**

L'oscillateur local, d'une puissance de 10 mW, est reparti entre la chaîne d'émission et de réception grâce à un Wilkinson.

**** Réception***

La chaîne de réception comporte 3 étages; Deux pour l'amplifications (T1, IC9) et un mélangeur (IC8). Ce dernier est sous-polarisé afin de le rendre non linéaire, ce qui permet d'obtenir un mélange, entre la RF et l'OL, optimum. La résistance R5, dans le drain de T1, permet d'assurer sa stabilité.

Pour permettre de diminuer le facteur de bruit, il est important de soigner la réalisation de la cavité PB3 (Fig. 2). Celle-ci étant sous-dimensionnée, les pertes d'insertion sont loin d'être négligeables.

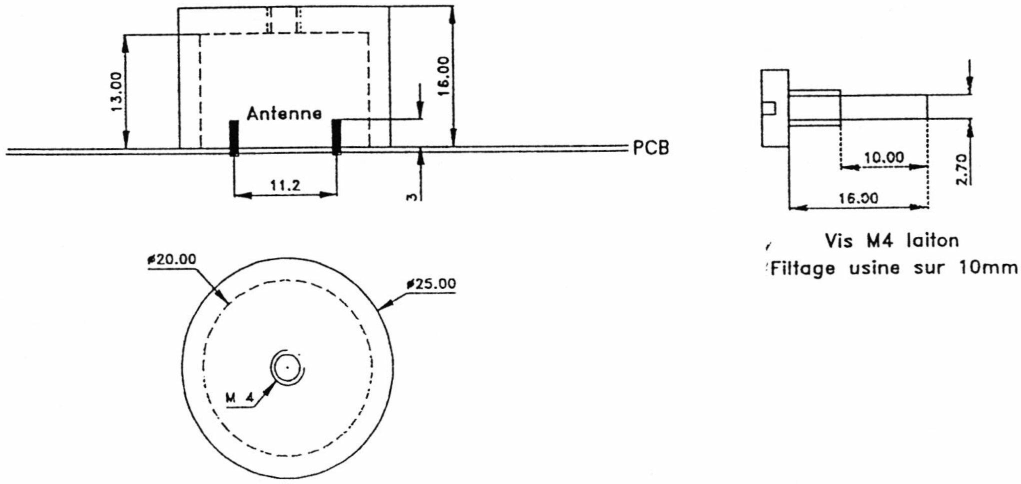



**Fig. 2 : Cavité 6cm**

**Emission*

La chaîne d'émission comporte elle aussi 3 étages, deux pour l'amplifications (IC5,IC6) et un mélangeur (IC7). Le produit de mélange désiré est filtré une première fois par PB2, amplifié par IC6, de nouveau filtré par PB1 puis amplifié à nouveau par IC5 (Ce dernier pourra être remplacé par un autre MMIC ayant un point de compression plus élevé, en fonction de l'évolution de la technologie). Le réglage de la fréquence de résonance de PB1 et PB2 pourra être fait avec une vis laiton M4 classique, le filetage étant conservé, contrairement à PB3.

## DESCRIPTION

**Circuit imprimé* **(Fig. 3)**

Le circuit imprimé est réalisé sur du DUROID d'épaisseur 0.79mm et de permittivité 2.55.

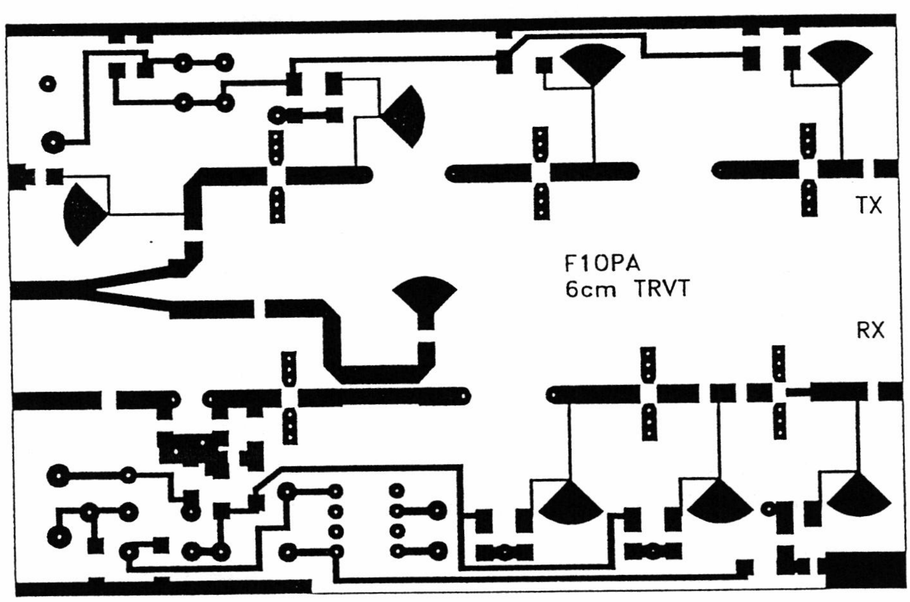



**Fig 3 : Circuit imprimé**

*** *Implantation des composants* (Fig 4, Fig 5)**

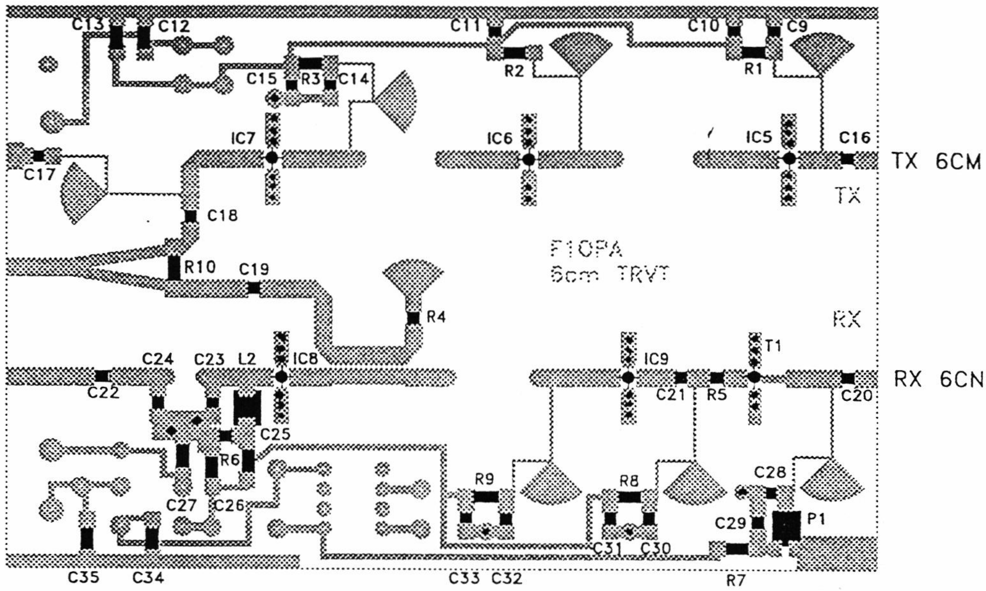



**Fig 4 : Coté pistes**

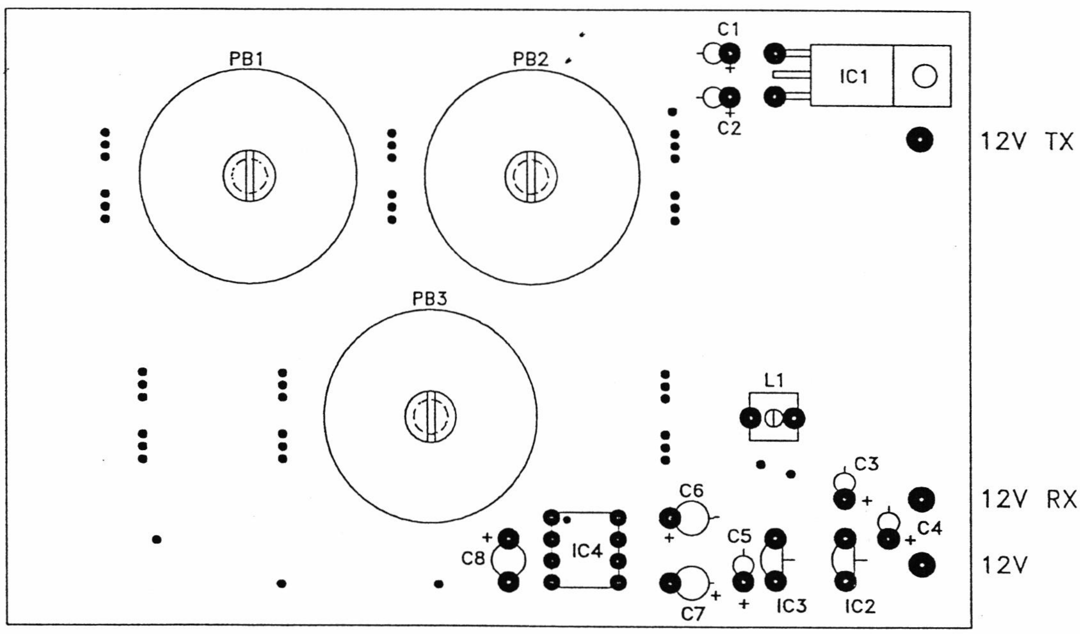



**Fig 5 : Coté composants**

L'ensemble est prévu pour rentrer dans une boite 111*74*30. Tous les composants sont classiques à l'éxception de C20 et C21 qui sont prévus pour fonctionner à cette fréquence.

Afin d'éviter les risques d'oscillation, il est important de soigner la mise à la masse , celle-ci est réalisée grâce à des vias.

Pour souder facilement les cavités, il est nécessaire de les faire chauffer quelques minutes à 300°C sur une plaque chauffante.

T1 étant fragile, il est préférable de le souder en dernier en faisant attention aux décharges statiques.

Le couvercle, coté pistes doit être recouvert de mousse absorbante afin d'éviter les effets de boitier.

## MISE AU POINT

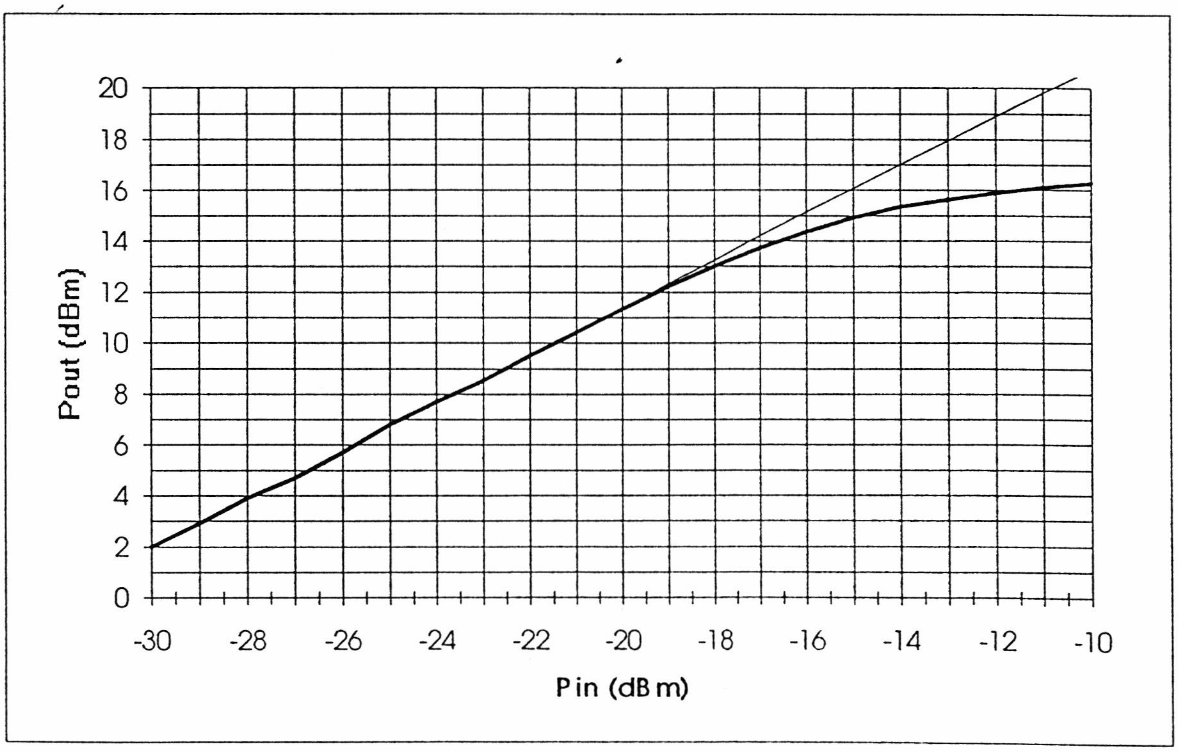
** Emission*

S'assurer tout d'abord que les différents MMIC sont correctement polarisés.

| | V (V) | I (mA) |
|---|---|---|
| **IC5** | **3.52** | **37.5** |
| **IC6** | **3.52** | **37.5** |
| **IC7** | **3.63** | **48.3** |

A part l'accord des cavités PB1 et PB2, il n'y a pas de réglage particulier.

J'obtiens les résultats suivant pour Pout=f(Pin) :



Le point de compression à 1dB est légèrement inférieur à 15dBm. Une mesure à l'analyseur de spectre indique que la raie la plus proche (OL) est à -35dB.

***Réception***

Régler grâce à P1 le point de polarisation de T1 et s'assurer que les autre étages sont bien polarisés.

| | V (V) | I (mA) |
|---|---|---|
| **T1** | **Vd=2V** | **Id=10mA** |
| **IC8** | **3.50** | **15.9** |
| **IC9** | **3.54** | **28** |

Grâce à une source sur 5760MHz on règle PB3 de façon à avoir l'amplitude maximum sur la FI, ensuite grâce à trois stubs on optimise le réglage de la chaîne de réception (Fig. 6).

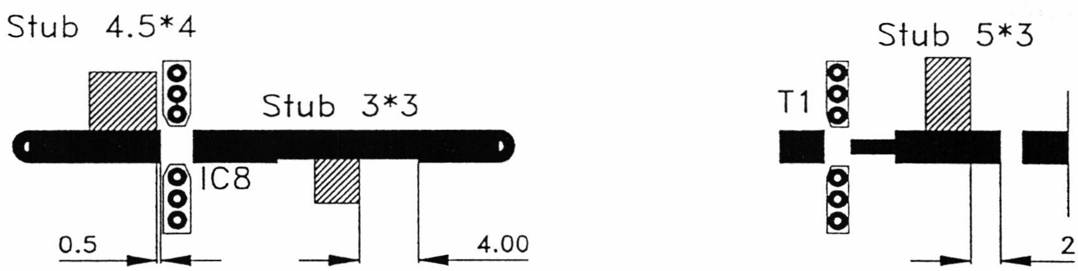
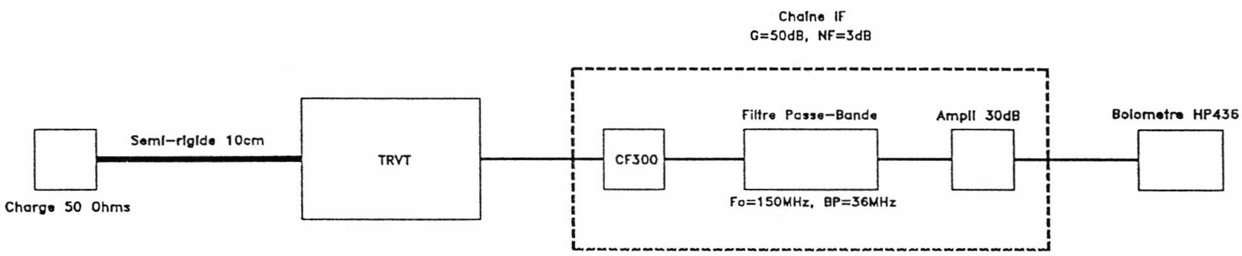



**Fig. 6 : Positionnement des stubs**

A l'aide du banc de mesure de bruit HP8970A + HP346B, je trouve un facteur de bruit de 1.65dB et un gain de 20dB. J'ai également réalisé deux autre types de mesure afin de vérifier ce résultat.

Mesure de gain : A l'aide d'un synthétiseur j'injecte une porteuse de fréquence 5760MHz et d'amplitude -40dBm, j'obtiens sur la FI un niveau de -22dBm. Le gain est donc de 18dB.

Mesure de bruit : Je réalisé le montage suivant




En refroidissant la charge 50 Ohms à 77K, grâce à de l'azote liquide, et en mesurant la différence de niveau sur le bolomètre, je trouve une température de bruit de 150K soit 1.78dB.

**Thot=296K Tcold=77K**
**Phot=481nW Pcold=245nW**
**Y=Phot/Pcold Y=1.963**

**Tsys=(Thot-(Y*Tcold))/(Y-1)**
**Tsys=150K**

**NF(dB)=10*log(1+(Tsys/296))**
**NF(dB)=1.78dB**

Cette mesure vérifie le résultat précèdent. Celle-ci étant réalisée sur 36MHz de bande passante, il est donc normal de trouver un résultat légèrement supérieur à la mesure réalisée sur le banc de mesure de bruit.

Le montage parait être sain, il ne semble pas avoir tendance à osciller et ceci quelque soit l'impédance d'entrée.

Pour le confort d'écoute, il est possible de placer un atténuateur de 10dB à l'entrée du récepteur 2m afin de ne pas avoir une déviation du S-metre au repos. Le facteur de bruit n'est augmenté que de 0.1 à 0.2dB.

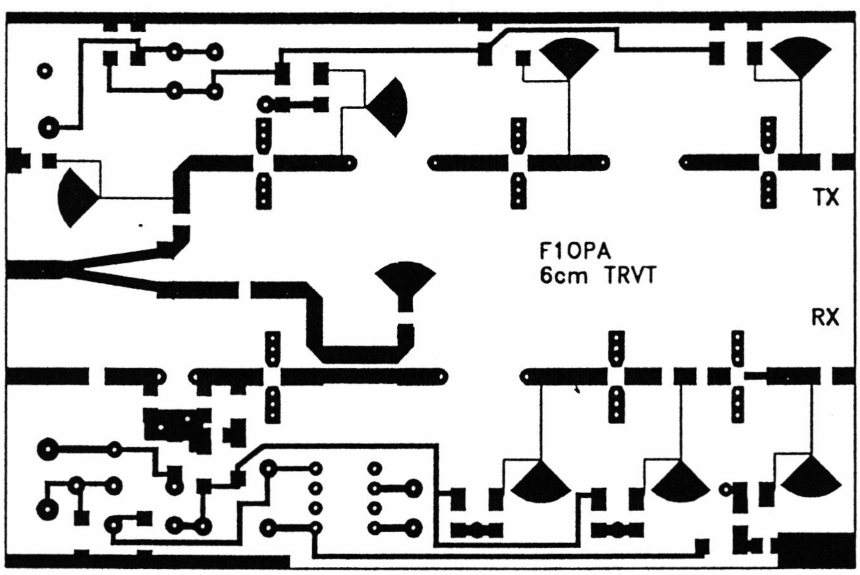
## CONCLUSIONS

Pour résumer on obtient : Pout (1dB) = 15dBm
NF < 1.7dB, G = 18dB

Le dessin du circuit imprimé a été réalisé sur AutoCad Light 2.0, si des personnes sont intéressées je peux fournir une copie du fichier.

J'espère que cette réalisation permettra d'augmenter le nombre d'OM QRV sur cette bande.

Un oscillateur est en cours de réalisation.



HEWLETT® PACKARD

# 2-18 GHz Ultra Low Noise Pseudomorphic HEMT

## Technical Data

ATF-36077

### Features
- PHEMT Technology
- Ultra-Low Noise Figure:
  0.5 dB Typical at 12 GHz
  0.3 dB Typical at 4 GHz
- High Associated Gain:
  12 dB Typical at 12 GHz
  17 dB Typical at 4 GHz
- Cost Effective Low Parasitic Ceramic Microstrip Package
- Tape-and-Reel Packing Option Available

### Applications
- 12 GHz DBS LNB (Low Noise Block)
- 4 GHz TVRO LNB (Low Noise Block)
- Ultra-Sensitive Low Noise Amplifiers

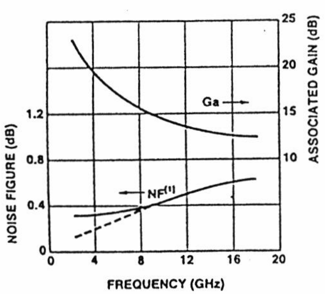
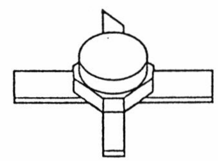


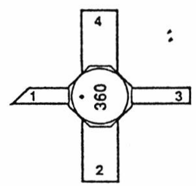

Figure 1. ATF-36077 Optimum Noise Figure and Associated Gain vs. Frequency for $V_{DS}$ = 1.5 V, $I_D$ = 10 mA.

Note: 1. See Noise Parameter Table.

### 77 Package



### Pin Configuration




### Pin Description
1 Gate
2 Source
3 Drain
4 Source

### Description
Hewlett-Packard's ATF-36077 is an ultra-low-noise Pseudomorphic High Electron Mobility Transistor (PHEMT), packaged in a cost effective, low parasitic, surface-mountable ceramic package. Properly matched, this transistor will provide typical 12 GHz noise figures of 0.5 dB, or typical 4 GHz noise figures of 0.3 dB. Additionally, the ATF-36077 has very low noise resistance, reducing the sensitivity of noise performance to variations in input impedance match, making the design of broadband low noise amplifiers much easier. The premium sensitivity of the ATF-36077 makes this device the ideal choice for use in the first stage of extremely low noise cascades. The repeatable performance and consistency make it appropriate for use in Ku-band Direct Broadcast Satellite (DBS) Television systems, C-band Television Receive Only (TVRO) LNAs, or other low noise amplifiers operating in the 2-18 GHz frequency range.

This GaAs PHEMT device has a nominal 0.2 micron gate length with a total gate periphery (width) of 200 microns. Proven gold based metalization systems and nitride passivation assure rugged, reliable devices.

### ATF-36077 Absolute Maximum Ratings

| Symbol | Parameter | Units | Absolute Maximum[1] |
|---|---|---|---|
| $V_{DS}$ | Drain – Source Voltage | V | +3 |
| $V_{GS}$ | Gate – Source Voltage | V | -3 |
| $I_D$ | Drain Current | mA | $I_{dss}$ |
| $P_T$ | Total Power Dissipation[3] | mW | 180 |
| $P_{in\ max}$ | RF Input Power | dBm | +10 |
| $T_{ch}$ | Channel Temperature | °C | 150 |
| $T_{STG}$ | Storage Temperature | °C | -65 to 150 |

Thermal Resistance[2,3]: $\Theta_{ch\text{-}c}$ = 60°C/W

Notes:
1. Operation of this device above any one of these parameters may cause permanent damage.
2. Measured at $P_{diss}$ = 15 mW and $T_{ch}$ = 100°C.
3. Derate at 16.7 mW/°C for $T_C$ > 139°C.

### ATF-36077 Electrical Specifications, $T_C$ = 25°C, $Z_O$ = 50 Ω
$V_{ds}$ = 1.5 V, $I_{ds}$ = 10 mA, (unless otherwise noted).

| Symbol | Parameters and Test Conditions | | Units | Min. | Typ. | Max. |
|---|---|---|---|---|---|---|
| NF | Noise Figure[1] | f = 12.0 GHz | dB | | 0.5 | 0.6 |
| $G_a$ | Gain at NF[1] | f = 12.0 GHz | dB | 11.0 | 12.0 | |
| $g_m$ | Transconductance | $V_{DS}$ = 1.5 V, $V_{GS}$ = 0 V | mS | 50 | 55 | |
| $I_{dss}$ | Saturated Drain Current | $V_{DS}$ = 1.5 V, $V_{GS}$ = 0 V | mA | 15 | 25 | 60 |
| $V_{p\ 10\%}$ | Pinchoff Voltage | $V_{DS}$ = 1.5 V, $I_{DS}$ = 10% of $I_{dss}$ | V | -1.0 | -0.35 | -0.15 |

Note:
1. Measured in a fixed tuned environment with Γ source = 0.54 at 156°; Γ load = 0.48 at 167°.

### ATF-36077 Characterization Information, $T_C$ = 25°C, $Z_O$ = 50 Ω
$V_{ds}$ = 1.5 V, $I_{ds}$ = 10 mA, (unless otherwise noted).

| Symbol | Parameters and Test Conditions | | Units | Typ. |
|---|---|---|---|---|
| NF | Noise Figure (Tuned Circuit) | f = 4 GHz<br>f = 12 GHz | dB<br>dB | 0.3[2]<br>0.5 |
| $G_a$ | Gain at Noise Figure (Tuned Circuit) | f = 4 GHz<br>f = 12 GHz | dB<br>dB | 17<br>12 |
| $S_{12\ off}$ | Reverse Isolation | f = 12 GHz, $V_{DS}$ = 1.5 V, $V_{GS}$ = -2 V | dB | 14 |
| $P_{1dB}$ | Output Power at 1 dB Gain Compression | f = 4 GHz<br>f = 12 GHz | dBm<br>dBm | 5<br>5 |
| $V_{GS\ 10\ mA}$ | Gate to Source Voltage for $I_{DS}$ = 10 mA | $V_{DS}$ = 1.5 V | V | -0.2 |

Note:
2. See noise parameter table.

## ATF-36077 Typical Scattering Parameters, Common Source, $Z_O$ = 50 Ω, $V_{DS}$ = 1.5 V, $I_D$ = 10 mA

| Freq. GHz | S11 Mag. | S11 Ang. | S21 dB | S21 Mag. | S21 Ang. | S12 dB | S12 Mag. | S12 Ang. | S22 Mag. | S22 Ang. |
|---|---|---|---|---|---|---|---|---|---|---|
| 1.0 | 0.99 | -17 | 14.00 | 5.010 | 163 | -36.08 | 0.016 | 78 | 0.60 | -14 |
| 2.0 | 0.97 | -33 | 13.81 | 4.904 | 147 | -30.33 | 0.030 | 66 | 0.59 | -28 |
| 3.0 | 0.94 | -49 | 13.53 | 4.745 | 132 | -27.25 | 0.043 | 54 | 0.57 | -41 |
| 4.0 | 0.90 | -65 | 13.17 | 4.556 | 116 | -25.32 | 0.054 | 43 | 0.55 | -54 |
| 5.0 | 0.86 | -79 | 12.78 | 4.357 | 102 | -24.04 | 0.063 | 33 | 0.53 | -66 |
| 6.0 | 0.82 | -93 | 12.39 | 4.162 | 88 | -23.17 | 0.069 | 24 | 0.50 | -78 |
| 7.0 | 0.78 | -107 | 12.00 | 3.981 | 75 | -22.58 | 0.074 | 16 | 0.48 | -89 |
| 8.0 | 0.75 | -120 | 11.64 | 3.820 | 62 | -22.17 | 0.078 | 8 | 0.46 | -99 |
| 9.0 | 0.72 | -133 | 11.32 | 3.682 | 49 | -21.90 | 0.080 | 1 | 0.44 | -109 |
| 10.0 | 0.69 | -146 | 11.04 | 3.566 | 37 | -21.71 | 0.082 | -6 | 0.42 | -119 |
| 11.0 | 0.66 | -159 | 10.81 | 3.473 | 25 | -21.57 | 0.083 | -13 | 0.40 | -129 |
| 12.0 | 0.63 | -172 | 10.63 | 3.401 | 13 | -21.44 | 0.085 | -19 | 0.38 | -139 |
| 13.0 | 0.61 | 175 | 10.50 | 3.349 | 1 | -21.32 | 0.086 | -25 | 0.37 | -149 |
| 14.0 | 0.60 | 161 | 10.41 | 3.315 | -12 | -21.19 | 0.087 | -32 | 0.35 | -160 |
| 15.0 | 0.58 | 147 | 10.36 | 3.296 | -24 | -21.04 | 0.089 | -39 | 0.33 | -171 |
| 16.0 | 0.57 | 131 | 10.34 | 3.289 | -37 | -20.87 | 0.091 | -47 | 0.31 | 177 |
| 17.0 | 0.56 | 114 | 10.34 | 3.289 | -50 | -20.69 | 0.092 | -55 | 0.29 | 164 |
| 18.0 | 0.57 | 97 | 10.35 | 3.291 | -64 | -20.53 | 0.094 | -65 | 0.26 | 148 |

## ATF-36077 Typical "Off" Scattering Parameters, Common Source, $Z_O$ = 50 Ω, $V_{DS}$ = 1.5 V, $I_D$ = 0 mA, $V_{GS}$ = -2 V

| Freq. GHz | S11 Mag. | S11 Ang. | S21 dB | S21 Mag. | S21 Ang. | S21 dB | S21 Mag. | S21 Ang. | S22 Mag. | S22 Ang. |
|---|---|---|---|---|---|---|---|---|---|---|
| 11.0 | 0.96 | -139 | -14.2 | 0.19 | -43 | -14.2 | 0.19 | -43 | 0.97 | -125 |
| 12.0 | 0.95 | -152 | -14.0 | 0.20 | -56 | -14.0 | 0.20 | -56 | 0.97 | -137 |
| 13.0 | 0.94 | -166 | -13.8 | 0.20 | -69 | -13.8 | 0.20 | -68 | 0.96 | -149 |

## ATF-36077 Typical Noise Parameters, Common Source, $Z_O$ = 50 Ω, $V_{DS}$ = 1.5 V, $I_D$ = 10 mA

| Freq. GHz | $F_{min}$[1] dB | $\Gamma_{opt}$ Mag. | $\Gamma_{opt}$ Ang. | $R_n/Z_o$ |
|---|---|---|---|---|
| 2 | 0.30 | 0.90 | 25 | 0.20 |
| 4 | 0.30 | 0.81 | 51 | 0.17 |
| 6 | 0.30 | 0.73 | 76 | 0.13 |
| 8 | 0.37 | 0.66 | 102 | 0.09 |
| 10 | 0.44 | 0.60 | 129 | 0.05 |
| 12 | 0.50 | 0.54 | 156 | 0.03 |
| 14 | 0.56 | 0.48 | -174 | 0.02 |
| 16 | 0.61 | 0.43 | -139 | 0.05 |
| 18 | 0.65 | 0.39 | -100 | 0.09 |

Note:
1. The $F_{min}$ values at 2,4, and 6 GHz have been adjusted to reflect expected circuit losses that will be encountered when matching to the optimum reflection coefficient ($\Gamma_{opt}$) at these frequencies. The theoretical $F_{min}$ values for these frequencies are: 0.10 dB at 2 GHz, 0.20 dB at 4 GHz, and 0.29 dB at 6 GHz. Noise parameters are derived from associated s parameters, packaged device measurements at 12 GHz, and die level measurements from 6 to 18 GHz.

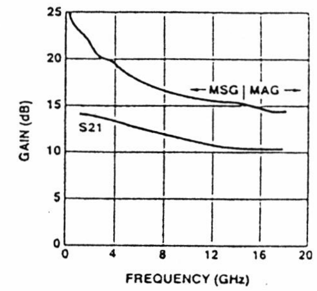
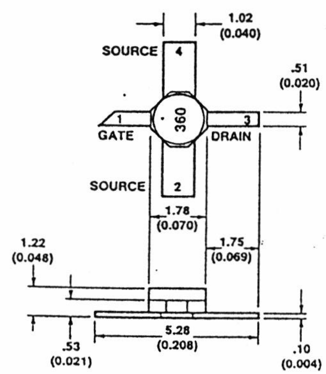



Figure 2. Maximum Available Gain, Maximum Stable Gain and Insertion Power Gain vs. Frequency. $V_{DS}$ = 1.5 V, $I_D$ = 10 mA.

## Package Dimensions




TYPICAL DIMENSIONS ARE IN MILLIMETERS (INCHES).

## Part Number Ordering Information

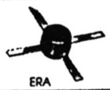
| Part Number | No. of Devices | Container |
|---|---|---|
| ATF-36077-TR1[2] | 1000 | 7" Reel |
| ATF-36077-STR | 1 | strip |

Note:
2. For more information, see "Tape and Reel Packaging for Semiconductor Devices," in "Communications Components" Designer's Catalog. The 77 package uses to the same tape cavity as does the 76 package.

# Mini-Circuits®

## SPECIFICATIONS:

at 25°C



ERA

ERA-SM

| MODEL NO. | FREQ. GHz $f_L$–$f_u$ | GAIN, dB Typical over frequency, GHz 0.1 | 1 | 2 | 3 | 4 | 6 | 8 | Min. @ 2 GHz | Flatness DC-2 GHz | MAXIMUM POWER, dBm at 2 GHz Output (1 dB Comp.) Typ. | Min. | Input (no dmg.) | DYNAMIC RANGE at 2 GHz NF dB Typ. | IP3 dBm Typ. | VSWR Typ. (1) In DC-3 GHz | In 3-$f_u$ GHz | Out DC-3 GHz | Out 3-$f_u$ GHz | MAXIMUM RATING I mA | P mW | DC POWER at Pin 3 Current (mA) | Volt. Typ. | THERMAL RESISTANCE θjc typ. °C/W | Case Style Note A | CONN. | Price $ Qty. (30) |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| ERA-1 | DC-8 | 12.2 | 12.1 | 11.8 | 11.5 | 11.3 | 11.0 | 10.2 | 9 | ±0.3 | 11.7 | 9.7 | 15 | 5.3 | 26 | 1.6 | 1.8 | 1.5 | 1.9 | 75 | 330 | 40 | 3.6 | 455 | VV105 | cb | 1.37 |
| ERA-2 | DC-6 | 16.2 | 16.0 | 15.6 | 15.1 | 14.6 | 14.0 | — | 12 | ±0.3 | 12.8 | 11 | 15 | 4.7 | 26 | 1.4 | 1.4 | 1.4 | 1.6 | 75 | 330 | 40 | 3.6 | 455 | VV105 | cb | 1.52 |
| ERA-3 | DC-3 | 22.9 | 22.2 | 20.8 | 19.2 | — | — | — | 17 | ±1.1 | 12.1 | 9 | 13 | 3.8 | 23 | 1.7 | — | 1.7 | — | 75 | 330 | 35 | 3.5 | 432 | VV105 | cb | 1.67 |
| ERA-4 | DC-4 | 13.8 | 13.7 | 13.5 | 13.3 | 13.0 | — | — | 11 | ±0.2 | 17.0 | 15 | 20 | 5.5 | 32.5 | 1.6 | 1.6 | 1.4 | 1.6 | 120 | 650 | 65 | 5.0 | 278 | VV105 | cb | 3.85 |
| ERA-5 | DC-4 | 20.2 | 19.8 | 18.8 | 17.7 | 16.4 | — | — | 16 | ±0.75 | 18.4 | 16.5 | 13 | 4.5 | 33 | 1.2 | 1.2 | 1.3 | 1.5 | 120 | 650 | 65 | 4.9 | 278 | VV105 | cb | 3.85 |
| ERA-6 | DC-4 | 11.1 | 11.1 | 11.3 | 11.5 | 11.3 | — | — | 10 | ±0.2 | 18.5 | 16.5 | 20 | 8.4 | 36.5 | 1.3 | 1.3 | 1.6 | 1.6 | 120 | 850 | 70 | 5.5 | 220 | VV105 | cb | 3.85 |
| ERA-1SM | DC-8 | 12.3 | 12.1 | 11.8 | 11.2 | 10.8 | 10.4 | 9.2 | 9 | ±0.3 | 11.3 | 9.3 | 15 | 5.5 | 26 | 1.6 | 1.9 | 1.5 | 1.9 | 75 | 330 | 40 | 3.6 | 460 | WW107 | cb | 1.42 |
| ERA-2SM | DC-6 | 16.2 | 15.8 | 15.2 | 14.4 | 13.6 | 13.0 | — | 12 | ±0.5 | 12.4 | 10.5 | 15 | 4.6 | 26 | 1.5 | 1.6 | 1.5 | 1.7 | 75 | 330 | 40 | 3.6 | 460 | WW107 | cb | 1.57 |
| ERA-3SM | DC-3 | 22.8 | 21.8 | 20.2 | 18.4 | — | — | — | 16 | ±1.3 | 11.5 | 9 | 13 | 3.8 | 23 | 1.5 | -- | 1.5 | — | 75 | 330 | 35 | 3.5 | 437 | WW107 | cb | 1.72 |
| ERA-4SM | DC-4 | 14.0 | 13.8 | 13.5 | 13.2 | 12.7 | — | — | 11 | ±0.3 | 16.8 | 15 | 20 | 5.2 | 33 | 1.6 | 1.6 | 1.3 | 1.5 | 120 | 650 | 65 | 5.0 | 283 | WW107 | cb | 3.90 |
| ERA-5SM | DC-4 | 20.2 | 19.5 | 18.5 | 17.3 | 16.2 | — | — | 16 | ±1.0 | 18.4 | 16.5 | 13 | 4.3 | 32.5 | 1.3 | 1.3 | 1.2 | 1.3 | 120 | 650 | 65 | 4.9 | 283 | WW107 | cb | 3.90 |
| ERA-6SM | DC-4 | 11.2 | 11.2 | 11.3 | 11.4 | 11.2 | — | — | 10 | ±0.2 | 17.9 | 16 | 20 | 8.4 | 36 | 1.3 | 1.3 | 1.6 | 1.5 | 120 | 850 | 70 | 5.5 | 225 | WW107 | cb | 3.90 |

NOTES:
- * at 1 GHz for ERA 4, 5, 6, 4SM, 5SM, 6SM.
- ** $f_u$ is the upper frequency limit for each model as shown in the table.
- o Low frequency cutoff determined by external coupling capacitors.

A. Units are non-hermetic unless otherwise noted. For details on case dimensions & finishes see "Case Styles & Outline Dimensions".
B. Prices and Specifications subject to change without notice.
1. Model number designated by alphanumeric code marking.
2. ERA-SM models available on tape and reel.

**absolute maximum ratings**
operating temperature: -45°C to 85°C
storage temperature: -65° to 150°C
device voltage:
3.0V min., 4.1V max. for ERA 1,2,3
4.2V min., 5.5V max. for ERA 4,5
5.0V min., 6.0V max. for ERA-6

### DESIGNERS KITS AVAILABLE

| Kit No. | Model Type | No. of Units in Kit | Description | Price $ per kit |
|---|---|---|---|---|
| K1-ERA | ERA | 30 | 10 of each 1,2,3 | 49.95 |
| K2-ERA | ERA | 20 | 10 of each 4,5 | 69.95 |
| K1-ERASM | ERA-SM | 30 | 10 of each 1SM, 2SM,3SM | 49.95 |
| K2-ERASM | ERA-SM | 20 | 10 of each 4SM, 5SM | 69.95 |
| K3-ERASM | ERA-SM | 30 | 10 of each 4SM, 5SM, 6SM | 99.95 |

### CASE STYLES/ outline dimensions (inch/mm)

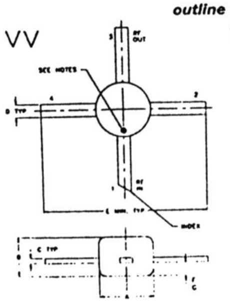
| case no. | A | B | C | D | E | F | G | H | wt. grams. | NOTES* |
|---|---|---|---|---|---|---|---|---|---|---|
| VV105 | .085<br>2.16 | .060<br>1.52 | .008<br>.20 | .020<br>.51 | .256<br>6.50 | .012<br>.30 | .025<br>.64 | | .015 | A13, C4, E2 |
| WW107 | .020<br>.51 | .10<br>2.54 | .020<br>.51 | .092<br>2.34 | .085<br>2.16 | .060<br>1.52 | .008<br>.20 | .026<br>.66 | .015 | A13, C4, E2, F4 |

tolerance .x±.1 .xx±.03 .xxx±.015 inch oz. = grams x.0353

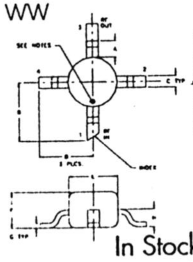
NOTES*:
A13. Case material: plastic. Lead finish: Tin or tin-lead plate.
C4. RF input lead (1) identified by diagonally cut lead; the cut may be 45°(ref) in either direction. It may also have an additional orientation mark. Model dash number identified by color dot or alphanumeric code on case.
E2. Lead width ±.010; lead thickness ±.005 inch.
F4. Tape and reel packaging available. See Tape and Reel Packaging Information for details. To order Tape & Reel version add -TR suffix to model.

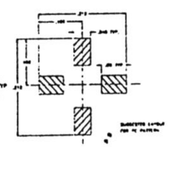
### outline drawings

VV



WW





**marking identification**

| Model | Alphanumeric Code |
|---|---|
| ERA-1, ERA-1SM | E1 |
| ERA-2, ERA-2SM | E2 |
| ERA-3, ERA-3SM | E3 |
| ERA-4, ERA-4SM | E4 |
| ERA-5, ERA-5SM | E5 |
| ERA-6, ERA-6SM | E6 |

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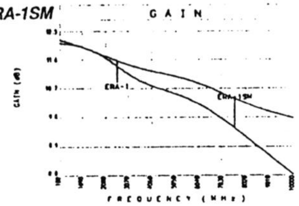
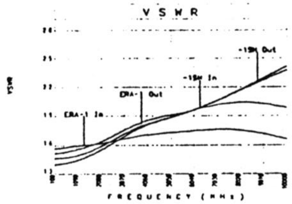
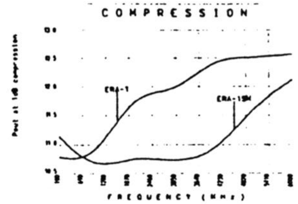
3

# S-PARAMETERS GAIN, GAIN COMPRESSION, RETURN LOSS, ISOLATION

## ERA-1SM

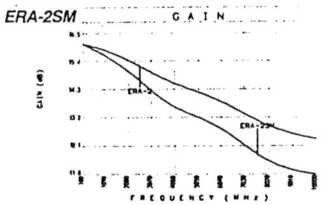


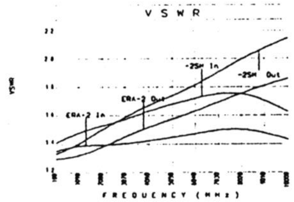



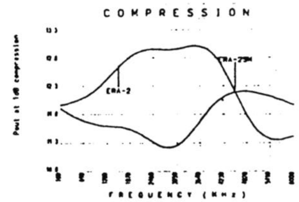




| Freq. MHz | $S_{11}$ (Input Return Loss) dB | Mag | Angle | $S_{21}$ (Power Gain) dB | Mag | Angle | $S_{12}$ (Isolation Out-In) dB | Mag | Angle | $S_{22}$ (Output Return Loss) dB | Mag | Angle | K |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 100 | -13.97 | 0.20 | 177.96 | 12.31 | 4.13 | 170.76 | -16.97 | 0.14 | -2.91 | -15.84 | 0.16 | 172.55 | 1.06 |
| 550 | -13.43 | 0.21 | 159.71 | 12.12 | 4.04 | 137.76 | -16.60 | 0.15 | -30.72 | -15.42 | 0.17 | 158.11 | 1.04 |
| 1000 | -12.92 | 0.23 | 144.17 | 11.97 | 3.97 | 104.55 | -16.66 | 0.15 | -55.46 | -14.62 | 0.19 | 142.15 | 1.04 |
| 2000 | -11.63 | 0.26 | 115.37 | 11.50 | 3.76 | 31.11 | -16.70 | 0.15 | -111.57 | -12.59 | 0.23 | 118.14 | 1.01 |
| 3000 | -10.30 | 0.31 | 94.66 | 10.94 | 3.52 | -42.64 | -16.75 | 0.15 | -168.73 | -10.90 | 0.29 | 102.41 | 0.98 |
| 4000 | -9.47 | 0.34 | 82.30 | 10.33 | 3.28 | -114.36 | -16.87 | 0.14 | 134.07 | -9.44 | 0.34 | 87.77 | 0.96 |
| 5000 | -9.17 | 0.35 | 74.43 | 9.53 | 3.00 | 175.55 | -17.16 | 0.14 | 75.93 | -8.35 | 0.38 | 69.43 | 0.98 |
| 6000 | -9.34 | 0.34 | 69.53 | 8.94 | 2.80 | 107.96 | -17.39 | 0.14 | 17.51 | -7.42 | 0.43 | 54.63 | 1.00 |
| 7220 | -10.05 | 0.31 | 63.58 | 9.04 | 2.83 | 26.04 | -18.12 | 0.12 | -55.87 | -6.81 | 0.46 | 33.66 | 1.05 |
| 8000 | -9.80 | 0.32 | 60.25 | 9.10 | 2.85 | -25.22 | -18.78 | 0.12 | -101.98 | -6.39 | 0.48 | 23.31 | 1.06 |

## ERA-2SM

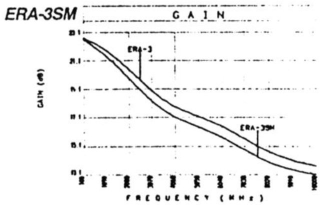


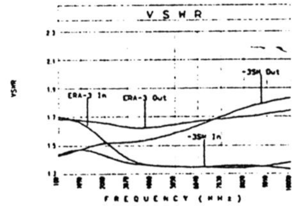



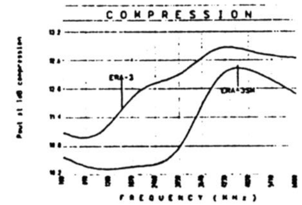




| Freq. MHz | $S_{11}$ (Input Return Loss) dB | Mag | Angle | $S_{21}$ (Power Gain) dB | Mag | Angle | $S_{12}$ (Isolation Out-In) dB | Mag | Angle | $S_{22}$ (Output Return Loss) dB | Mag | Angle | K |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 100 | -17.46 | 0.13 | 176.77 | 16.14 | 6.41 | 170.10 | -19.36 | 0.11 | -2.02 | -19.03 | 0.11 | 174.06 | 1.03 |
| 550 | -16.15 | 0.16 | 152.26 | 15.95 | 6.27 | 135.76 | -19.03 | 0.11 | -28.55 | -18.19 | 0.12 | 154.95 | 1.02 |
| 1000 | -15.45 | 0.17 | 132.82 | 15.69 | 6.09 | 100.80 | -19.05 | 0.11 | -52.12 | -17.07 | 0.14 | 138.34 | 1.02 |
| 2000 | -13.66 | 0.21 | 99.88 | 14.90 | 5.56 | 23.41 | -18.33 | 0.12 | -105.24 | -14.85 | 0.18 | 115.14 | 0.99 |
| 3000 | -12.34 | 0.24 | 79.89 | 13.99 | 5.01 | -50.95 | -18.64 | 0.12 | -160.52 | -13.10 | 0.22 | 102.29 | 1.00 |
| 4000 | -11.85 | 0.26 | 71.57 | 13.03 | 4.48 | -123.74 | -18.52 | 0.12 | 144.60 | -11.53 | 0.27 | 89.80 | 1.02 |
| 5000 | -11.74 | 0.26 | 69.51 | 11.86 | 3.92 | 166.38 | -18.68 | 0.12 | 87.38 | -10.34 | 0.30 | 71.96 | 1.08 |
| 6000 | -12.16 | 0.25 | 69.51 | 11.11 | 3.59 | 98.14 | -18.67 | 0.12 | 29.47 | -9.27 | 0.34 | 58.13 | 1.11 |
| 7220 | -13.03 | 0.22 | 76.69 | 11.07 | 3.58 | 15.13 | -19.28 | 0.11 | -41.09 | -8.96 | 0.36 | 36.99 | 1.18 |
| 8000 | -11.73 | 0.26 | 79.33 | 10.97 | 3.54 | -36.97 | -19.93 | 0.10 | -85.53 | -8.48 | 0.38 | 27.38 | 1.20 |

## ERA-3SM




| Freq. MHz | $S_{11}$ (Input Return Loss) dB | Mag | Angle | $S_{21}$ (Power Gain) dB | Mag | Angle | $S_{12}$ (Isolation Out-In) dB | Mag | Angle | $S_{22}$ (Output Return Loss) dB | Mag | Angle | K |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 100 | -13.57 | 0.21 | 174.98 | 22.57 | 13.44 | 168.49 | -24.83 | 0.06 | -6.88 | -13.54 | 0.21 | 172.92 | 0.93 |
| 550 | -13.16 | 0.22 | 143.84 | 22.13 | 12.78 | 128.25 | -24.15 | 0.06 | -20.53 | -13.45 | 0.21 | 148.84 | 0.92 |
| 1000 | -13.21 | 0.22 | 119.72 | 21.46 | 11.83 | 88.76 | -23.91 | 0.06 | -39.52 | -13.20 | 0.22 | 127.78 | 0.93 |
| 2000 | -13.91 | 0.20 | 80.56 | 19.44 | 9.38 | 5.82 | -22.50 | 0.07 | -88.03 | -12.83 | 0.23 | 99.76 | 0.95 |
| 3000 | -14.90 | 0.18 | 63.12 | 17.48 | 7.48 | -70.08 | -21.40 | 0.09 | -140.31 | -12.59 | 0.23 | 86.91 | 0.99 |
| 4000 | -15.99 | 0.16 | 66.26 | 15.69 | 6.09 | -142.98 | -20.38 | 0.10 | 164.98 | -11.83 | 0.26 | 78.08 | 1.03 |
| 5000 | -15.68 | 0.16 | 78.95 | 14.00 | 5.01 | 147.59 | -20.07 | 0.10 | 108.25 | -11.05 | 0.28 | 62.38 | 1.10 |
| 6000 | -15.06 | 0.18 | 89.08 | 12.77 | 4.35 | 81.29 | -19.70 | 0.10 | 50.92 | -9.95 | 0.32 | 52.96 | 1.14 |
| 7220 | -12.95 | 0.23 | 106.02 | 12.45 | 4.19 | -2.14 | -19.68 | 0.10 | -17.83 | -10.09 | 0.31 | 36.16 | 1.13 |
| 8000 | -10.12 | 0.31 | 97.46 | 12.06 | 4.01 | -54.58 | -19.83 | 0.10 | -60.30 | -9.58 | 0.33 | 29.45 | 1.08 |

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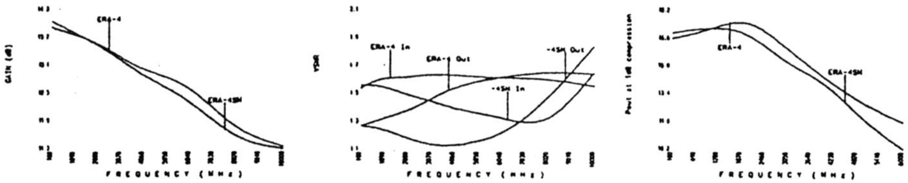
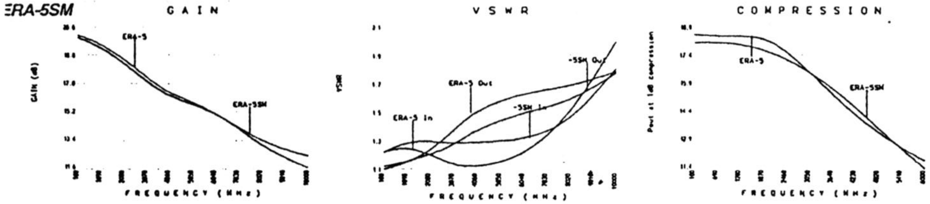
### ERA-4SM



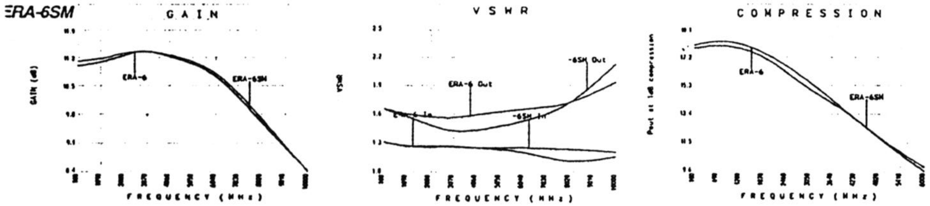

| Freq. MHz | $S_{11}$ (Input Return Loss) dB | Mag | Angle | $S_{21}$ (Power Gain) dB | Mag | Angle | $S_{12}$ (Isolation Out-In) dB | Mag | Angle | $S_{22}$ (Output Return Loss) dB | Mag | Angle | K |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 100 | -12.64 | 0.23 | 177.59 | 13.67 | 4.83 | 170.61 | -18.09 | 0.12 | -5.31 | -17.67 | 0.13 | 178.50 | 1.04 |
| 550 | -12.36 | 0.24 | 158.88 | 13.56 | 4.76 | 136.16 | -18.17 | 0.12 | -29.89 | -16.88 | 0.14 | 162.80 | 1.04 |
| 1000 | -12.10 | 0.25 | 143.21 | 13.47 | 4.72 | 101.77 | -18.16 | 0.12 | -54.80 | -16.29 | 0.15 | 149.98 | 1.04 |
| 2000 | -11.54 | 0.26 | 111.96 | 13.16 | 4.55 | 23.71 | -18.21 | 0.12 | -109.29 | -14.66 | 0.18 | 129.88 | 1.03 |
| 3000 | -11.21 | 0.28 | 88.25 | 12.85 | 4.39 | -53.45 | -18.30 | 0.12 | -165.76 | -13.27 | 0.22 | 116.80 | 1.03 |
| 4000 | -11.88 | 0.25 | 76.83 | 12.41 | 4.17 | -130.95 | -18.66 | 0.12 | 138.93 | -12.24 | 0.24 | 103.89 | 1.08 |
| 5000 | -13.04 | 0.22 | 79.31 | 11.63 | 3.82 | 152.70 | -19.18 | 0.11 | 81.11 | -12.25 | 0.24 | 86.11 | 1.22 |
| 6000 | -13.34 | 0.22 | 93.52 | 11.03 | 3.56 | 77.52 | -19.69 | 0.10 | 26.67 | -12.46 | 0.24 | 76.84 | 1.35 |
| 7220 | -10.27 | 0.31 | 108.89 | 10.63 | 3.40 | -17.86 | -20.58 | 0.09 | -35.88 | -13.29 | 0.22 | 80.82 | 1.45 |
| 8000 | -7.77 | 0.41 | 98.82 | 9.81 | 3.09 | -76.73 | -20.76 | 0.09 | -75.38 | -11.39 | 0.27 | 82.14 | 1.39 |

### ERA-5SM




| Freq. MHz | $S_{11}$ (Input Return Loss) dB | Mag | Angle | $S_{21}$ (Power Gain) dB | Mag | Angle | $S_{12}$ (Isolation Out-In) dB | Mag | Angle | $S_{22}$ (Output Return Loss) dB | Mag | Angle | K |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 100 | -14.91 | 0.18 | 172.80 | 20.08 | 10.09 | 169.05 | -23.72 | 0.07 | 1.81 | -17.76 | 0.13 | 173.95 | 1.03 |
| 550 | -14.61 | 0.19 | 149.79 | 19.80 | 9.77 | 130.66 | -22.47 | 0.08 | -25.08 | -17.80 | 0.13 | 158.90 | 0.99 |
| 1000 | -14.54 | 0.19 | 127.39 | 19.43 | 9.36 | 92.07 | -22.47 | 0.08 | -46.48 | -17.18 | 0.14 | 144.80 | 1.00 |
| 2000 | -15.20 | 0.17 | 88.80 | 18.21 | 8.14 | 7.80 | -21.96 | 0.08 | -96.71 | -15.98 | 0.16 | 126.69 | 1.02 |
| 3000 | -16.90 | 0.14 | 72.32 | 16.81 | 6.93 | -72.65 | -21.39 | 0.09 | -148.00 | -14.68 | 0.18 | 119.64 | 1.07 |
| 4000 | -18.41 | 0.12 | 90.96 | 15.31 | 5.83 | -150.97 | -20.92 | 0.09 | 160.47 | -13.07 | 0.22 | 111.07 | 1.13 |
| 5000 | -15.79 | 0.16 | 113.66 | 13.69 | 4.84 | 134.42 | -20.88 | 0.09 | 106.02 | -12.89 | 0.23 | 96.01 | 1.24 |
| 6000 | -11.86 | 0.26 | 120.29 | 12.47 | 4.20 | 61.41 | -20.42 | 0.10 | 52.41 | -11.48 | 0.27 | 89.96 | 1.22 |
| 7220 | -7.68 | 0.41 | 114.34 | 11.43 | 3.73 | -31.64 | -19.67 | 0.10 | -10.06 | -10.13 | 0.31 | 87.00 | 1.03 |
| 8000 | -5.99 | 0.50 | 98.67 | 10.30 | 3.27 | -87.65 | -19.50 | 0.11 | -53.25 | -8.56 | 0.37 | 79.20 | 0.91 |

### ERA-6SM




| Freq. MHz | $S_{11}$ (Input Return Loss) dB | Mag | Angle | $S_{21}$ (Power Gain) dB | Mag | Angle | $S_{12}$ (Isolation Out-In) dB | Mag | Angle | $S_{22}$ (Output Return Loss) dB | Mag | Angle | K |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 100 | -16.75 | 0.15 | -2.94 | 11.09 | 3.59 | 170.58 | -17.19 | 0.14 | -4.79 | -11.83 | 0.26 | -1.41 | 1.13 |
| 550 | -16.91 | 0.14 | -0.03 | 11.01 | 3.55 | 135.73 | -16.54 | 0.15 | -32.00 | -12.06 | 0.25 | -14.25 | 1.10 |
| 1000 | -17.32 | 0.14 | 2.75 | 11.04 | 3.56 | 100.28 | -16.63 | 0.15 | -57.70 | -12.78 | 0.23 | -25.84 | 1.12 |
| 2000 | -18.03 | 0.13 | 12.27 | 11.04 | 3.56 | 19.44 | -16.80 | 0.14 | -116.46 | -15.24 | 0.17 | -62.30 | 1.16 |
| 3000 | -17.99 | 0.13 | 26.56 | 10.92 | 3.52 | -62.65 | -17.17 | 0.14 | -175.73 | -16.71 | 0.15 | -125.67 | 1.21 |
| 4000 | -18.20 | 0.12 | 43.82 | 10.39 | 3.31 | -145.69 | -17.69 | 0.13 | 126.08 | -14.15 | 0.20 | 173.19 | 1.29 |
| 5000 | -17.56 | 0.13 | 64.89 | 9.28 | 2.91 | 132.93 | -18.59 | 0.12 | 66.95 | -12.24 | 0.24 | 137.21 | 1.49 |
| 6000 | -16.31 | 0.15 | 80.31 | 8.14 | 2.55 | 53.42 | -19.44 | 0.11 | 9.86 | -9.70 | 0.33 | 111.65 | 1.69 |
| 7220 | -13.36 | 0.21 | 94.70 | 6.60 | 2.14 | -44.65 | -20.58 | 0.09 | -56.02 | -7.21 | 0.44 | 90.13 | 1.94 |
| 8000 | -10.91 | 0.28 | 86.50 | 5.10 | 1.80 | -102.76 | -21.17 | 0.09 | -99.65 | -5.94 | 0.50 | 75.87 | 2.11 |

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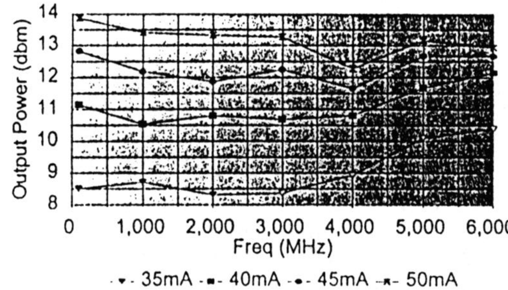
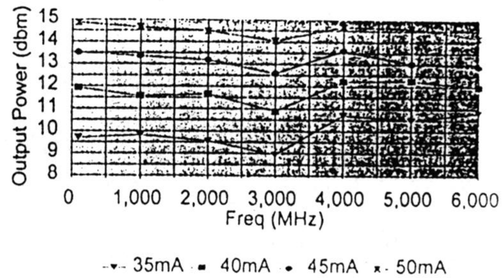
## OUTPUT POWER (1dB COMPRESSION) VS. CURRENT

Figure 6

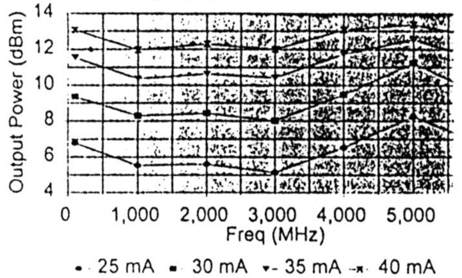
**ERA-1SM 1dB Compression**
At 35,40,45,50 mA Bias Currents



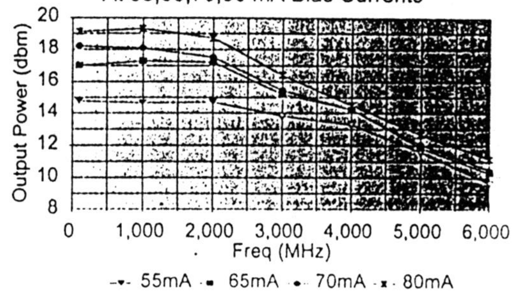

**ERA-2SM 1dB Compression**
At 35,40,45,50 mA Bias Currents



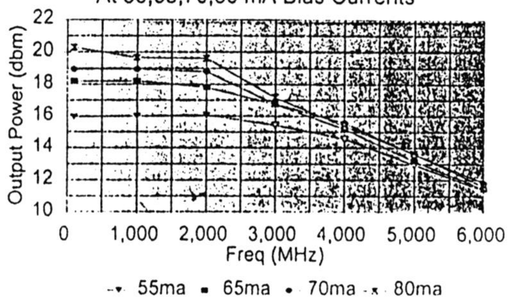

**ERA-3SM 1dB Compression**
At 25,30,35,40 mA Bias Currents



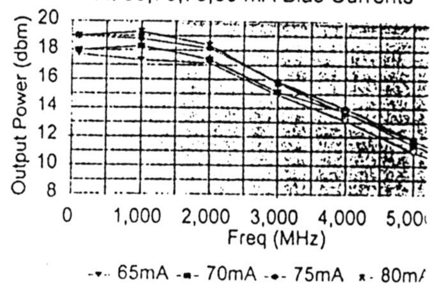

**ERA-4SM 1dB Compression**
At 55,65,70,80 mA Bias Currents




**ERA-5SM 1dB Compression**
At 55,65,70,80 mA Bias Currents




**ERA-6SM 1dB Compression**
At 65,70,75,80 mA Bias Currents



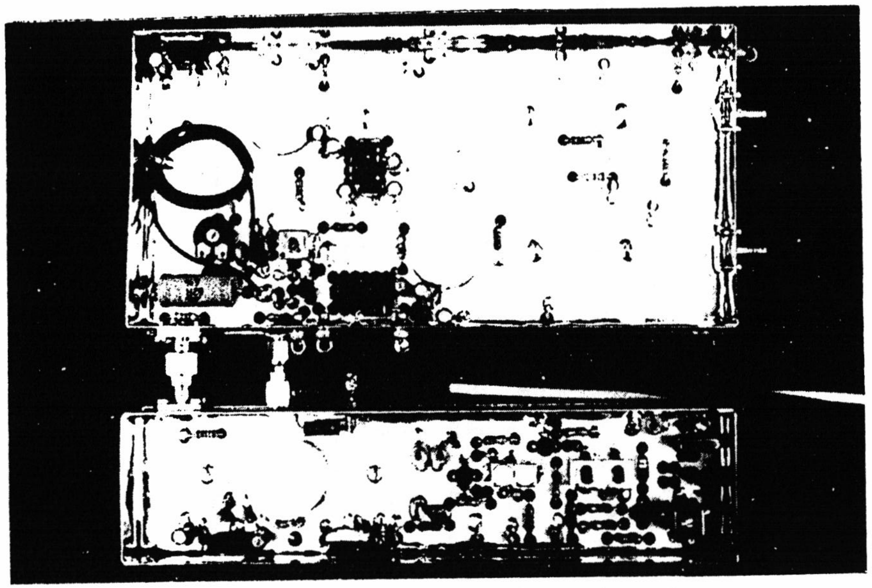
# Transverter for 5.7 GHz

Michael Kuhne, DB6NT
Birkenweg 15, W-8674 Naila 2

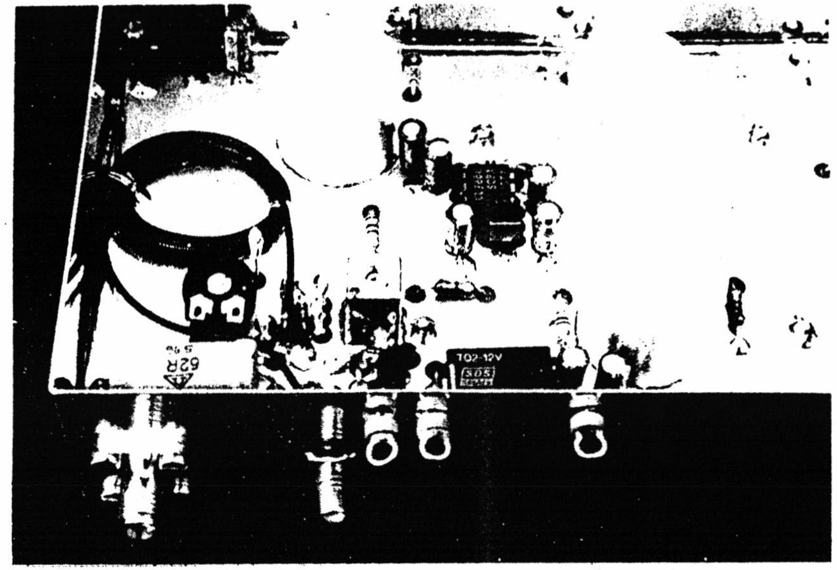
DUBUS 3/1991

(Part I)

**Abstract:** This transverter 5.760GHZ/0.144GHZ is made on one single board with exception of the LO. The LO produces 10 mW on 5.616 GHz and will described in Part II. The transverter has a very clean 200 mW output. The input power on 144 MHz should be 0.5 - 3 W. The noise figure of the RX is less than 2 dB and it has typically 23 dB gain. The transverter has been developed very closely according to the design of the 10 GHz transverter described before.



Bild/Figure 1: 5.7 GHz Transverter



Bild/Figure 2: 5.7 GHz Transverter (Close View)

## 1. Design

The goal of this development was to develop a simple single board 5.7 GHZ transverter with an external oscillator on 5.616 GHz. Without need for tuning a noise figure of less than 2 dB can be achieved for the RX and a clean output power of 200 mW on 5.760Ghz.

A wilkinson type hybrid divides the external LO-power on 5.616GHz for the RX and TX mixers respectively. These are simple single balanced active mixers with GaAs-FET triodes and gate injection.

The TX signal is filtered by a resonator and further amplified in a second stage. In front of the RX-mixer there are two low noise stages with a MGF1303 and a MGF1302 allowing for some 20 DB of RF-gain. The IF is switched by a silicon diode switch from receive to transmit. This is activated by a DC-voltage present on the IF-line (delivered by a IC202 for example).

## 5,7GHz TRANSVERTER — DB 6 NT

6.91

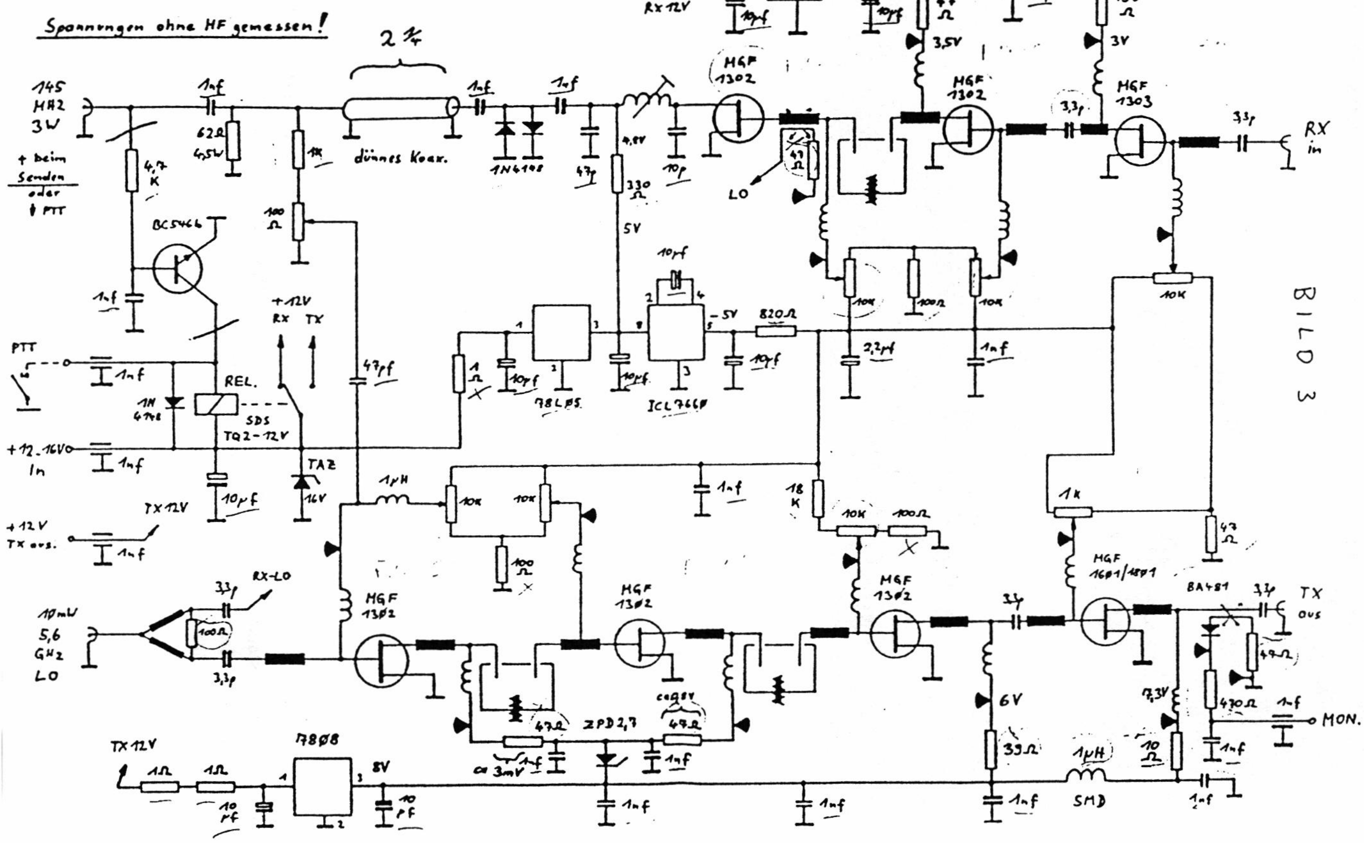


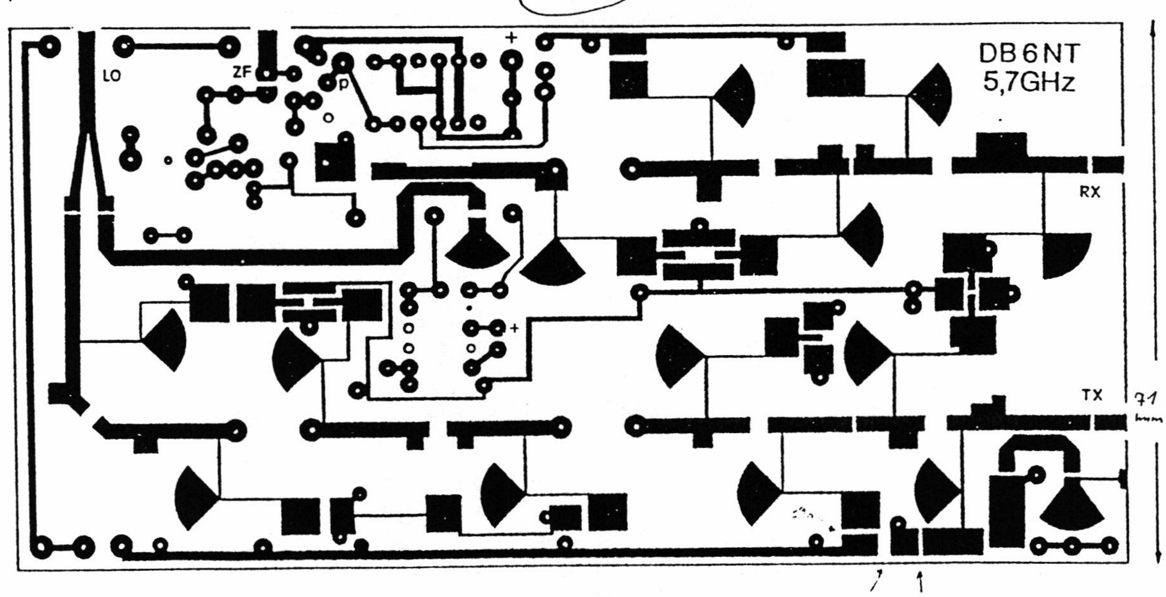

BILD 3

0,79Ø

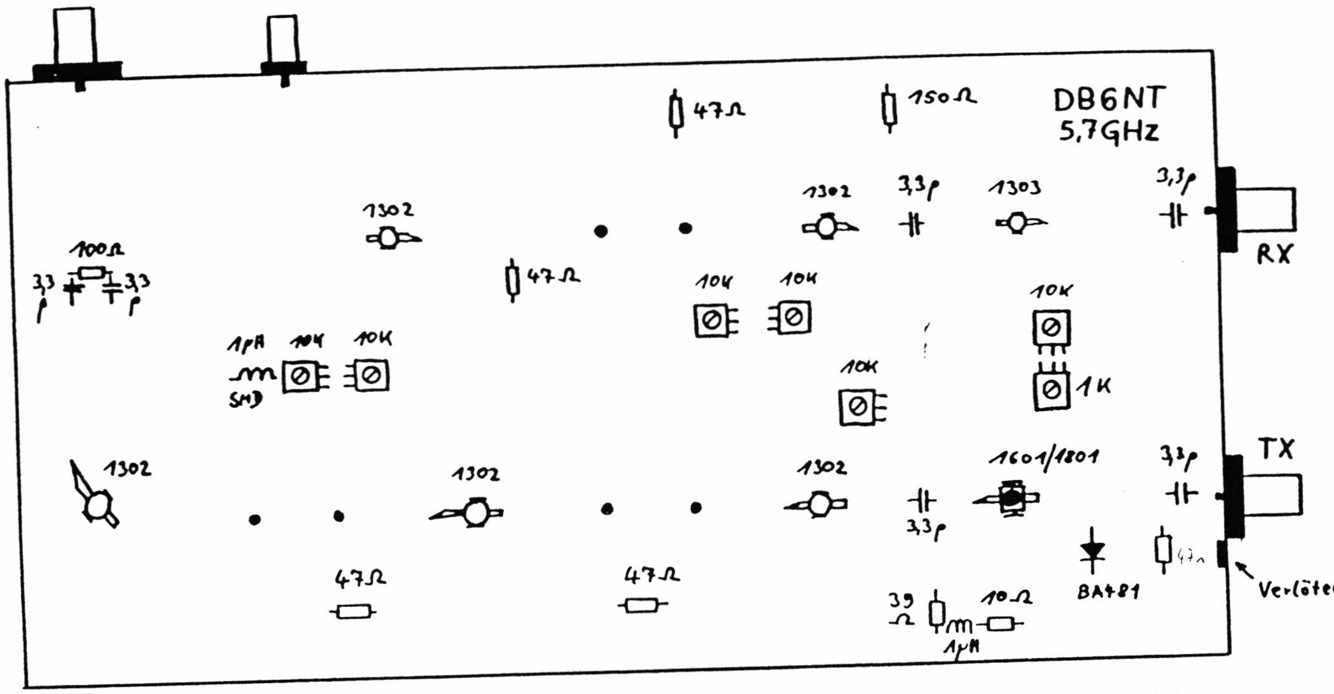
M = 2 : 1

DB6NT
6.91

145,5mm



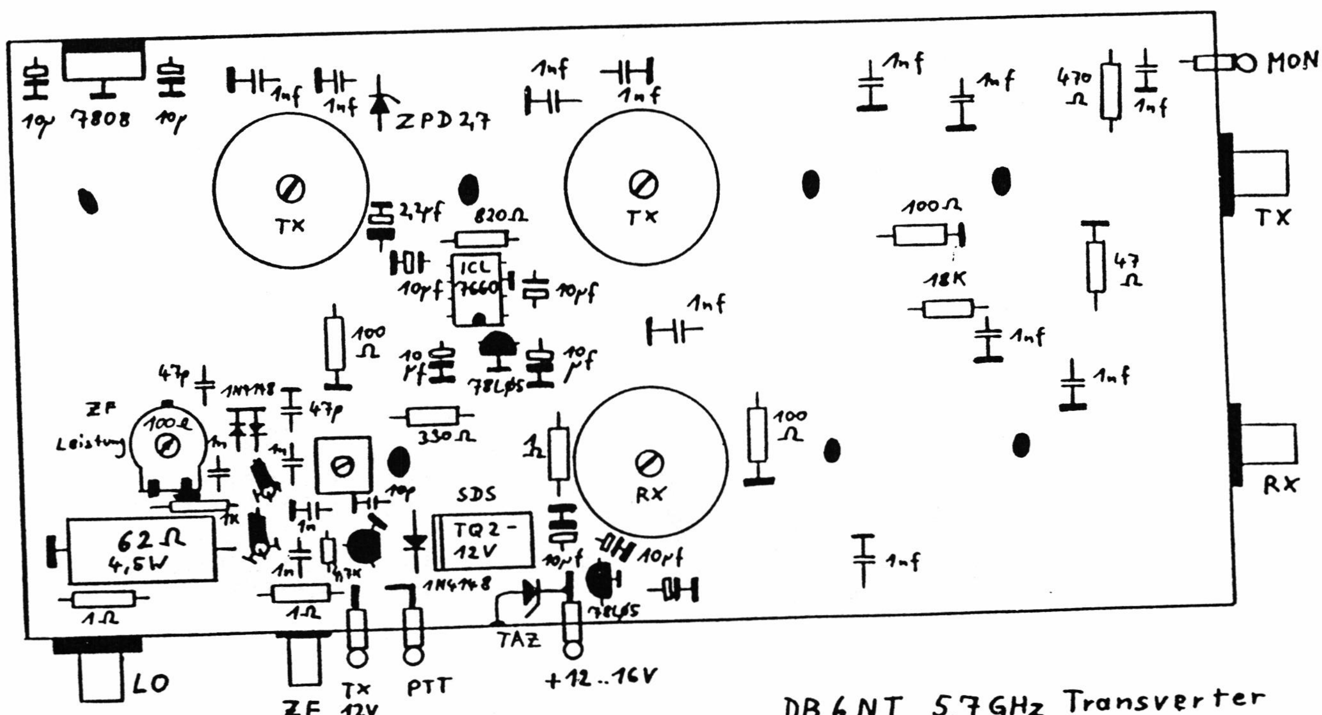

Bild/Figure 4: Layout of 5.7 GHz Transverter




Bild/Figure 5: Parts Layout bottom side

DB 6NT 5,7 GHz Transverter

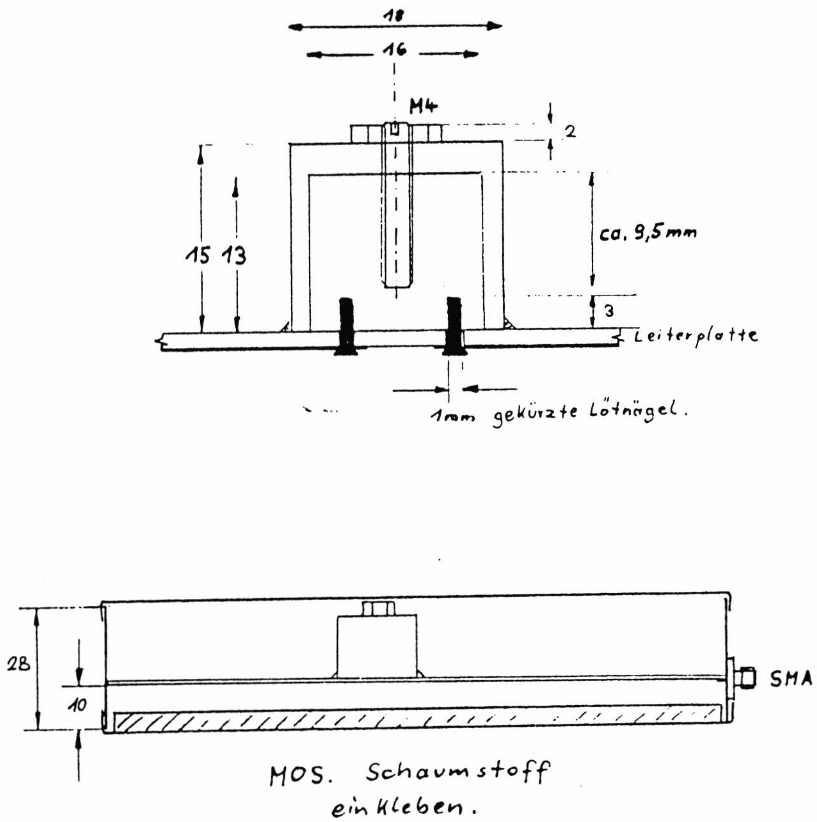
6.91




Bild/Figure 6: Parts Layout top side

DB 6NT 5,7 GHz Transverter

6.91




Bild/Figure 7: Construction Details

## 2. Construction and Tuning

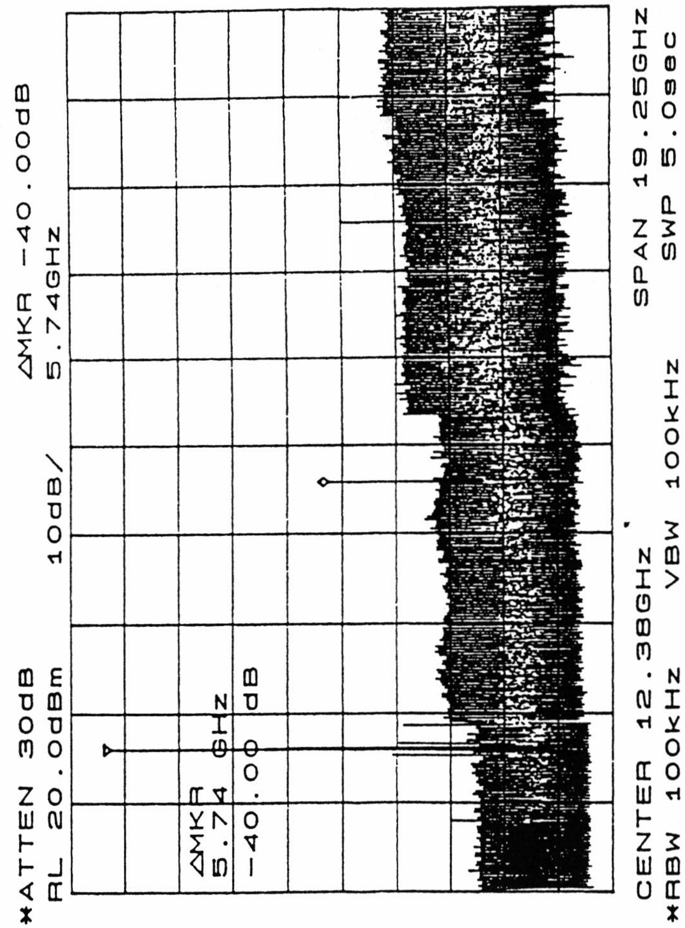
For construction follow exactly the details given for the construction of the similar 10 GHz transverter •.

Tuning can be performed with the aid of only two instruments:

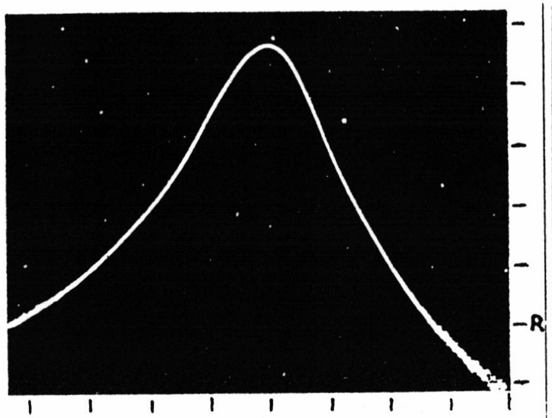
1. Powermeter for 5.76 GHz
2. Voltmeter

---

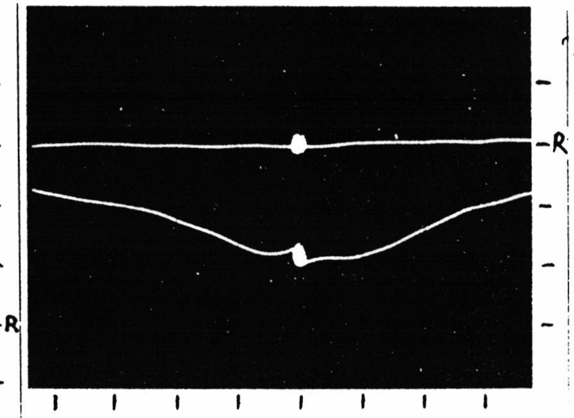
• *Michael Kuhne, DB6NT, Jürgen Dahms, DC0DA, "Einfacher 10 GHz Transverter", DUBUS 1/1991, pp. 3 - 14*




Bild/Figure 8: Measurement Results for Output Spectrum

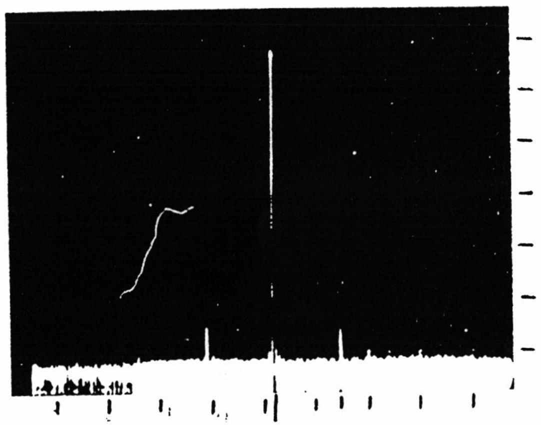


RX-Gain V=5dB/DIV. H=10MHz/DIV.



Eingangsanpassung

RL-IN V=5dB/DIV. H=100MHz/DIV.



Ausgangssignal H=100MHz/DIV. H=10dB/DIV.

Messwerte der Prototypen

| Aufbau | : | A | B | C | |
|---|---|---|---|---|---|
| Output | : | 320 | 310 | 310mW | Sat. |
| Nebenwellen | : | -42 | -48 | -52 | dB |
| NF | : | 2.0 | 1.8 | 1.7 | dB |
| RX-Gain | : | 23 | 24 | 23 | dB |
| RX-Spiegelfrequenzunterdrückung | | | | > -45dB | |

Verw. Messgeräte: AIL 707, AIL 7360
HP 346C, HP 435, HP 8485A, HP 8620
HP 11666A, HP 11664, HP 8755B.

Bild/Figure 9: Measurement Results for 5.7GHz Transverter

After connection of a suitable antenna or dummy load all currents and voltages given in the circuit diagram (Figure 3) should be adjusted with the relevant SMD-pots. This basic tuning should be carried out without any LO-power or IF-power.

## 2.1 RX-Tuning

Connect LO and 144 MHz transceiver. Switch on LO. Drain voltage on RX-Mixer should drop about 0.1V. Turn tuning screw of the RX-resonator from the top slowly downwards. You will observe two settings with an increase in output noise level. The first one - seen from the top position of the tuning screw - is the right one. The second setting corresponds to the image frequency of 5472MHz. Next mixer current and IF-Filter should be tuned to maximum noise level.

## 2.2 TX-Tuning

Switch transceiver to transmit in CW without keying down. After switching on LO the voltage drop at drain resistor of TX-mixer should increase from 5 mV to about 300mV. Switch off LO and key down 2m-transceiver. Adjust 100Ω pot in the IF-input for a voltage drop of 150mV at the drain resistor of TX-mixer.

Switch on LO again and leave 2m-transceiver keyed down. The you can measure a voltage drop at the drain resistor of first TX-amp. Tune first resonator in TX. There are three tuning positions which peak the first TX-amp current. The top position is the right one again. The lower positions correspond to the LO and the image frequency. The right position can be checked by switching off signal at 144 MHz. Current should drop immidiately.

Measure current in MGF1801 drain circuit by observing the voltage drop at drain resistor. Tuning the second TX-resonator should lead to an increase of current. An output powermeter should indicate RF-power now. By fine tuning all resonators and the IF-power the TX-ouptut power can be optimized. A power level of 200 mW should be possible.

## 3. Results

TX-Ouput spectrum can be seen in figure 8. LO and $F_{lo}+2xF_{zf}$ are down about 55 dB. Image frequency $(F_{lo}-F_{zf})$ is down more than 60 dB (Figure 9). Saturation output power is in excess of 300 mW.

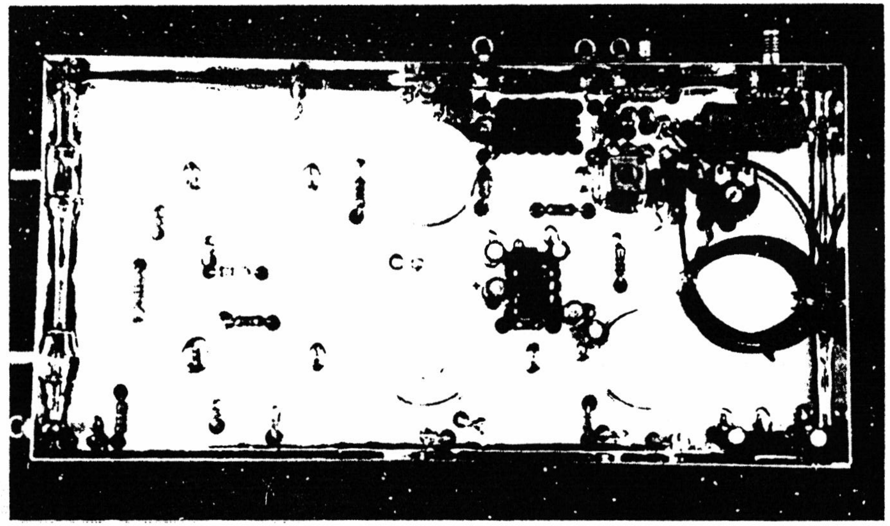
Noise figure of RX is typically 1.85 dB. gain is about 23 dB. Image frequency suppression is more than 45 dB.

## 4. Teileliste/Parts List

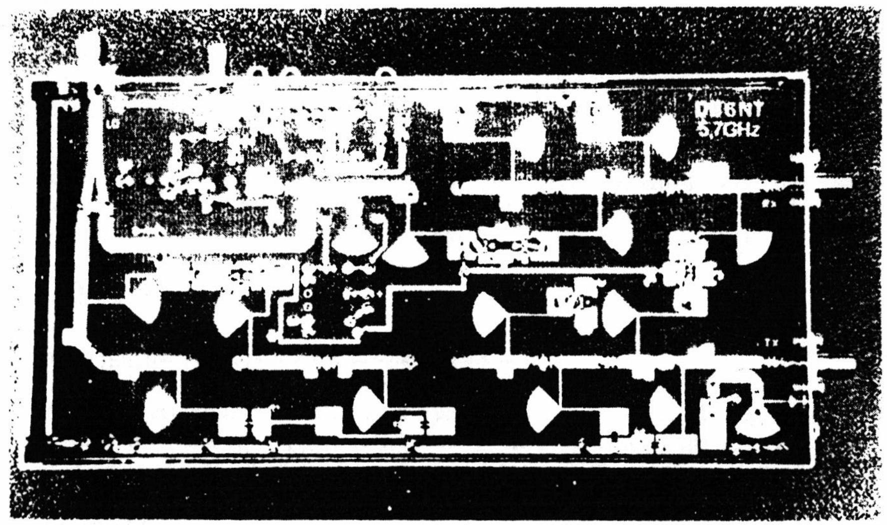
| Anzahl | Bezeichnung | Bauform | Bezugsquelle |
|---|---|---|---|
| 12 | Widerst. 0.25 W | 0207 | Diverse |
| 1 | Widerst. 4.5 W 62 Ohm | WK8 | Bürklin 29E368 |
| 9 | SMD-R | 1206 | Diverse |
| 6 | SMD-Poti 10k | SMD | Diverse |
| 1 | SMD-Poti 1k | | Diverse |
| 1 | SMD-Poti 100 | | Diverse |
| 15 | Keramik-C 1nF | EGPU | Diverse |
| 2 | Keramik-C 47 pF | EGPU | Diverse |
| 1 | Keramik-C 10 pF | EGPU | Diverse |
| 6 | Keramik-C 3.3 pF | EGPU | Diverse |
| 9 | Elkos 10 uF/16 V | 4.5 x 8 mm | Diverse |
| 1 | Elkos 2.2 uF/16 V | 4.5 x 8 mm | Diverse |
| 4 | Duko 1 nF by pass | Lötbar | Diverse |
| 2 | Spule 1 uH | SMD | Diverse |
| 1 | Spule 115 nH NEOSID | 5061 | Diverse |
| 1 | Dünnes Koax 36 cm RG174 o.ä. | | Diverse |
| 1 | Relais SDS | TQ2-12V | Bürklin 30G7556 |
| 1 | TAZ Diode | 1N6276 | Diverse |
| 1 | Z-Diode | ZPD2.7 | Diverse |
| 1 | Schottky diode | BA481 | Diverse |
| 3 | SI-Diode | 1N4148 | Diverse |
| 1 | SI-NPN | BC546B | Diverse |
| 1 | Inverter | ICL7660 | Diverse |
| 1 | Regler 8 V | MC7808 | Diverse |
| 1 | Regler 5 V | MC78L05 | Diverse |
| 5 | GaAs-FET | MGF1302 | |
| 1 | GaAs-FET | MGF1601/1801 | |
| 1 | GaAs-FET | MGF1303 | |
| 3 | Koaxbuchse | SMA | |
| 1 | Koaxbuchse | SMC | |
| 3 | Resonatoren | | |
| 1 | Weißblechgehäuse 74 x 148 x 30 | | |
| 1 | Teflon PCB ULTRALAM2000 0.78mm | Er=2.5 | Dirk Fischer, Neuer Graben, 46 Dortmund 1, Tel.:(++49)231105572 |

## 5. Remarks

The companion LO for 5.616GHz and a 9 cm transverter will be published in DUBUS 4/1991.



Bild/Figure 10: Top View on PCB



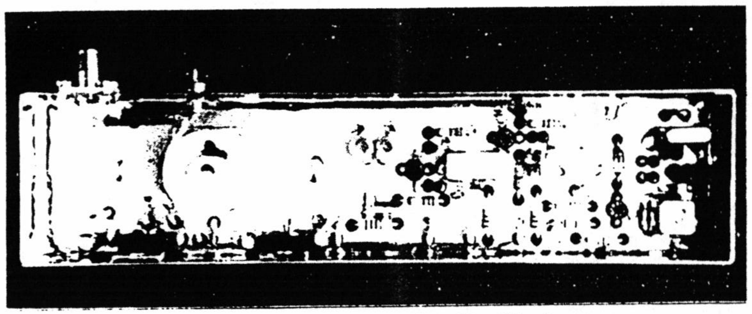

Bild/Figure 11: Bottom View on PCB

# 5.7 GHz Transverter: LO

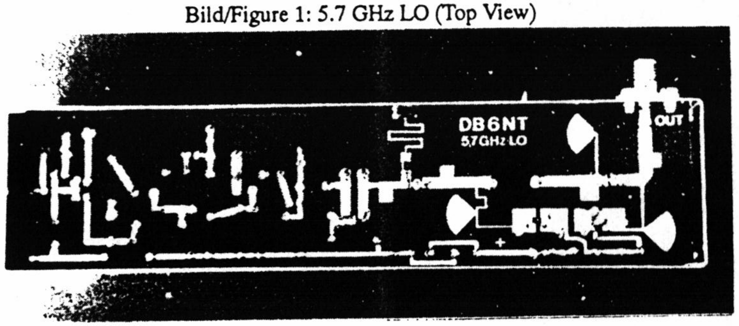
Michael Kuhne, DB6NT
Birkenweg 15, W-8674 Naila 2

(Part II)

DUBUS 4/1991



Bild/Figure 1: 5.7 GHz LO (Top View)



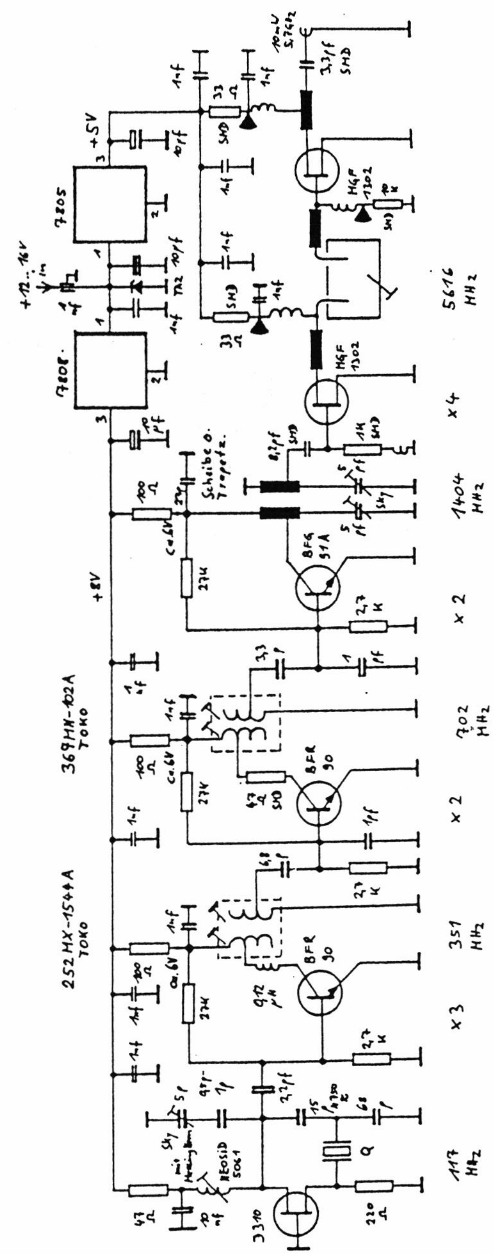

Bild/Figure 2: 5.7 GHz LO (Bottom View)

## 1. Design

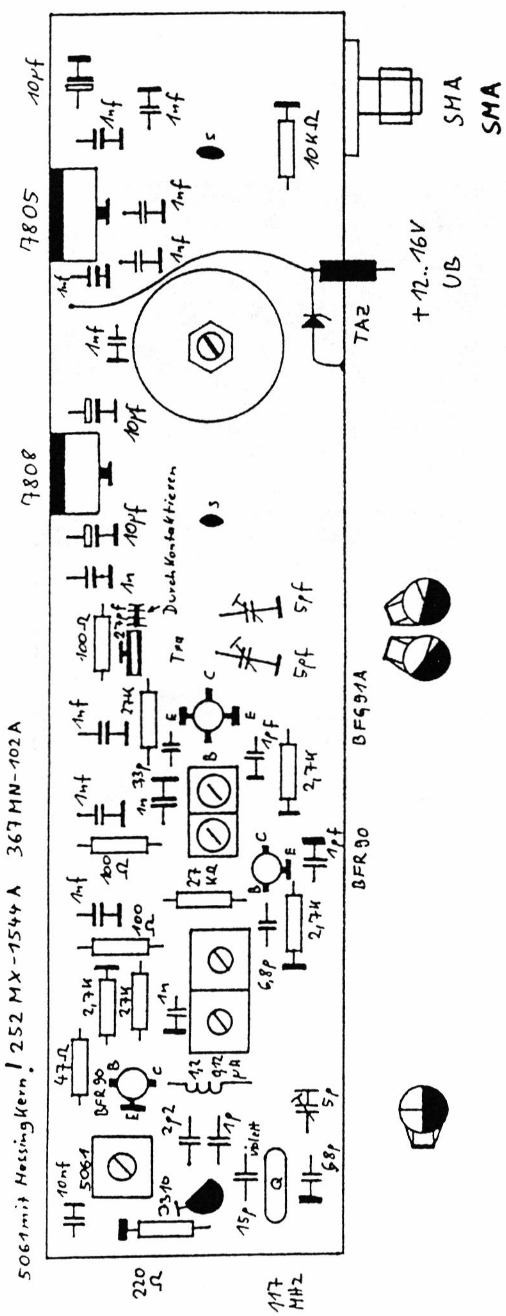
The Crystal-Oscillator on 117 MHz is designed around the J310 FET in the usual source feedback-circuit. The following stages are compromised of a tripler with a BFR90, a doubler with a BFR90, a doubler with a BFG91A, a doubler with a MGF1302 and a final stage with a MGF1302. Ready made Helical-Filters provide the selectivity on 351 and 702 MHz, a tuned stripline-filter serves on 1404 MHz and a single-pole resonator provides the final selection of the 5616 MHz signal. The oscillator coil is modified with a brass core for better stability and a loosely coupled SKY-trimmer provides a fine tuning facility.

5,7 GHz FREQUENZAUFBEREITUNG DB 6 NT

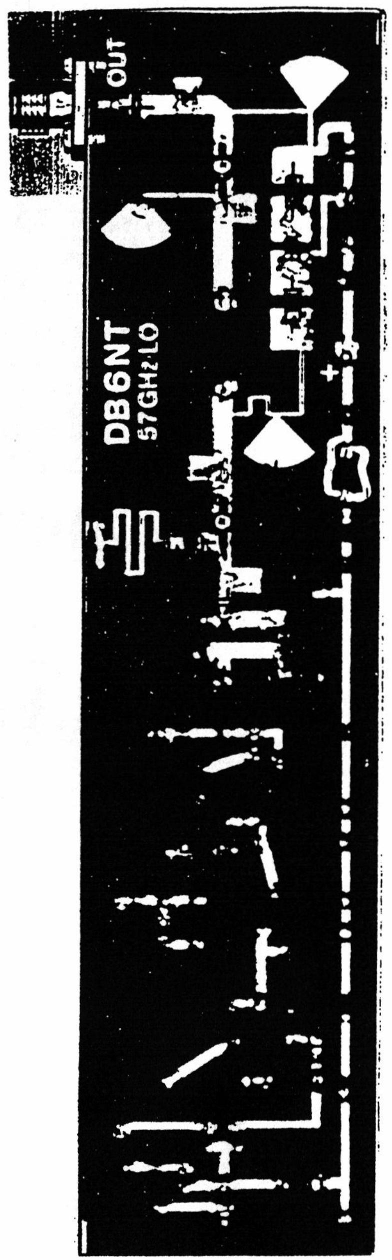
7.91




Bild/Figure 3: Circuit of 5.7GHz LO




Bild/Figure 4: Parts Layout




Bild/Figure 5: Parts Layout of 5.7GHz LO

HL = Hohlniete zur Durchkontaktierung DK

## 2. Construction and Tuning

After connection of the supply-voltage first the xtal-oscillator has to be initialized. Indication for its oscillation is a rise in collector current of first tripler (BFR90), which is equivalent to a decrease of its collector voltage. After turnon of the oscillator this voltage should be around 6 V. The fine tuning control (SKY-Trimmer) is in its mid position. Tuning of the first helical-filter leads to an increase in collector current of the doubler stage. The same procedure holds for the second doubler and helical filter. Adjusting the SKY-trimmers of the 1404 stripline filter to its normal position (see parts layout in figure 6) should follow in a current decrease in the FET-quadrupler, which turns out as a voltage increase in the drain circuit.

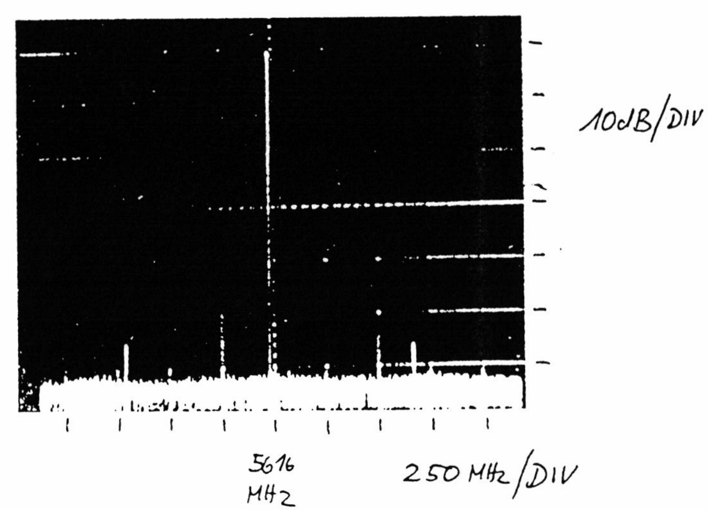
Now all preceding filters are tuned for minimum current in the quadrupler FET. After that the resonator tuning screw can be adjusted for maximum output on 5616 or to minmum current in the last FET-stage.

The target value for output power is 10 mW. If its more than 15 mW the power can be dcreased by increasing the resistors in the drain-circuits.

On the other hand power can be increased to 50 mW by increasing the gate resitor of the quadrupler to 4.7k and decreasing the drain resistors to 15 Ohms.

## 3. Results

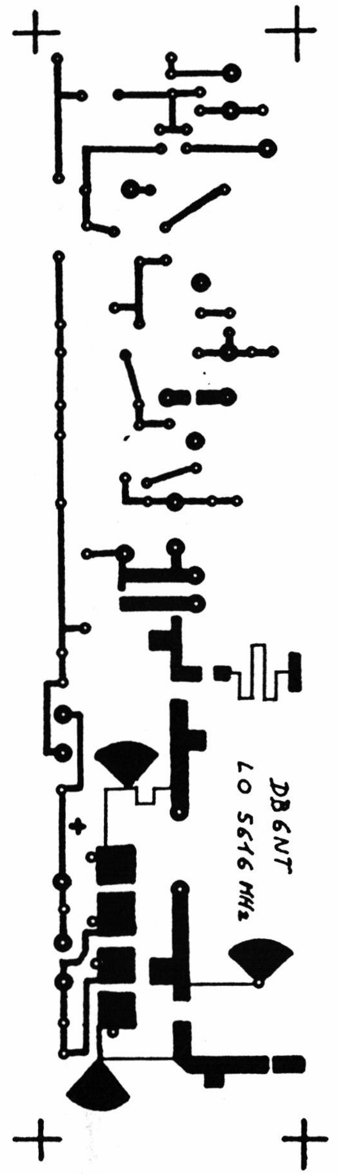
Figure 6 shows the output spectrum of LO with two spurious signals +- 704 MHz which have a level of -55 dBc.

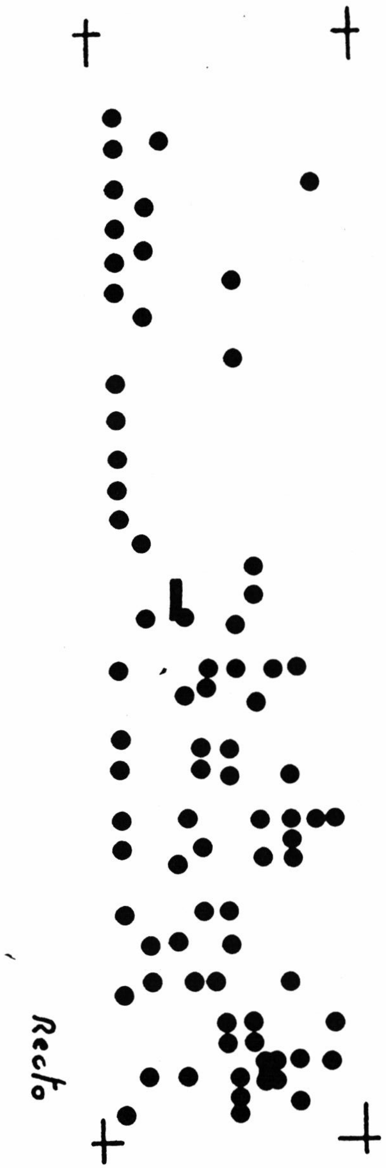



Bild/Figure 6: Measurement Results for Output Spectrum

## 4. Teileliste/Parts List

| Anzahl | Bezeichnung | Bauform | Bezugsquelle |
|---|---|---|---|
| 12 | Widerst. 0.25 W | 0207 | Diverse |
| 4 | Widerst. SMD | 1206 | Diverse |
| 13 | Keramik-C 1nF | EGPU | Diverse |
| 1 | Keramik-C 10 nF | EGPU | Diverse |
| 3 | Keramik-C 1 pF NP0 | EGPU | Diverse |
| 1 | Keramik-C 2.2 pF NP0 | EGPU | Diverse |
| 1 | Keramik-C 3.3 pF NP0 | EGPU | Diverse |
| 1 | Keramik-C 6,8 pF NP0 | EGPU | Diverse |
| 1 | Keramik-C 15 pF N750 | EGPU | Diverse |
| 1 | Keramik-C 68 pF N750 | EGPU | Diverse |
| 1 | SMD-C 3.3 pF | 0805 | Diverse |
| 1 | SMD-C 8.2 pF | 0805 | Diverse |
| 1 | Trapez-C 27 pF | | Diverse |
| 3 | Trimmer SKY 5 pF | Grün | Diverse |
| 3 | Elkos 10 uF/16 V | 4.5 x 8 mm | Diverse |
| 1 | Duko 1 nF | Lötbar | Diverse |
| 1 | Drossel 0,12 uH | 0207 | Diverse |
| 1 | Spule 115 nH NEOSID | 5061 | Diverse |
| 1 | Helixfilter 367MN-102A | TOKO | Componex/Düsseldorf |
| 1 | Helixfilter 252MX-1544A | TOKO | Componex/Düsseldorf |
| 1 | TAZ Diode | 1N6276 | Diverse |
| 1 | Quarz 117 MHz | HC18U | Diverse |
| 1 | Regler 8 V | MC7808 | Diverse |
| 1 | Regler 5 V | MC7805 | Diverse |
| 1 | FET | J310 | Diverse |
| 2 | Transistor | BFR90 | Diverse |
| 1 | Transistor | BFG91A | Diverse |
| 2 | GaAs-FET | MGF1302 | Mitsubishi |
| 1 | Koaxbuchse (Stecker) | SMA | |
| 1 | Resonator | | Eigenbau |
| 1 | Weißblechgehäuse 35 x 148 x 30 | | |
| 1 | Teflon PCB ULTRALAM2000 | Er=2.5, 2 x 0,035 mm Cu, 0,78 mm | Dirk Fischer, Neuer Graben 83, 4600 Dortmund 1, Tel.:(++49)(0)231-105752 |



## Building a 6-cm Transverter

Here are some ideas and circuits you can use to build your high-performance 6-cm transverter. I'll describe my latest mixer and band-pass filter designs and provide references for building the entire transverter. I'll also discuss my approach to integrating everything into a useable system.

*5760-MHz Mixer*

The mixer for 6 cm is quite similar to the one I designed for 10 GHz, using a hybrid splitter and a pair of $^6/_4$-wavelength ring mixers etched on Rogers 15-mil 5880 Duroid. 6 cm is perhaps the ideal band for using these mixers. The size is just about right—on 3 cm the rings are a little too tiny and on 9 cm they are getting a bit too big. More important, use of a relatively low 144-MHz IF makes it easy to design the hybrid ring for good performance at both the signal and oscillator frequencies. I published a design for a 5616-MHz LO multiplier chain in the May 1993 issue of *QEX* that will easily drive these diode mixers. You feed in the output of a 561.6-MHz no-tune LO and the chain puts out +7 or +15 dBm, depending on which output option you choose. I typically use +9 or +10 dBm to drive a pair of mixers with HSMS-8202 diodes. Using a higher IF for better image rejection often makes it necessary to compromise between good LO rejection and low conversion loss, since the $^6/_4$-wavelength ring is frequency sensitive.

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While the Hewlett Packard HSMS-2822 diodes sort of work, I don't recommend them if you can get HSMS-8202 Ku-band diodes. I get 1.5 to 3.8 dB less loss with the Ku-band diodes. More important, it wasn't necessary to tune the mixer to get good performance. One mixer I built showed only 5.3 dB of conversion loss without tuning. The input 1-dB compression point was +1.3 dBm. If you have a spectrum analyzer, one way to evaluate the diodes is to compare the upper and lower mixing sidebands of an untuned and unfiltered mixer. With good diodes, the two should be nearly identical, while diodes with excess stray reactances may result in significant differences of a dB or two in conversion loss. Of course, if you don't mind tuning the mixer with bits of copper foil, the 2822s are useable even at 10 GHz. I'd use +13 to +15 dBm of LO drive with the 2822s.

The radial stubs for the mixer and splitter are of different sizes. I made the one for the splitter smaller to compensate for the lead inductance of real resistors. At 6 cm, even tiny chip resistors can have a significant amount of stray inductance.

It may be worthwhile to add quarter-wave RF chokes between the splitter outputs and ground. 400 mils of 30-gauge wire should be about right for 5616 MHz. This will isolate the mixers from each other at the IF. Of course, the lack of isolation could make testing easier, since you could use the transmit IF signal to tune up the receive filter with a spectrum analyzer. On the other hand, noise from the transmit IF circuitry could prevent your receiver from obtaining the expected low noise figure. In one 10-GHz transverter I built, this noise doubled the NF from 1 to 2 dB.

*Mixer Construction*

After etching the board on 15-mil 5880 Duroid, I trim the board with a shear into a rectangle, leaving copper foil out to the edges. This allows me to solder 0.50 × 0.025-inch brass strips around the mixer board to form a frame suitable for adding a cover. These strips are drilled and tapped for five SMA connectors. SMA connectors are overkill for the IF connection, but they result in a compact and RF-tight assembly. By using 2-hole flange connectors for the IF, you can offset the center pins so they clear the ground foils. I've also used Teflon feed-throughs, but these compromise the RF integrity of the assembly. It is possible to improve performance slightly by tuning the mixers with small pieces of copper foil, but this shouldn't be necessary. My May/June 1993 *QST* 10-GHz article discusses mixer tuning.

*5760-MHz Band-Pass Filter*

The low IF used by amateurs makes filtering a challenge. Often, amateurs resort to waveguide filters to generate a clean signal. This works, but 6-cm waveguide is a little big for my taste. (Plus, it isn't the most commonly found stuff, particularly at New England flea markets. I think it's around, but who wants to cart around heavy metal objects that few people want?) A pipe-cap filter is pretty easy, but as some amateurs have discovered, a single pipe-cap filter gives marginal performance, especially if you are fortunate enough to find a surplus amplifier capable of generating significant power.

By carefully evaluating the plots by Kent Britain in the 1988 *Microwave Update* and making an educated guess using my knowledge of filters, I came up with a two-resonator filter that seems rather easy to duplicate—if you have the proper equipment to tune it up. I tried improving on my initial guess by varying the probes, and small changes weren't too critical, particularly when compared to trying to adjust a single resonator design for lots of spurious attenuation. The design shown in Fig 1 has a bandwidth of 47 MHz with 3 dB of loss. Using a 145-MHz IF, unwanted mixing products are at least 43 dB down. This should be adequate if you are using low-side injection with

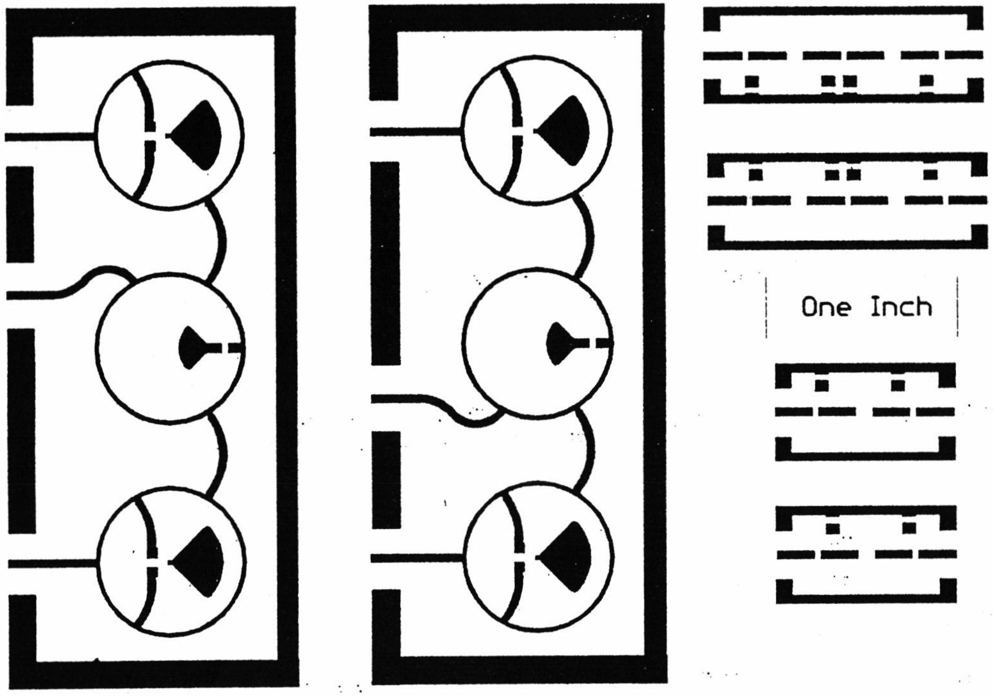



Fig 1—Etching pattern of mixer and MMIC amplifier boards. Board material is 15-mil Rogers RT/duroid 5880 with a dielectric constant of 2.2.

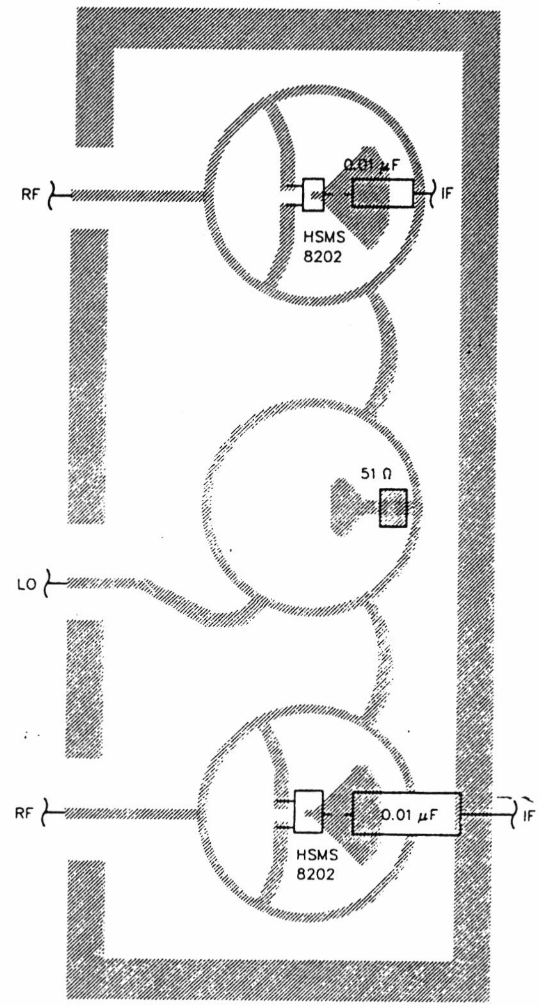



Fig 2—Parts-placement diagram for the mixer board.

typical surplus amplifiers.

Like most of you, I don't have a network analyzer to use to sweep the filters. Instead, I used a mixer/LO to upconvert a VHF/UHF signal generator. By terminating the mixer with an isolator, I get measurements that seem to make sense. I've tried in the past to make measurements with a multiplier/filter driven by the signal generator, but this required much more work and yielded rather fuzzy results. With the mixer, I was able to cover 400 MHz around 5760 with less than a dB of variation in spectrum analyzer response; it wasn't even necessary to use the correction factor needed with the multiplier setup.

The advantage of using more resonators is the resulting steeper skirts that better attenuate the unwanted mixing products—I found the Q of the resonators I built wasn't high enough to clean up the signal with just a single resonator. Using corrosion-resistant nickel-plated brass screws is part of the reason, but I don't have a source for small quantities of silver-plated screws of the right size. Brass and nickel are significantly less conductive than copper and silver, so losses are noticeably higher if you use tuning screws made of these materials. By the time I narrowed the bandwidth of the filter enough to get lots of stop-band attenuation, the losses became excessive. In addition, with a really narrow filter you need to start worrying about effects like mechanical and temperature stability. It isn't hard to see how a filter with flexible Teflon board supporting some probe pins could be easily detuned.

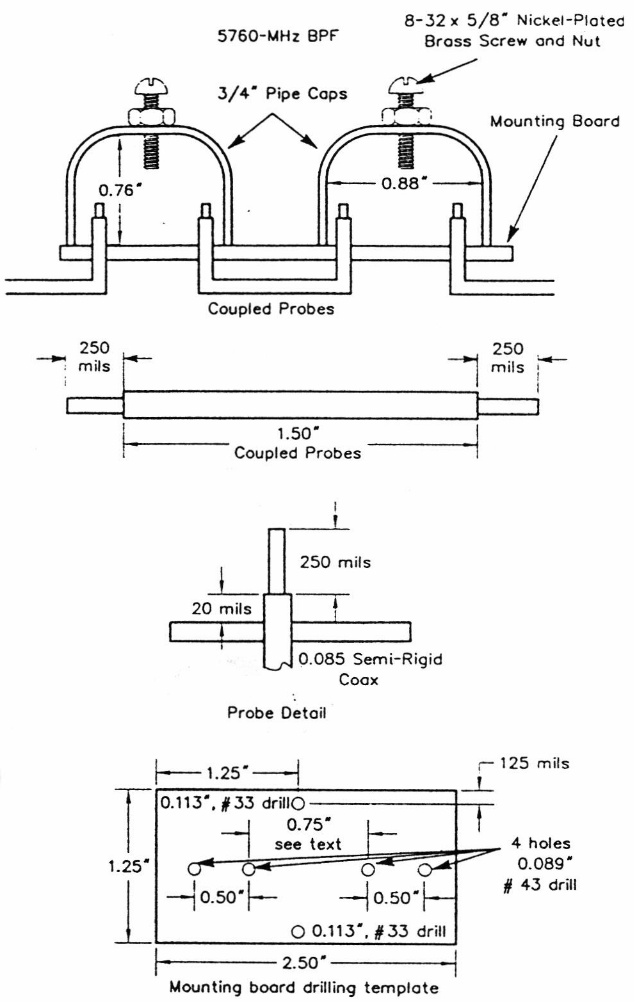



**Fig 3—Coupled 5760-MHz pipe-cap band-pass filter. This filter provides a 47-MHz bandwidth with 3 dB of insertion loss.**

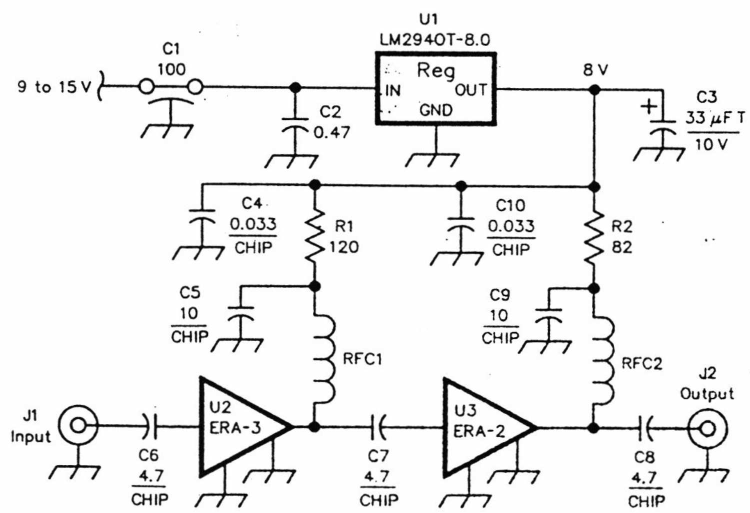



**Fig 4—Schematic diagram for the MMIC amplifier.**

**RFC1, RFC2—3 turns no. 28 enam wire closewound, 0.062-inch inside diameter.**

**U1—National 2940T-8.0 low-drop-out regulator.**
**U2—ERA-3 Mini-Circuits MMIC.**
**U3—ERA-2 Mini-Circuits MMIC.**

### *Filter Construction Details*

The filters use a pair of ¾-inch copper pipe caps. The one I measured actually had an inside diameter of 0.88 inches. They are tapped 8-32 at the top, though copper isn't the best material for holding threads. (You could solder brass nuts to the top, but then you would need longer screws than the 8-32 × ⅝-inch screws I used.) I estimate that the screws extend about 0.35 inches into the cavity. Polishing up the inside of the caps is a good idea—a smooth surface helps raise the Q of the resonators.

I've found that tightening the lock nut pulls out the screw slightly, raising the resonant frequency. Conversely, tightening the screw against the copper threads of the pipe cap moves the screw in a little, lowering the resonant frequency. Done just right, you can end up with a precisely tuned resonator with the lock nut at just the right tension. This is how I normally tune up my waveguides and cavities. Alternately, you might consider a better tuning mechanism that doesn't require as much skill. I saw an interesting Gunn oscillator design that used dielectric rods attached to piston tuning assemblies.

The resonators are coupled together with a 2.0-inch length of 0.085-inch diameter semi-rigid coax. While I didn't experiment with different lengths, I recommend you stick with this length of coax. 250 mils of shield are removed from each end of the coax, so that the dielectric is exposed. The dielectric is left on to protect the center conductor. It probably makes sense to bend the coax first, then drill the coupling holes for the resonators through the mounting board. You can then vary the 0.75-inch spacing to suit the coax, as it isn't critical. This may be easier than precisely bending the coax to match the spacing. The 0.500-inch spacing between the probes in the resonators *should* be maintained, unless you want to experiment with a new design. I made the mounting board out of unetched 1/16-inch-thick double-sided circuit board. The poor thermal conductivity of the fiberglass is an asset—you can solder the pipe caps to the board without unsoldering the probes. Including the #33 mounting holes is a good idea even if you don't intend to use them immediately, as adding them later might take a bit more work.

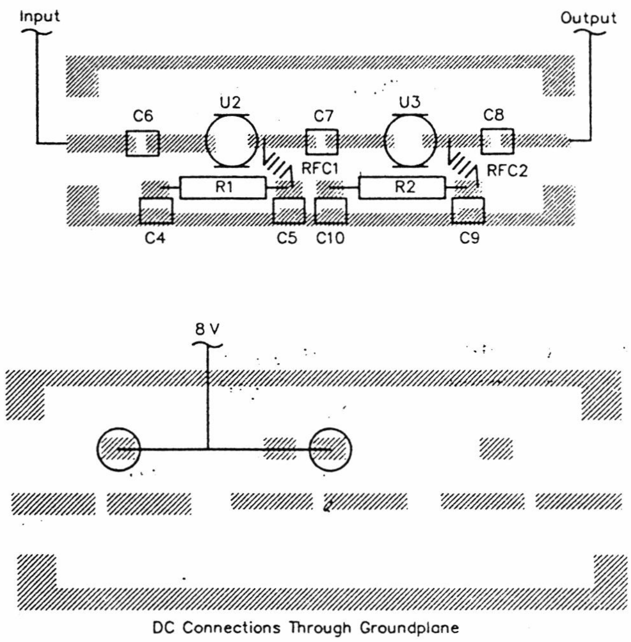



**Fig 5—Parts-placement diagram for the MMIC amplifier.**

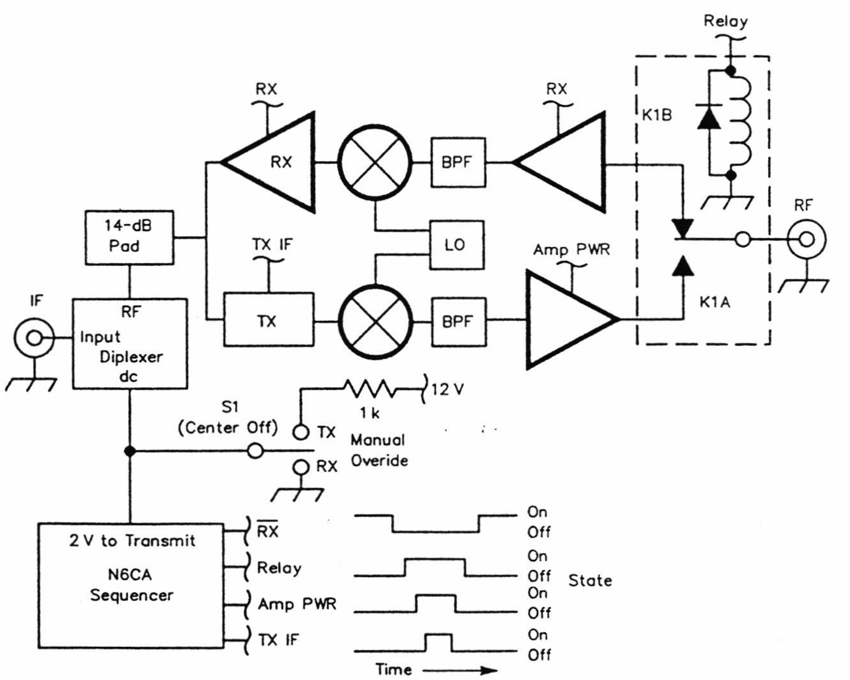



**Fig 6—Using the N6CA sequencer board with a transverter and T/R relay.**

### *5760-MHz Amplifiers*

Transmit amplifier response plays a big part in the cleanliness of your microwave signal. Most surplus amplifiers useable for 5.76-GHz amateur work are designed for operation at 5.9 to 6.4 GHz. Not surprisingly, those using high-side local oscillators require more filtering, since the amplifiers are actually optimized for the LO frequency. This often isn't a problem with retuned or homebrew amplifiers, since the tuning typically results in a narrowband amplifier with rejection off the tuned frequency. But, many surplus amps do work reasonably well in the amateur band without any tuning, so many people do use them "as is."

The new Mini-Circuits ERA MMICs are just what we need to take the filtered signal and amplify it up to a level adequate for driving TWTAs or surplus amplifiers. The cascade of ERA-2/3 MMICs shown in Fig 2 has 26 dB of gain and +14 dBm of linear output. To prevent unwanted feedback, the amplifiers are only 0.50 inches wide. This results in a cutoff frequency of 11.8 GHz—high enough to offer significant attenuation over the bandwidth of the MMICs. A much wider enclosure invites waveguide propagation unless hard-to-find microwave absorber material is used. The simple MMIC circuitry makes this easy to accomplish. Amplifiers using GaAs FETs are often much wider, in order to accommodate the low-loss bias circuitry. I've also included the etching pattern for an amplifier using just a single MMIC, for applications that don't require the gain of two MMICs.

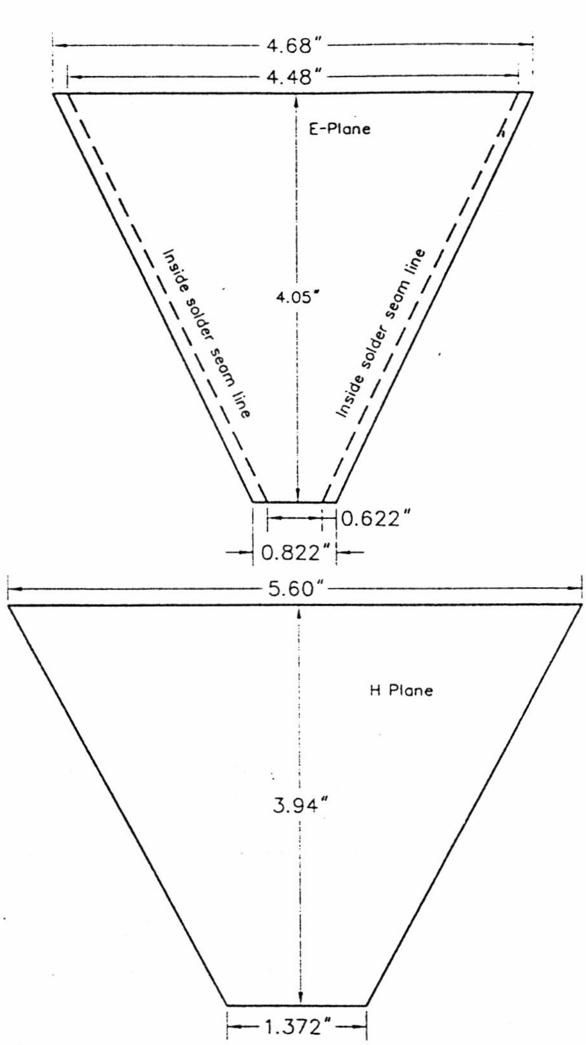



**Fig 7—15.8-dBi 5760-MHz horn dimensions. The horn solders to WR-137 waveguide. The dotted line marks the inside solder seam.**

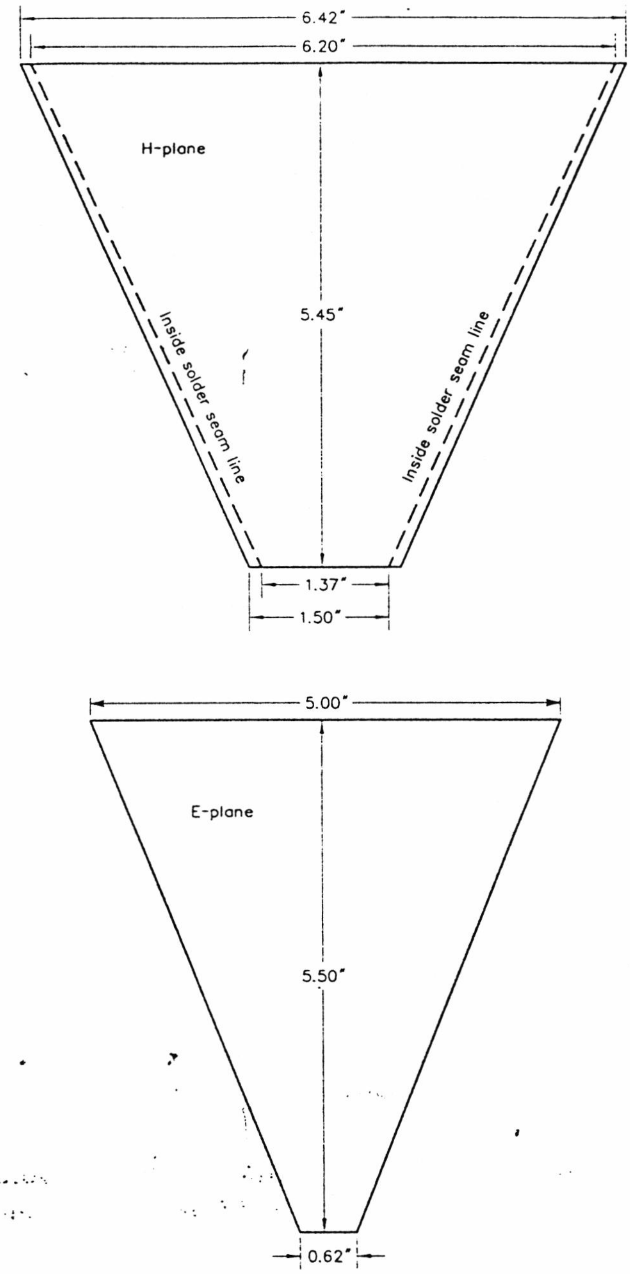



**Fig 8—17-dBi 5760-MHz horn dimensions. The horn solders to WR-137 waveguide. The dotted line marks the inside solder seam.**

### *MMIC Amplifier Construction Details*

After etching the board on 15-mil 5880 Duroid ($\varepsilon_r$=2.2), I trimmed the board to 0.50 × 1.45 inches. Next, I drilled holes for the power leads and carefully cleared away the copper ground plane around these holes with a large drill bit. Practicing with some scrap Duroid and different drill bits is highly recommended—you don't want the bit to "grab" and ruin the board. I use dial calipers for laying out the brass strips that form the frame around the board, just as with the mixer board. The frame also holds the connectors and feedthrough capacitor. I ended up mounting the capacitor on the side of the box since there was no room to mount it on the output end plate. There is a convention that the dc input connector ought to be mounted next to the output connector, but there are numerous exceptions to this rule.

I punch a 94-mil hole for each of the MMICs to sit in, then bend the grounding leads flat against the MMIC and stick them through one of the holes. After the input and output leads are flush against the board, I bend the grounding leads flush against the copper ground plane and solder them down. Then I attach the other surface-mounted parts. This differs from the usual practice of mounting the semiconductors last. I do it this way to ensure the best possible ground lead connections, which is critical for proper microwave performance. Finally, I wire up the regulator on the ground plane side of the board.

There are two choices for a high-quality LNA on this band. The first is the design by Ranier, DJ9BV, in the March 1996 *QEX*. The second is the one I published in the September 1994 *QEX*. Ranier's has a slightly lower noise figure, but the 13 dB of typical gain isn't quite enough to overcome the noise generated by most 6-cm mixers, so a second stage will be required. My two-stage design with 22 dB of gain is about right for a terrestrial station. An EME station free of interference might effectively use as much as 30 dB of preamp gain ahead of the mixer to maintain a low system noise figure.

*System Integration*

Many of my transverters successfully use the Chip Angle sequencing circuit found in the *ARRL Handbook*. As the diagram in Fig 6 shows, I avoid possible relay damage by sequencing both the RF drive to the transmit mixer and the dc power to the transmit amplifiers. While you can still end up hot-switching the relay if the relay sticks and then releases at the wrong time, I think this is so rare that I haven't bothered to design a suitable interlock. (I suppose you could sense the dc continuity of the relay and use this to indicate relay closure.) The September 1995 *QEX* has schematics that show the sequencer and IF circuitry in more detail.

I've found that this technique works just fine with semi-break-in transceivers since the transverter doesn't really care what the transceiver is doing (except for the dc control signal, obviously). Even if the transceiver is transmitting when the converter is receiving, everything is still operating acceptably—the 14-dB pad and switching-diode loss protects the receive MMIC from excessive RF. Thus, it even makes sense to use a center-off toggle switch to provide a manual override. This allows you to plug in any low-power IF radio and manually force the transverter into the proper state.

Those of us who travel 100 miles from home to operate from a tall mountain appreciate this feature. This is also handy for testing the transverter with test equipment, though there is a potential for an unexpected problem. Some signal generators cut back the RF if they see dc voltage on the output connector, so you might think there is a problem with the transmit converter when there really isn't; the transmit converter isn't putting out power simply because it isn't seeing much drive.

This scheme works well with low-power solid-state amplifiers because they are easy to turn on and off with power FETs. You can buy P-channel FETs with relatively low on-resistance—with even better devices appearing as time goes on. A simpler approach is to use the on/off control pin of the National LM2941 adjustable regulator, which offers a low drop-out voltage and delivers an amp. I'm not sure how well it will work with TWTAs, which often have warm up times and can't be turned on and off instantly, at least with inexpensive techniques.

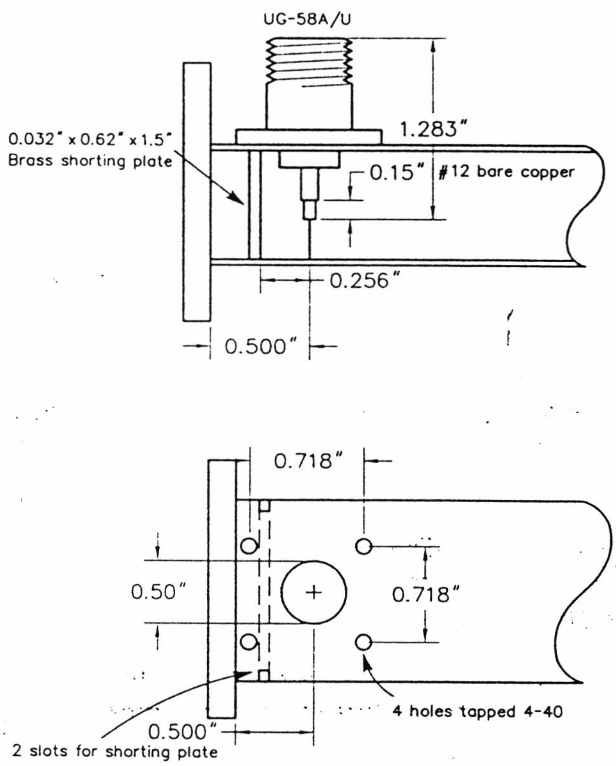



**Fig 9—WR-137-to-N connector transition.**

Much less critical is the receive sequencing. Usually, the delay is so long that it doesn't matter whether the receive converter is turned on before or after the transmitter! In checking my previous work, it looks like I've done it successfully both ways. Turning the receive converter off first makes sense if you want to conserve as much current as possible, a consideration when running from batteries. I've not done any testing to see whether preamps are more resistant to high levels of RF with or without power applied, but the answer may determine the best approach with marginal relay isolation.

*Connecting the Blocks*

The coax and waveguide techniques that work well at 10 GHz work even better at 6 GHz, so most people don't have any problems. However, building a 10-GHz station isn't really a prerequisite to getting on 10 GHz, though almost everyone on 6 GHz also has 10 GHz. Finding 6-GHz surplus equipment is a lot easier than finding stuff suitable for narrowband 10 GHz, though sources do seem to be drying up.

At 6 GHz, you can pretty much toss out the idea of short pigtails; they just don't work with commonly found coaxial cables. You might be able to make the technique work with tiny 0.035 or 0.047-inch diameter semi-rigid coax, but I've never seen the stuff available inexpensively. By soldering the shield directly to the ground plane, you can get very short connections. The solid shield also helps, since you don't have to worry as much about stray wires shorting out the cable. Even at 2.3 GHz, pigtails seem to work only for low-power circuitry; I've had little success getting them to work above plastic MMIC power levels.

Normally, I connect all the assemblies together with SMA connectors and use N connectors for the antenna hardware. Testing cables and adapters at RF is a good idea—there is cheap hardware around that won't work well at this frequency. I've joked that one particularly cheap adapter could be used as an image stripping filter, due to its rather pronounced notch response. While I've left the BNC connectors on our spectrum analyzer when doing quick checks, I don't recommend using them at 6 GHz.

At 6 GHz, your coaxial cable can actually be too big. If it supports more than one propagation mode, you can have signals propagated by different modes canceling each other out, resulting in extremely high losses. Andrew Corp suggests 5.0 GHz as the maximum frequency for LDF5-50A 7/8-inch Heliax. Of course, if you sweep the cable, you'll probably find frequencies above 5 GHz at which the cable works just fine. This is why many people are able to use 1/2-inch Heliax at 10.368 GHz: the frequency is in one of the clear windows above the cable's single-mode limit. I've had no trouble using either RG-213/U or 1/2-inch Superflex on 6 cm, although RG-213/U is rather lossy.

*Horn Antennas*

I included the simple antennas of Figs 7 and 8 to show an easy technique for mounting a small horn to a mast. Many textbooks and articles aren't clear how you accomplish this task. I use pieces of waveguide that come with unusual flanges. The flanges provide rugged attachment points for screws. To me, this is a lot easier than trying to fabricate some sort of bracket or welding the horn to a suitable mounting plate. This also works well for slot antennas. Besides, I have no other use for the mounting flanges. The horns have predicted gains of 15.8 and 17.0 dBi, respectively.

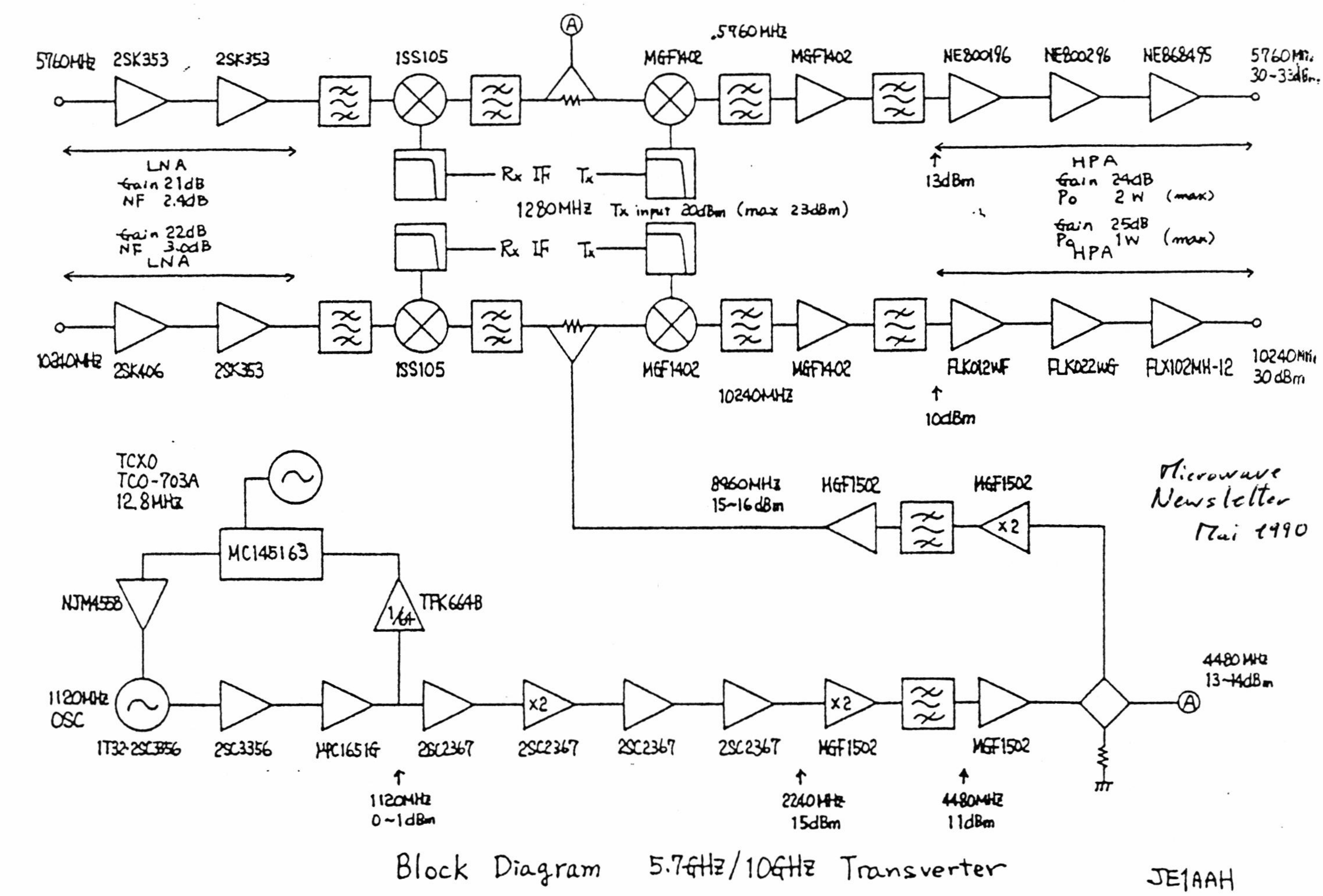
I made the phase centers equal by making the horn H-plane width smaller while keeping the horn length and E-plane widths constant. Having them equal is useful if you intend to feed a lens. While you could make the E-plane width wider, you actually lose a little bit of gain, even though the horn is bigger. You could say that the transition to free space is occurring too quickly for maximum gain. This is why horn gains above 23 dBi are rare—dish antennas become much more practical than a very long horn. With the 17 dBi horn, I also made the horn a little longer.

*Horn Construction*

I made the smaller horn out of 1/16-inch unetched double-sided fiberglass circuit board. For light weight, I made the 17-dBi horn of thinner 0.025-inch G-10 circuit board, though it's not as strong. For durability, I find it important to tape the joints with copper foil. Otherwise, people borrowing the horns return them with broken solder joints. To protect the copper from corrosion, I painted them with clear acrylic spray paint.

Since most people use coax on this band, Fig 9 shows an N-to-WR-137 transition. I used an Amphenol 82-97 UG-58A/U connector. These connectors have a center contact that press fits into the Teflon. A different connector may require a bit of experimentation, since the center contact forms part of the probe. Bare #12 copper wire is easily obtained by stripping ordinary house wire sold in hardware stores. I used a hacksaw to cut a pair of slots for the brass shorting plate. The 0.256-inch dimension is from the center of the probe to the inside surface of the shorting plate. I slid the snugly fitting 1.5 × 0.622 × 0.032-inch shorting plate through the slots and soldered it in place with a propane torch.

I soldered the horn directly to the waveguide with copper tape. A soldering iron is useful for tacking the tape into position. Then I used a propane torch to do the final soldering, since the waveguide needs quite a bit of heat to properly melt the solder. The small horn and large horns measured 14 dB and 30 dB return loss, respectively. I consider 14 dB adequate, though a purist would add tuning screws or vary the probe length for a better impedance match. People normally put screws on the centerline of the waveguide, where they have the most effect. Usually, two or three screws placed a quarter or eighth of a waveguide wavelength apart will match nearly any load. I wouldn't worry too much about not having a set-up to measure SWR—I made a couple of 200-mile 10-GHz contacts before I finally got a precision directional coupler and reduced the SWR of my dish feed below 2:1. □□




Block Diagram 5.7GHz/10GHz Transverter JE1AAH

Peter Vogl, DL1RQ



# A 6cm Transverter using Stripline Technology

## As presented at the 1990 VHF-UHF conference in Munich

This article describes five building block modules for a 6cm transverter with an IF of 435MHz.

The use of stripline techniques and surface-mount devices (SMDs) in quantity enables a very compact unit to be built with a high degree of reproducibility and reliability.

The achievable noise figure of 1.5dB and power output of 1.5W represent the state of the art in amateur technology.

## 1. POWER SUPPLY AND CONTROL MODULE DL1RQ NT6-1

### 1.1. Description

The front panel (fig. 1.1) is constructed together with the power supply and control PCB as a compact module, in order to save as much room as possible.

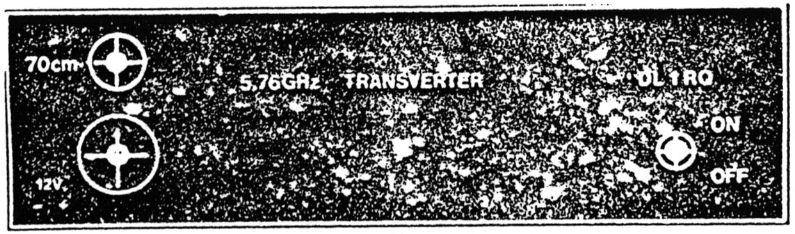



Fig.1.1 Front panel for DL1RQ NT6-1: 145 x 40mm
Material: Epoxy PCB 1.5mm thick, copper-clad both sides

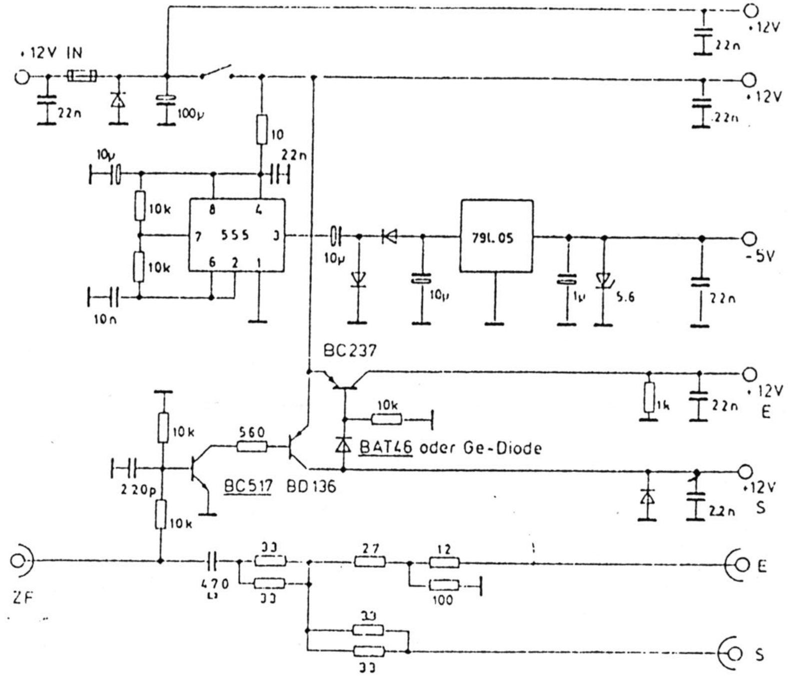



Fig.1.2 Circuit of DL1RQ NT6-1: (BAT46 may be any Germanium Diode)
E = RX; S=TX; ZF = IF

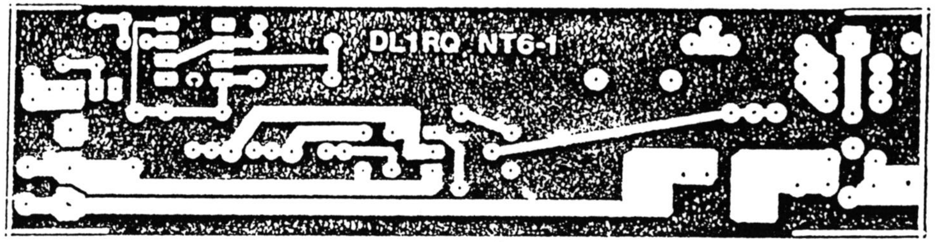



Fig.1.3 Power Supply and Control Module DL1RQ NT6-1
122 x 30mm; Material: Epoxy PCB 1.5mm thick, copper-clad both sides

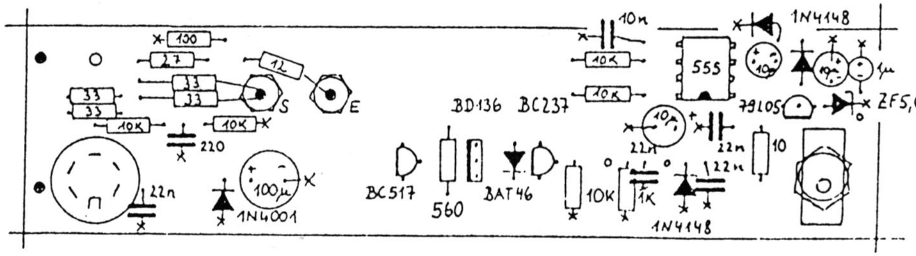



Fig1.4 Position of components on DL1RQ NT6-1


Fig.1.5 Components on conductor side of DL1RQ NT6-1
The SMC connectors are mounted from the conductor side. The fuse holder and the SMD capacitor 470pF are soldered onto the conductor side

Fig.1.6 View of the component side of DL1RQ NT6-1

The circuit (fig.1.2) is protected against short circuit and reverse polarity and provides a DC-DC converter for the negative supply -5V, a transistor-controlled switch for the 12V supply voltage as well as a resistive divider with 6dB reduction in power load for the send channel and 10dB reduction for the receive side. The measurements of the front panel were chosen to correspond with the width of the FT-790 driver used by the author.

The height of the panel is laid out to match commercially-available side profiles (1). Figs. 1.3 to 1.6 show the printed circuit board DL1RQ NT6-1 and the component overlay in drawing and photographs.

## 2. OSCILLATOR MODULE FOR 1331.25MHz - DL1RQ 012

A special oscillator module was not designed for the 6cm transverter, instead the respected and much copied oscillator module DL1RQ 012 was put into service again. This unit has already been published in (2) and (3). For the sake of completeness of this article it will be described again, in a slightly improved version. You should be aware that it is built in a tinplate box with lid, 37 x 111 x 30mm, which makes it of larger profile than the other 6cm modules.

### 2.1. Description

The oscillator module is built out of four stages (fig. 2.1). Beginning with the accepted and low-noise crystal oscillator with a P8002 (or J300) transistor, the frequency is tripled in the following stage using a BFY90.

Another tripler stage with a BFR96 follows and this drives a BFR96 doubler via a two-element filter. The output frequency is coupled through a three-element filter.

Output power is in excess of 40mW, more than enough to drive the module DL1RQ 023.

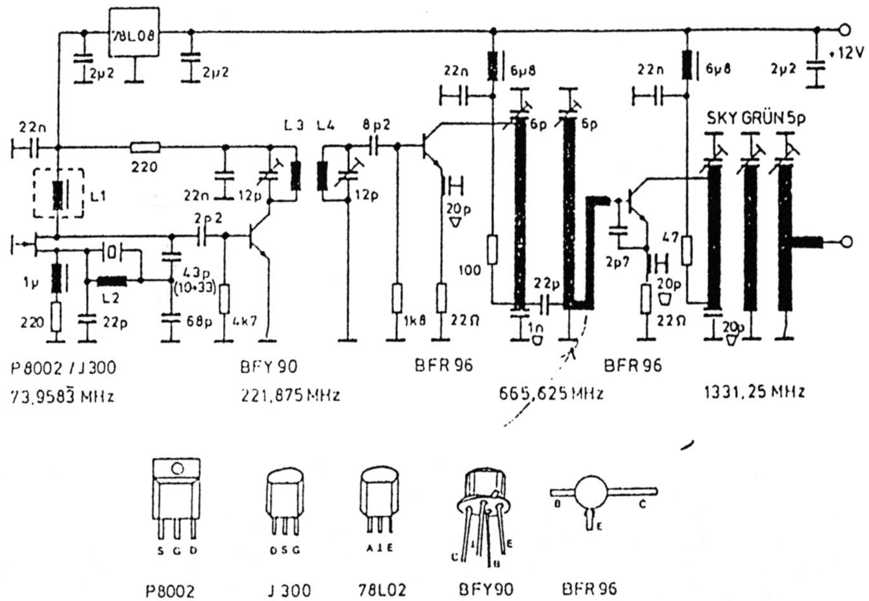



Fig.2.1 Circuit of DL1RQ 012
L1: BV5061 Neosid 115nH brown/blue
L2: 13.5cm varnished copper 0.15mm diameter on 10k resistor
L3: 2.5 turns varnished copper 1.00mm diameter air-wound on 5mm former
L4: as L3
Sky trimmers - 5pF green

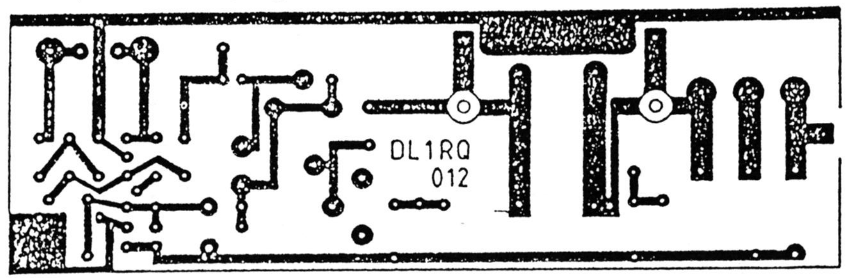



Fig.2.2 PCB for oscillator module DL1RQ 012
108 x 34mm; Material: Epoxy PCB 1.5mm thick, copper-clad both sides

### 2.2. Construction of DL1RQ 012

PCB, component layout and a sample construction can be seen in figs. 2.2 to 2.4.

Through contacts are indicated on the component overlay by an X in a circle and are formed either by component leads or by small pieces of 1.5mm copper wire. Components to be grounded, where no drilled holes have been provided, are soldered direct to the ground plane. This is indicated by the symbol * on the overlay plan.

The height at which the PCB is mounted in the tinplate enclosure depends on how tall the crystal used is and on the height of the crystal thermostat's insulation.

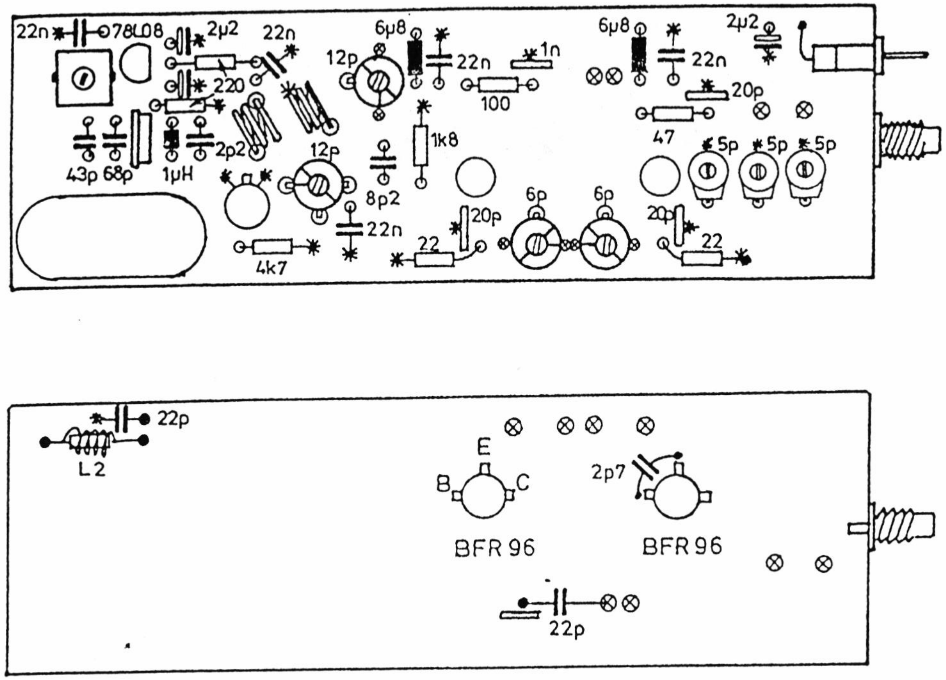



Fig.2.3 Component Layout of DL1RQ 012

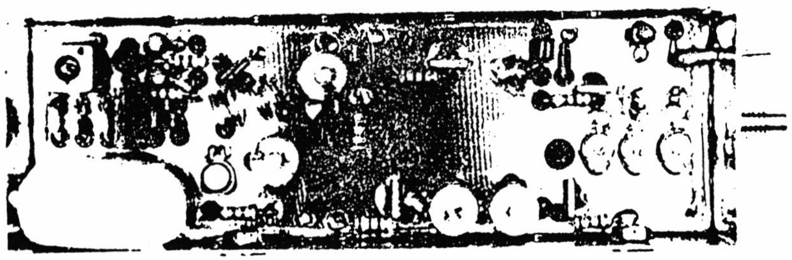


Fig.2.4 **View of the completed module in tinplate enclosure.**
**Lid measurements: 37 x 111 x 30mm**

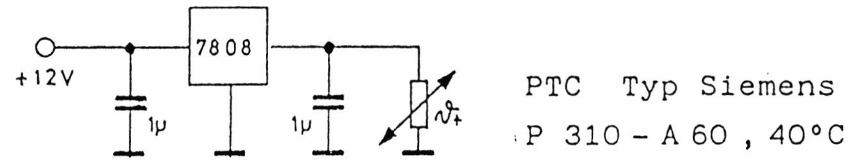



Fig.2.5 Circuit of a simple 60 degrees C thermostat
Hint: The equilibrium temperature lies *above* the indicated doubling temperature of the resistance value of the PTC resistor.

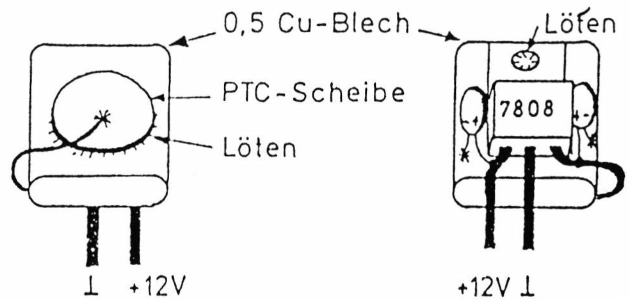



Fig.2.6 Construction details of crystal thermostat.
Cu-Blech = copper sheet material; Löten = solder; Scheibe = disc

## 2.3. Advice on simple crystal thermostats

Circuit and construction are indicated in figs. 2.5 and 2.6. To avoid thermal instability you need a particularly "intimate" contact between the PTC disc and the copper sheet.

The constructor soldered the PTC disc flat onto the copper sheet. Unfortunately the discs are prone to developing hair-line cracks when heated rapidly, and this leads to significant instability. The use of good flux and pre-heating of the PTC disc are recommended.

The thermostat is placed up against the crystal using plenty of heat-sink paste and surrounded by as much expanded polystyrene as possible.

# 3. TRANSISTOR QUADRUPLER 1331.25/5325MHz DL1RQ 023

## 3.1. Description

The quadrupler consists of two doubler circuits connected in cascade. The first doubler, using a BFR91, displays very good input matching. It operates with a fixed-tuned stripline circuit at 2662.5MHz. The second doubler is equipped with an MGF1302 and after a stripline filter delivers around 40mW output on 5325MHz. By means of a 10dB stripline coupler an oscillator level of approx. 4mW is fed out to the receiver module.

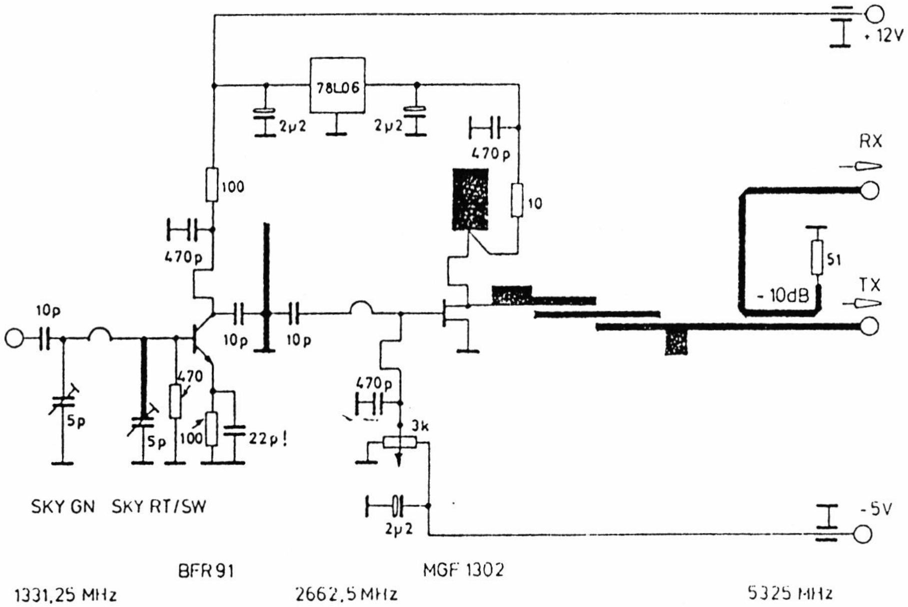



Fig.3.1 Circuit of DL1RQ 023
GN = green; RT/SW = red/black

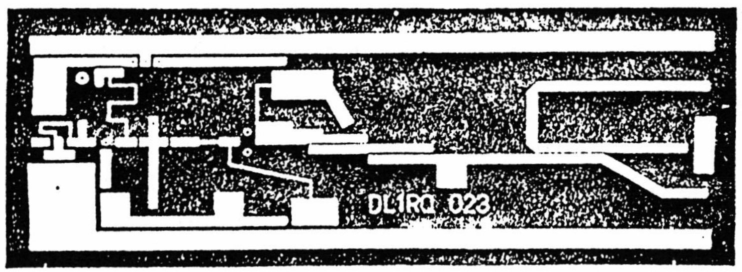



Fig.3.2 PCB for Quadrupler DL1RQ 023
108 x 34mm; MateriaL: RT-Duroid d-5870; 0.439mm

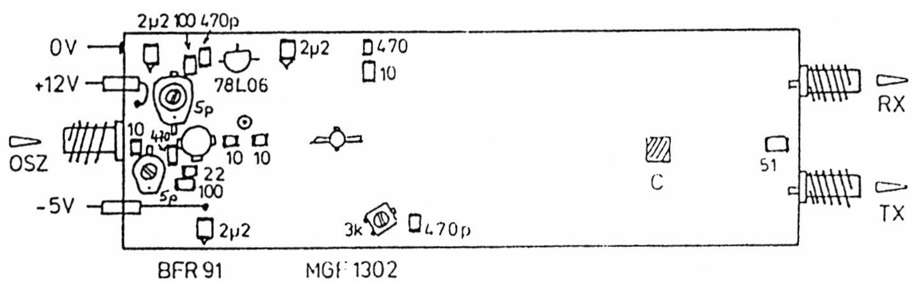



Fig.3.3 Component layout of DL1RQ 023
OSZ = oscillator

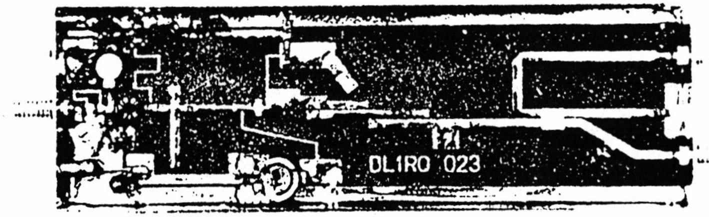



Fig.3.4 The completed module in tinplate enclosure.
Lid measurements 37 x 111 x 13

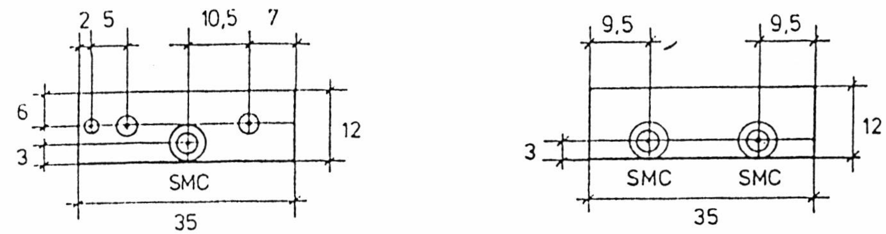



Fig.3.5 Dimensioned sketch of the end panels of DL1RQ 023

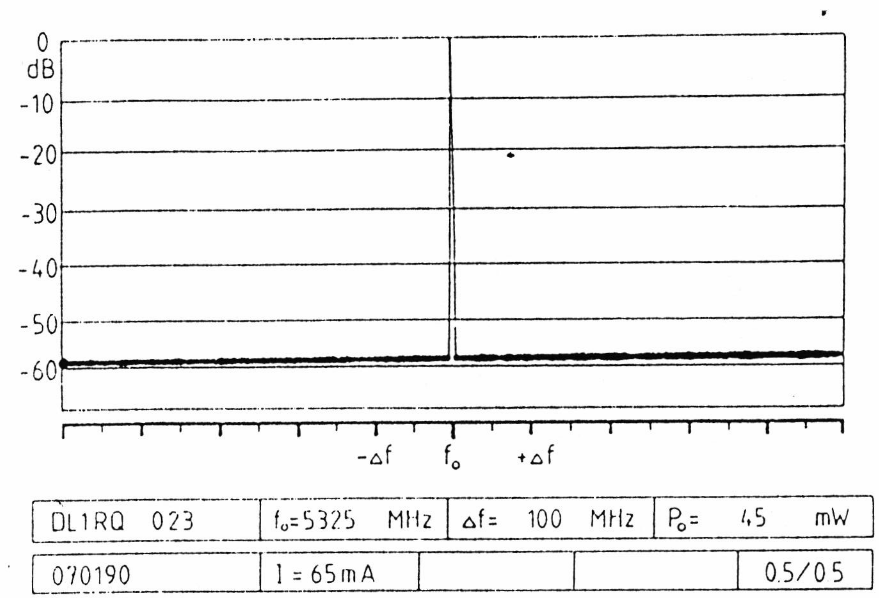



Fig.3.6 Output spectrum of the quadrupler DL1RQ 023

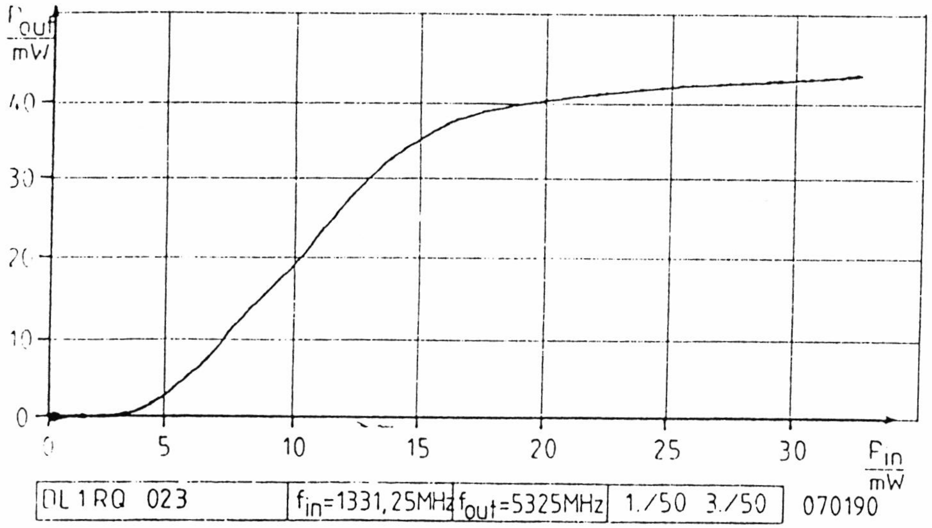



Fig.3.7 Relationship between input power at 1331.25MHz and output power at 5325MHz

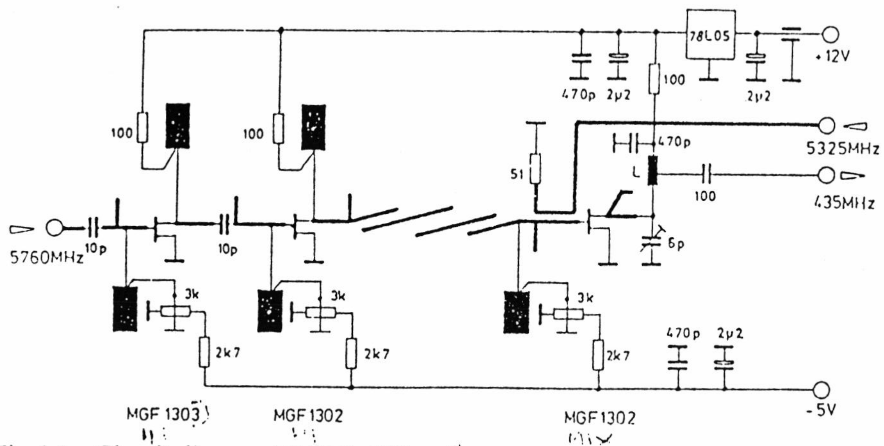



Fig.4.1 Circuit diagram DL1RQ 670

L: 2 turns of 0.5mm silver plated copper wire on 3mm former.
All components SMD pattern except voltage regulator 78L05 and 6pF trimmer (Valvo/Mullard grey).
10pF coupling cap's: Valvo 2222 851/853 (D104.520) 2 x 1.25 x 0.51mm

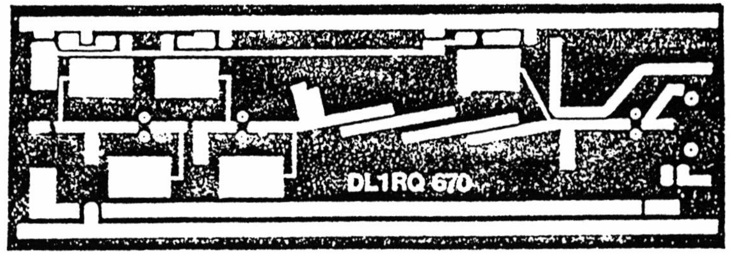



Fig.4.2 PCB DL1RQ 670
108 x 34; Material: RT-Duroid D-5870; 0.439mm

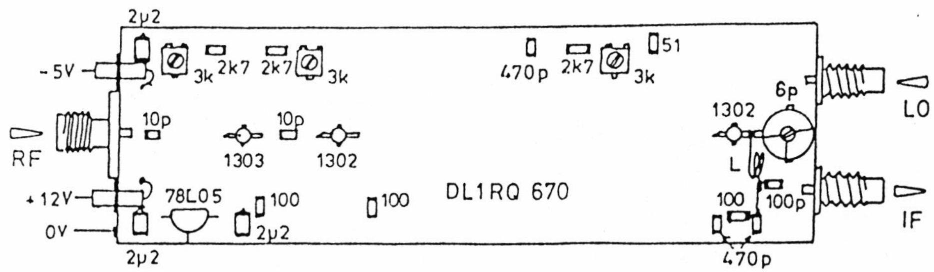



Fig.4.3 Component layout of DL1RQ 670

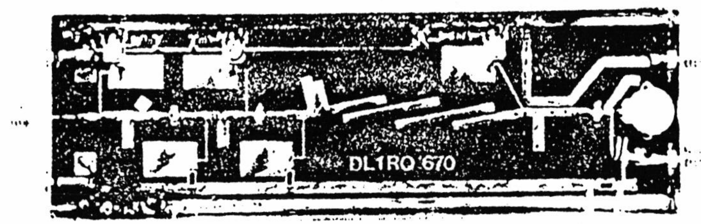



Fig.4.4 The completed module in tinplate enclosure.
Lid measurements 37 x 111 x 13

### 3.2 Construction of DL1RQ 023

PCB, component layout and a sample construction can be seen in figs. 3.2 to 3.4.

C is a 4 x 4mm piece of single-sided RT-Duroid PCB material 0.439mm thick and fixed with a small dab of universal glue (UHU, etc.)

Transistor BFR91 is mounted in the fashion of an SMD component. The cold end of the stripline resonator at the collector of the

groundplane side. The source connections of the MGF1302 are bent downwards, passed through holes and soldered to the groundplane.

A tinplate enclosure with the (lid!) dimensions of 37 x 111 x 13 is not commercially available to the knowledge of the writer. Accordingly just the lid of an available box with the dimensions 37 x 111 x 30 was used [type wg3113 from (4)]. The side walls 12mm tall were home-made to match the originals.

The SMC connectors are soldered from outside the end panels (fig. 3.5). The Teflon insulation which projects inside should be removed right back to the wall of the enclosure and the contact pin shortened somewhat.

The Teflon PCB should be trimmed so that the tracks line up with the pins of the connectors, then soldered to them. After this the edges of the PCB can be made "watertight" by soldering all round to the walls of the

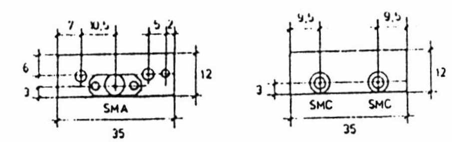



Fig.4.5 **Dimensioned sketch of the end panels of DL1RQ 670**

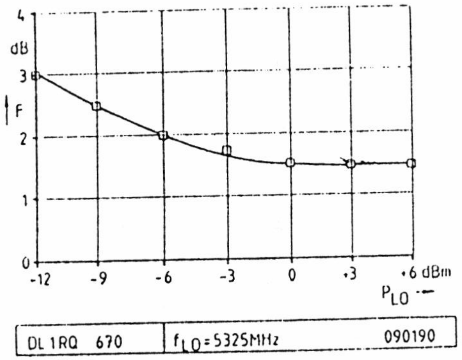



Fig.4.6 **Noise measurements for the receiver module DL1RQ 670 in relationship to the oscillator perfomance**

enclosure; this applies not only to the groundplane underside but also to the side strips or islands on the track side.

### 3.3. Alignment hints for DL1RQ 023

The following should be set up first:

The trimmer in the BFR91's base circuit should be fully turned "out", the input trimmer turned inwards a quarter. The gate-bias trimmer of the MGF1302 should be adjusted to its middle position.

After this the input trimmer should be adjusted for best input matching. Using the base circuit and the gate-bias, tune for maximum output power. In this operation, turn the base circuit in as little as possible and watch the drain current by the voltage drop on the drain resistor!

### 3.4 Measured values for the frequency quadrupler DL1RQ 023

Figs. 3.6 and 3.7 shown what should be achieved with this module.

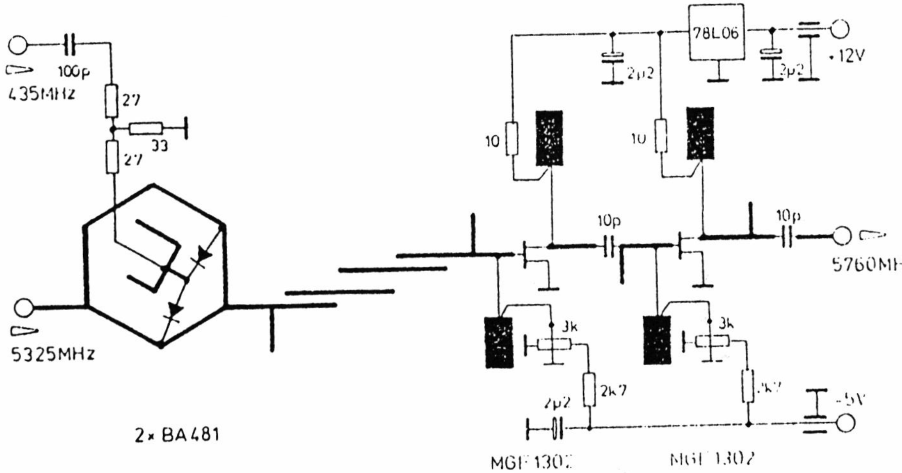



Fig.5.1 **Circuit diagram of DL1RQ 706**

## 4. RECEIVER MODULE DL1RQ 670

### 4.1. Description

The receiver module (fig. 4.1) contains a two-stage preamplifier with transistors MGF1303 and MGF1302. Following is a two-pole stripline filter for selecting the receive frequency of 5670MHz. The mixer stage is realised with an MGF1302 transistor and in this form is already proven in the writer's 10GHz concept (5). The noise measurement of the receive module remains constant at 1.5dB from 1mW oscillator loading to above 10mW.

### 4.2. Construction of the receive module DL1RQ670

Figs. 4.2 to 4.5 show all the details.

### 4.3. Alignment hints and measurements

The trimmer of the 435MHz circuit should be tuned for maximum noise in the 70cm receiver connected (turned about halfway inwards). On the trimmers for the gate voltage the following drain currents are measured:

Preamplifier stage MGF1303: approx. 12mA, corresponds to 1.2V on drain resistor Preamplifier stage MGF1302: approx. 15mA, corresponds to 1.5V on drain resistor Mixer stage MGF1302: approx. 1.5mA, corresponds to 0.15V on drain resistor.

The currents of the first stage and the mixer stage should ideally be optimised on a noise measurement station, but checks on several lab. samples indicated that the techniques described achieved close to ideal noise performance (fig. 4.6).

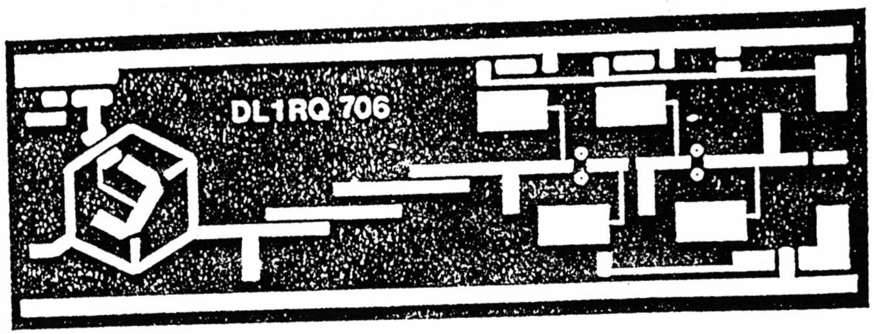



Fig.5.2 **PCB DL1RQ 706**
**108 x 34mm; Material: RT-Duroid D-5870; 0.439mm**

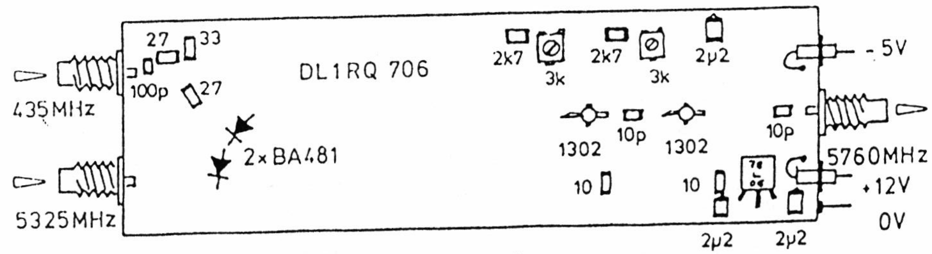



Fig.5.3.1 **Component layout of DL1RQ 706**

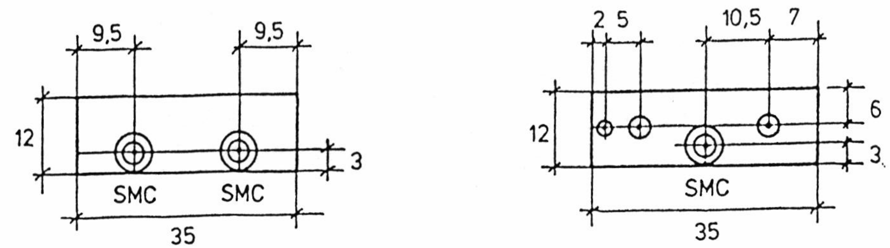



Fig.5.3.2 End plate dimensions for DL1RQ 706

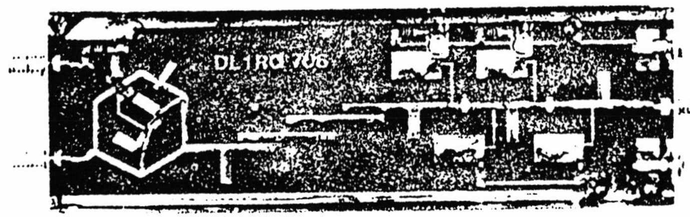



Fig.5.4 The completed module in tinplate enclosure.
Lid measurements 37 x 111 x 13.

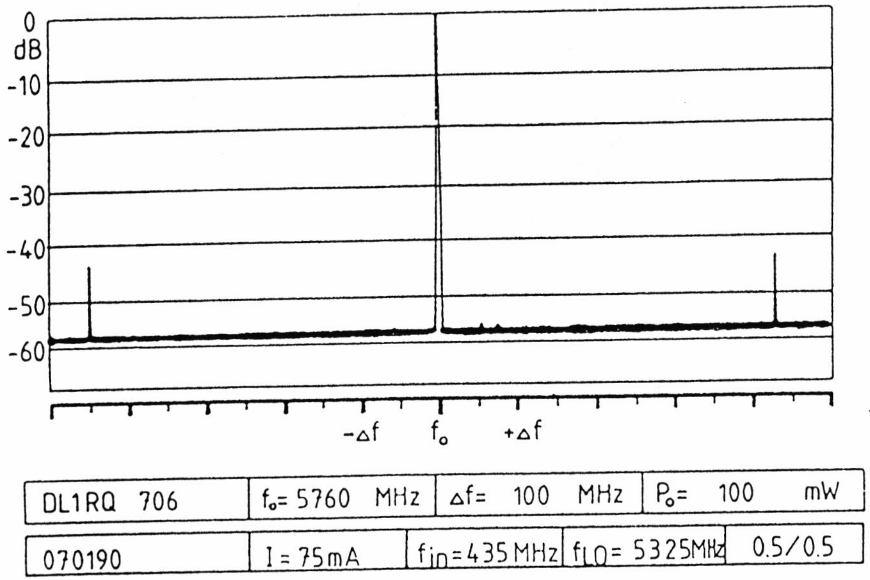



Fig.5.5 Output spectrum of the transmit module DL1RQ 706

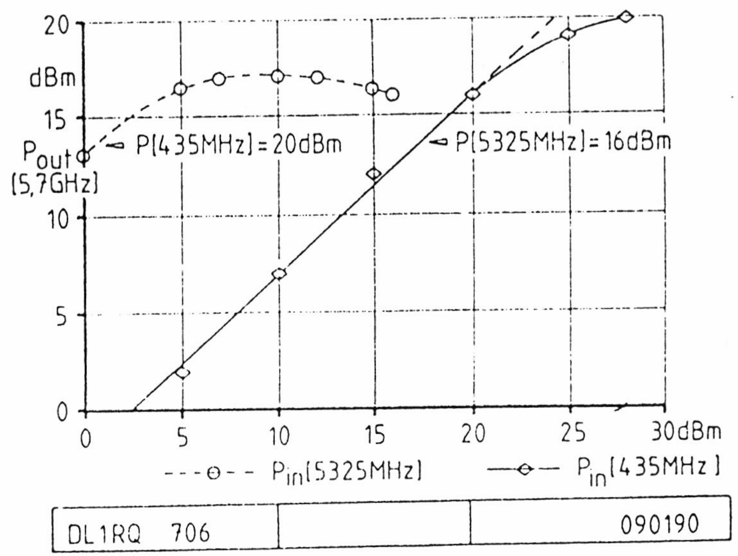



Fig.5.6 Oscillator level and linearity behaviour ofthe transmit module (with 6dB divider from NT6-1)

## 5. TRANSMITTER MODULE DL1RQ 706

### 5.1. Description

The transmitter module (fig. 5.1) contains a push-pull ring-hybrid mixer (rat-race balanced mixer) with two diodes of the type BA481. The 70cm signal which has already been attenuated 6dB in the driver module is reduced once more here by 10dB before it reaches the mixer diodes.

The oscillator power is divided in the ring and fed to the diodes in push-pull. Under ideal conditions there should be good suppression of the oscillator on the output of the mixer.

A stripline filter connected behind this suppresses the image frequency and provides additional suppression of the oscillator. Following this is a two-stage amplifier which raises the mixer signal to a maximum of 100mW.

Figs. 5.2 to 5.4 give the details necessary for construction.

### 5.2. Alignment hints for the transmitter module DL1RQ 706 and measurements

The only alignment is adjusting the quiescent current of the two transistors:

First stage MGF1302: approx. 30-40mW, corresponds to 0.3-0.4V on drain resistor
Second stage MGF1302: approx. 40mA, corresponds to 0.4V on drain resistor.

Figs. 5.5 and 5.6 indicate first-class linearity and adequate suppression of undesired mixer products.

Arising from these measurements it is evident that the mixer presented here is very tolerant of the oscillator levels offered to it.

Peter Vogl, DL1RQ



# A 6cm Transverter using Stripline Technology. Part-2 (conclucion)

As presented at the 1990 VHF-UHF conference in Munich

## 6. FINAL STAGE DL1RQ 661

### 6.1. Description

At the input of the final stage module (Fig.6.1) we find a further stripline filter, which suppresses all unwanted frequencies to below the 60dB level. My own preparatory tests indicate the filter's attenuation in through-pass mode to be approx. 2.5dB.

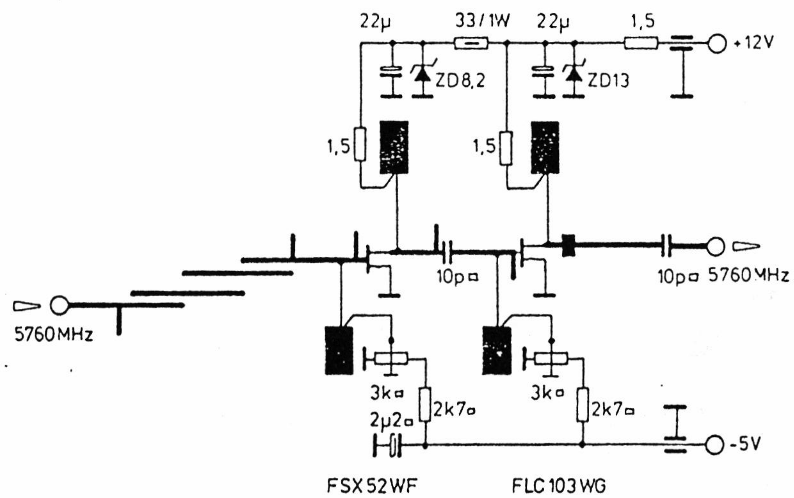



Fig.6.1 Circuit of the final amplifier DL1RQ 661

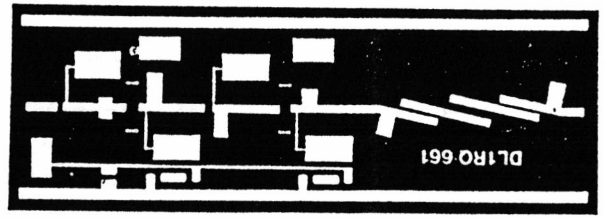



Fig.6.2 PCB layout of final amplifier DL1RQ 661
108 x 34mm; material: RT-Duroid D-5870; 0.439mm

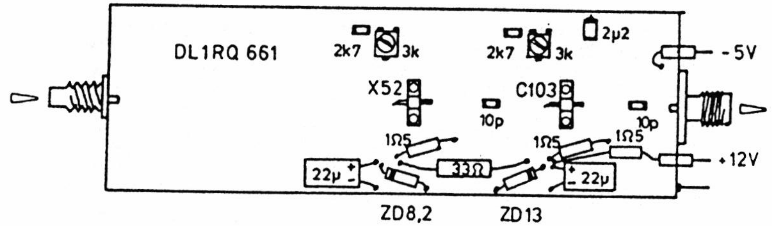



Fig.6.3 Component Layout DL1RQ 661

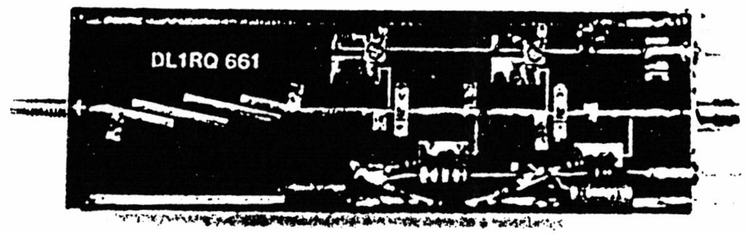



Fig.6.4 Completed module DL1RQ 661 in tinplate enclosure.
Lid measurements 37 x 111 x 13mm

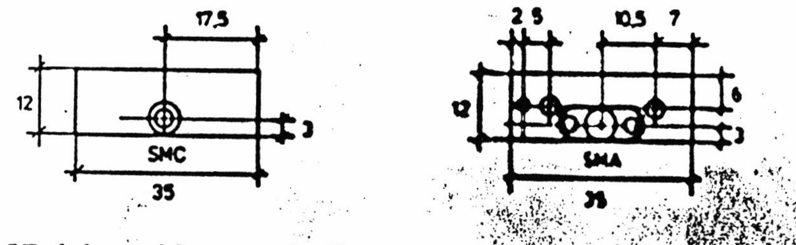



Fig.6.5 End plate and Connector details DL1RQ 661

Following the filter is a two-stage power amplifier. The medium power GaAsFET FSX52WF by Fujitsu is used for the driver stage. The final stage is Fujitsu's power GaAsFET FLC103WG.

The module achieves 17.5dB gain, filter attenuation included. Achievable output power is a maximum of 1.6 Watts.

## 6.2. Constructional information

When mounting the power transistors on the PCB (Fig's.6.2 and 6.3) two points should be noted in particular:

a) as ideal as possible RF transfer from the groundplane of the PCB to the source flange of the transistors;

b) as ideal as possible conduction path for the heat produced.

The author found the following construction convenient. The PCB is soldered with its groundplane onto a 0.5 or 1mm thick piece of copper. Good tinning in advance and plenty of pressure during the soldering operation are necessary to avoid air bubbles.

After soldering together a 2.5mm milling bit is used to mill out two oval slots for the transistors in the Teflon material. The milling must cease at the depth when the copper of the heatsink is visible. After the PCB treated in this way has been soldered "water-tight" in the enclosure all components up to the transistors are put in place. The last-mentioned are finally fastened in the slots with appropriate screws. The transistors were soldered in by the author with a low-temperature iron (140 degrees) under closely controlled temperature conditions.

If the copper surface is vertical this should suffice in operation for cooling. If, however, the module is fixed flat onto a chassis, the gap between it and a matching copper plate must be filled with heatsink compound to make up the difference. In no circumstances should the heatsink plate touch against a cover.

## 6.3 Alignment information and test results

The quiescent currents of the transistors should be set as follows:

Driver stage FSX52WF: approx. 70mA, corresponding to 105mV on the 1.5 ohm drain resistor.

Final stage FLC103WG: approx. 240mA, corresponding to 360mV on the 1.5 ohm drain resistor.

Fig's.6.6, 6.7 and 6.8 are measurement charts of test prototypes.

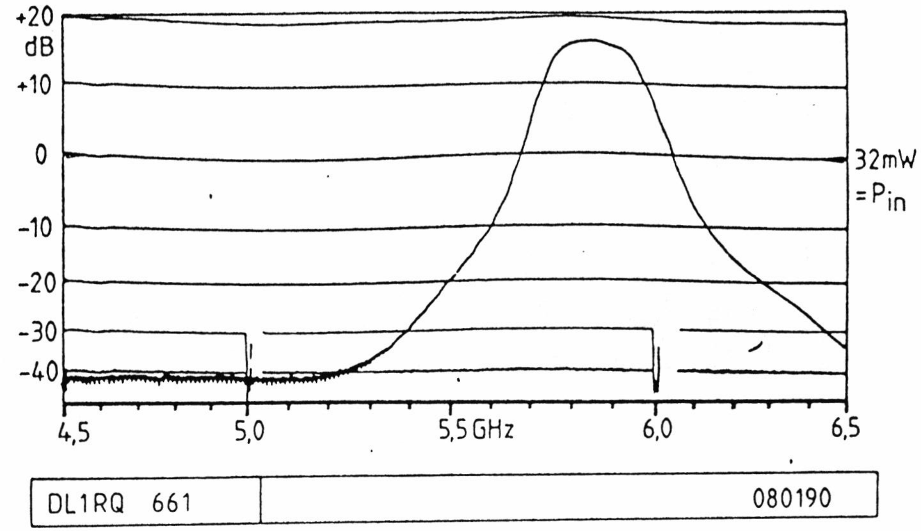



Fig.6.6 Frequency respons e of the final amplifier DL1RQ 661

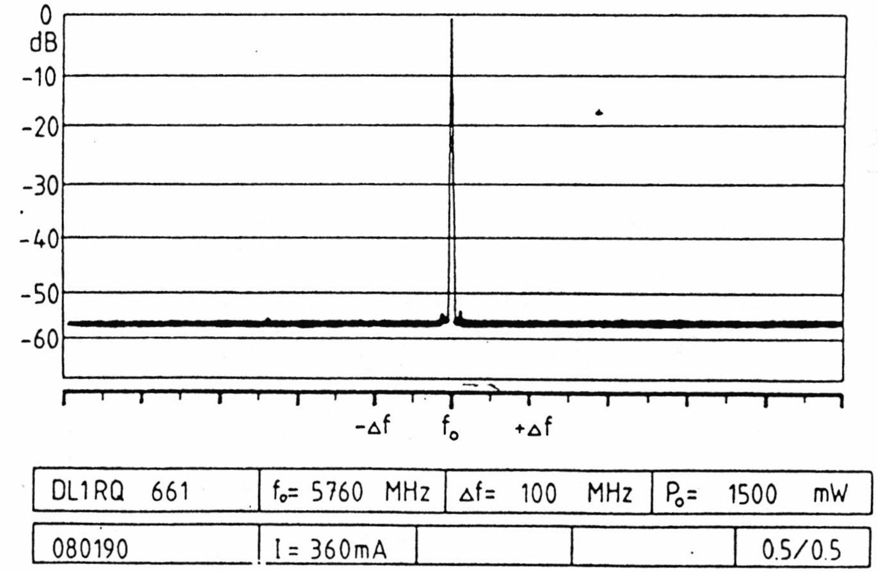



Fig.6.7 Output Spectrum of the final amplifier DL1RQ 661

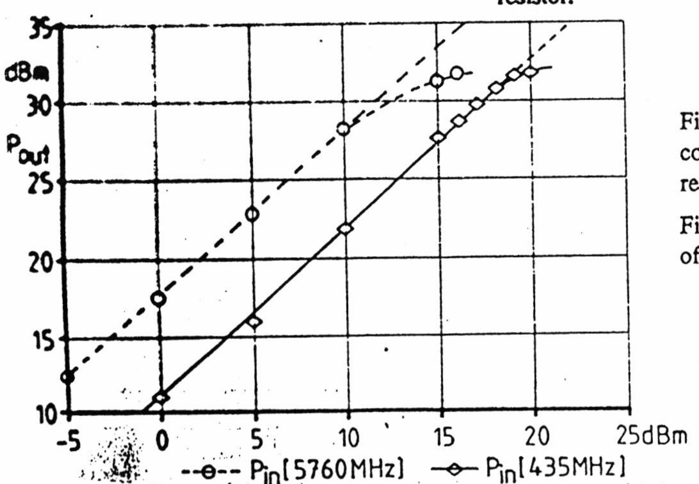
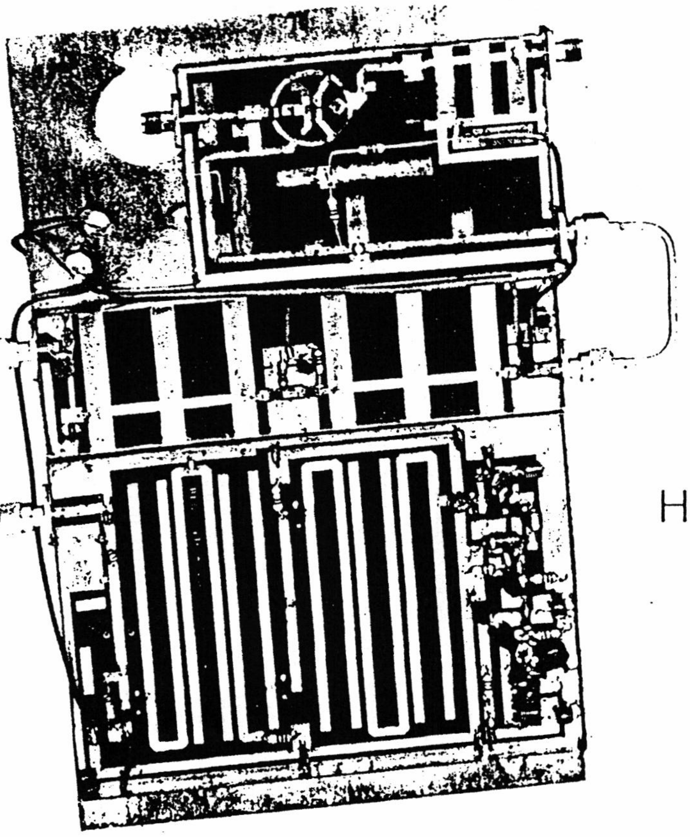



**Fig.6.8 Linearity behaviour of the final amplifiet module DL1RQ 661 (pecked line) and the complete transverter from 435MHz to 5760MHz (solid line)**



# A Single-Board Bilateral 5760-MHz Transverter

Here's a project to get you started on an exotic microwave band with little investment in time and money, and *no* test equipment.

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QST 10-90 C5

The KK7B 5760-MHz transverter (top board), along with its companion LO (bottom two boards) offers a no-tune approach to microwave operation.

Single-board, no-tune transverter designs for the 902, 1296, 2304 and 3456-MHz bands have been published in recent years.[1-4] These boards follow a common theme: They use printed-circuit filters and inexpensive plastic monolithic microwave integrated circuit (MMIC) gain blocks to achieve good performance at low cost. The use of printed-circuit filters and broadband MMICs also eliminates the need for RF alignment or microwave test equipment for proper operation. Low cost and ease of assembly and operation have tempted many amateurs to experiment with the microwave bands.

This article describes a single-board transverter for 5760 MHz. Design and construction are similar to that of the 3.4-GHz transverter described by Jim Davey, WA8NLC, in June 1989 *QST*.[5] I won't repeat many of the design and construction techniques described in Jim's article, so a review of that information will be helpful if you're unfamiliar with the single-board transverter concept.[6]

In addition to the transverter board described here, you'll need a local oscillator (LO), 1296-MHz IF radio and antenna. Information on completing your 5760-MHz station is presented later.

## Design Considerations

In attempting to push the single-board design concept to the 5760-MHz band, two difficulties were encountered:

- The performance of currently available silicon MMICs used in the transmitter and receiver gain stages rapidly deteriorates at this frequency.
- No-tune printed-circuit filters provide insufficient image rejection and LO rejection for use with a 144-MHz IF transceiver.

The problem of image and LO rejection was solved by designing the 5760-MHz board for an IF of 1296 MHz. A single three-section printed-circuit filter with 10% bandwidth provides image and LO rejection of more than 30 dB with a passband insertion loss of less than 1 dB.

The problem of obtaining suitable devices for the gain stages has two attractive solutions. One solution is to simply set the project aside for six months until inexpensive plastic gallium-arsenide (GaAs) MMICs are available. The other solution, presented here, is to integrate everything except the gain stages into a single circuit board, with each port matched to 50 ohms. Then you are free to use any external gain blocks that may become available.

The basic transverter board—without additional gain stages at 5760 MHz—provides a transmit signal of −6 dBm (250 μW) at the 1-dB compression point (−4 dBm saturated) and a receiver noise figure of 9 dB (assuming that the 1296-MHz IF rig has a 2-dB receiver noise figure). This is acceptable performance for line-of-sight contacts over distances of many miles when used with a small dish antenna. Leaving the gain stages off the transverter board solves another problem: You don't need to worry

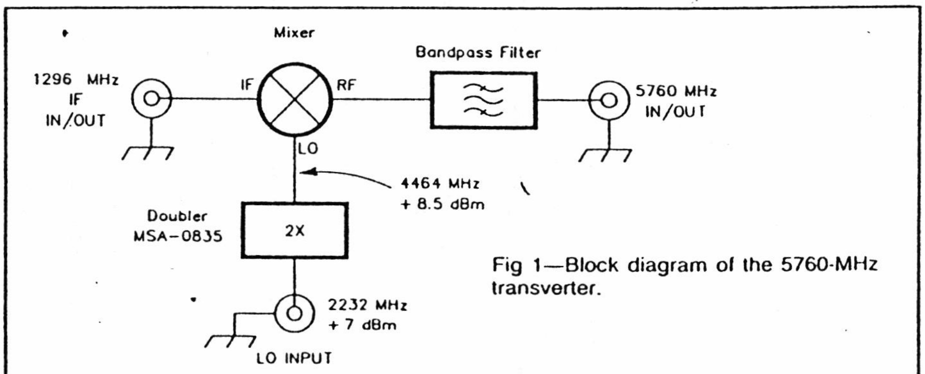


Fig 1—Block diagram of the 5760-MHz transverter.