

A SATELLITE RECEIVING FRONT-END FOR 2400MHZ

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1. Introduction

The communication links with amateur-radio satellites are slowly moving from VHF and UHF to microwave frequencies, especially to the 2400-2450MHz amateur satellite band. The overcrowding of the UHF and especially VHF amateur frequency bands is however not the only reason to move the links to higher frequencies, there are other reasons as well.

In a typical radio link between a satellite and a ground station, the satellite antenna pattern and corresponding antenna gain are defined according to the illuminated area on the Earth's surface, while the ground station antenna is usually limited in size for cost reasons. If one examines the above radio link performance, one finds out that the link path loss between the satellite and the ground station is frequency independent, if the satellite antenna pattern and the ground station antenna size (area) are kept constant.

A radio-link performance is measured as the available signal-to-noise ratio. If one keeps both the TX power and the path loss constant, the received signal level becomes frequency independent and the noise level then drives the frequency choice. The natural noise level is usually expressed as an equivalent noise temperature, since the noise power is linearly proportional to the absolute temperature of the radiating body.

The natural sky noise temperature may exceed one million degrees Kelvin in the HF bands (even without man-made interferences), drops down to a few hundred Kelvins in the 144MHz band and gets below 20K between 1GHz and 10GHz. Above 10GHz the sky noise temperature starts rising towards 300K, since the Earth's atmosphere becomes lossy (opaque) in the millimeter-wave band. From the above it is clear that the frequency range between 1GHz and 10GHz is naturally the most suitable for satellite communications and especially for the downlinks to the ground stations, since the ground station receiving antennas are pointed to the (very) cold sky.

Unfortunately, we radio-amateurs are only allowed to use part of our 23cm band between 1260MHz and 1270MHz for uplinks to amateur satellites. Both uplinks and downlinks are allowed in the next higher 13cm frequency band between 2400MHz and 2450MHz. Therefore the 13cm frequency band has already been used by several amateur satellites. OSCAR-7, UOSAT-OSCAR-9, UOSAT-OSCAR-11, PACSAT-AO-16 and DOVE-OSCAR-17 have beacons in this frequency band while AO-13 and ARSENE have linear transponders. Almost all future amateur satellites will also carry beacons and transponders in the 2.4GHz frequency band.

Past experience has shown that amateur satellite

communications are mainly disturbed by military radar in the 70cm and 23cm bands and by microwave ovens in the 13cm band. The disturbs from microwave ovens are limited both in the duration and in the daytime they are likely to appear. According to my own experience, the disturbs originating from a microwave oven are efficiently neutralized by a conventional SSB-receiver "noise-blanker" circuit.

The only drawback of using microwave frequencies for satellite communications is the increased Doppler shift when compared to VHF or UHF. The quickly-changing Doppler shift is a nuisance especially with low-Earth-orbit satellites while it is tolerable with high-orbit satellites like AO-13 or ARSENE. Using more appropriate types of modulation, like FM in place of SSB or appropriate digital modulations one could easily overcome the drawback of the increased Doppler shift even with low-Earth-orbit satellites.

Unfortunately, most of the amateur equipment actually available for 2.4GHz seems both far too complex and perhaps of obsolete design. The reason for this is probably that 13cm equipment was developed out of existing 23cm equipment, while the same 23cm equipment was developed out of even earlier 70cm equipment and so on. This has lead to poor technical solutions, using inappropriate components, that are both complex and unreliable. Therefore I tried to find a simple, easily reproducible receiver front-end design for 2.4GHz and the results of these experiments are shown in this article.

In this article I am going to describe a complete 2.4GHz receiving front-end as shown on Fig. 1. The 2.4GHz receiver includes a suitable antenna installed on an azimuth & elevation rotator, a low-noise antenna amplifier, a receiving downconverter and a 144MHz (first IF) SSB receiver or transceiver. To minimize any degradation of the noise figure, the LNA is installed directly on the antenna. On the other hand, thanks to the LNA, the receiving downconverter may be installed anywhere between the LNA and the 144MHz receiver. Both the LNA and the downconverter receive the supply voltage through the signal cable to simplify the interconnections.

2. Antenna for 2.4GHz satellite reception

In all satellite communications it is always desirable to have a very sensitive receiver in the ground station. In this way the minimum amount of transmitter power can be used and the same satellite transponder can be used by a large number of stations at the same time. The receiver sensitivity can be improved using a larger antenna, but the latter is more expensive and also requires a more powerful, more accurate and faster antenna rotator.

To receive radio-amateur satellites in the 2.4GHz band, a reasonable choice is an antenna with a diameter between 80cm and 1m. Such an antenna can still be carried and steered by a conventional amateur antenna rotator (like a Kenpro KR5600), the antenna beam is still wide enough for the mechanical inaccuracies of cheap antenna rotators and the rotation speed of these rotators is still fast enough to allow the tracking of low-Earth-orbit satellites. The reception of currently

active satellites (AO-13, PACSAT-1 and DOVE-1) is possible with even smaller antennas and less sensitive receivers. In the future even more powerful 2.4GHz satellite beacons and transponders are expected.

A reflector antenna, like a parabolic dish with an appropriate feed, usually becomes practical when the required antenna diameter exceeds 4 or 5 wavelengths. Slow-wave structure antennas like Yagis or helices are only practical for smaller antenna gains. Considering the problem of correctly feeding a larger number of smaller antennas in the 2.4GHz band and the unwanted side lobes of such an array that degrade the antenna noise temperature, a reflector antenna becomes practical at a diameter of around 50cm.

A parabolic reflector that is accurate enough for 2.4GHz operation can be manufactured at home by screwing or riveting together a number of aluminum sheet petals. On the other hand, parabolic dishes in the range between 80cm and 1m diameter are easily available on the market for 12GHz satellite TV reception. Since future amateur satellites are also going to use higher frequencies up to 10GHz or 24GHz, I recommend a commercially-made dish due to its better surface accuracy.

Parabolic dishes of the same size may look quite different. Besides the dish diameter there is yet another important parameter of a dish: the focal-to-diameter (f/D) ratio. Commercially available dishes for satellite TV usually fall into two categories: symmetrical "deep" dishes with a f/D around 0.4 and unsymmetrical "flat" dishes with an "offset" feed and a f/D around 0.7. For the operation at 2.4GHz I recommend a conventional symmetrical "deep" dish with a $f/D=0.4$, since it is easier to build the corresponding feed for the operation at 2.4GHz.

The polarization of a reflector antenna depends exclusively on the feed used. Since the position of a satellite is continuously changing with respect to a ground station, circular polarization is frequently used in satellite communications. Most amateur satellites use right-hand circular polarization (RHCP). When building a circularly-polarized antenna one should consider that the sense of a circularly polarized wave is inverted on each reflection. A reflector antenna including a single reflector therefore requires a LHCP feed to produce a RHCP wave in the far field!

A simple circularly-polarized feed is shown on Fig. 2. The feed shown on Fig. 2 is suitable for a "deep" dish with a $f/D=0.4$. The feed includes a short two-turn helix with its own reflector, an impedance-matching transformer and a weather-proof box for the LNA to be installed as close to the feed as possible.

The helix itself is made of 8mm wide copper tape, supported by insulating rods (PVC or better teflon). The supporting rods are bolted together with self-locking screws. To avoid disturbing the operation of the helix, the rods should be of the smallest practical cross section and the screws should be as short as possible. The helix operates as an end-fire antenna and in this operating mode the polarization corresponds to the sense of the winding: LHCP is obtained by

winding the helix as a left-hand screw.

The reflector makes one single block together with the weatherproof case for the LNA and is built from 0.5mm thick zinc-plated steel sheet. This material can be easily formed and then soldered together using conventional soft-solder techniques. As first, a disk of 120mm diameter is cut out and the required holes are drilled in the disk. Then the inner (90mm diameter) ring is soldered to the disk and finally the outer (120mm diameter) tube is soldered to the disk. Both rings form a quarter-wavelength deep annular choke used to suppress the side lobes of the feed and thus improve the noise temperature of the antenna.

The impedance of the helix is in the 100-140ohm range. A quarter-wavelength transformer is required to match the feed to 50ohms. Since at 2.4GHz a quarter-wavelength transformer is rather short (around 2cm) and the required coax can not be found easily on the market, the best solution is to make your own transformer as shown on Fig. 2.

In the 2.4GHz frequency range SMA connectors are usually used for low power levels. SMA connectors are usually installed on a semi-rigid cable with a teflon dielectric like UT141. Such a cable uses a copper tube with an outer diameter of around 3.6mm (0.141") as the outer conductor. UT141 has an impedance of 50ohms, semirigid cables with other impedances are not easily found on the market.

One possible way to increase the characteristic impedance of a coaxial cable is to increase the diameter of the outer conductor (shielding). In practice the original UT141 shield is removed over a length of about 30mm and the remaining central conductor and insulation are inserted in a thicker copper tube. A M4 brass nut is screwed on the remaining original shield and then soldered to the new copper tube to work as a transition. Since the original cable dielectric is still too thin, a teflon or polyethylene sleeve should be pulled over the original dielectric to fill the gap between the original dielectric and the new copper tube.

At 2.4GHz a quarter wavelength in teflon or polyethylene is around 22mm. As the new shield a copper tube with an inner diameter of 6mm and an outer diameter of 8mm can be used. The copper tube is cut to 22mm, tinned on both ends and soldered to the rear part of the reflector. As long as the copper tube is still warm of the soldering operation, a piece of RG213 polyethylene dielectric is forced into the copper tube. When the copper tube cools down, any excess polyethylene can be cut away and the central hole is enlarged to 3.5mm with a drill tip. Finally, the prepared UT141 cable end is inserted in the hole in the polyethylene and the M4 brass nut is soldered to the copper tube.

The feed reflector is used as a weatherproof box for the LNA as well. Although the LNA is built in its own little box, the latter is not weatherproof and requires additional protection. To be able to install or handle the LNA, the excess, unmodified UT141 cable should be around 10cm long and fitted with a male angular SMA connector at the other end.

The cover of the outer box is fitted with three self-locking screws. The output cable is routed through the

side (cylindrical) wall of the outer box. Experience has shown that any "waterproof" enclosure is a very efficient moisture collector, so rather than a watertight box I recommend a ventilation hole positioned so that it always looks down, even when the antenna is turned in azimuth and elevation.

3. Two-stage HEMT/GaAsFET LNA for 2.4GHz

The usual amateur procedure to design and build a microwave low-noise preamplifier includes the selection of the active devices based on cost and availability rather than electrical performances and the design of an (un)suitable microstrip circuit based on extensive computations. Unfortunately very little experimentation has been done with the real amplifier circuit and even less experiments have been made to use other construction techniques than microstrip.

Most of the available GaAs FETs and improved devices like HEMTs have been designed for operation at 12GHz in a 50ohm environment in a microstrip circuit of satellite-TV downconverter. The same devices can offer even better performances at lower frequencies. However, if one wants to make use of the improved device performance at 2.4GHz, the devices have to be accurately matched to much higher impedances than 50ohms. The required impedance transformations require very narrow microstrip lines or other more complex microstrip circuits and all of these are very lossy even on high-quality teflon printed-circuit boards.

From the above it is clear that microstrip is not a suitable technology to build a 2.4GHz LNA with conventional GaAsFETs. Further, at lower frequencies, GaAsFET amplifiers may have stability problems and a careful design of the amplifier matching and bias networks is required. Finally, there are a few practical considerations, like overload problems from very strong out-of-band signals, like the ground-station's own uplink transmitter.

The design of the LNA for 2.4GHz is based on a very successful design of a two-stage GaAsFET LNA for L-band (1000MHz - 1700MHz), already published in UKW-Berichte (VHF-Communications) 3/1991, pages 163-169. Two stages provide enough gain to overcome the losses in the cable to the indoor receiving downconverter. Rather than using microstrip technology, the LNA is built using air lines that have much higher characteristic impedances and even lumped components. Suitable line impedances enable very low-loss impedance transformations resulting in a very low overall noise figure of less than 0.4dB when using a CFY65 HEMT in the first stage.

The circuit diagram of the two-stage HEMT/GaAsFET LNA for 2.4GHz is shown on Fig. 3. Matching a GaAsFET to a 50ohm environment can be achieved easily using series inductors at 2.4GHz. Only a few if any additional components are required, like the input capacitor C*. The series inductors are simply short pieces of silver-plated wire or component leads. Such matching circuits have one or two orders of magnitude smaller losses than an equivalent microstrip circuit!

Since the parasitics are mainly capacitances, a GaAsFET

amplifier stage will only oscillate at low frequencies if both the input and the output are terminated into inductive loads with sufficiently high Q_s . In a two-stage amplifier stability can be achieved with a carefully designed interstage matching network. In particular, in a two-stage GaAsFET amplifier any inductors or RF chokes towards ground or supply rails should be avoided in the interstage matching network, since they may represent a high- Q inductor at some frequency, sometimes far away from the designed operating frequency of the amplifier.

The 2.4GHz LNA is built in a small box made of 0.5mm thick brass sheet (see Fig. 4), which is 50mm long, 20mm wide and 15mm high. The box is small enough to prevent resonances below 7GHz. However, HEMTs are able to oscillate at much higher frequencies too. Such oscillations are difficult to detect since the other semiconductors used in the receiver are not able to detect nor to convert these spurious signals to lower frequencies. HEMT oscillations above 10 or 15GHz are therefore only observed as amplifier gain variations when the box is closed. Of course the solution for these oscillations is a piece of microwave absorber (black "antistatic" foam) positioned over the second stage to avoid degrading the noise figure at 2.4GHz.

For operation at 2.4GHz a female SMA connector should be used on the input while a female BNC (UG1094) is still good enough on the output. Of course inexpensive connectors for computer networks are to be strictly avoided! Both connectors are soldered to the LNA box to minimize the parasitics. One should also avoid to use "CB quality" RG58 cable to connect the LNA to the downconverter. Much better choices are RG223 (double shield, silver plated) or RG142 (teflon dielectric) or RG214 (for longer cable runs).

The whole LNA circuit is supported by six leadless ceramic disc capacitors. The nominal value of these capacitors (470pF) is unimportant. It is much more important to use only single-layer ceramic leadless capacitors, of either round (disc) or trapezoidal shape. When soldering these capacitors one should take care to add enough solder between the capacitor metalization and the brass sheet bottom to ensure good solder wetting and prevent the cracking of the brittle capacitor due to case sheet bending. WARNING! Avoid using the so called "chip" or "SMD" multilayer capacitors, that behave in strange way at microwave frequencies.

The two interstage coupling capacitors (1nF) are conventional ceramic disc capacitors with wire leads. The parasitic inductance of the leads of these capacitors is actually forming the inductors L3, L4 and L5. The 33uF tantalum capacitor is used to suppress any spikes on the supply rail that could damage the microwave semiconductors. All of the resistors are conventional 1/8W resistors with wire leads. The resistors marked with a "*" are adjusted for the correct bias of the two FETs.

The quarter wavelength chokes L1 and L6 are made out of 4cm of 0.15mm thick enameled copper wire. The piece of wire is first tinned for about 5mm at each end, the untinned part is then wrapped around an 1mm drill tip to obtain a coil, the actual number of turns being unimportant.

L2 is made out of 0.6mm thick silver-plated copper wire. In practice a wire taken out of the central conductor of a short piece of RG214 cable is used. For the operation at 2.4GHz L2 is an almost straight piece of wire about 15mm long. L3 and L4 are simply around 5mm long 1nF capacitor leads as well as L5. The spacing of L2, L3 and L4 above ground may be adjusted during alignment for best results.

The last components to be installed are the GaAs FETs. After this operation the source bias resistors may be adjusted. The amplifier should be connected to a variable power supply initially set to around 7V. The voltage should be slowly increased as source resistors are being added to keep the voltage across the HEMT around 3V and the voltage across the GaAsFET around 4.5V, since HEMTs operate at lower voltages than GaAsFETs. Of course the resistors should only be soldered with the power off. Typically, the source voltage is around 1V above ground potential after the resistors are adjusted. To avoid parasitic oscillations during DC voltage measurements, it is recommended to terminate both the input and the output to 50ohm RF loads.

The RF alignment should start without the capacitor C*. The amplifier input should be connected to a noise source and the amplifier output to a downconverter and a receiver with a sensitive S-meter or a noise-figure test set. L3, L4 and L5 should be adjusted for the highest gain. Finally one adds C*: a small piece of copper sheet of the size of about 7mmX10mm. C* is adjusted by bending the copper sheet for the best noise figure. L2 may be adjusted for the same reason.

The HEMT may be replaced with a cheaper GaAsFET. The noise figure will be just a few tenths of a dB worse using an inexpensive MGF1302 in the first stage. Any GaAsFET may be used in the second stage: a replacement for the obsolete CFY18 is CFY25 or CFY35 (SMD) or MGF1302. Using different semiconductors the optimum values of the inductors and C* may change a little bit.

GaAsFETs and HEMTs usually have four leads. Out of these four leads, two opposite leads are the source terminals and these are usually easy to find. It is more difficult to find out which one of the two remaining leads is the drain terminal and which one is the gate terminal. Some GaAsFETs have one of the four leads cut at 45 degrees. The latter is the drain terminal with Siemens GaAsFETs and the gate terminal with Mitsubishi GaAsFETs. On the other hand, the CFY65 HEMT has the gate marked with a small red dot. Finally, with the new inexpensive SMD semiconductors I recommend using an analog ohmmeter (in the range ohmX100 or ohmX1000) to find out the terminal allocation!

Satellite communications usually require a full-duplex operation of the ground station, receiving and transmitting at the same time to be able to monitor one's own modulation. Since the described LNA is a wideband amplifier with no frequency-selective components and the antenna feed is also wideband, disturbs from the nearby uplink transmitter are possible. Practical operation via Mode-S transponders of AO-13 and ARSENE has shown that with a very close spacing of the two antennas (1m), the 2.4GHz receiver sensitivity is

degraded at 435MHz transmitter power levels above 5W. Thanks to the very variable power of SSB modulation and the propagation delay one can thus easily monitor his own modulation.

4. Receiving downconverter for 2.4GHz

A large number of receiving downconverters for amateur microwave frequency bands has been described in the literature in the last 10 or 15 years. On the other hand, all of the described downconverters are very similar, just like the technology did not make any progress in the last 15 years. All of the described downconverters have a GaAsFET on the input, a crystal oscillator somewhere between 80 and 110MHz, an "infinitely" long chain of frequency-multiplier stages, some filters made in microstrip technology and finally a mixer with Schottky diodes or yet another GaAsFET.

Such converter designs have several weak points. The first is the crystal oscillator. Above 80MHz the fifth overtone resonance has to be used although the crystal itself is thinner than 0.1mm. Thin crystals suffer a faster aging and break quickly from a mechanical shock. On the other hand, the operation of a crystal oscillator at the 5th overtone resonance is usually far from reliable.

Frequency-multiplier stages are even worse. Using vacuum tubes or bipolar transistors one can build simple and reliable multiplier stages up to perhaps 500MHz. Above this frequency harmonics are mainly generated by the collector-base junction varactor in a bipolar transistor. The operation of a varactor or step-recovery-diode multiplier requires much higher signal levels, close to the destruction levels for the transistors used as frequency multipliers. The operation of such multiplier stages is of course quite unreliable, the transistors may fail or just degrade their performances after some time and very seldom one is able to reproduce the circuit and achieve the same electrical performances as mentioned by the author in his original article.

Finally, most downconverter designs use high-Q circuits, tuned with unsuitable trimmers, both in the multiplier chain and in the RF stages. Tuning such a downconverter requires a large amount of both patience and instrumentation. Unfortunately, the tuning has to be performed once again after any shocks applied to the downconverter or even after wide temperature excursions.

Of course it makes sense to look at alternate designs. Together with the introduction of satellite TV, inexpensive single-chip PLL synthesizers appeared on the market for frequencies up to 1.5GHz or even 2.5GHz. One of the first devices was the Siemens SDA3202, now there are many similar chips produced by different manufacturers. The SDA3202 was successfully used by amateurs in wideband transmitters and receivers, like FM ATV on the 23cm band. Unfortunately these single-chip synthesizers are not suitable for SSB converters or transverters, since the division modulus are too large and the PLL gain is too low. Simply speaking, such a PLL is too slow to stabilize quick random variations of the VCO frequency

for SSB or CW narrowband operation.

A synthesizer for a SSB microwave converter therefore has to be built from a number of digital integrated circuits, to avoid the unreliable frequency-multiplier chain. The block diagram of the receiving converter for 2.4GHz was already shown on Fig. 1. The detailed circuit diagram is shown on Fig. 5 and Fig. 6: the RF part on Fig. 5 and the PLL logic on Fig. 6.

The receiving converter includes a RF amplifier stage with a CFY19 GaAsFET, a harmonic mixer with two schottky diodes BA481 and an IF amplifier stage with a BFR90. The synthesizer produces a signal at 1128MHz, since the harmonic mixer generates itself the second harmonic at 2256MHz as required for the downconversion from 2400MHz to 144MHz. The synthesizer includes a VCO operating at 1128MHz (BFR91 and BB105), a buffer stage (BFR91), a fast ECL prescaler U664, some additional dividers (74HC393), a 9MHz crystal oscillator and a frequency/phase comparator (74HC74 and 74HC00).

The input RF amplifier with the GaAsFET CFY19 is required to overcome the losses in the 2.4GHz bandpass filter (L4, L5, L6 and L7) and mask the high noise figure of the harmonic mixer. The latter is using two antiparallel diodes to convert the incoming RF signal with the second harmonic of the local oscillator. Since the input RF frequency (2.4GHz) and the local oscillator frequency (1128MHz) are in an approximate ratio of 2:1, the mixer circuit results very simple and requires just two resonant microstrips (L8 and L9). The latter are dimensioned to represent a quarter wavelength for the VCO frequency and a half wavelength for the input RF signal.

Harmonic mixers require very good mixer diodes, since the same diodes provide both the frequency doubling and input signal conversion. Common Schottky diodes like BA481, HP2900 or similar are not very suitable for this application and the result is a rather high noise figure of the mixer alone in the range between 12 and 15dB. Including the losses in the 2.4GHz bandpass and the gain and noise of the CFY19, the overall downconverter noise figure is in the 6 to 7dB range.

Much better results can be obtained using new components like the Ku-band double Schottky diode BAT14-099 (SMD plastic package): the mixer noise figure drops down to only 6-7dB and the overall downconverter noise figure down to only 3dB!

The mixer is followed immediately by an IF amplifier stage (BFR90), so that even a "deaf" 144MHz IF receiver can no longer spoil the downconverter noise figure. The IF amplifier does not include any tuned circuits, since no selectivity is required. Since there are no multiplier stages, there are also no spurious frequencies that could disturb the operation of the IF amplifier or even bring the IF stage into saturation. There is just a low-pass filter on the input (L19 and L20) to reject the VCO signal at 1128MHz.

The downconverter receives the required supply voltage through its output IF coaxial connector. On the other hand, the downconverter supplies the LNA with the same supply voltage through its input RF coaxial connector. All this simplifies the installation, since there are no additional supply cables.

A SSB synthesizer requires a VCO with a top phase-noise performance. The VCO includes an amplifier stage (BFR91) and an interdigital bandpass filter (L13, L14 and L15) feedback. The VCO frequency is adjusted by tuning the central finger (L14) of the interdigital filter with the varicap diode BB105. The latter can shift the VCO frequency by about 10% of the central operating frequency. The real tuning range is further restricted by the control voltage between 0 and 5V.

The operation of the described VCO is so stable that the VCO supply voltage does not need to be stabilized. The VCO is followed by a buffer stage with another BFR91. The buffered VCO signal is then fed to the harmonic mixer through a coupler and a small amount of the VCO signal power (about 1%) is fed to the input of the ECL prescaler U664. The coupler also allows a wideband termination (56ohm resistor) for the prescaler input.

The PLL divider should operate reliably up to the maximum VCO frequency of about 1200MHz. A fast ECL divider is therefore required for the first divider stages. The integrated circuit U664 includes an input signal preamplifier and chain of 6 flip-flops with no feedback, so that the overall division modulo is equal to 64. The output frequency of the U664 prescaler is around 17.6MHz. This frequency is still too high for the frequency/phase comparator and further division is necessary.

In a PLL synthesizer for a SSB converter the smallest possible division ratios and the highest comparison frequency are desired to speed-up the PLL and reduce the phase noise of the VCO. The main limitation is the frequency/phase comparator. The most popular circuit is the charge-pump comparator and the latter is limited to about one tenth of the maximum clock frequency of the logic family used. Since 74HCxx family logic circuits have a maximum clock frequency of about 40MHz, the highest practical comparison frequency is around 4MHz.

The prescaler output signal is therefore further divided by 8 by one half of the 74HC393 while the second half of the same chip is used to divide the frequency of the reference crystal oscillator. The comparison frequency is around 2.2MHz to operate the circuit with a considerable safety margin. The output of the frequency/phase comparator is a charge-pump circuit with two fast Schottky diodes BAT47, allowing an output voltage swing from about 0.5V up to 4.5V.

Most charge-pump frequency/phase comparators have a hysteresis or backlash problem. The latter may be a major source of phase noise in a PLL frequency synthesizer. In the described circuit, the backlash problem is solved by having the charge-pump circuit (diodes BAT47) faster than the feedback network (1/4 74HC00, pins 8,9,10) to the two D-flip-flops. The response of such a comparator is still nonlinear around the locking point, but the backlash problem is eliminated, resulting in an excellent synthesizer frequency stability.

The receiving downconverter for 2.4GHz is also practically built as two separate modules. The RF part is built on a two-sided printed-circuit board in microstrip technology, of the size of 120mmX40mm, as shown on Fig. 7.

The PLL logic is built on a single-sided printed-circuit board, of the size of 80mmX40mm, as shown on Fig. 8. The location of the components in the RF part is shown on Fig. 9, while the location of the components in the PLL logic is shown on Fig. 10.

Both modules need to be well shielded and separated one from another, so that the spurious signals generated in the PLL logic can not get into the low level RF stages, especially into the 144MHz IF amplifier. Each module is therefore soldered into its own metal enclosure. The latter includes a 30mm high frame made from 0.5mm thick brass sheet and the printed-circuit board is soldered into this frame 10mm above the bottom, as shown on Fig. 11. The printed-circuit board ground plane should be soldered along all four sides to form a "watertight" enclosure for RF signals. The cover can be made of 0.5mm thick aluminum sheet and screwed on the frame with four self-locking screws on the marked places.

In the RF module, the bottom of the shielded enclosure is represented by the microstrip ground plane. The form of the RF module is intentionally made elongated so that the shielded enclosure resonances are shifted well above the operating frequencies of the circuit and microwave absorbers are usually not needed. The supply voltages and other low frequency connections (VCO control voltage) are fed through 220pF (or more) feed-through capacitors, soldered on the microstrip ground plane and in the walls of the enclosure. The 33uF tantalum capacitors and the ferrite-bead chokes FP are also located under the microstrip printed circuit.

Some components are installed in 6mm diameter holes in the RF printed-circuit board: the transistors BFR90 and BFR91, the varicap diode BB105, the 56ohm resistor and the two GaAsFET source-bypass, leadless-disc 470pF capacitors. All of these 6mm holes are then covered with a piece of thin copper sheet of about 7mmX7mm soldered to the ground plane of the microstrip circuit.

The choice of the GaAsFET is not critical, since the noise figure is mainly spoiled by the mixer and the GaAsFET is correspondingly biased and matched for the highest gain and not for the best noise figure of the RF amplifier alone. In the three prototypes built CFY18, CFY19 and CFY30 were tested with similar results except for some changes in the input matching components L3 and C*. For CFY19 and CFY30 L3 has one turn of silver-plated copper wire of 0.6mm diameter wound on an internal diameter of 2mm. C* is a piece of thin copper sheet of about 10mmX7mm, soldered to the hot end of L2. L1 is a quarter-wavelength choke for 2.4GHz, built in the same way as the two chokes in the LNA.

Schottky diodes HP2900, BA481 and BAT14-099 were tried in the mixer. Of course the BAT14-099 provides a considerably better noise figure and lower conversion loss, besides simpler fine-tuning of the mixer circuit. Fortunately these new devices are not expensive at all and are slowly becoming available also in small quantities on the radio-amateur market. L19 is also a quarter-wavelength choke, but it has to operate both at 2.4GHz and at 1128MHz, so its length has to be chosen somewhere between these two wavelengths.

Suitable ECL prescalers are available from many manufacturers, since these devices were used in TV tuners a few years ago. The Telefunken U664 seems to be the easiest to get as a TV replacement part. The circuit also works with the Siemens SDA2101, which has a lower frequency limit and a less sensitive input. Unfortunately the SDA2211 could not be tested since unavailable, although this circuit probably has the best RF performance (1.8GHz) and lower power consumption. All of the mentioned ECL prescalers are packaged in the same 8-pin package and are pin compatible.

The PLL logic is built in a similar metal enclosure, although because of a single-sided board the enclosure does not have a shielded bottom. The feed-through capacitors are only built in the enclosure walls. The 7805 regulator (TO220 version) is bolted to the brass frame for heat sinking. The only remaining critical component are the charge-pump diodes. These must be Schottky diodes for speed reasons, to avoid backlash problems in the frequency/phase comparator. General purpose switching Schottky diodes like BAT47, HSCH1001 or similar devices are suitable for this purpose.

The checkout and alignment of the downconverter should start with the selection of the reference crystal. Since most of the amateur-radio satellites operate in the subband 2400.000MHz to 2402.000MHz, it makes sense to convert this frequency band to 144-146MHz using a local oscillator at 2256MHz. The latter frequency requires a crystal for 8812.5kHz in the PLL. Only ARSENE operates at a higher frequency around 2446.5MHz. To convert its signal to 144.5MHz, a local oscillator at 2302MHz is required. The latter needs a crystal for 8992.19kHz in the PLL.

Both required crystals can be found among standard Citizen-Band radio crystals, of course operated on their fundamental resonance. For 8812.5kHz one can use the CB channel #1 RX crystal (26.510MHz) that resonates around 8837kHz in its fundamental mode. The crystal frequency is still about 24kHz too high, but fortunately the fundamental resonance of a crystal can be easily shifted downwards with a series inductor (L^*). Due to the relatively large frequency shift required, the value of L^* is also quite large, up to 30uH, and some experimenting is necessary in any case.

On the other hand, it is much easier to obtain 8992.19kHz from a CB channel #2 TX crystal (26.975kHz), that usually requires just minor frequency adjustments either with a series inductor L^* or with a series trimmer capacitor, but never both at the same time! When using inexpensive CB crystals one should always check that the oscillator operates reliably on a single crystal mode. For better frequency stability the CB crystals should be replaced with custom-made crystals, specified for a series resonance at the desired frequency and fine-tuned with a small series inductor L^* .

In the next step the complete PLL should be assembled and the VCO should be checked out. The operation of the VCO can be observed as a DC voltage decrease on both BFR91 bases, of course measured through suitable RF chokes. The VCO should be fine-tuned by adjusting the length of L14. Usually L14 has to be made a little longer by adding a small piece of copper

foil on its hot end (towards the transistor BFR91). L14 should be adjusted to obtain a PLL control voltage of 2.5V in the locked state of the PLL. When the PLL achieves lock, the pulses on LOCK T.P. become very narrow and the DC voltage component approaches zero. Even this voltage should only be checked through a RF choke, since conventional AVO meters usually behave in a crazy way with a RF of just 2.2MHz applied to their terminals!

Once the PLL operates reliably, the GaAsFET is soldered in the circuit and its source bias resistors are adjusted just like in the LNA. Then a noise source should be connected to the input of the downconverter and a receiver with a sensitive S-meter to the output. All of the following adjustments are simply made for the highest gain or the strongest signal on the S-meter. The first thing to check is whether the receiver is able to hear the noise source at all! Next the RF input match should be optimized by changing C*: moving and bending the small piece of copper sheet.

When the RF input match is optimized one can start working on the harmonic mixer. In the latter L8 and L9 can be optimized by adding small pieces of copper foil (about 7mmX7mm) in different places along these two resonant lines. The optimum position and size of these pieces of copper foil depend very much on the diodes used in the mixer. Suitable diodes like BAT14-099 usually require little tuning, but on the other hand it is much more difficult to obtain a decent noise figure out of BA481 or HP2900 diodes.

Finally the 2.4GHz bandpass may be optimized by adjusting the length of L5 and L6. These two resonators are however already adjusted to the correct length and hardly require any additional adjustments, at least if their cold ends are correctly grounded using a 1mm diameter copper wire to interconnect both sides of the microstrip board. If adjusting L5 or L6, proceed very carefully and never add or remove more than 0.5mm at a time! In any case, these microstrip resonators do not have a particularly high Q, especially since the board material is lossy, 1.6mm thick glassfiber-epoxy.

5. Assembly of a 2.4GHz receiving ground station

At the end of this article I find it necessary to spend a few words about the assembly of a ground station from the components already described. The single components are not necessarily built in the same order as described in this article, nor all of them may not be required in a particular groundstation. Using better components in the downconverter one could avoid the LNA or alternatively use a much smaller antenna, like a long helix.

Both the LNA and the receiving downconverter are designed to get the supply voltage through the same RF cables. The LNA receives its supply voltage through its output connector while the downconverter has the supply voltage on both RF connectors. In this way all of the system components are connected with a single RF cable and this is usually very practical when making a practical installation on the roof or on the antenna tower.

As a 144MHz IF receiver radio amateurs usually use a 2m SSB transceiver. SSB transceivers usually do not have available the supply voltage on the RF connector. In addition it is very easy to destroy a receiving downconverter connected to a transceiver, if the latter is accidentally switched to transmission. Therefore I recommend a protected supply circuit for the downconverter as shown on Fig. 12. This simple circuit can be built in a small metal box and the few components soldered directly among the three connectors. The circuit introduces some RF signal loss because of the 33ohm resistor, but on the other hand this resistor protects both the downconverter and the transmitter if the latter is accidentally enabled. The 33ohm resistor is rated at 1/2W power dissipation and this should be enough for short overloads, although in both my prototypes these resistors already changed their color!

If the LNA is omitted, then the downconverter should be installed as close as possible to the antenna and the supply voltage should be removed from the downconverter input, to avoid shorts in the antenna. On the other hand, when using the LNA, the downconverter can be installed indoors. The LNA gain is sufficient to cover a loss of about 10dB in the cable to the downconverter corresponding to about 20m of good quality cable like RG214. An indoor downconverter is less subjected to temperature variations and its frequency is more stable as if it were installed outdoor, directly on the antenna.

The whole receiving system as described in this article provides a system noise temperature in the range between 60K and 70K. In other words, if the antenna is turned towards the warm ground, the noise level should increase by 6 or 7dB. When the antenna is turned towards the cold sky, about half of the noise is provided by the antenna itself due to the sky temperature and the dish feed spillover collecting warm-ground noise. The remaining half of the noise power, 30-35K, is the receiver noise corresponding to an overall noise figure between 0.4 and 0.5dB. It is believed that this figure represents today the amateur technology limit, but anyway not much can be gained going beyond due to the natural noise collected by the antenna.

Using a 90cm (3 foot) diameter parabolic dish the described receiving system provides more than 10dB of margin over what is required to receive the currently active satellites in the 2.4GHz band: AO-16 and DOVE beacons and the AO-13 linear transponder. This margin is very welcome when the satellites are in an unfavorable position on the sky with the antennas oriented away from us. On the other hand this means that the above-mentioned satellites can be received with smaller antennas and even without a LNA at the same time!

Weaker signals were only received from UO-11 and ARSENE. UO-11 probably has a defective transmitter that has only been switched on on very rare occasions while ARSENE has a low gain S-band antenna. Although the ARSENE signal was 10-15dB below the AO-13 signal, the described receiver with a 90cm dish was sufficient for normal SSB contacts, to listen to my own SSB echo with just about 500W EIRP on 70cm and even to hear the ARSENE transponder noise, when the kilowatt stations were

off, of course...

At the end, lets hope that P3D and its high-power S-band transmitter get safely to the desired orbit!

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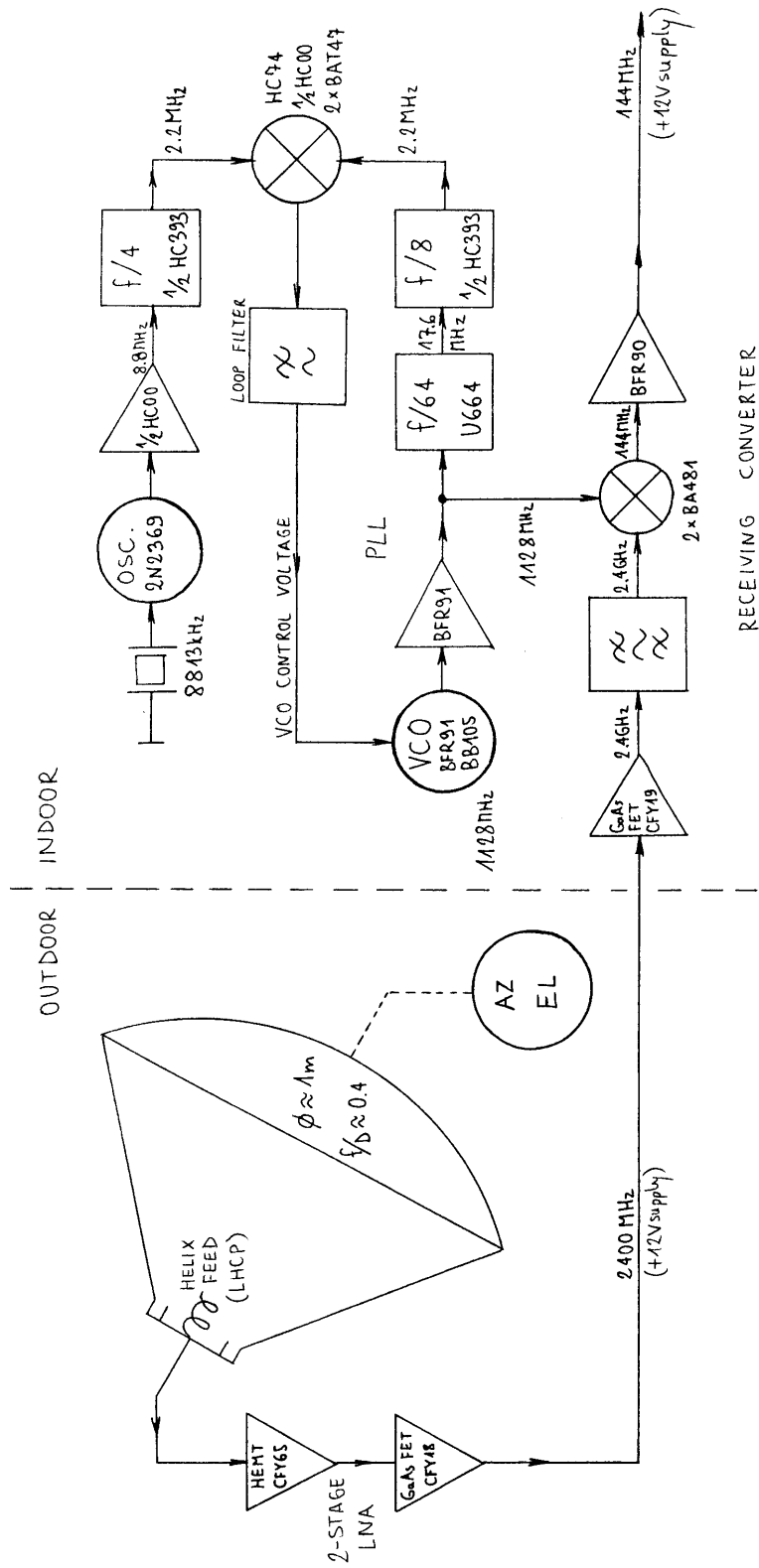


Fig. 1. - 2.4 GHz front - end block diagram.

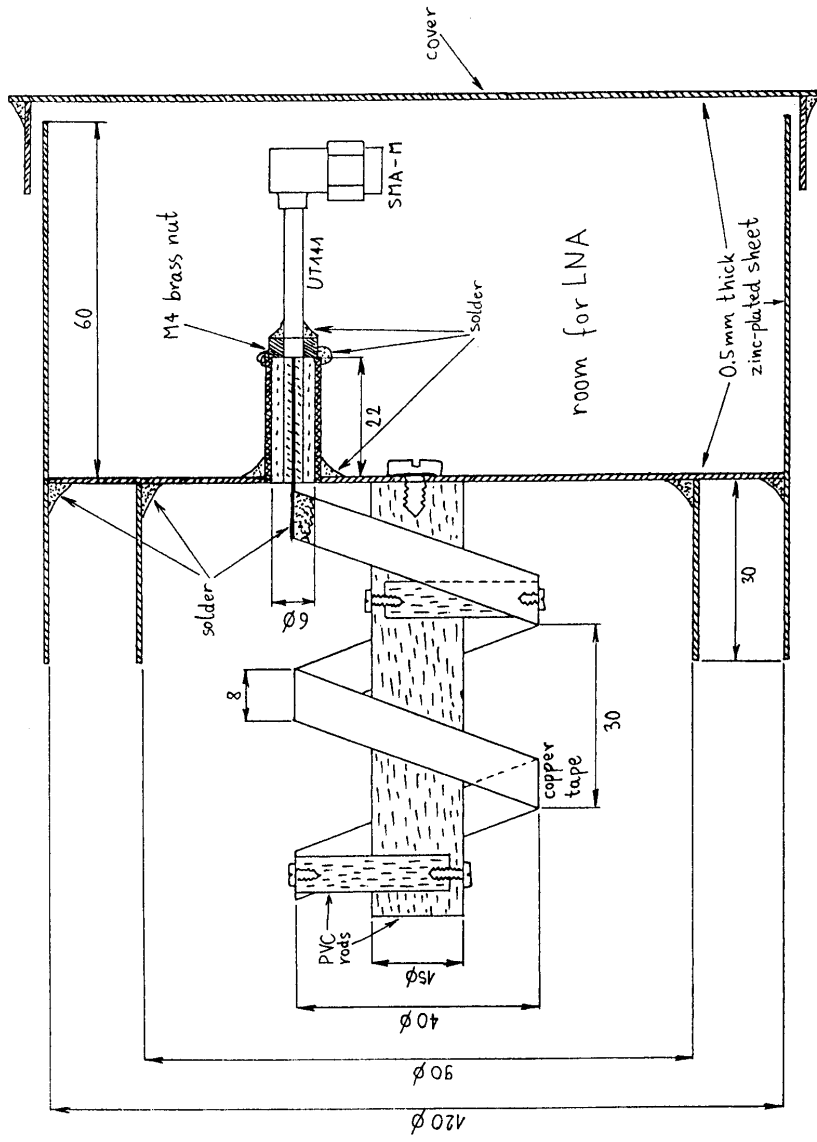


Fig. 2. - Circularly-polarized feed for 2.4GHz (LHCP), for a $f/D = 0.4$ dish.

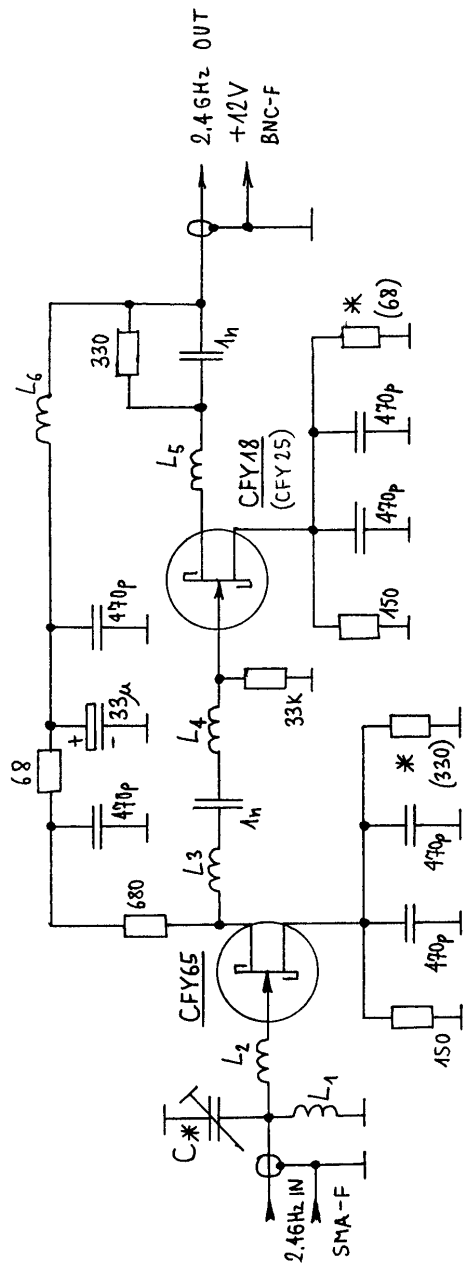


Fig. 3. - Two-stage HEMT/GaAs FET LNA for 2.4 GHz.

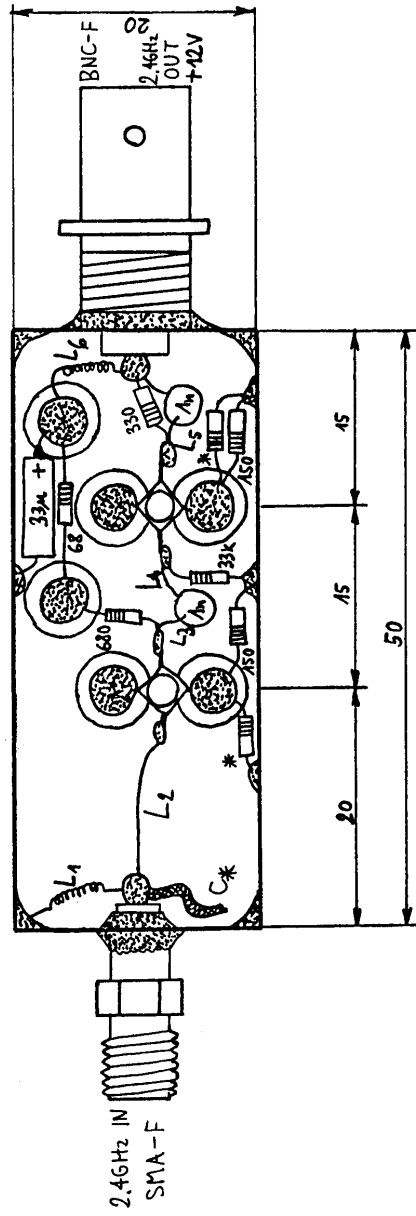


Fig. 4. - 2.4GHz LNA construction and parts layout.

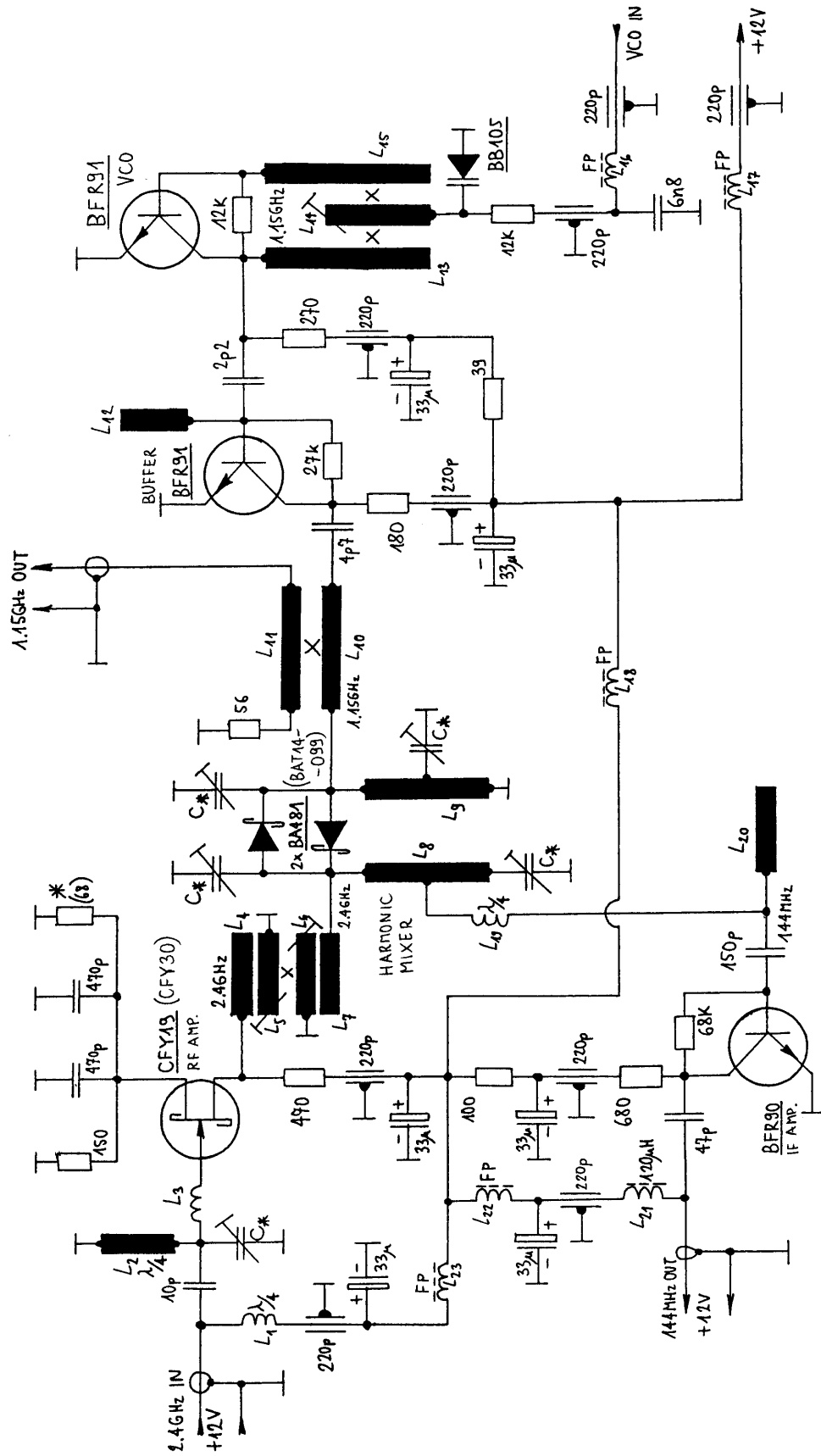


Fig. 5.- Receiving converter for 2.4GHz - RF part.

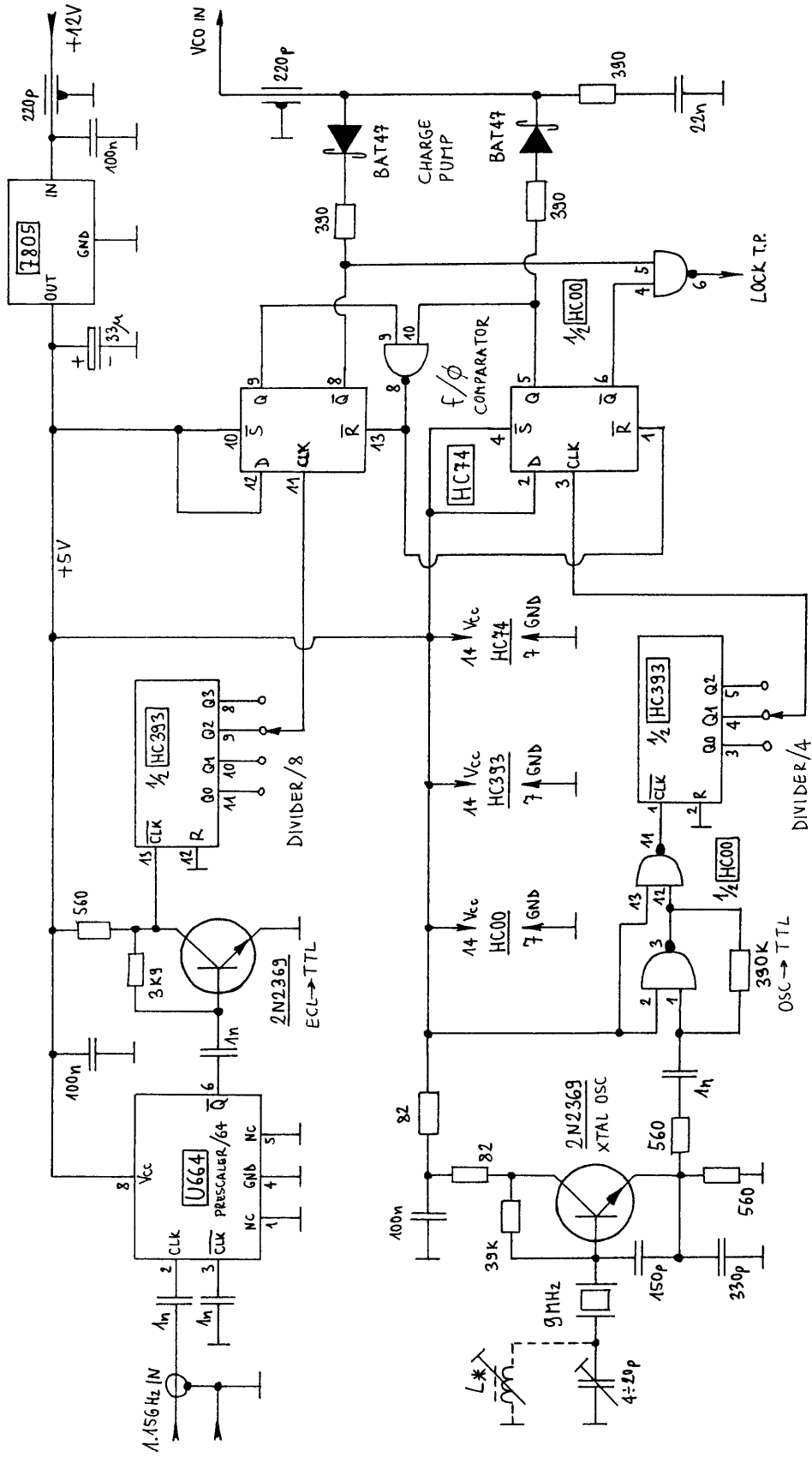


Fig. 6.- Receiving converter for 2.46GHz - PLL logic.

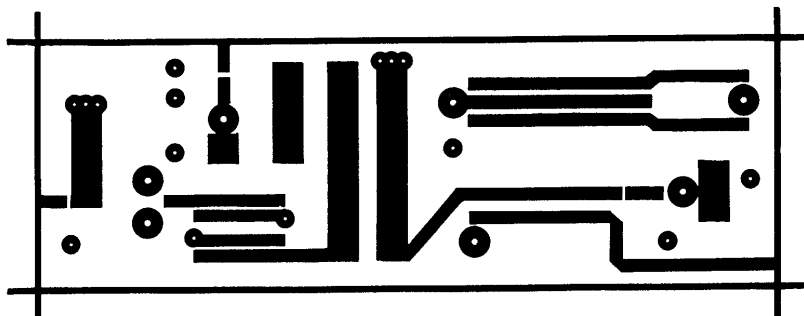


Fig. 7.- Printed-circuit board for the 2.4 GHz converter, RF part, double-sided 1.6 mm thick glassfiber-epoxy, top view, the other side is not etched!



Fig. 8. - Printed-circuit board for the 2.4 GHz converter, PLL logic, single-sided, bottom view.

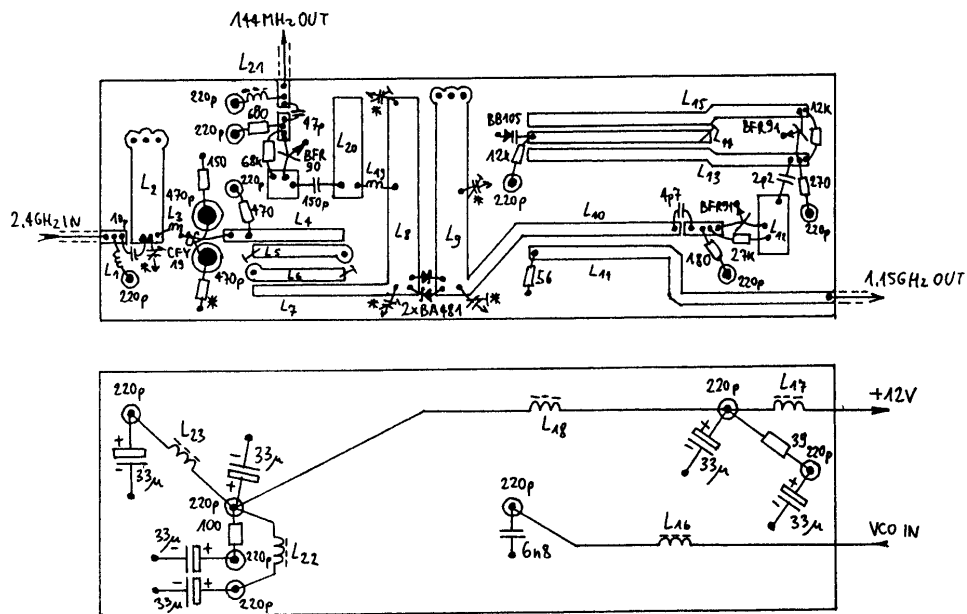


Fig. 9 - 2.4 GHz converter, RF part component layout.

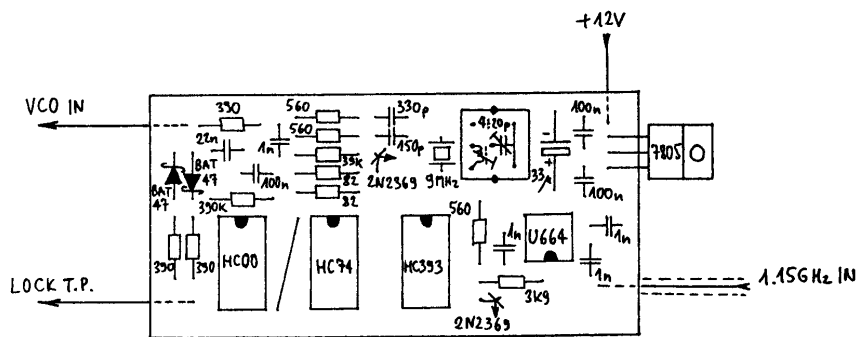


Fig. 10 - 2.4 GHz converter, PLL logic component layout.

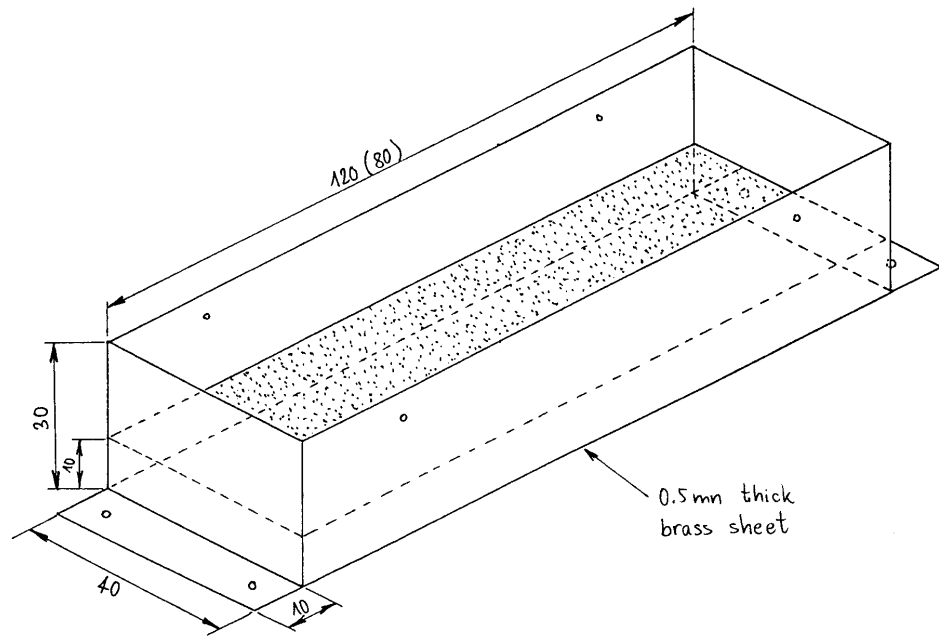


Fig. 11 - 2.4GHz converter module enclosure.

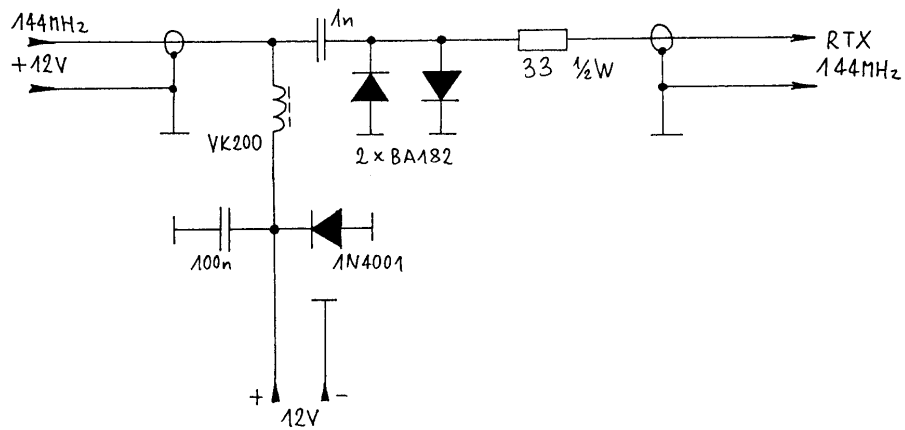


Fig. 12 - Protected supply circuit.