Microstrip transverters for 23 and 13 cm

1. Introduction

The microstrip technology is certainly well known to anybody working in the RF or microwave field since mostof both professional and amateur equipment is at least partially built using this technology. Due to its widespread use the name "microstrip" itself does not really tell much about the construction of a certain piece of equipment, ranging from simple transmission lines and matching transformers built in microstrip and combined with components manufactured in other technologies (cavity, coaxial, waveguide) to complete circuits including single resonators, complex filters, power splitters, couplers, chokes, matching transformers, tuning stubs, small capacitors and even antennas! In the production of professional equipment the microstrip technology brings a significant reduction of the manufacturing time and consequently a reduction of the overall cost. In addition to this a number of theoretical tools, measuring instruments and computer programs were developed to reduce both the design time and the production line tuning of the single circuits.

Unfortunately most amateurs do not have access to the expensive professional instrumentation or computer-aided-design (CAD) tools. On the other hand, many very common design problems can not be easily solved by theoretical tools or CAD programs like (real world) lossy laminates having an anisotropic dielectric constant ε or semiconductor devices operating in their nonlinear region (mixers, varactor and transistor multipliers, power amplifiers). Of course there are other even less predictable facts like the influence of the various shields and/or the resonances of the metal case actually containing the microstrip circuit. Practical experiments are thus necessary in any case, even with the best CAD program.

Fortunately we amateurs only have moderate requirements like narrowband operation or gain tolerances. Since most of the components used in our designs are usually not sufficiently characterized at microwave frequencies in their data sheets, like cheap plastic case transistors or conventional glassfiber-epoxy laminate, the logical design procedure is to roughly calculate or estimate the circuit parameters and then practically optimize the circuit performance.

Microstrip circuits are usually built as a double sided printed circuit board. The transmission lines and other microstrip components are all etched on one side of the PCB. The other side is not etched since it acts as a ground plane for the transmission lines and other components. Since the distance between the transmission lines and the ground plane (thickness of the laminate) and the widths of the lines are small compared to the wavelength and to other circuit dimensions, it is assumed that most of the electric and magnetic field is constrained to the close proximity of the transmission line. Since the magnetic and electric field intensities decrease rapidly with distance, microstrip circuits usually do not require any shields or additional ground planes. Additional metal planes or even closed metal boxes generally only have a very small influence on the circuit. Unfortunately closed metal boxes have self resonances with very high Q-factors. At these particular frequencies they can introduce considerable unwanted couplings even between phisically distant microstrip transmission lines. There are many efficient solutions for such problems and some of them will be shown later in this article. Actually it is necessary to understand that improper shielding may even introduce new problems at microwave frequencies!

Selective circuits at microwave frequencies may be implemented using $\lambda/4$ or $\lambda/2$ microstrip resonators as stand alone resonators or arranged in more or less complex filters. In the amateur literature two basically different designs are described. The first uses full size fixed tuned resonators and therefore requires very close tolerances of the PCB laminate and the circuit pattern etched onto it. Due to the manufacturing tolerances the loaded Q-factor of the single resonators has to be kept low and a large number of resonators are required to obtain the desired spurious frequencies rejection. A large number of resonators in series calls for a low loss, expensive teflon laminate as substrate material. Practical experimenting is difficult and costly.

The other design approach employs significantly shorted λ/4 resonators (acting practically as coils) by capacitive trimmers. Suitable trimmers allow a very broad tuning range. Unfortunately this also means that the circuit may be easily tuned on the wrong mixer sideband, harmonic or other spurious frequency! Since the trimmer does not act only as a tuning element but it also provides a significant part of the capacity required in the circuit, the tuning may become very sharp and critical and the mechanical stability (post tuning drift) may not be sufficient. This is especially true when using cheap trimmers not originally designed for microwave frequencies close to their minimum capacity. Suitable microwave trimmers are at least an order of magnitude more expensive and are not easily available. In any case the circuit contains an unpredictable variable - the parasitic reactances of the trimmers used - and this makes the duplication in amateur conditions considerably more difficult.

In the transverters described in this article a different solution was successfully tested. The filters in the transverters are made of single or coupled full length λ/4 resonators. The λ/4 microstrips are etched very close to the final dimensions and the tuning is performed by adjusting the length of the resonating strips at the hot end (see also fig. 1). Cutting the hot end of the strip produces large frequency variations and is of course an irreversible operation. A fine frequency adjustment can be obtained by soldering a short length of 1 mm Ø silver plated copper wire at the hot end of a 2 or 2.5 mm wide strip (characteristic impedance 60 or 50 Ω respectively on an 1.6 mm thick glassfiber epoxy laminate).

Of course a reliable method has to be used to detect the actual resonant frequencies of the microstrips. A very simple method is to use a small dielectric rod (plastic screwdriver for RF ferrite cores) and approach it to the hot end of the microstrip. The presence of the dielectric rod causes a decrease of the resonant frequency of the microstrip. Monitoring the output of the circuit it can be immediately found out whether the microstrip resonator is too short – the presence of the dielectric rod increases the output signal, or whether it is too long – – the presence of the dielectric rod decreases the output signal.

Where larger capacity variations are required due to the lower loaded Q (like in the 1296 MHz power amplifier) a small piece of thin copper plate is used in place of the silver plated wire to tune the circuit.

In both transverters for 1296 MHz and 2304/2320 MHz the required RF selectivity is not concentrated in a single multiresonator filter but it is distributed among the RF amplifier stages both in the receive signal path and in the transmit signal path, mainly in the form of two resonator filters (see fig. 2) which are used at the same time as matching devices between two amplifier stages.

To obtain a useable value of coupling between two $\lambda/4$ resonators there are two basic arrangements: both microstrips parallel and oriented in the same direction (fig. 2a) and both microstrips parallel but oriented in opposite directions (fig. 2b). The coupling capacitors allow a more convenient selection of the taps on the microstrips and provide also DC decoupling of the amplifier stages.

Designing the transverters particular care was taken to use exclusively cheap and easily available materials and components without degrading the overall performance or the reproducibility. Both transverters are built on low cost glassfiber epoxy FR4 laminate which already has noticeable losses at 2304 MHz and this is probably its frequenoy limit for high Q selective circuits. Except for the RF power amplifiers all the transistors are packaged in low cost plastic cases. <u>The reproducibility can only be enhanced by designing out the needs for critical components</u>

like chip capacitors or microwave trimmers. All the critical RF grounds are therefore directly connected to the ground plane on the other side of the PCB or to "printed" capacitors. The remaining capacitors are conventional ceramic disc (max. diameter 5mm) or pearl types with wire leads, even those used to couple the microstrip resonators, since no difference could be measured in the electrical performances when replacing them with the more expensive and fragile chip capacitors.

2. Block diagrams

The block diagrams of the 23 cm and 13 cm microstrip transverters are shown on figures 3 and 4 respectively. Both transverters are of modular construction, each "box" on the block diagrams representing a single module built on its own printed circuit board. Separate mixers are being used in the transmit and receive signal paths. Since the single ended bipolar transistor mixers are termination sensitive, each converter has its own last LO multiplier stage or stages.

Both transverters include a solid state RF antenna switch with PIN diodes to replace expensive and potentially unreliable coaxial relays. The VOX module is used to interface the transverters to any conventional 144 MHz base transceiver having a common transmit/receive antenna connector. The VOX module includes an RF detector driving a solid state DC supply switch, a receive IF preamp at 144 MHz with a base station TX protection circuit and a power attenuator to reduce the base station TX power to feed the transmit converter. Of course the operation of the VOX module must be "transparent": it must not limit the operational performance of the transverter in any circumstances. On the other hand the VOX module simplifies the operation and increases the reliability, since a single connection cable is used between the base RTX and the transverter and the circuit of the transverter can not be damaged by a wrong connection or a faulty cable.

The 23 cm transverter has a single local oscillator module, since the 1296 MHz segment is being used for narrowband operation in most countries. The LO power splitting at 576 MHz is made with a simple capacitive divider (fig. 16.) The 13 cm transverter has two local oscillator modules since only the 2304 MHz segment is allowed in some countries (Italy) and only the 2320 MHz segment is allowed in some other countries (Germany). Fortunately in Yugoslavia and in many other countries both segments are allowed and transverters covering both subbands are required to be compatible with all possible correspondents. A diode switch is required in this case to switch between the outputs of the LO modules and the LO inputs of the converters. If operation in a single subband is only required the LO module output may be connected as in the 23 cm transverter.

Finally the modular construction allows a number of variations, including modules built in other technologies and/or transmit only and receive only converters primarily for the satellite uplink band around 1270 MHz and downlink band around 2400 MHz respectively.

3. Receive converters

The receive converters for 23 cm (fig. 5) and 13 cm (fig. 6) are basically of identical design except for the obvious changes due to the almost 2:1 frequency ratio. Both converters have two RF amplifier stages (T_1 and T_2), the main RF selectivity being given by the two resonator interstage filters. The input resonator only provides a broad selectivity to reject far away interference, since its insertion loss has to be kept low to avoid noise figure degradation.

The receive mixers employ a single bipolar transistor. Both LO and RF signals are applied to the base of the mixer transistor through $\lambda/4$ lines (L_{10} and L_{16} on fig. 5, L_{10} and L_{22} on fig. 6) The purpose of these lines is to transform the relatively low out-of-passband impedance of the filters into a high impedance at the base of the mixer transistor. In this way the RF filter does not attenuate the LO signal and the LO filter does not attenuate the RF signal. To increase the conversion gain of a transistor mixer, the base must be efficiently grounded for the output frequency and the collector must be efficiently grounded for the input frequencies. The base of the receive mixer transistor is virtually grounded for 144 MHz through the 10 nF capacitor. The collector is connected to a low pass Π filter/impedance matching network tuned to 144 MHz. The first capacitor of the filter is printed on the PCB to minimize its parasitic inductivity.

The 23 cm converter includes a single frequency doubler stage (T_3) to obtain the 1152 MHz signal from 576 MHz. The 13 cm transverter needs two frequency doubler stages (T_3 and T_4) to obtain first 1080 (1088) MHz and finally 2160 (2176) MHz from the original 540 (544) MHz signal. Transistor multiplier stages have a similar requirement as mixer stages concerning the input and output impedances: the base should see a low impedance for the output frequency and the collector should see a low impedance for the input frequency. The function of the two stubs L_{16} and L_{17} on fig 6 ($\lambda/4$ at 2160 MHz including the parasitic inductivity of the transistor package) is to provide a short circuit for the output frequency of the multiplier stage. At lower frequencies a capacitor between base and emitter is usually sufficient (low impedance lines L_{11} on fig. 5 and L_{11} on fig. 6).

The 23 cm converter reaches an overall noise figure of around 3 dB. Since the RF performances of the transistors used fall off rapidly with increasing frequency, the performance of the 13 cm converter is considerably worse, the overall noise figure being around 7 dB. This performance can also be reached by a far simplier interdigital cavity diode converter, however the manufacture of the interdigital cavity requires a considerable amount of work and the mixer and multiplier diodes are not easily available and they actually cost more than all the plastic case transistors used in the microstrip converter. Of course it is possible to use better transistors since the tuning elements already present in the circuit enable a correct matching for almost any bipolar microwave transistor.

4. Transmit converter and power amplifier for 23 cm

The transmit converter for 1296 (1270) MHz is shown on fig. 7. The frequency doubler (T_1) from 576 MHz to 1152 MHz is very similar to that in the receive converter. The transmit mixer (T_2) is a single ended configuration using a single bipolar transistor. Both LO and 144 MHz IF signals are applied to the base of the mixer transistor. An additional 10 dB attenuator is placed in the IF signal path since it is more convenient to perform the base station TX signal attenuation in two consecutive steps avoiding some otherwise critical connections.

The transmit mixer is followed by two selective RF amplifier stages (T_3 and T_4) at 1296 MHz. The totally five $\lambda/4$ resonators are completely sufficient to attenuate all unwanted signals like the LO at 1152 MHz and other unwanted products generated in the mixer stage. The second amplifier stage supplies about 20 mW of power at 1296 MHz and the transmit converter can already be used as a low power transmitter in the 23 cm band.

The transmit power amplifier for 1296 (1270) MHz is shown on fig. 8. It includes three amplifier stages to increase the output power to around 1,5 W. The main function of the microstrips is to provide interstage matching with minimal insertion loss. The first two amplifier stages use BFR96 transistors, which can provide 6÷7 dB power gain at 1296 MHz depending on the output power level and bias conditions. The first BFR96 (T_1) operates in class AB supplying about 100 mW to the second BFR96 (T_2). This transistor increases the power level to about 400 mW. This is probably the maximum <u>safe</u> power level a plastic case transistor like the BFR96 can supply. For higher power levels more expensive transmission transistors are required, packaged in metal-ceramic cases with a stud or flange for heat dissipation. The transistor used in the third amplifier stage (T_3), 2N5944, does not provide a very high gain (about 5 dB), but it is quite rugged since it was designed for transmitter operation. Since this transistor is internally matched for operation in the 70 cm band, its input impedance at 1296 MHz has a very high reactive component, compensated with L_8 and L_9. L_7 is an air wound $\lambda/4$ choke since a single printed microstrip $\lambda/4$ choke was not sufficient.

5. Transmit converter and power amplifiers for 13cm

The transmit converter for 2304 (2320) MHz is shown on fig. 9. The two frequency multiplier stages (T_1 and T_2) from 540 (544) MHz to 2160 (2176) MHz are very similar to those in the receive converter. The transmit mixer (T_3) is practically identical to that for the 23 cm band, including the 144 MHz IF attenuator. However due to the higher frequency the transistors have a lower gain and more amplifier stages are required. The residual LO signal and other unwanted mixing products are relatively less distant from the desired signal and therefore more filtering is required. Unfortunately laminate losses become significant at 2.3 GHz and some gain is also necessary to overcome the losses in the microstrip resonators.

The two selective RF amplifier stages (T_4 and T_5) following the mixer provide about half of the selectivity required (attenuation of unwanted signals) and increase the wanted 2304/2320 MHz signal level to about 5 mW.

This signal feeds the selective transmit power amplifier for 2304/2320 MHz, shown on fig. 10. This amplifier consists of four amplifier stages. The first two stages (T_1 and T_2) provide the remaining selectivity and about 10 dB of gain therefore increasing the useful signal level to about 50 mW. The following two stages (T_3 and T_4) employ BFR96 transistors. With a careful input matching these can supply about 3 dB of gain per stage and about 200 mW of power at 2304/2320 MHz.

Note that all the amplifier transistors are polarized in class A to obtain the maximum possible gain. When bipolar transistors are operated in class A close to their maximum useable frequency and at high signal levels, it is very common to observe a "negative rectification" phenomenon: with the drive power applied the DC current gain decreases and consecutively the collector DC current also decreases. This is actually just the opposite of what we are accustomed to when working with RF amplifiers in class AB or B at lower frequencies!

The 200 mW available from the last BFR96 transistor are already sufficient for a low power transverter. In this case only a RF antenna switch such as that described in the following section needs to be added to complete the microwave part of the transverter.

However, in the case a slightly higher output power and somewhat better receiver sensitivity are desired, a transmit power amplifier and a receive preamplifier are required. To avoid interconnection losses both stages are integrated together with a PIN diode RF antenna switch onto a single printed circuit board (see fig. 11). The RF power transistor (T_1) BFQ 34 requires a quite complex matching network to allow using similar microstrip tuning elements as in lower level stages. T_2 BC213 is a bias regulator for the RF power transistor. It stabilizes the operating point of the RF transistor around the optimum value of 140 ÷ 150 mA of DC collector current to counter the "negative rectification" problem. The obtainabe output power, subtracting the losses in the PIN antenna switch, is in the 500 mW range at the antenna connector.

The receive preamp improves the receive converter noise figure by about 1,5 dB when equipped with the relatively cheap transistor BFQ69.

6. RF antenna switches for 23 cm and 13 cm

Microwave coaxial relays are still very expensive components due to the high amount of skilled mechanical work required for their constraction. Since they are electromechanical components, they are also subjected to wear out and are thus potentially unreliable. Fortunately for low power transmission only and moderate insertion loss and cross talk requirements a solid state replacement is readily available. Popular PIN diodes like the BA379 can be used to switch 5 to 10 W of RF power depending on the circuit configuration of the switch and number of diodes used. The maximum switched power is limited both by the power dissipation rating and breakdown voltage of the single diodes. The BA379 has however another very interesting property: the PIN diode structure is very slow also to turn on. Therefore RF voltages of sufficiently high frequency can not switch the diode on even if the positive halfwave amplitude farly exceeds the diode turn on voltage of about 0.7V. In our particular application this means that these diodes do not need any reverse DC bias in the non conducting state even if a RF voltage of more than 20 V_{pp} is applied to these diodes during transmission.

Due to the residual diode resistance in the on state and other parasitics the insertion loss of the RF antenna switch shown on fig. 12 is around 0.5 dB in the 23 cm band and around 1 dB in the 13 cm band. Accurate insertion loss measurements are difficult due to the mismatches at coax to microstrip transitions, microstrip radiation and other causes. The crosstalk attenuation is sufficient for the application shown limiting the 13 cm TX power to below 1W to avoid RX front end damage.

The RF antenna switch is controlled by the two supply voltages +12VRX and +12VTX switched by the VOX module. The silicon diodes 1N4148 in series with the supply of the "shunt" PIN diodes are required to speed up the switching, since the supply voltages do not fall immediately to zero after a transmit/receive or receive/transmit switchover.

7. Local oscillator modules

Two different local oscillator modules were developed essentially to allow for different types of crystals to be used. One local oscillator module is designed for low frequency third overtone crystals for 32 MHz or 45 MHz (see fig. 13) while the other (see fig. 14) is designed for high frequency fifth overtone crystals for 96 MHz or 90 MHz. Both local oscillator modules supply the same output power of about 10 mW in the 576 MHz, 540 MHz or 544 MHz frequency range. It is however preferred to have the actual frequencies sligtly lower. If the beginning of the microwave narrowband segment coincides with the beginning of the 144 MHz band it is very likely to have disturbs from high power stations operating in the 2m amateur band since most commercial base transceivers are not shielded enough. Further, most modern PLL synthesizer transceivers do not allow any tuning below the beginning of the 144 MHz band and if the transverter local oscillator drifts in the correct direction, part of the microwave band is no longer accessible! It is therefore recomended to convert the first 400 kHz of the microwave bands where almost all narrowband activity takes place in the 144.600 ÷ 145.000 MHz subband which remains very quiet even during peak contest activity.

It is relatively easy to design reliable oscillators with low frequency third overtone crystals. The oscillator shown on fig. 13 should start to oscillate without any tuning, the function of the inductor L_1 in the emitter of the oscillator transistor (T_1) is to inhibit oscillations at the fundamental resonance frequency of the crystal. The oscillator is designed for parallel resonance specified crystals and with some crystals a higher value capacitive trimmer may be required.

The collector of the oscillator transistor should be grounded for the oscillator frequency. However connecting the collector of T_1 to a circuit resonant at an oscillator harmonic frequency this can be conveniently extracted thus saving a separate multiplier stage. The following two stages are conventional frequency multipliers. L_4, L_5, L_6 and L_7 are all printed on the circuit board. The inductive coupling between L_4 and L_5 is sufficient so that usually no coupling capacitor is required. The supply voltage of the oscillator and following stage is stabilized for obvious reasons.

90 MHz and 96 MHz crystals usually operate on the fifth overtone resonance. Suitable oscillators are more critical especially if a reasonable fine frequency adjustment range has to be provided. The oscillator shown on fig. 14 is designed for series resonance specified crystals. In any case it is necessary to accurately tune the capacitive trimmer associated to L_1 to obtain a stable oscillation at the desired frequency.

The oscillator is followed by two multiplier stages, the only difference from the previous module being in the two inductors L_2 and L_3 resonating at 288 MHz, which are air wound self supporting coils in place of the printed inductors. Again the supply voltage of the oscillator is stabilized to 7.5V.

As already mentioned when describing the block diagrams, a diode LO switch is required to connect two local oscillator modules to the transmit and receive converters (see fig. 15). Note that there are no DC decoupling capacitors on the signal lines since such capacitors are already present on the outputs of the LO modules and on the inputs of the receive and transmit converters. If a single LO module is sufficient a simplier solution can be used as shown on fig 16, however to obtain a suitable power split it is necessary to play with the value of the capacitor and the lengths of the cables feeding the LO signal.

8. VOX module

The circuit diagram of the VOX module is shown on fig. 17. Most of the base station transmitter output power is dissipated on the load made of 270Ω ½W resistors. A small fraction is however rectified by the OA95 diode. The obtained DC signal switches on T_2 (BC 237), which in turn discharges relatively rapidly the 4.7μF capacitor through the 1kΩ resistor. The charging of the capacitor is considerably slower due to the 150kΩ resistor. These two resistors practically determine the time constants of the VOX circuit: a fast switch on of the transmit section and a delayed return to the receive mode to allow for speech pauses in SSB operation.

The voltage on the 4.7μF VOX time constant capacitor drives a DC amplifier made up of CMOS inverters from IC_1 (4049 UB). The DC amplifier has a built in hysteresis and drives the two PNP supply switch transistors T_3 (BC213) and T_4 (TIP 32). A circuit with PNP transistors was selected since it offers a really minimum voltage fall across the switching transistors. Of course care should be taken not to accidentally short the +12VRX or +12VTX lines towards ground since the supply switch transistors will be damaged!

The VOX module also includes an attenuator to reduce the base TX signal level for the TX mixer. Note however that the 13 cm mixer requires a higher level 2m signal than the 23 cm mixer and therefore some resistors need to be changed (values in brackets on fig. 17) The attenuator resistors are selected so that about 1W of RF power at 144 MHz fully drives the transverter, the base station TX load is designed for a 3W base station output.

Since the VOX circuit can not know in advance when the base station TX will be switched on, a protection circuit in the form of a power limiter is required to avoid burning out the receive converter when switching from reception to transmission. The action of the limiter with the two BA243 diodes is later reinforced with the application of a DC bias to the diodes. The function of the IF amp (T_1) is to compensate the losses in the protection circuit and base station TX load.

For operational convenience two LEDs may be installed from the +12VRX and +12VTX supply voltages through suitable series resistors to ground to indicate the status of the transverter.

9. Construction of the transverters

As already mentioned in the introduction all the microwave circuits of both transverters are built in microstrip technology on an 1.6 mm (1/16") thick glassfiber epoxy laminate FR4. The corresponding PCB masters are shown on figures 19, 20, 21, 22, 23, 24, 25, 26 and 27. Of course only the upper side is shown since the lower side copper cladding is not etched! Note that although the circuit diagram of the RF antenna switch is the same for both 23cm and 13 cm there are two different printed circuit boards due to the different frequencies involved!

Figures 32, 33, 34, 35, 36, 37, 38, 39 and 40 show the locations of the various components on the printed circuit boards. This information is however not sufficient, since it is not possible to describe the installation of a microwave component just with its electrical symbol. The various mounting details are shown on fig. 18.

Plastic case transistors are installed in 6mm diameter holes drilled at the marked positions. Note that all the transistor leads are connected in the shortest possible way, especially the emitter. It is the emitter parasitic inductivity that primarily limits the gain of a microwave transistor! Power transistors in the metal/ceramic stud package are installed in 10mm diameter holes. As first, two thin copper strips of the same width as the transistor emitter leads are soldered in position to provide efficient emitter grounding. Then an 1mm thick copper plate of at least $15 \times 25\ mm^2$ with a suitable hole in the center for the transistor stud is soldered into place. Finally the transistor leads are trimmed to the required length (do not forget to suitably mark the collector!) and the transistor is first screwed on, the nut locked with a lockwasher, and only then the transistor leads can be soldered.

The ceramic capacitors should be mounted with the shortest possible leads. In fact, the leads are first trimmed to about 1mm from the capacitor body. When the capacitor is soldered into place the remaining lead is completely covered with solder. In a perfectly built transverter no capacitor leads are visible! All the low value ceramic capacitors are disc types of 3÷5 mm diameter except for the 1pF capacitors which are pearl types. Higher value ceramic capacitors are multilayer types which also have very small dimensions.

The supply bypass capacitors, both ceramic and tantalum "drops", have one lead grounded through an 1mm hole drilled in the PCB. Since the actual position of these capacitors is not very critical, these holes are not marked on the printed circuit boards!

The resistors do not play any role in the microwave circuit and therefore their installation is not so critical. In any case it is necessary to think about the parasitic capacitances introduced by the resistor leads--these should be no longer than necessary.

Last but the most important, the grounding of the $\lambda/4$ microstrip resonators. This should be done exactly in the same way as in the prototypes otherwise both the resonant frequency and the coupling to the active devices in the circuit will change! At the marked position a 1.5mm diameter hole is drilled. A short length of 1mm diameter Cu Ag wire is inserted in the hole and well soldered at both sides. Note that the wire is placed just in the current maximum of the resonator and its parasitic inductance therefore has a noticeable effect on the resonator performances.

The various modules are interconnected using short lengths of thin teflon dielectric coaxial cable. Conventional polietilene dielectric cable could also be used, however it is considerably more difficult to solder since the dielectric melts! It is very important to connect the cable ends to the microstrip circuit exactly as shown in fig. 18. Most troubles like strange parasitic resonances are usually caused by a too high parasitic inductivity of the cable braid grounding. First a suitable length of cable is cut and both ends are prepared as shown including the tinning of both center conductors and braids. Then a small piece of copper or brass plate, about 8×8 mm^2 for thin PTFE cables (RG-188) and somewhat larger for RG-142 or even RG-58 cables, is first tinned and then soldered on the groundplane of the microstrip circuit. Finally the prepared cable end is soldered in position.

The local oscillator modules and the VOX module are built on conventional single sided printed circuit boards, whose masters are shown on figures 28, 29, 30 and 31. The corresponding components locations are shown on figures 41, 42, 43 and 44. All the components are installed in the conventional way except for the BFW 92 transistors which are mounted below the PCB to minimize the parasitic lead inductivities. Only high quality plastic foil dielectric trimmers should be used since ceramic types have inferior electrical and mechanical (stability!) performances. The 2÷10pF trimmers in the 540/576 MHz resonant circuits can be conveniently replaced with 2÷6pF types to allow a smoother adjustment.

The building order of the various modules should follow the alignment procedure. As first the relatively independent modules, such as the local oscillator modules, should be built, aligned and accurately tested before other dependent modules are being built. Of course if you have a lot of instrumentation accessible you may also invert the construction order, but this is usually not the case for amateurs!

After all modules have been built and tested they can be installed in a suitable metal container. No internal shields are actually necessary, however all the metal walls and covers should be spaced at least 3 cm from the nearest microstrip lines to avoid detuning and/or unwanted couplings.

10. Alignment and testing of the transverters

The principle of the alignment has already been described in the introduction. All the microstrip resonators have been etched very close to their final dimensions. Therefore only a fine tuning is required. If a certain resonator has to be shorted since its resonant frequency is too low, do never cut more than 0.5 mm a time! When soldering on or around microstrip resonators, remember that the glassfiber-epoxy dielectric constant ε has a positive temperature coefficient. Heating a microstrip resonator will decrease its resonant frequency. Testing and alignment should therefore only proceed after the circuit has cooled back to room temperature!

As first the local oscillator multiplier chains have to be aligned. All the trimmers and/or resonators are aligned for the maximum output at the required frequencies, which should be monitored with a suitable frequency counter and/or Lecher wires. Since the bipolar transistors must work in an nonlinear mode for efficient frequency multiplication or conversion, a shift in the DC bias conditions, in particular the DC base to emitter voltage should also be observed. Due to the rectification effect in the BE junction, the DC base to emitter voltage decreases with the application of the RF drive and may even become negative with RF voltages of a few V_{PP}. The BE junction of the following transistor thus acts as a RF probe just where needed - at the output of the stage to be aligned!

The low level stages in the receive converter should be aligned using a suitable signal source and a base 144 MHz receiver with a sensitive "S-meter". Since the microstrip resonators are already pretuned to the desired frequency, a wideband noise generator can be used as the signal source. All the resonators may be simply aligned for the maximum obtainable gain since the bipolar silicon transistors used in the transverters do not show any noticeable difference between the optimum noise figure match and the maximum gain match at frequencies above 1 GHz.

To align the transmit converter its output should be connected to a RF milliwattmeter (a 50 Ω load with a schottky diode detector). Since the LO signal is relatively very close to the wanted mixing product (especially in the 13 cm band) and the single ended mixer does not attenuate it, considerable care should be taken to adjust the microstrips to the correct frequency. Since the microstrips are etched close to their final dimensions, this simply means that the correct resonance must be found with minor adjustments only. Of course the output signal must disappear immediately when removing the 144 MHz drive signal. In any case, the famous Lecher wires are a very simple to build but accurate (better than 1%) frequency meter.

Once the transmit converter is adjusted and tested the power amplifier may be connected between the transmit converter and the RF milliwattmeter. Again all the adjustments are made to obtain the maximum output power. When adjusting the 23 cm power amp (fig. 8), the 10 Ω resistor in the collector circuit of T_2 should be increased to 47 Ω to avoid burning out T_2 (BFR96) if the resonant circuits are not yet well aligned. In the 13 cm selective power amplifier the bias resistors of the last two stages (T_3 and T_4) should be selected to obtain the maximum output power but considering also the maximum allowed power dissipation for the transistors used!

As already mentioned in the introduction, microstrip circuits in theory do not require any shielding. In practice some problems may happen when the single circuits are installed in a metal container and connected with coaxial cable. Every closed metal box has an infinite number of self resonances having a very high Q factor and the situation is further complicated due to the presence of various printed circuit boards and different length wires and cables. Considering the practical dimensions of the printed circuit boards and metal container, the lowest frequency resonance modes fall in the low GHz range. This is the reason why this problem is almost unknown to people working below 500 MHz while it is a very common problem for people working above 1 GHz.

Resonance problems can hardly be solved by shielding, in fact shields usually just introduce new resonances! Such problems can only be solved by reducing the Q of the particular resonant mode either by introducing some lossy material – absorber or by soldering a few non inductive resistors between the resonance "hot spots".

Absorbing material is most suitable to solve resonance problems caused by the metal box and by the large surface ground planes of the microstrip printed circuit boards. Its most efficient location is between two ground planes where the resonance effect takes place, for example between the bottom of the metal container and the ground plane of the microstrip circuit board. Of course the absorbing material should be installed in such a way to avoid damping the useful signals in the microstrip circuit! The availability and price of professional absorbing materials may represent a problem however a very cheap alternative is readily available on the market: the black conducting foam usually used for packing MOS integrated circuits and other ESD sensitive devices!

Discrete resistors can also be used in place of the absorbing foam, however they are more suitable to solve problems caused by phisically smaller components like coaxial cables connecting the various microstrip circuits together. The resonance does not take place inside the cable, between the center conductor and the braid, since the cable is connected to matched loads at both ends. The braid of the cable together with a metal ground plane builds another transmission line. Since the braid is grounded at both ends, resonances at $n \cdot \lambda/2$ will occur with a very high Q. Due to the finite parasitic braid grounding inductivity at both ends these resonances are coupled to the waves propagating inside the cable. Naturally poor braid grounding makes this problem considerably worse. Such resonances can efficiently be damped by soldering one or more 100 Ω resistors between the cable braid and the ground plane avoiding the voltage nodes of all the resonances to be suppressed.

Finally some cable lengths are critical in any case, particularly those feeding the local oscillator module signal to the microstrip multiplier stages in the transmit and receive converters. Since frequency multiplier stages are termination sensitive, their efficiency may drop to zero with some "unfortunate" cable lengths. Also the 13 cm transmit amplifier chain may require some cable length trimming to optimize the overall gain and output power.

11. Conclusion

The transverters described in this article have actually been operating for over two years and several prototypes were built. A number of tests were made to find out any potentially unreliable circuit configurations. The only major problem was caused by the frequency multiplier stages, particularly the transistors used in these stages. Beside the well known failures caused by current, voltage or dissipated power overloads another failure mechanism was observed when relatively high RF voltages are applied to the base of a RF transistor as it is usually necessary for efficient frequency multiplication. A parasitic schottky diode junction starts building in parallel with the BE junction of the transistor. Since the schottky junction knee voltage is of the order of Ø.3V while that of the BE junction remains around Ø.7 V, this process slowly decreases the current gain (β) of the transistor down to zero! The process is gradual and may be very slow: it may take a few weeks or even months of continuous operation before a performance degradation can be noticed. It was also found out that there are large variations in the sensitivity for this failure mode of the same transistor type supplied by different manufacturers. In fact it is very difficult to observe this failure mode with first class components supplied by reputable manufacturers like Siemens or Valvo without exceeding the transistor maximum ratings.

On the other hand, many different transistors can be used in the low level microwave stages since the two resonator microstrip filters allow a broad tuning range. Microwave ceramic case transistors ("micro-X" or "cerec" package) all provided considerably more gain and better noise figures than their plastic counterparts, but they are almost an order of magnitude more expensive. Fortunately two emitter lead transistors in plastic packages became available recently and these should match the microwave performance of the ceramic packaged transistors at a readily acceptable price.

Both transverters were designed to keep the alignment procedure as simple as possible. Therefore frequency trap resonators, balanced mixers and other complex circuits that require costly instrumentation for their testing and alignment were intentionally avoided. Of course these circuits can also be built in microstrip technology! Beside the transverter modules many other circuits have been built using the same microstrip technology: beacon transmitters for the 23cm and 13cm bands and receive converters for various satellite bands: the 1.7GHz meteorological satellite band, the 1540 MHz Marecs downlink band and the 1575.42 MHz Navstar frequency. In fact the first converters built using the technology described were tested on Meteosat signals at 1694.5 MHz.

In the practical use it was found out that the circuits built in the technology described are able to operate in extreme environmental conditions (both temperature and humidity) as are usually found in high mountain contest locations and that they can withstand very rough handling like falls on a rough rocky surface without detuning!

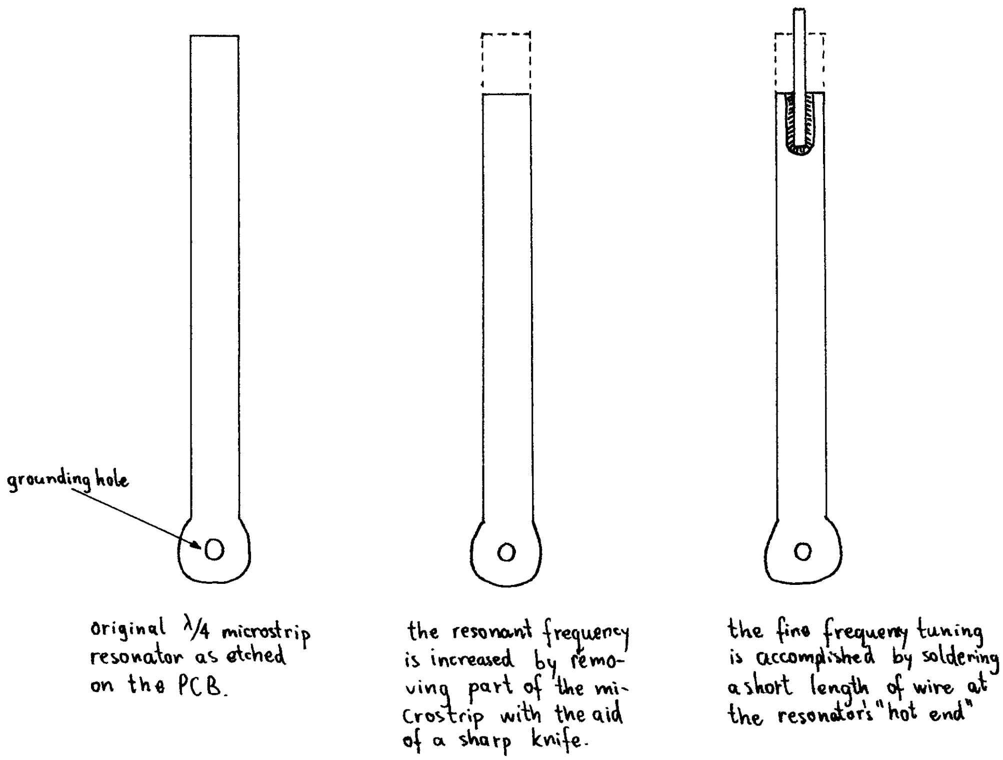


Fig. 1 - Tuning a λ/4 microstrip resonator

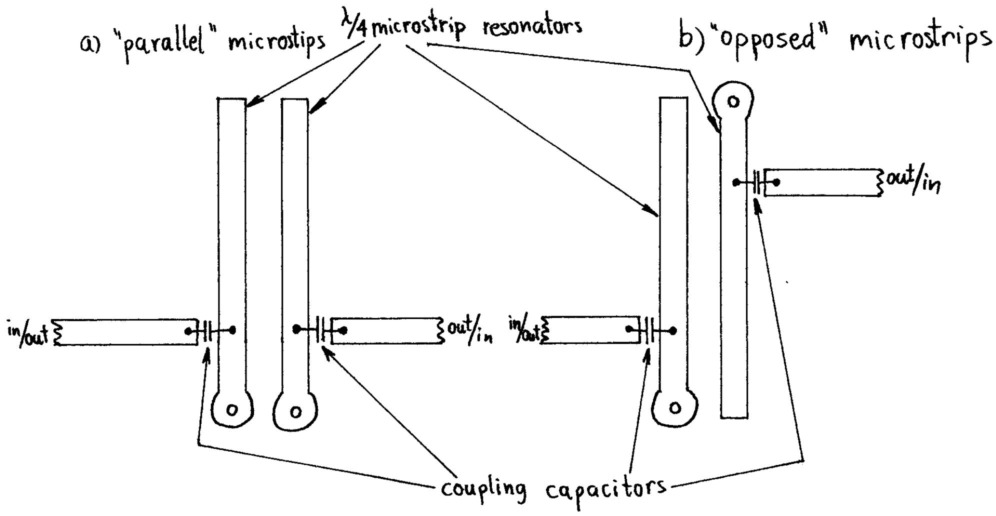


Fig. 2 - Two resonator filters in microstrip technology.

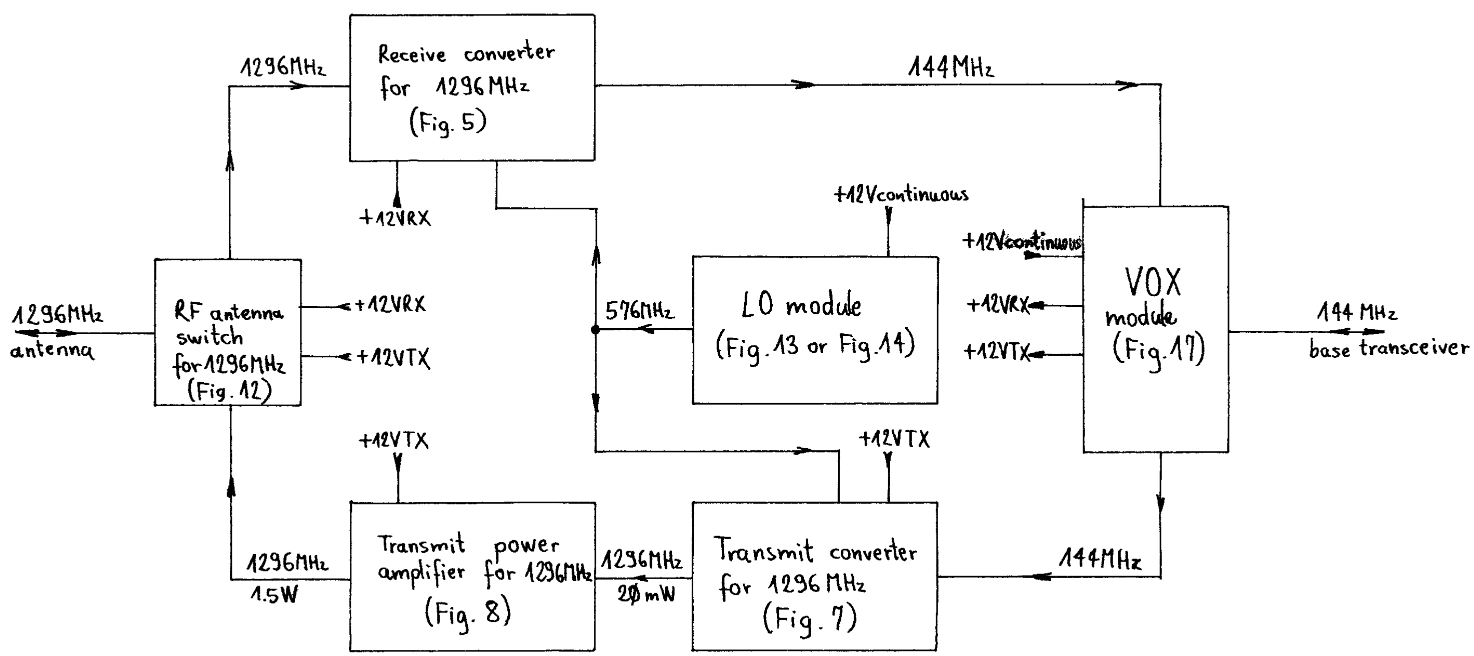


Fig. 3 – Block diagram of the 1296/144 MHz transverter.

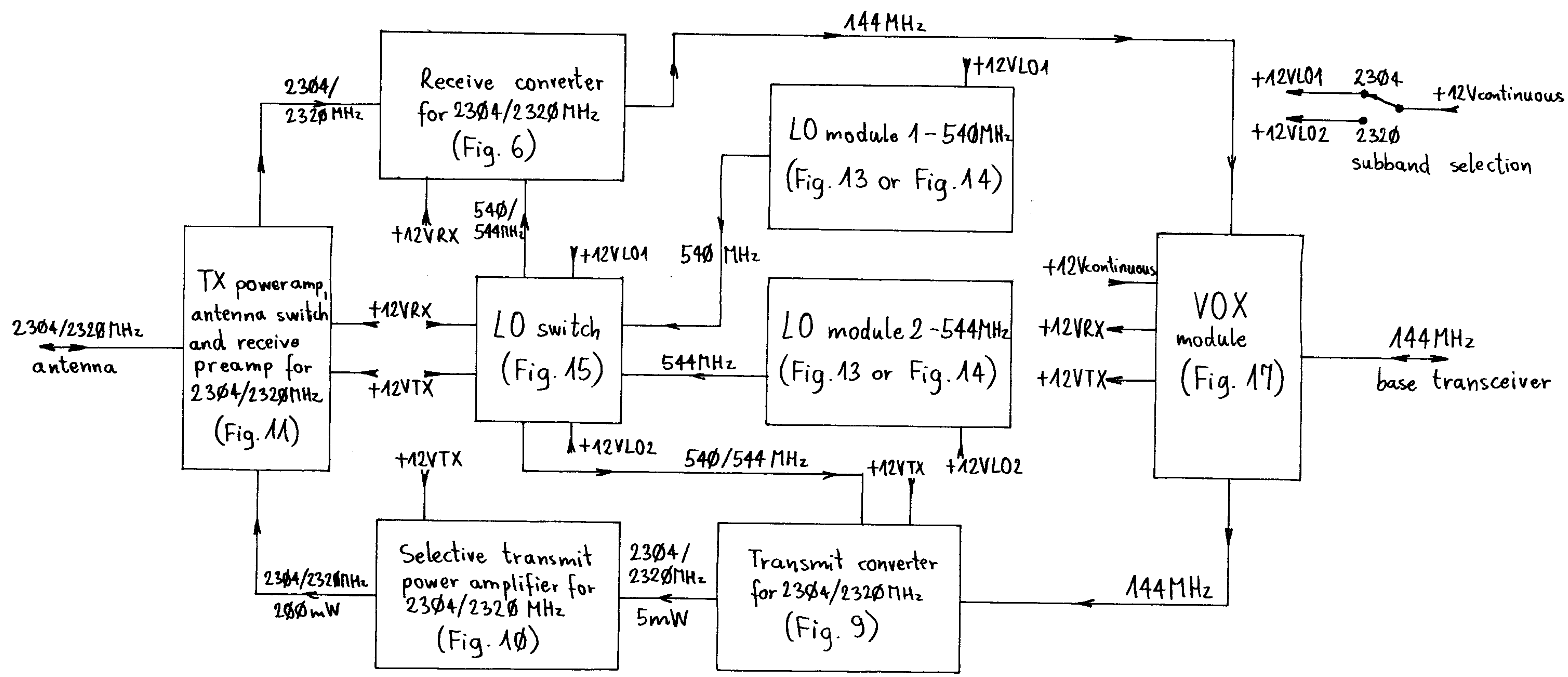


Fig. 4. - Block diagram of the 2304/2320/144 MHz transverter.

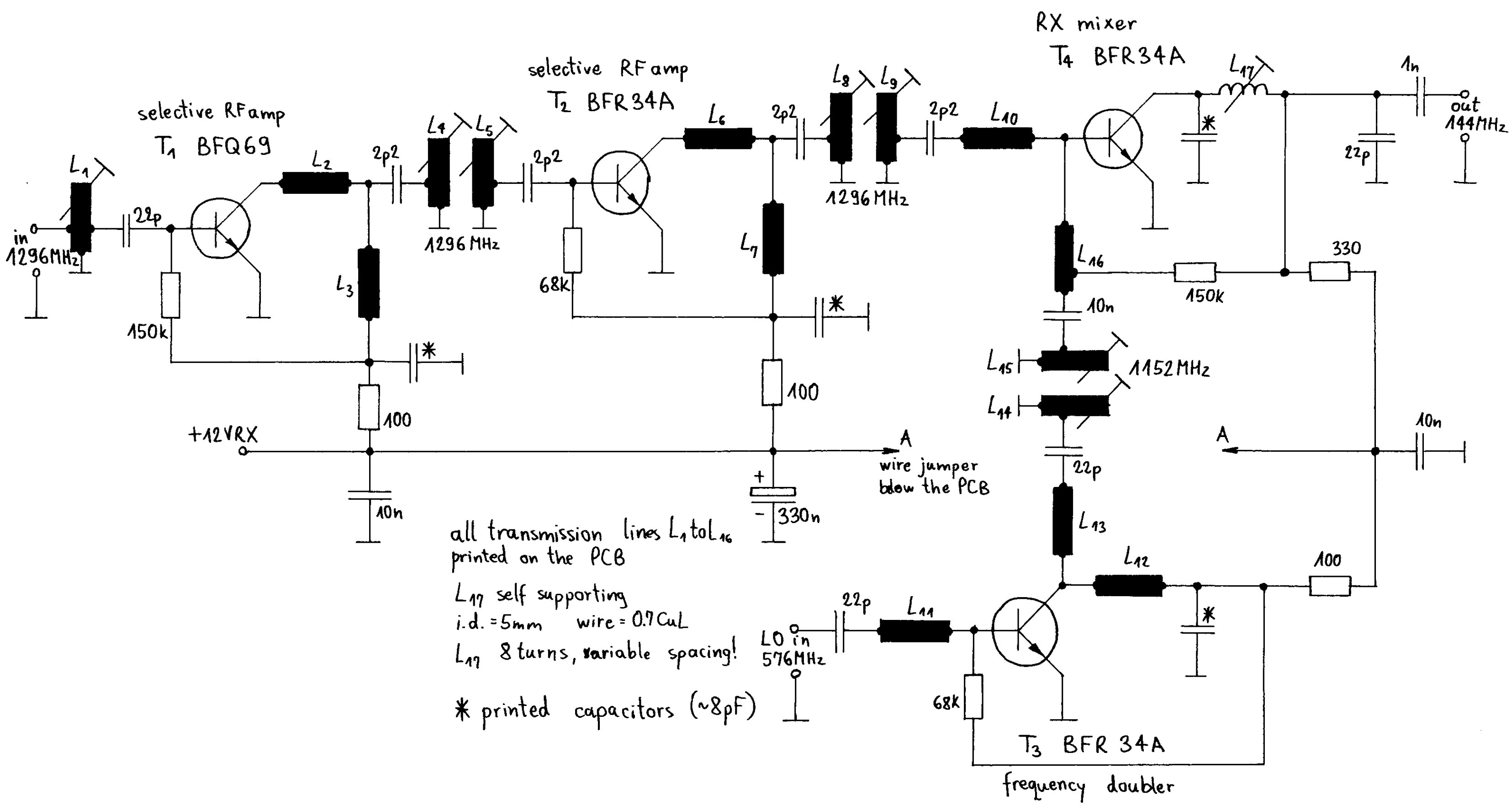


Fig. 5 – Receive converter for 1296 MHz.

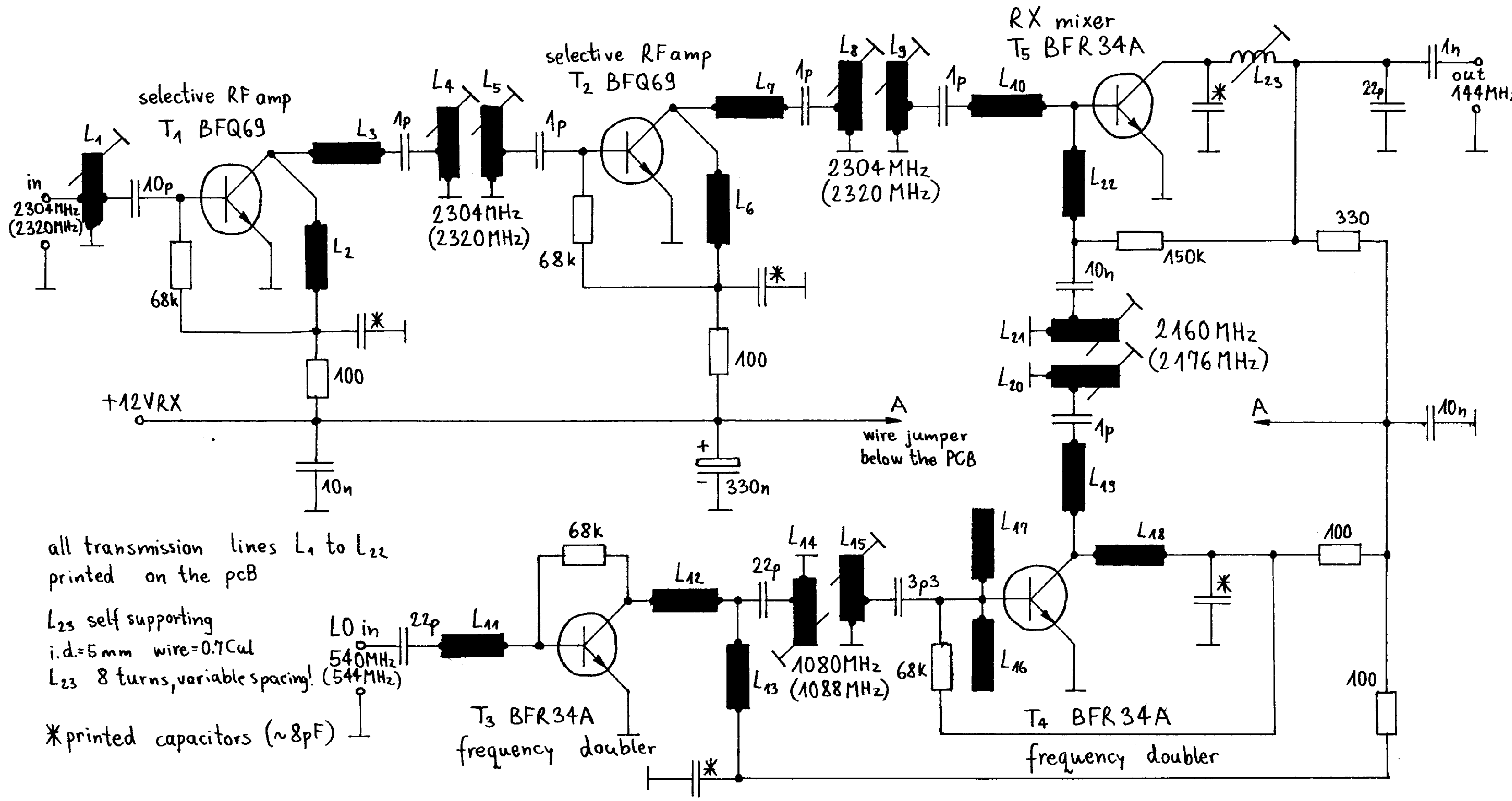


Fig. 6 - Receive converter for 2304/2320 MHz.

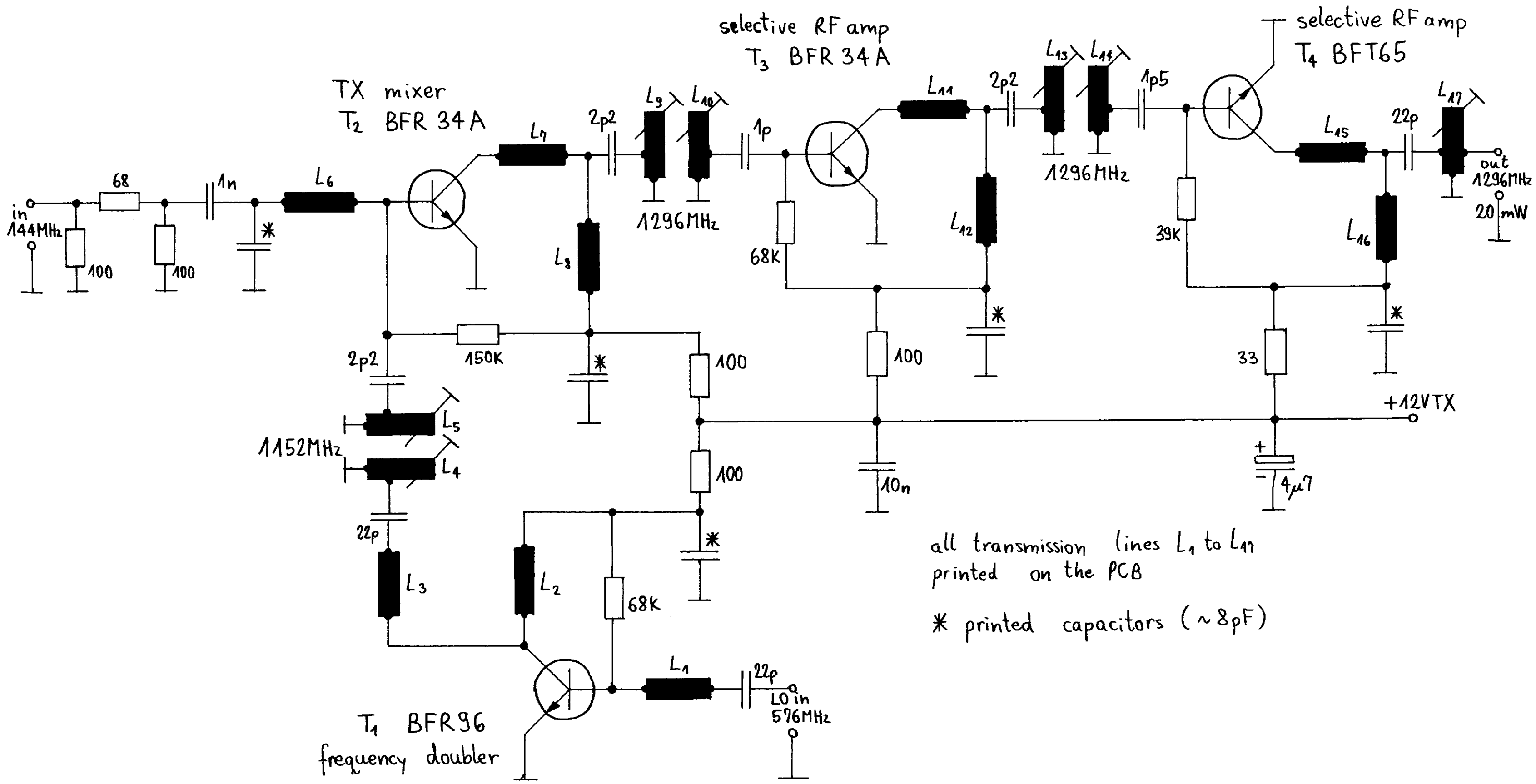


Fig. 7 —Transmit converter for 1296 (1270) MHz.

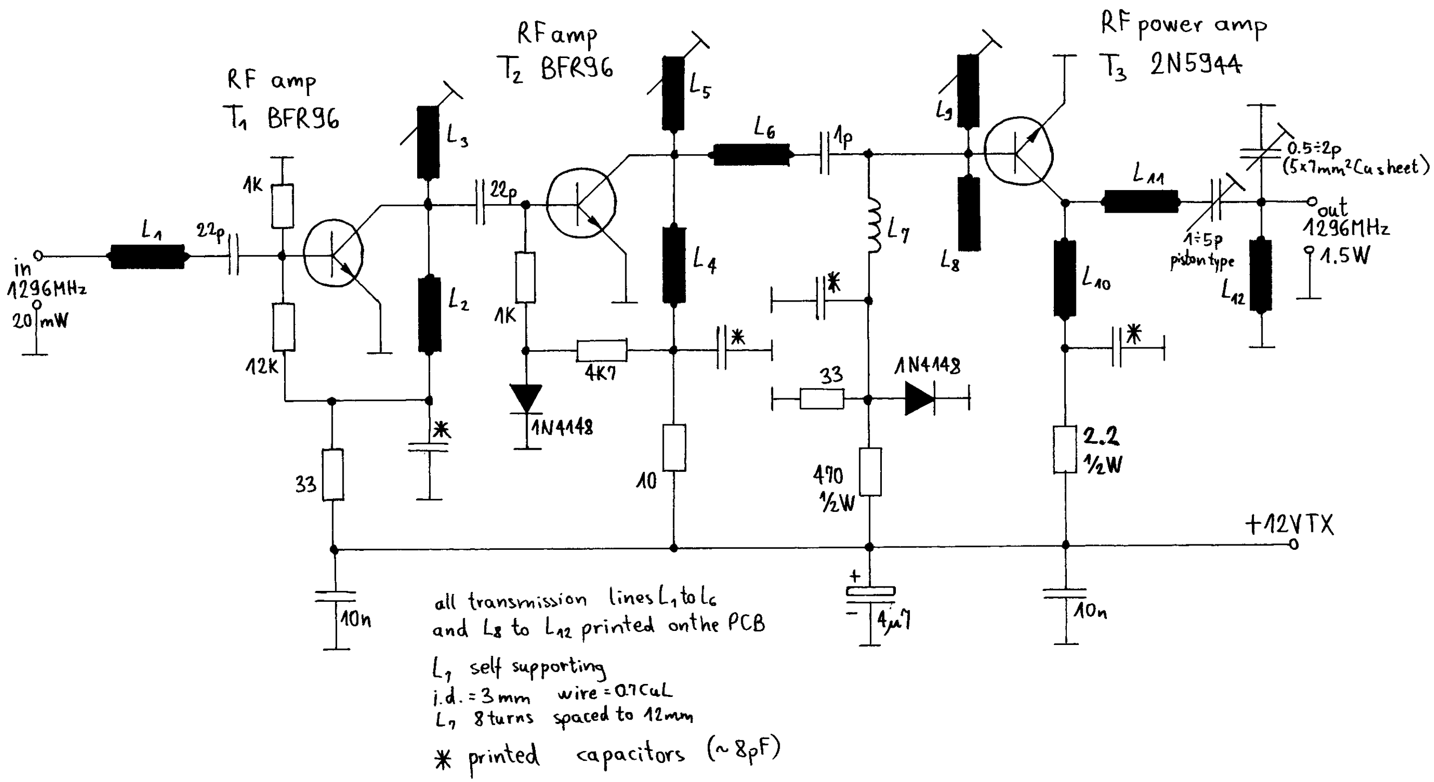


Fig. 8 — Transmit power amplifier for 1296 (1270) MHz.

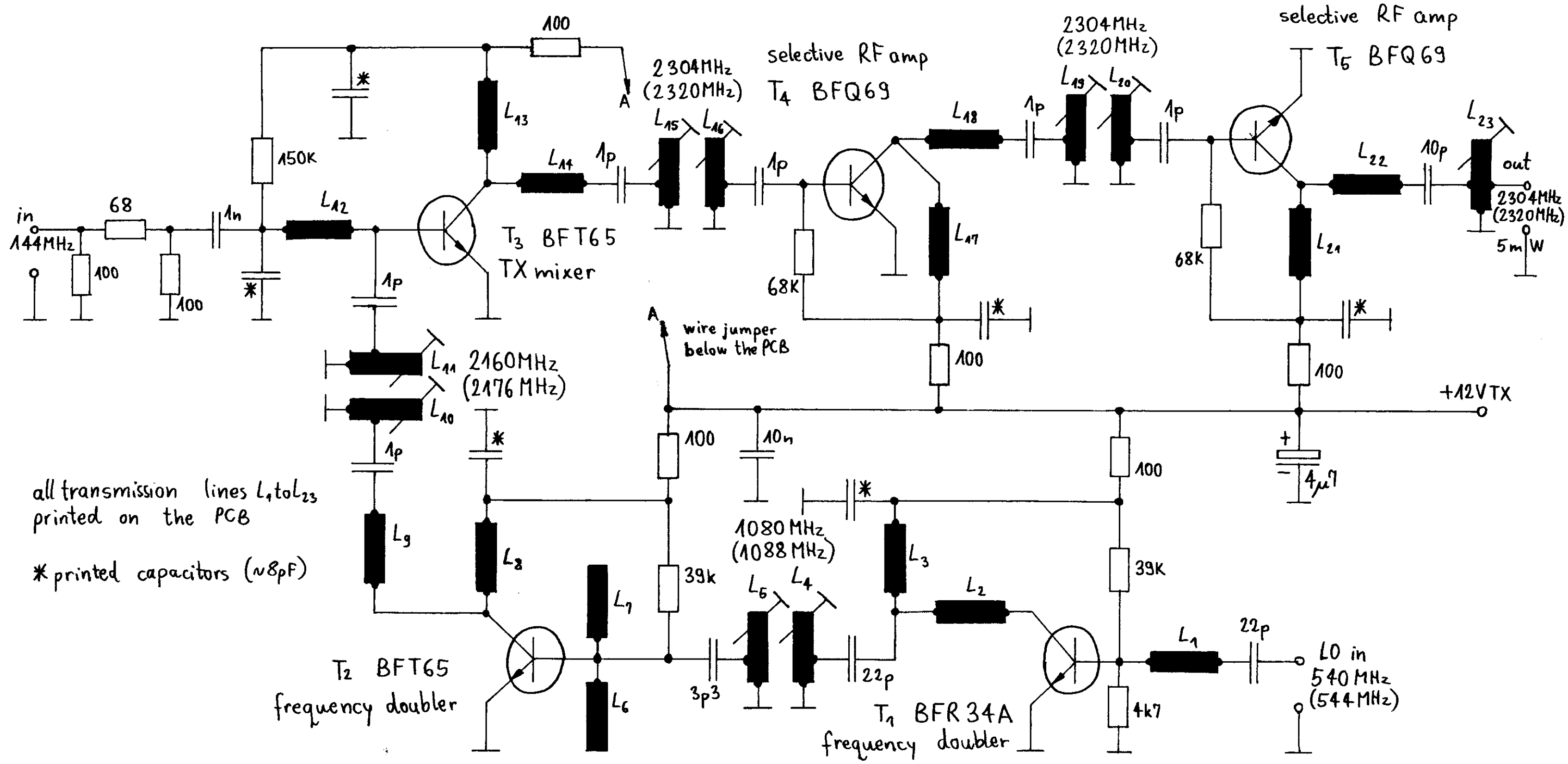


Fig. 9 - Transmit converter for 2304 / 2320 MHz.

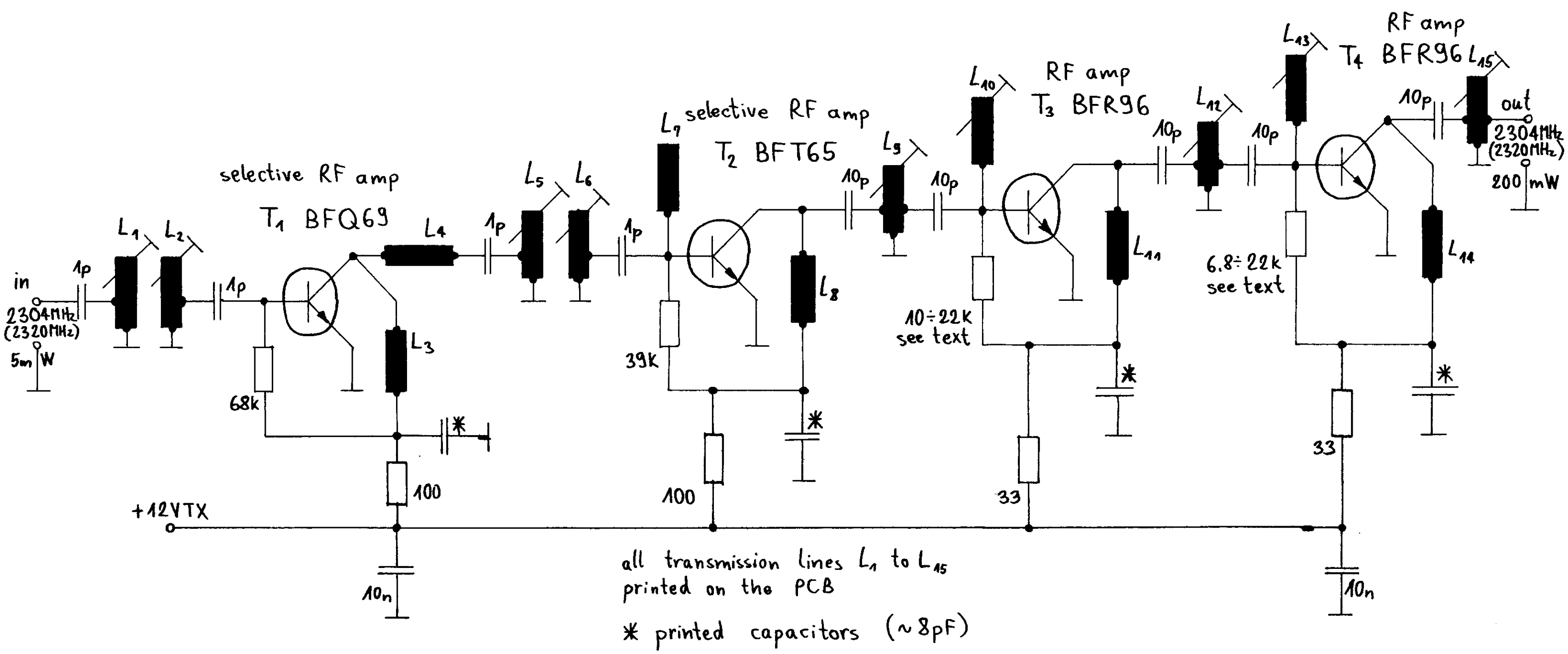


Fig. 10 – Selective transmit power amplifier for 2304/2320 MHz.

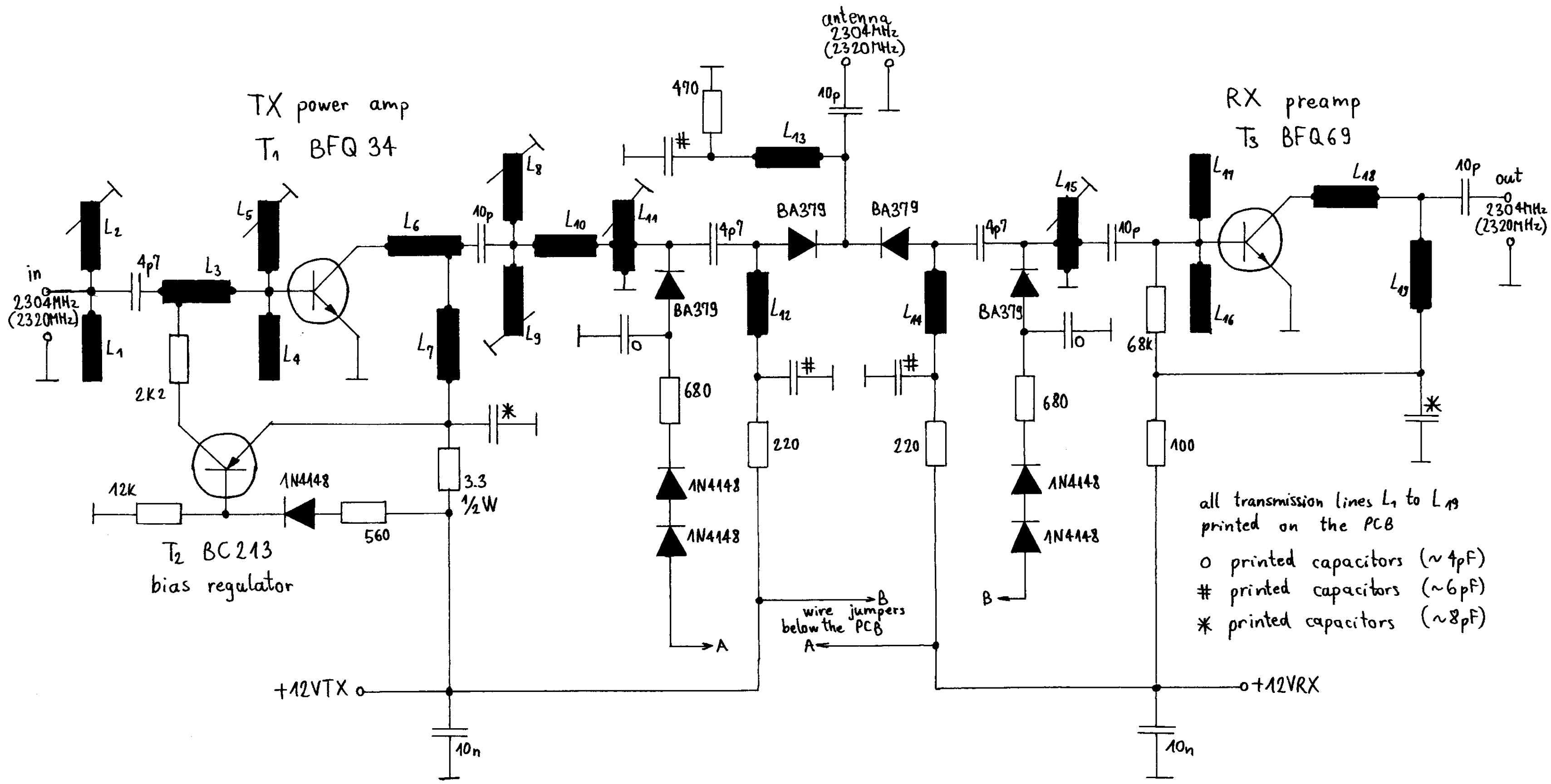
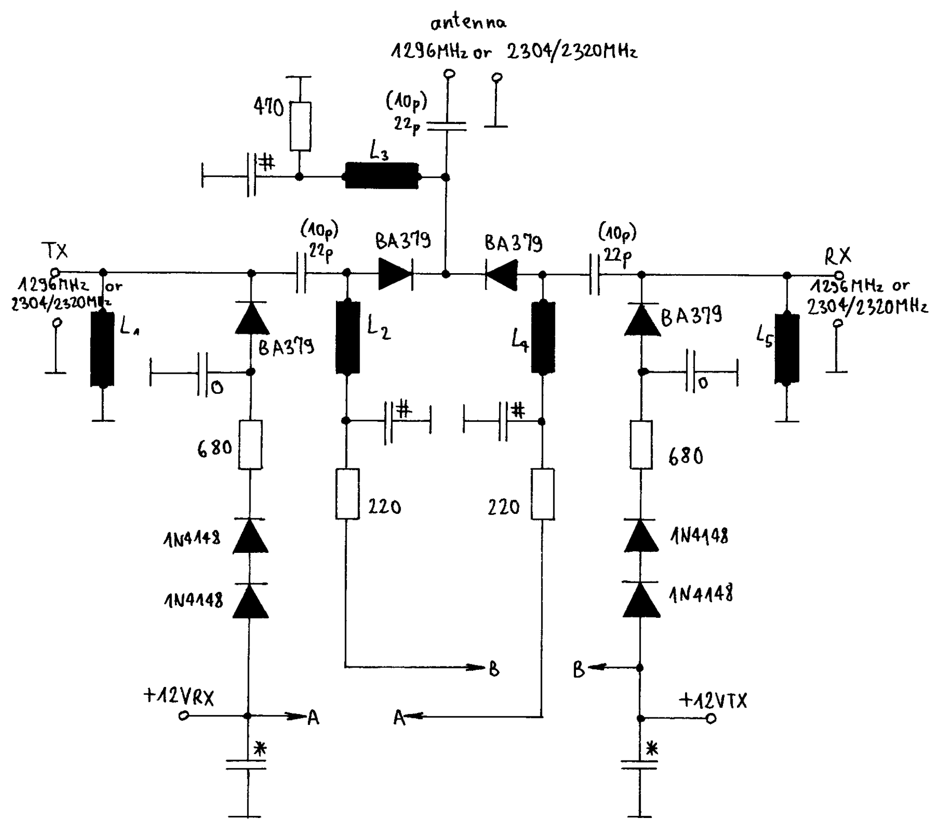


Fig. 11 – TX power amp, antenna switch and receive preamp for 2304/2320 MHz.



all transmission lines L_1 to L_5
printed on the PCB

printed capacitors	1296MHz	2304/2320 MHz
o	~8pF	~4pF
#	~8pF	~6pF
*	~8pF	~8pF

values in brackets for 2304/2320 MHz

Fig. 12 - RF antenna switch for 1296 MHz or 2304/2320 MHz.

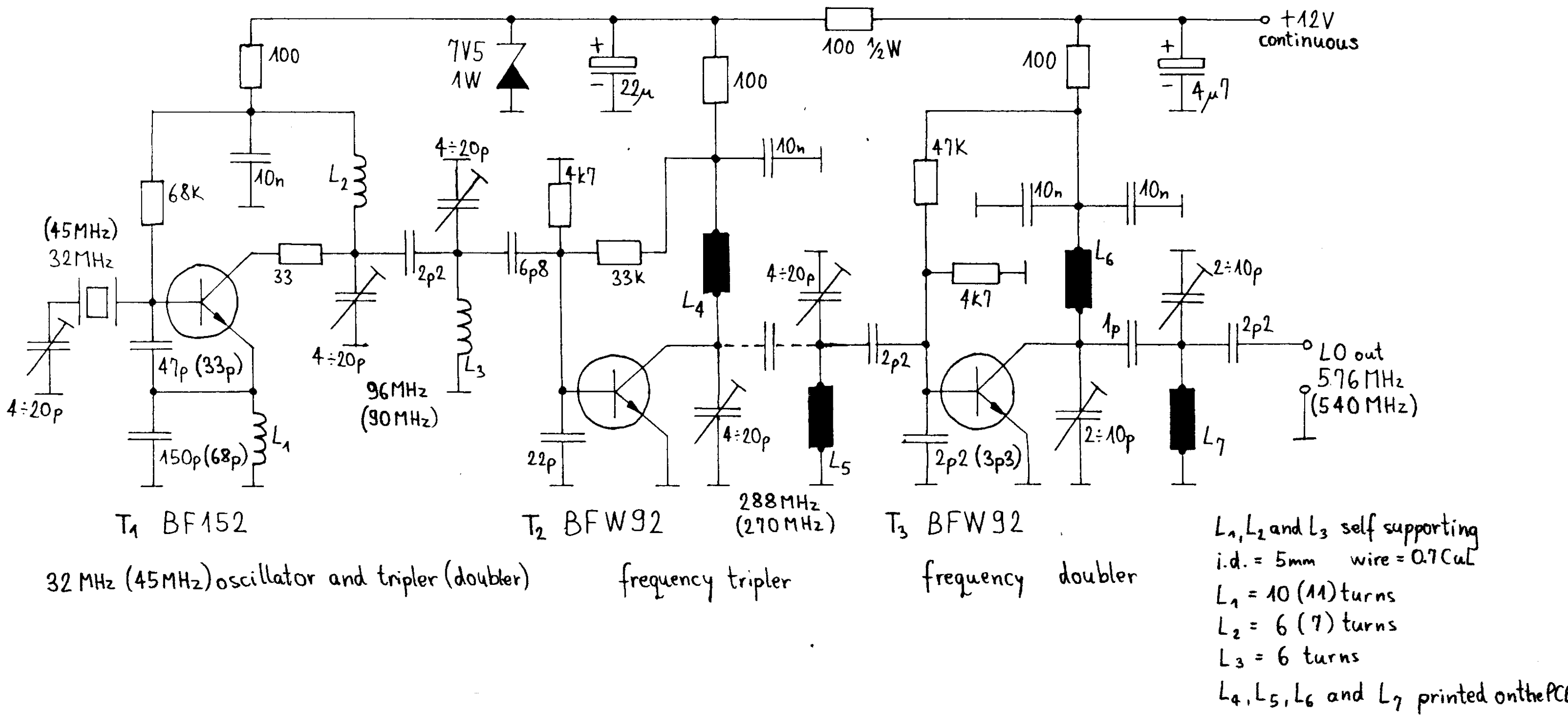


Fig. 13 – Local oscillator module for 32 MHz crystals (values in brackets for 45 MHz crystals).

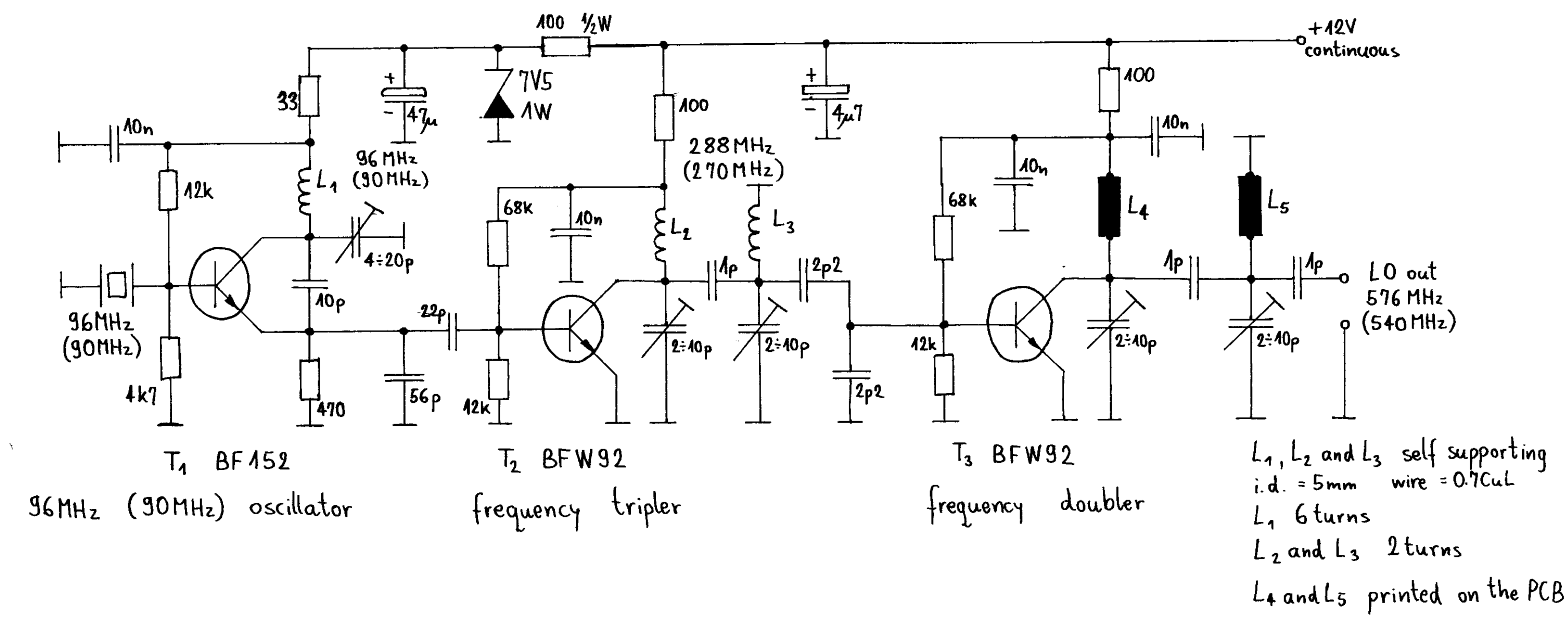


Fig. 14 – Local oscillator module for 96MHz crystals (values in brackets for 90MHz crystals)

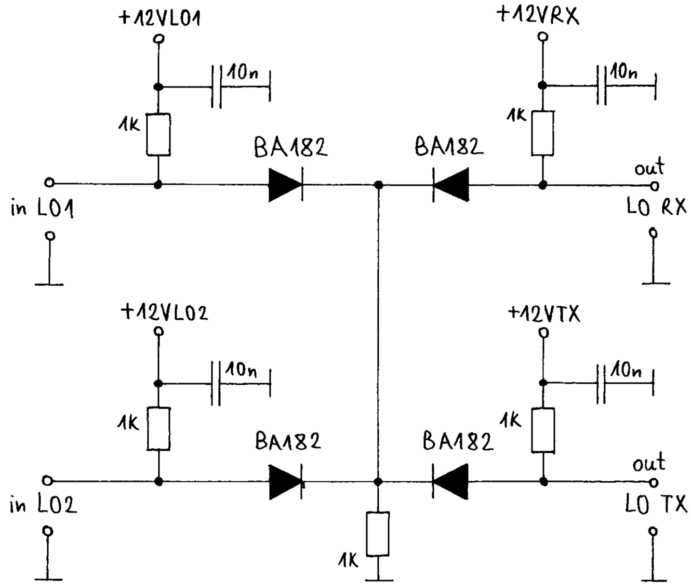


Fig. 15 – LO switch.

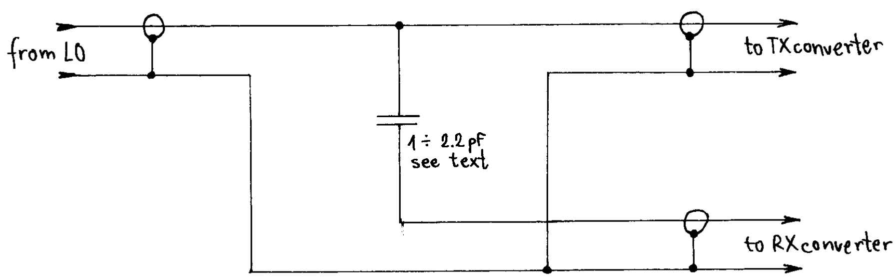


Fig. 16 – A simplier way of feeding the RX and TX converter with the LO signal.

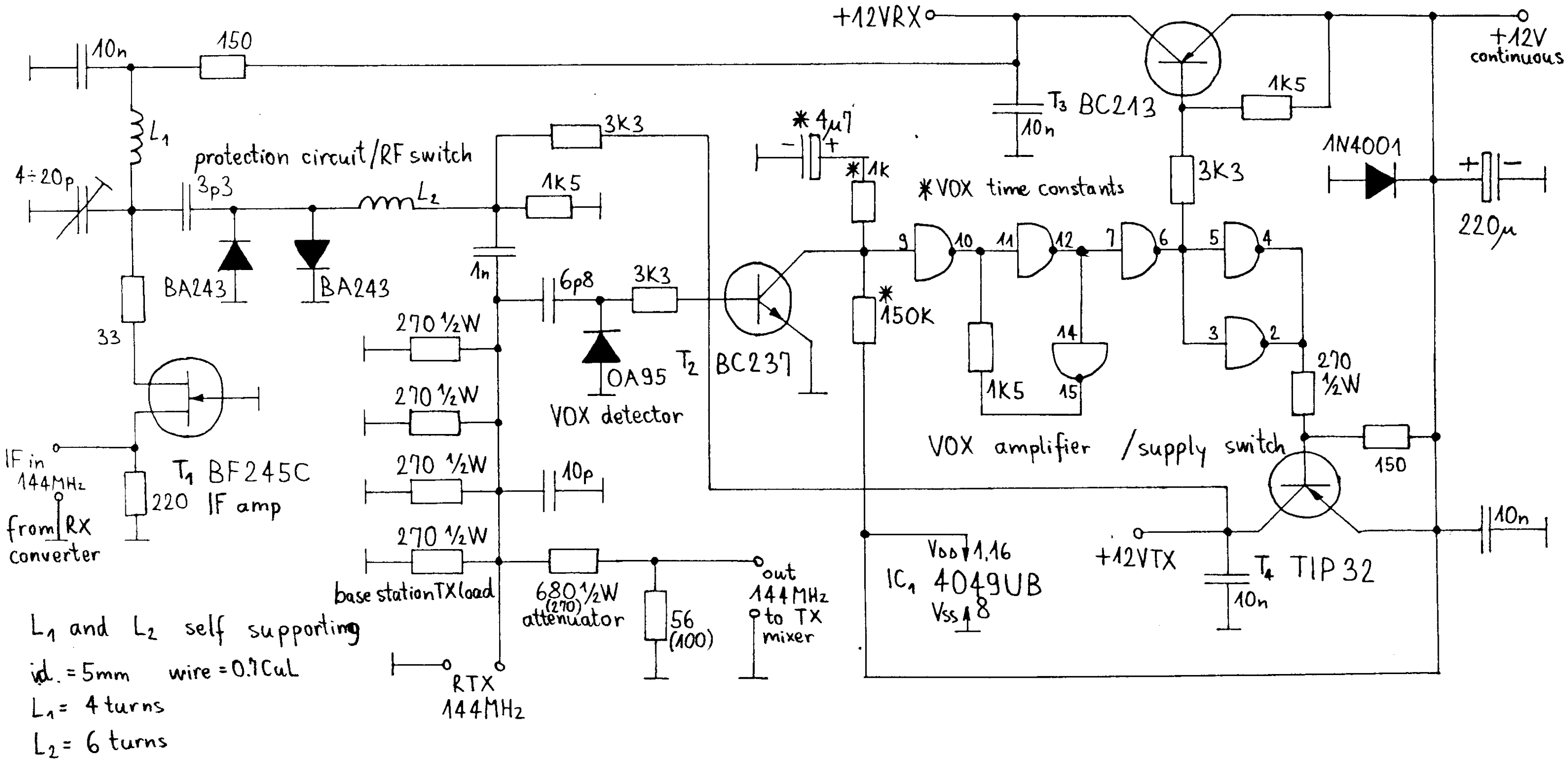


Fig. 17- VOX module (values in brackets for the 2304/2320 MHz transverter)

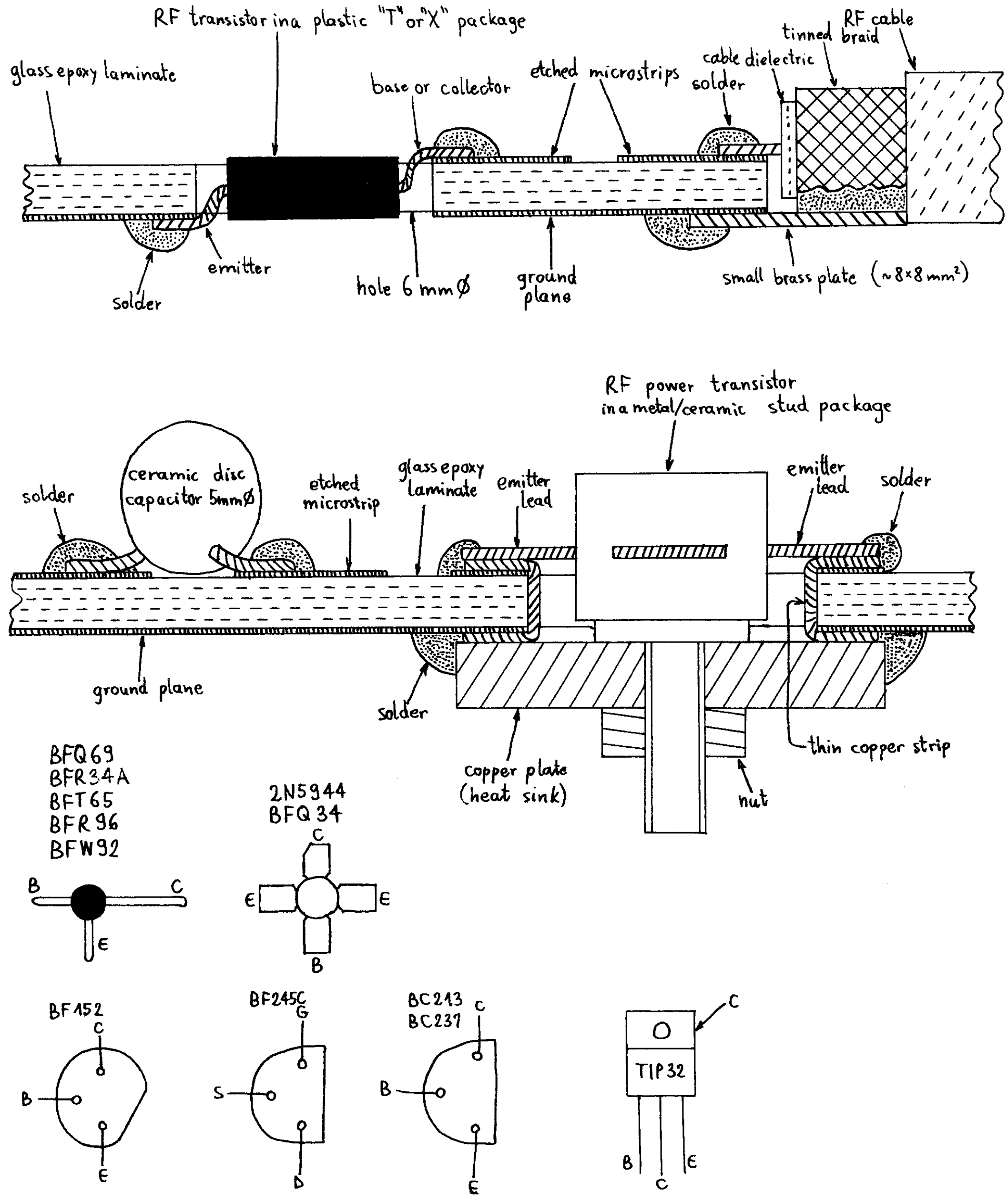


Fig. 18 - Mounting details of various components.

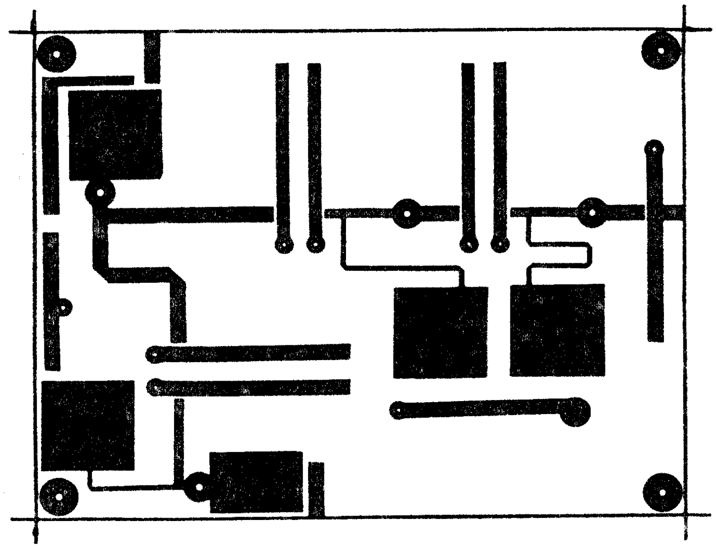


Fig. 19 - PCB of the receive converter for 1296MHz

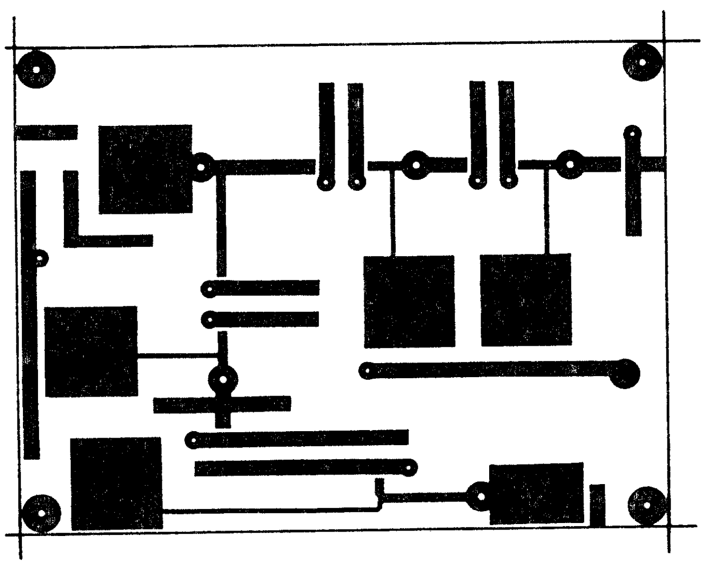


Fig. 20 - PCB of the receive converter for 2304/2320MHz.

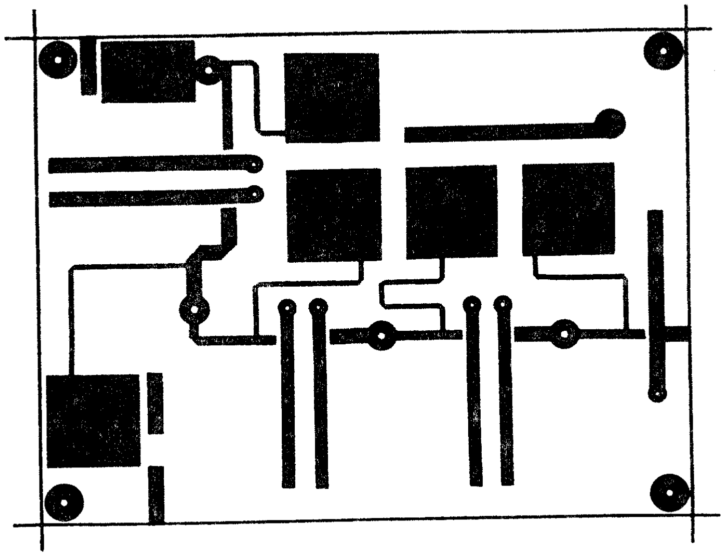


Fig. 21 -PCB of the transmit converter for
1296 (1270) MHz.

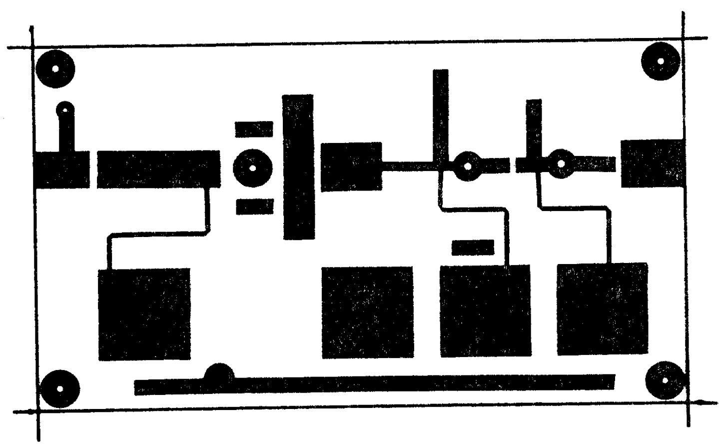


Fig. 22 - PCB of the transmit power amplifier
for 1296 (1270) MHz.

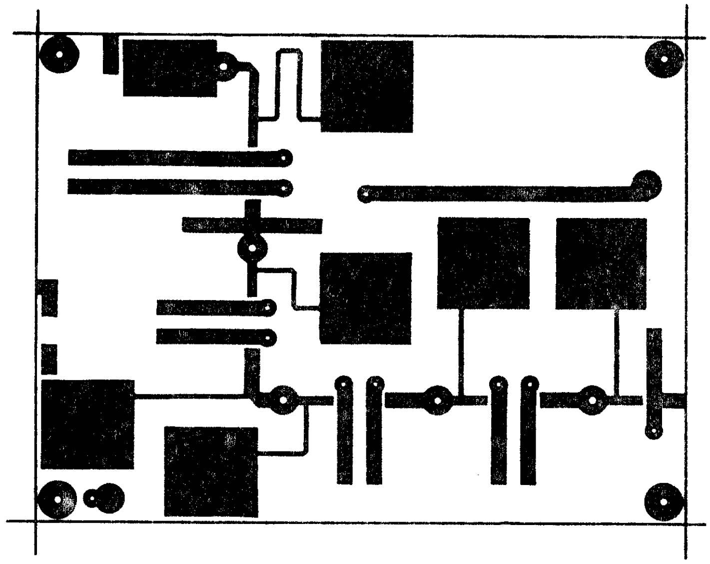


Fig. 23 - PCB of the transmit converter for 2304/2320 MHz

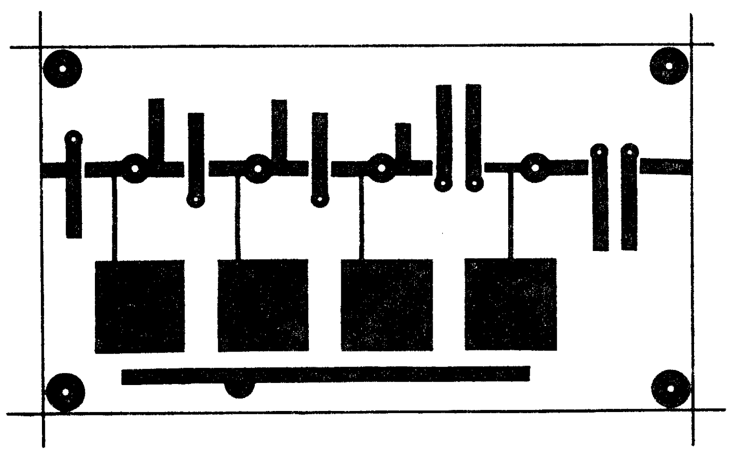


Fig. 24 - PCB of the selective transmit power ampli fier for 2304/2320 MHz.

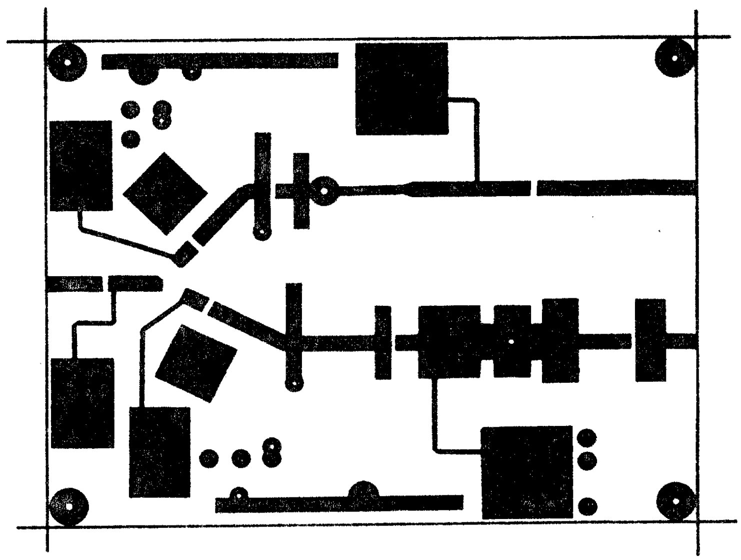


Fig. 25 –PCB of the TX power amp, antenna switch and receive preamp for 2304/2320 MHz.

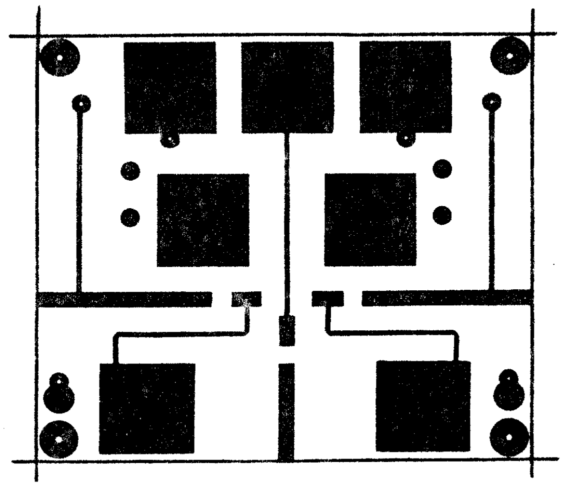


Fig. 26 - PCB of the RF antenna switch for 1296 MHz.

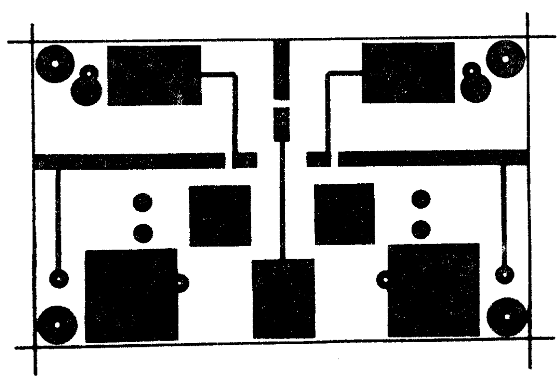


Fig. 27 - PCB of the RF antenna switch for 2304/2320 MHz.

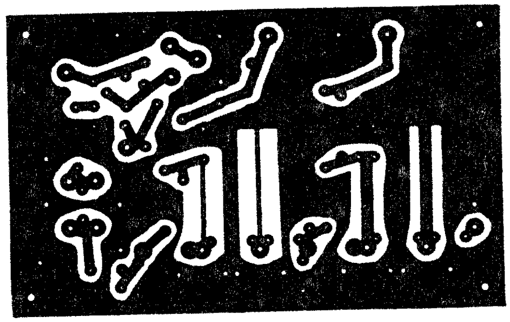


Fig. 28 - PCB of the LO module for 32 (45) MHz crystals

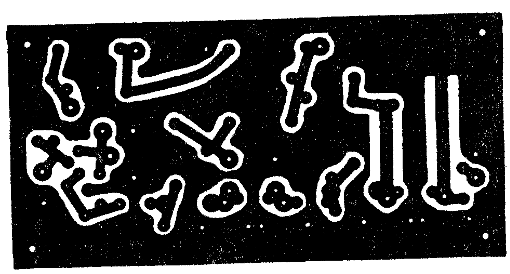


Fig. 29 - PCB of the LO module for 96 (90) MHz crystals.

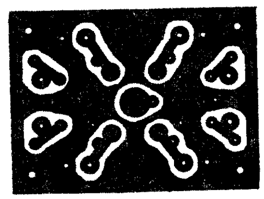


Fig. 30 - PCB of the LO switch.

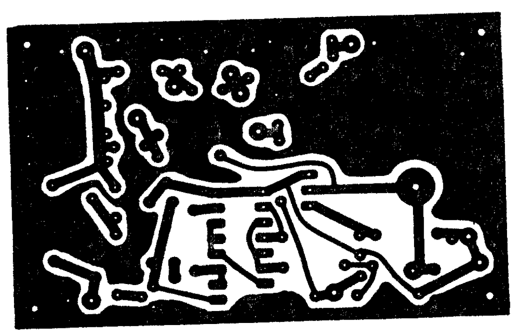


Fig. 31 - PCB of the VOX module.

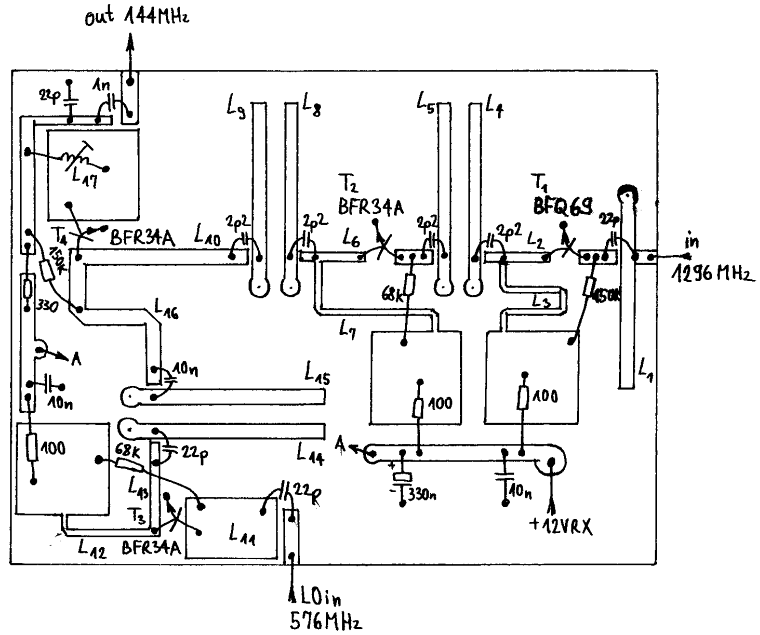


Fig. 32 - Components location of the receive converter for 1296 MHz.

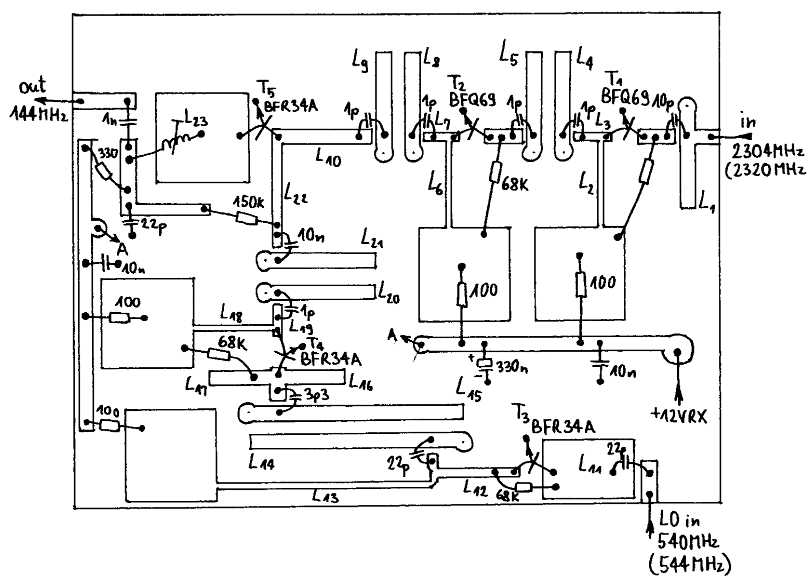


Fig. 33 - Components location of the receive converter for 2304/2320 MHz.

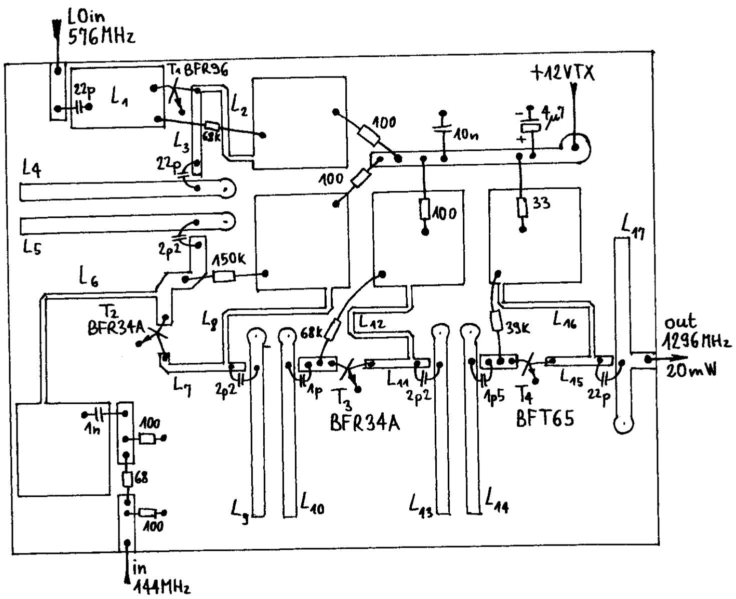


Fig. 34 - Components location of the transmit converter for 1296 (1270) MHz.

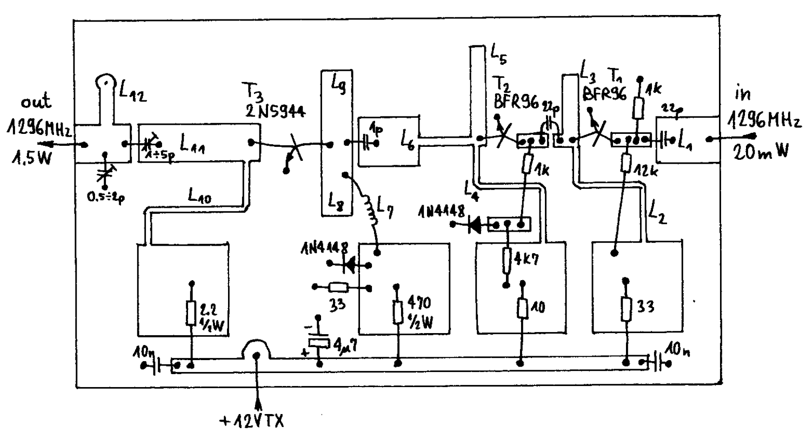


Fig. 35 - Components location of the transmit power amplifier for 1296 (1270) MHz.

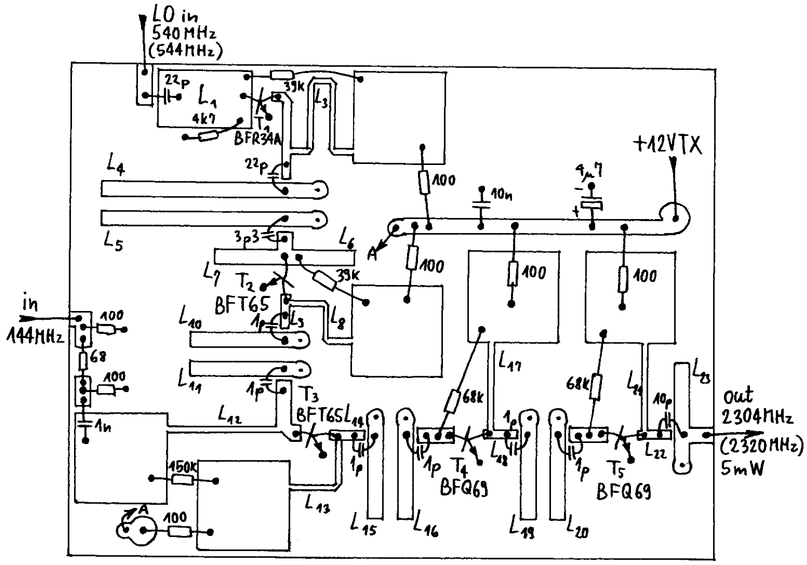


Fig. 36 - Components location of the transmit converter for 2304 /2320 MHz.

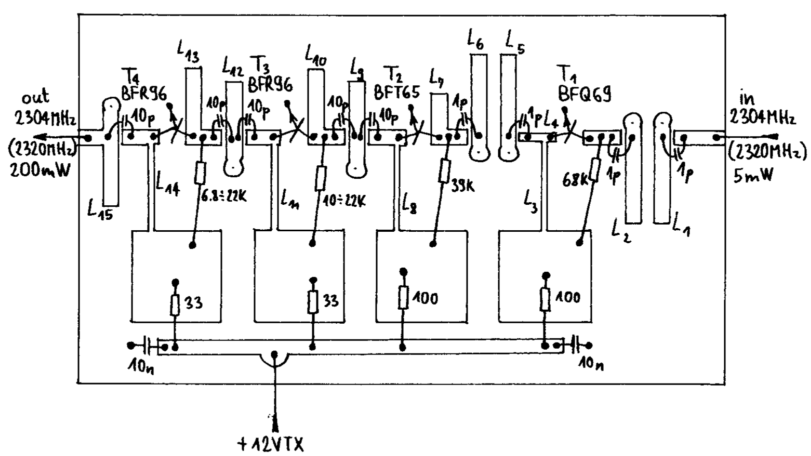


Fig. 37 - Components location of the selective transmit power amplifier for 2304/2320 MHz.

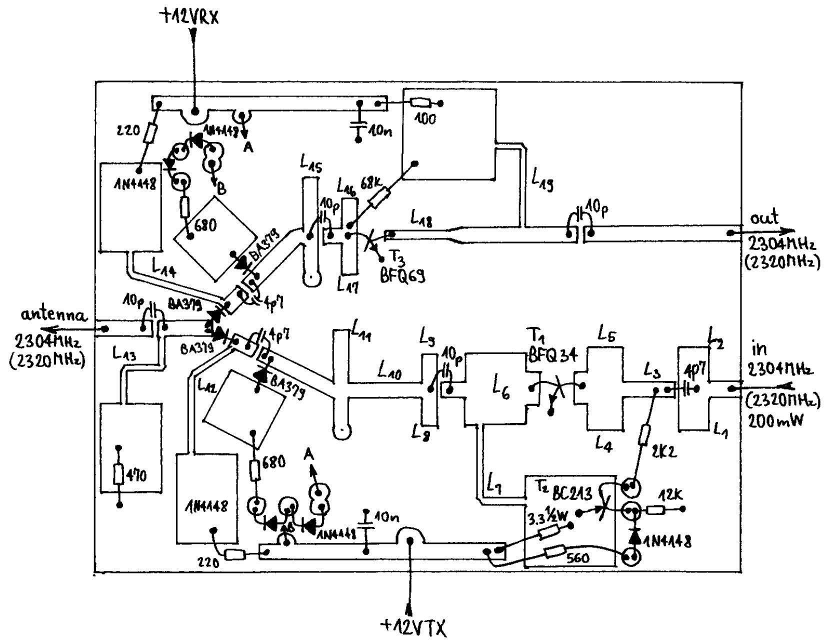


Fig. 38 – Components location of the TX power amp, antenna switch and receive preamp for 2304/2320 MHz.

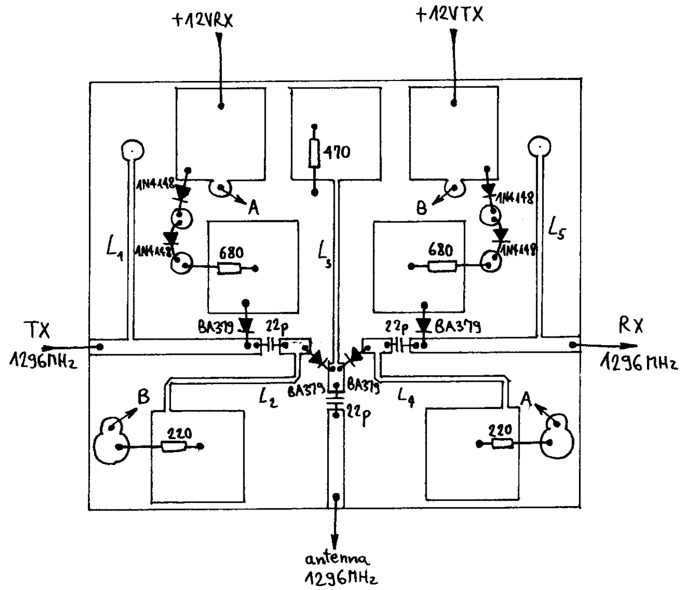


Fig. 39 -Components location of the RF antenna switch for 1296MHz.

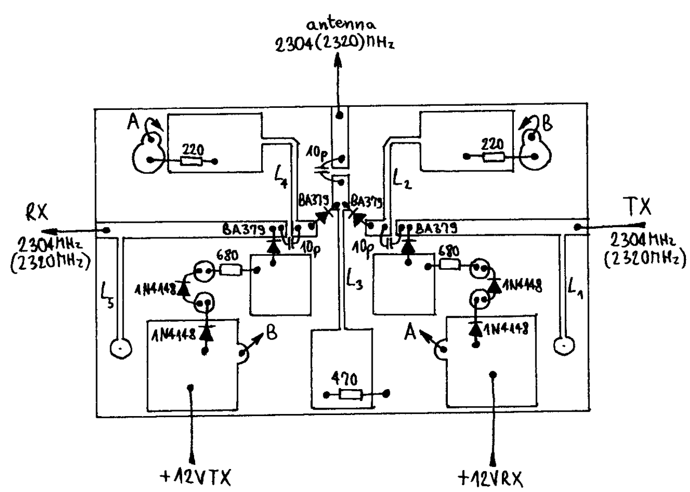
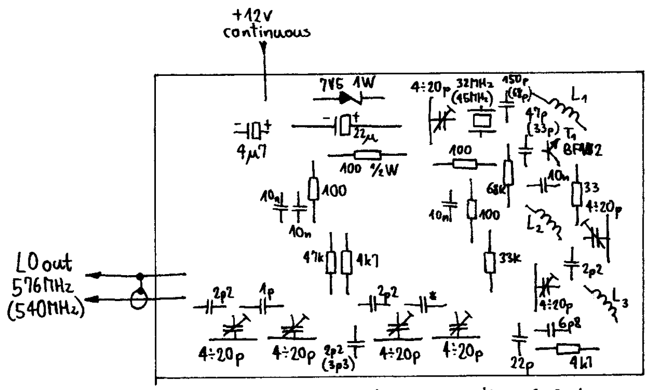
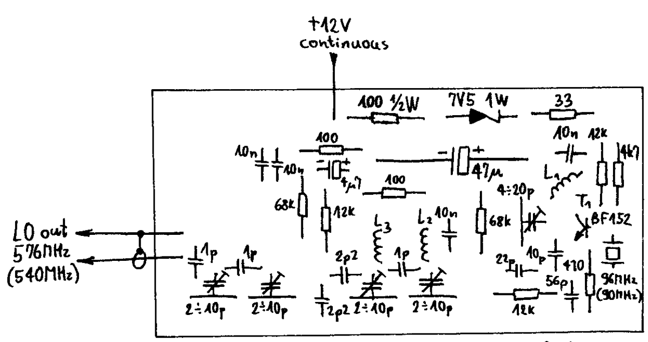


Fig. 40 - Components location of the RF antenna switch for 2304/2320 MHz.



T2 and T3 are mounted below the PCB to minimize lead inductivities!

Fig. 41 - Components location of the local oscillator module for 32 (45) MHz crystals.



T2 and T3 are mounted below the PCB to minimize lead inductivities!

Fig. 42 - Components location of the local oscillator module for 96 (90) MHz crystals.

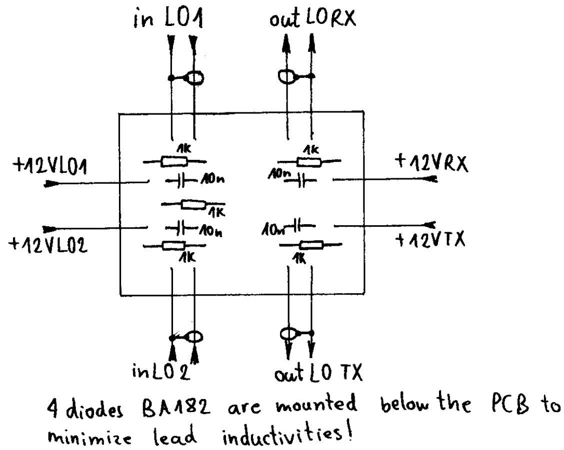


Fig. 43 - Components location of the LO switch.

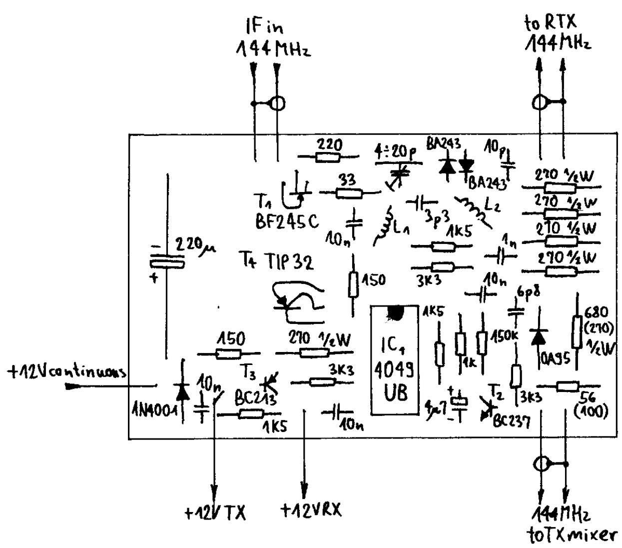


Fig. 44 - Components location of the VOX module.