use philips c's for coupling etc.

trimmers all Philips

hardline best to use

Matjaž Vidmar, YU 3 UMV now S53MV

Microstrip Transverters for 23 and 13 cm Part 1

1. INTRODUCTION

The microstrip technology is certainly well known to anybody working in the RF or microwave field, since most of both professional and amateur equipment is, at least, partially built using it. Due to its widespread use, the name "microstrip", itself does not really say much about its application in 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 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 cannot easily be solved by theoretical tools or CAD programs like (real world) lossy laminates having an anisotropic dielectric constant ε, or semiconductor devices operating in their non-linear region (mixers, varactor and transistor multipliers, power amplifiers). Of course there are other even less predictable factors, such as, the influence of various shields and / or the resonances of the metal case actually containing the microstrip circuit. Practical experiments are therefore necessary in any case, even with the best CAD program.

Fortunately, we amateurs only have moderate requirements such as narrow-band operation or gain tolerances. Since most of components used in our designs are usually not sufficiently characterized at microwave frequencies in data sheets, such as 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 physically 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 λ / 4 or λ / 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 frequency rejection. A large number of resonators in series calls for a low-loss, expensive teflon laminate as a 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, making the duplication in amateur conditions considerably more difficult.

In the transverters described in this article, a different solution was sucessfully 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 discovered 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 (as 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 multi resonator 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.

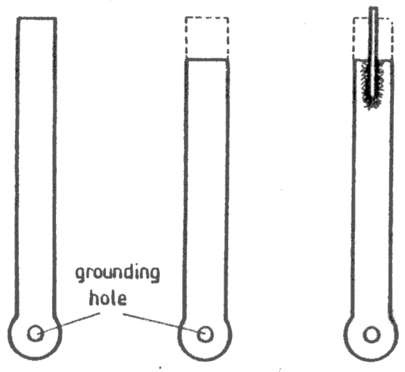


Fig. 1: Tuning a λ / 4 microstrip resonator

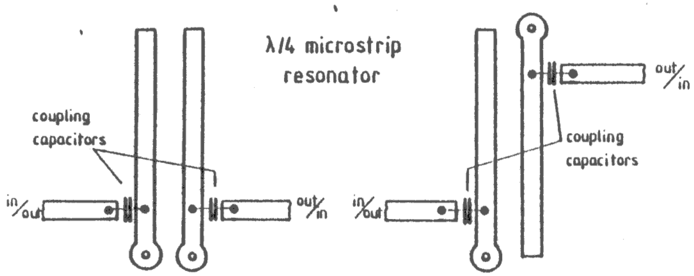
Left: Original λ / 4 microstrip resonator as etched on the PCB

Middle: The resonant frequency is increased by removing part of the microstrip with the aid of a sharp knife

Right: The fine frequency tuning is accomplished by soldering a short length of wire at the resonator's "hot end"

To obtain a usable value of coupling between two λ / 4 resonators there are two basic arrangements: both microstrips parallel and oriented in the same direction (**fig. 2 a**) and both microstrips parallel but oriented in opposite directions (**fig. 2 b**). 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 glass-fiber epoxy FR 4 laminate which has noticeable losses at 2304 MHz (this is probably its frequency limit for high Q selective circuits). Except for the RF power amplifiers all the transistors are packaged in low-cost plastic cases. **The reproducibility can only be enhanced by designing out the needs for critical components 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 5 mm) 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.



Figl. 2:
Two resonator filters in microstrip technology
Left: parallel microstrips
Right: opposed microstrips

2. BLOCK DIAGRAMS

The block diagrams of the 23 cm and 13 cm microstrip transverters are shown in **figs. 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 feeding 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 narrow-band 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 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 only, is required, the LO module output may be connected as in the 23 cm transverter.

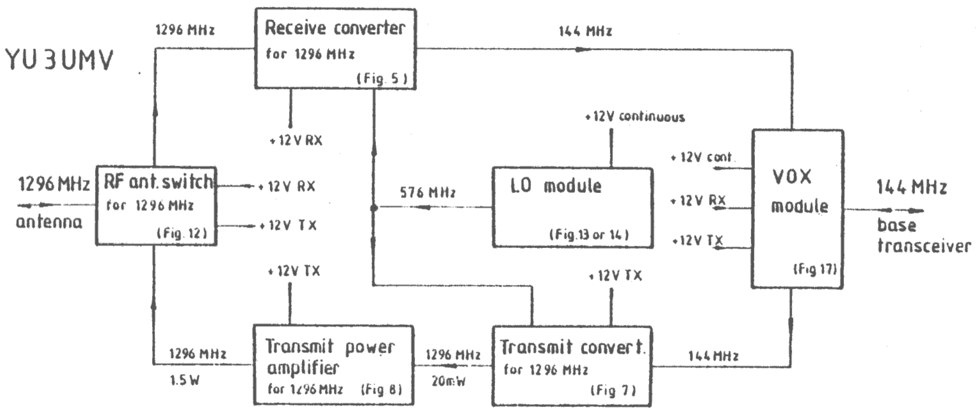


Fig. 3: Block diagram of the 1296 / 144 MHz transverter

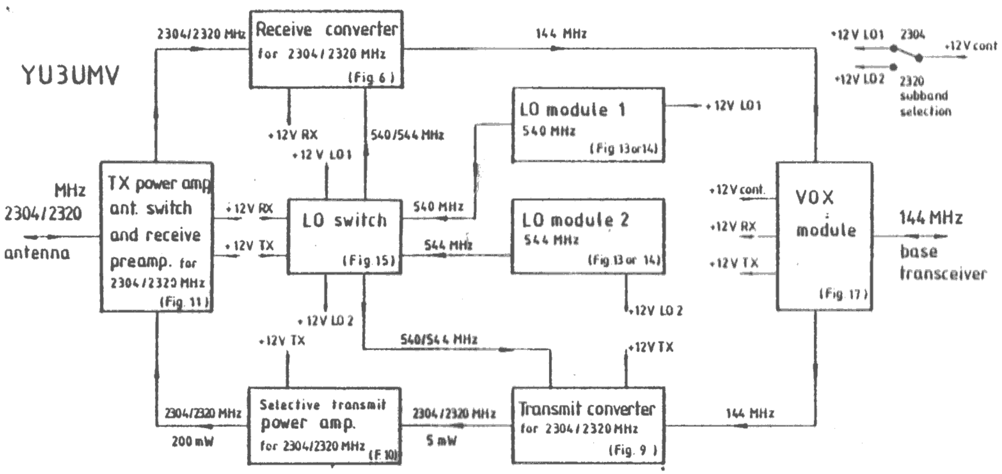


Fig. 4: Block diagram of the 2304 / 2320 / 144 MHz transverter

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 provided by the two resonator interstage filters. The input resonator only provides a broad selectivity to reject far removed 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 λ / 4 lines (L_{10} and L_{16} in fig. 5, L_{10} and L_{22} in 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 inductance.

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 (λ / 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 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 attained 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, 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 in 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

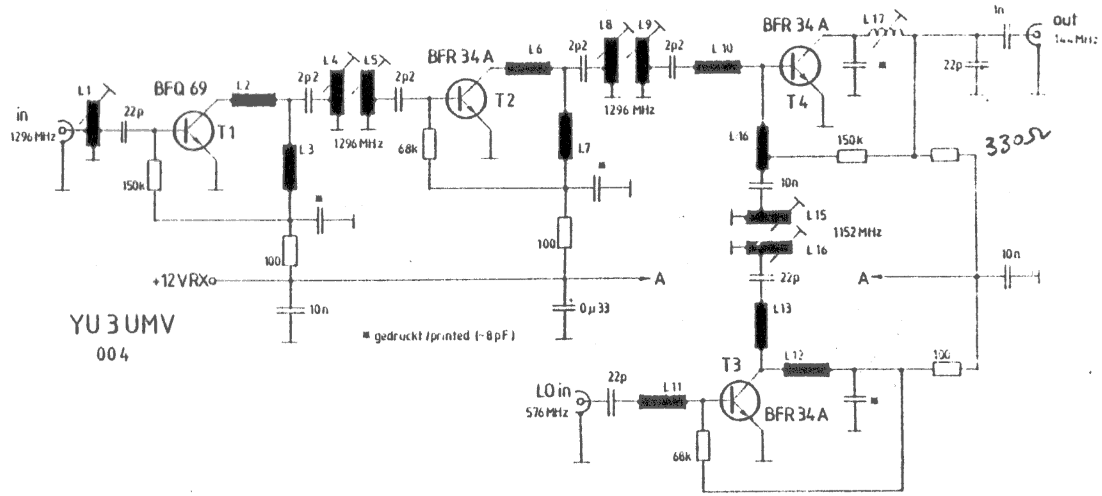


Fig. 5: Receive converter for 1296 MHz. All transmission lines L_1 to L_{16} are printed on the PCB. L_{17} self supporting, I. d. = 5 mm, wire = 0.7 CuI, L_{17} 8 turns, variable spacing!

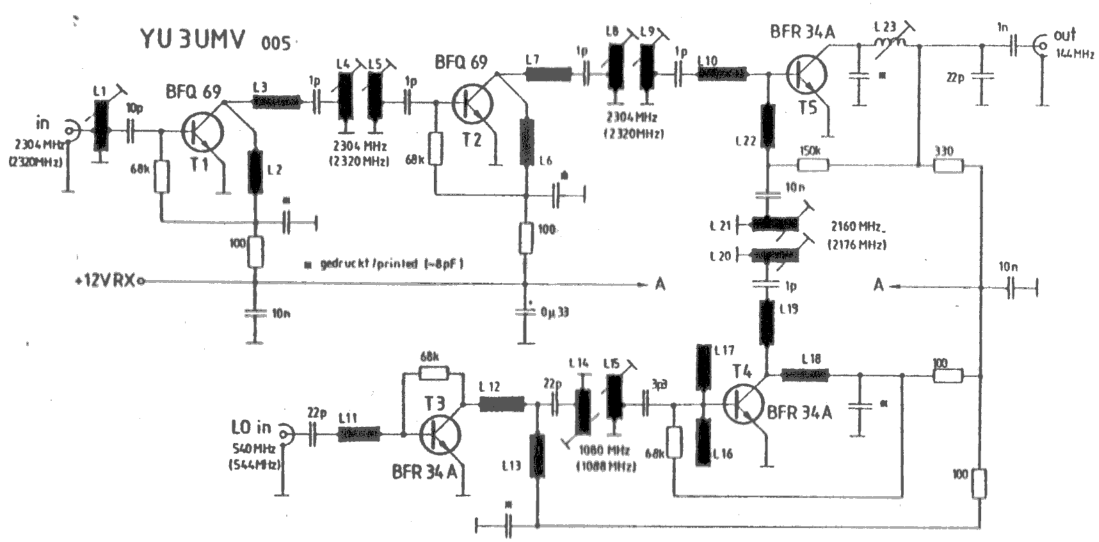


Fig. 6: Receive converter for 2304 / 2320 MHz. All transmission lines L_1 to L_{22} are printed on the PCB. L_{23} self supporting; i. d. = 5 mm, wire = 0.7 CuL, L_{23} 8 turns, variable spacing!

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 as it is more convenient to perform the base station TX-signal attenuation in two consecutive steps thus 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 five in total λ / 4 resonators are completely sufficient to attenuate all unwanted signals such as 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 in **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 BFR 96 transistors, which can provide 6 to 7 dB power gain at 1296 MHz depending on the output power level and bias conditions. The first BFR 96 (T_1) operates in class AB supplying about 100 mW to the secound BFR 96 (T_2). This transistor increases the power level to about 400 mW. This is probably the maximum **safe** power level a plastic-case transistor, like the BFR 96, 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), 2 N 5944, 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 λ / 4 choke since a single printed microstrip λ / 4 choke was not sufficient.

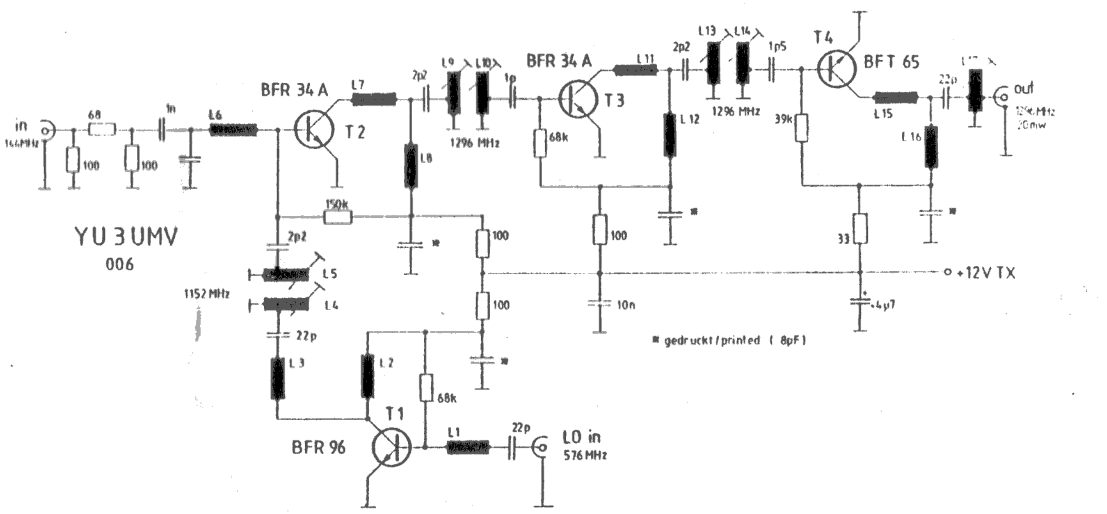


Fig. 7: Transmit converter for 1296 (1270) MHz. All transmission lines L_1 to L_{17} are printed on the PCB.

5. TRANSMIT CONVERTER AND POWER AMPLIFIERS FOR 13 cm

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 micro-strip 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 thereby increasing the useful level to about 50 mW.

The following two stages (T_3 and T_4) employ BFR 96 transistors. With 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 biased in class A to obtain the maximum possible gain. When bipolar transistors are operated in class A,

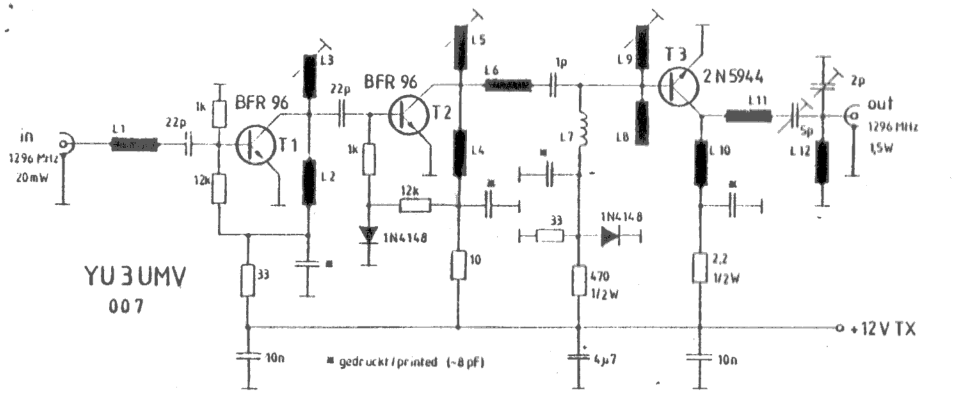


Fig. 8: Transmit power amplifier for 1296 (1270) MHz. All transmission lines L_1 to L_6 and L_8 to L_{12} are printed on the PCB. L_1 is self supporting, i. d. = 3 mm, wire = 0.7 CuL, L_7 8 turns, spaced to 12 mm

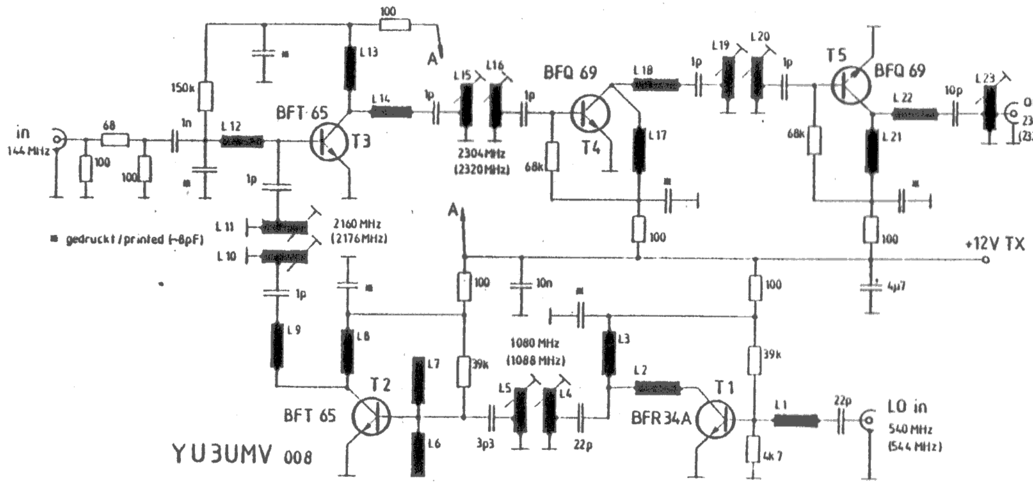


Fig. 9: Transmit converter for 2304 / 2320 MHz. All transmission lines L_1 to L_{23} are printed on the PCB

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 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 BFR 96 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 antennas switch, onto a single printed circuit board (see **fig. 11**). The RF power transistor (T_1) BFQ 34 requires a fairly complex matching network to allow the use of similar microstrip tuning elements as in the lower level stages. T_2 BC 213 is a bias regulator for the RF power transistor. It stabilizes the operating point of the RF transistor around the optimum value of 140 to 150 mA of DC collector current in order to counter the "negative rectification" problem. The obtainable 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 BFQ 69.

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 construction. Since they are electromechanical components, they are also subjected to wear and are thus potentially unreliable. Fortunately, for low-power transmission only, for moderate insertion loss and cross-talk requirements, a solid state replacement is readily available. Popular PIN diodes like the BA 379 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 BA 379 has, however, another very interesting property: the PIN-diode structure is also very slow to turn on. Therefore RF voltages

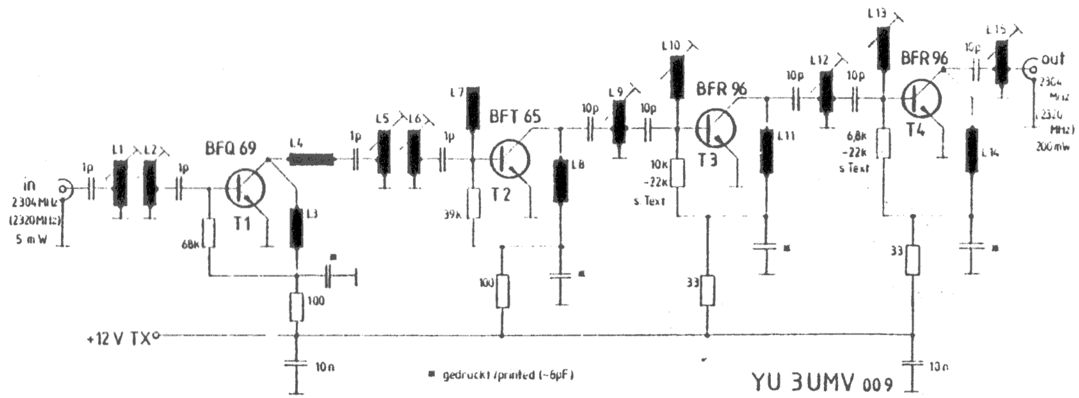


Fig. 10: Selective transmit power amplifier for 2304 / 2320 MHz. All transmission lines L_1 to L15 are printed on the PCB

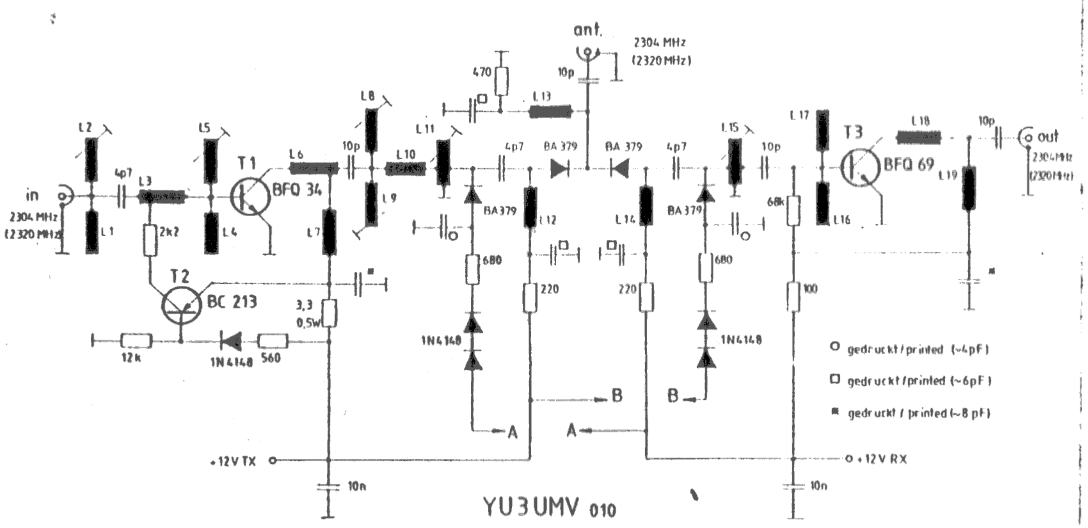


**Fig. 11: TX power amp, antenna switch and receive preamp 2304 / 2320 MHz.
All transmission lines L_1 to L_{19} are printed on the PCB**

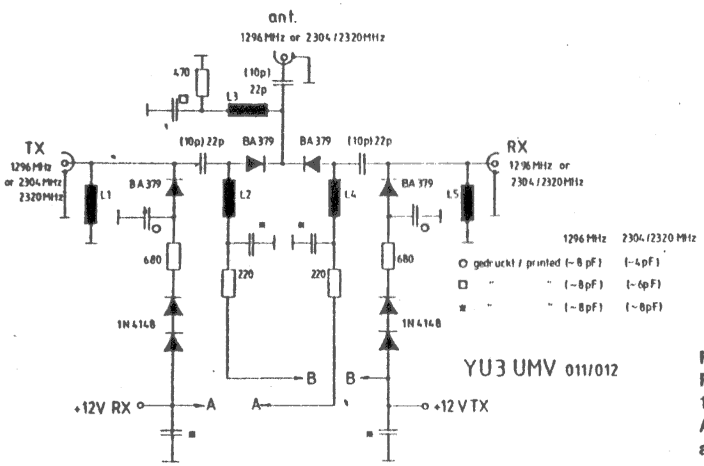


**Fig. 12:
RF antenna switch for
1296 MHz or 2304 / 2320 MHz.
All transmission lines L_1 to L_5 are printed on the PCB**

of sufficiently high frequency can not switch the diode on even if the positive halfwave amplitude greatly exceeds the diode turn-on voltage of about 0.7 V. In our particular application, this means that these diodes do not need any reserve DC bias in the non-conducting state, even if an 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 prasitics, the insertion loss of the RF antenna switch shown in **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 transition, microstrip radiation and other causes. The cross-talk attenuation is sufficient for the application shown, limiting the 13 cm TX power to below 1 W to avoid RX frontend damage.

The RF antenna switch is controlled by the two supply voltages + 12 V RX and + 12 V TX switched by the VOX module. The silicon diodes 1 N 4148 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.

Concluding part in de next edition.

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Matjaž Vidmar, YU 3 UMV

Microstrip Transverters for 23 and 13 cm Part 2

7. OSCILLATOR MODULES

Two different oscillator modules were developed, primarily in order to utilize various types of crystals. One of them (fig. 13) works with third overtone crystals in the range 32 to 45 MHz and the other unit was designed for fifth overtone crystals working at 90 or 96 MHz (fig. 14). Both types of modules supply a power output of 10 MW at 576 MHz, 540 MHz or 544 MHz.

The oscillator injection frequency should, in any case, always be chosen somewhat lower. If the start of the narrow-band operating range of the microwave band when translated, overlaps the start of the 144 MHz amateur band, then interference from high-power, two-meter stations is the likely outcome since most of the two-meter TRs available nowadays, and used as a base equipment, are not sufficiently screened. On top of this, most modern PLL synthesizer transceivers do not allow themselves to be tuned below 144.0. If the transverter crystal should drift in the correct direction, a portion of the microwave band would be

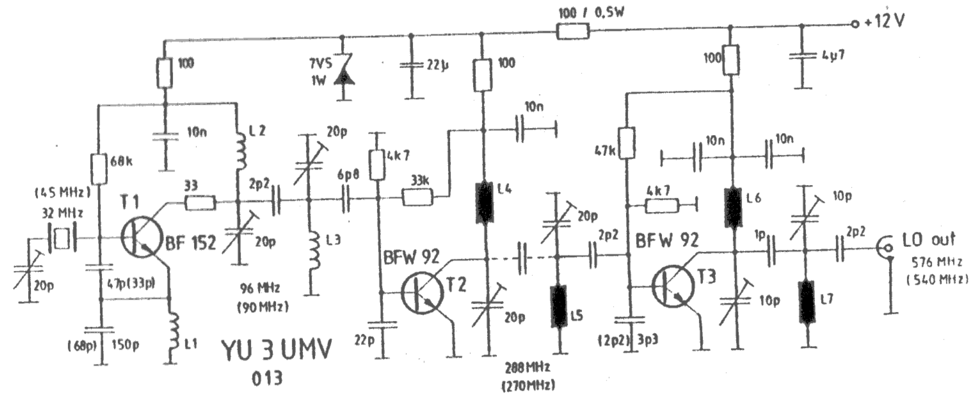


Fig. 13: Local oscillator module for 32 MHz crystals (45 MHz values in brackets)
L 1 to L 3: self supporting 0.7 mm CuL wire, 5 mm int. dia.
L 1: 10 (11) turns, L 2: 6 (7) turns, L 3: 6 turns
The inductors L 4 to L 7 etched.

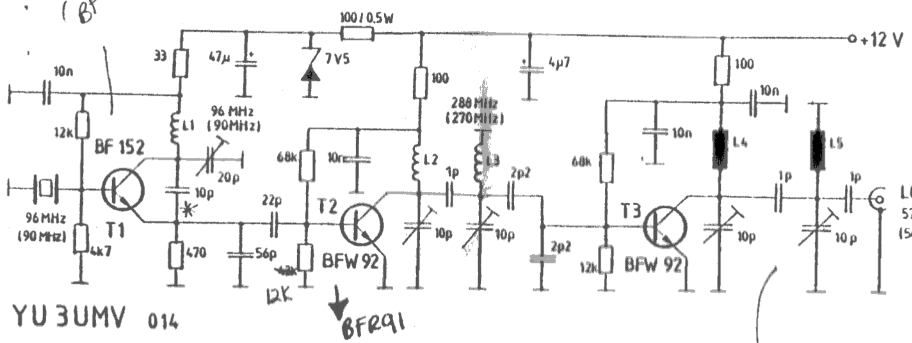


Fig. 14: Local oscillator module for 96 MHz crystals (90 MHz values in brackets).
L 1 to L 3: self supporting 0.7 mm CuL wire, 5 mm int. dia.
L 1: 6 turns, L 2, L 3: 2 turns
The inductors L 4 and L 5 are etched.

unattainable. Therefore, the first 400 kHz of the microwave band, the part normally used for narrow-band activity, should be translated to the 144.600 to 145.000 MHz range which is mainly free from hectic contest traffic.

Using third overtone crystals, a reliable oscillator can be fairly designed. The one shown in **fig. 13** should oscillate without any need for special adjustment. The inductance L 1 in the oscillator T 1 emitter prevents the crystal from oscillating at its natural frequency. The oscillator has been designed for parallel-resonance specified crystals in order that in both versions a higher value for the crystal trimmer capacitor will be required.

The collector of the crystal oscillator transistor should be at ground potential as far as the crystal frequency is concerned. That is conveniently arranged by including an harmonically tuned LC circuit in the collector which results in a fourth harmonic selection and thereby saving a multiplier stage.

Both the following stages are conventional frequency multipliers. The inductors 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 sufficiently close so as to obviate the need for a coupling capacitor. The supply voltage, both for the oscillator and for the following stage is, of course, highly stabilized. Crystals in the range about 90 MHz oscillate normally at the fifth overtone. Suitable oscillators are more critical especially when the crystal must be pulled in order to provide a specific fine-tuning range. The oscillator circuit shown in **fig. 14** was designed for series-resonance crystals. In any case, it is necessary that the trimmer for L 1 be very carefully tuned in order that a stable working condition for the oscillator at the desired frequency is achieved.

Also this oscillator is followed by two frequency multiplier stages but the difference to the formerly described unit is that the both inductors for 288 MHz, L 2 and L 3, are air-wound self-supporting coils instead of being etched from the PCB. Also, in this unit, the oscillator supply voltage is extra stabilized with a 7.5 V zenner diode.

As already mentioned when the block diagrams were described, a diode switch (**fig. 15**) is employed to effect the switching of the two oscillator units to the transmit and receive converters.

It should be mentioned that no DC blocking capacitors have been provided in the signal lead as

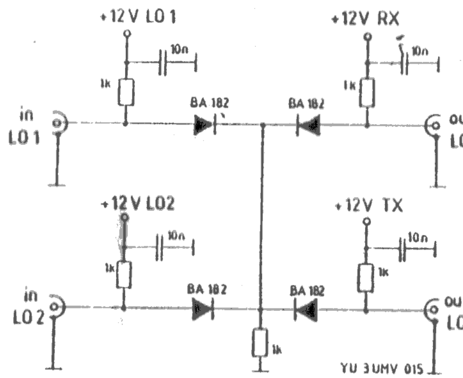


Fig. 15:
Oscillator module electronic switch

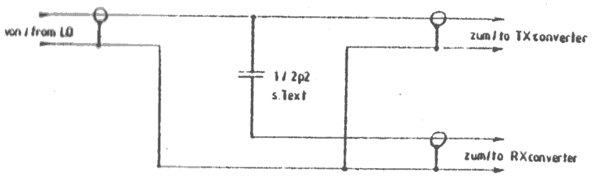


Fig. 16:
A simple method of splitting the oscillator signal to feed both transmit and receive mixers.

they already exist in the outputs of the oscillator modules and in the inputs of the transmit and receive converters.

If only one oscillator module is to be employed, the simpler circuit of **fig. 16** should be used. In order to achieve a usable power division, a certain experimentation with the value of capacitors and cable lengths may be necessary.

8. VOX MODULE

The circuit diagram for the voice operated switching (VOX) is shown in **fig. 17**. The largest proportion of the base station's output power is dissipated in an attenuator consisting of several resistors. The voltage across the attenuator is rectified by an OA 95, or similar germanium diode, for control purposes. The rectified voltage switches on T 2 which discharges the 4.7 µF capacitor through a 1 kΩ resistor in a relatively short period of time. The charging of this capacitor is much slower owing to the presence of the 150 kΩ resistor. These two resistors determine the time constants of the VOX circuity: a quick switch-on time for the transmitter and a delayed return to the receive condition in order to bridge natural pauses in SSB speech.

The voltage on the aforementioned 4.7 µF capacitor controles a 4049 UB (I 1) CMOS inverter DC amplifier. This amplifier has a built-in hysteresis and controls both PNP supply switches, T 3 and T 4. The PNP switches were chosen because of the low voltage drop across them as opposed to that of the normal employed switching transistors. Of course, the working + 12 VRX and + 12 VTX supply lines must not be grounded inadvertently, otherwise the switching transistors will be destroyed.

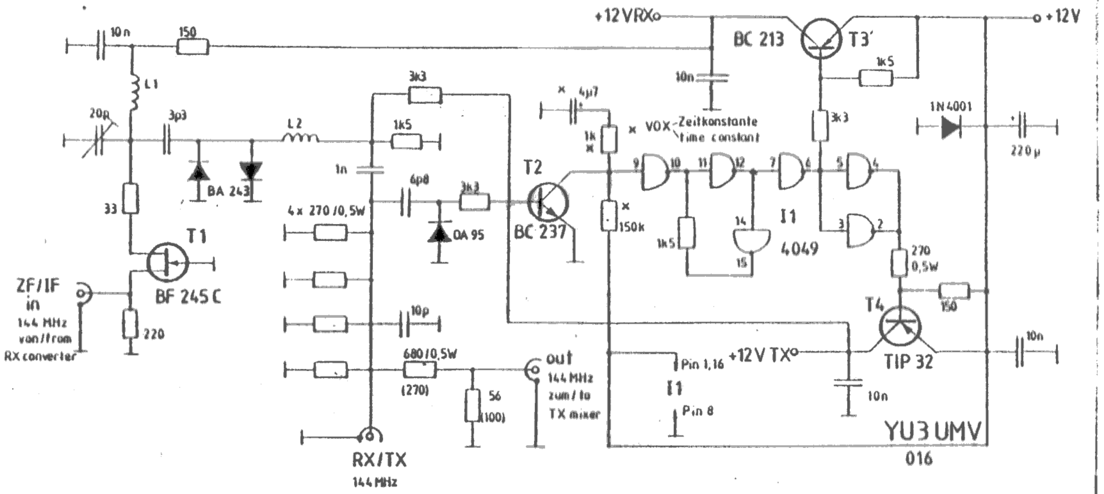


Fig. 17: The VOX module (values in brackets for the 2304 / 2320 MHz transverter)
L 1, L 2: self supporting 0.7 mm CuL wire, 5 mm int. dia.
L 1: 4 turns, L 2: 6 turns.

The VOX module also contains an attenuator which reduces the base station's power in order that it may be applied to the send converter mixer. It should be observed that the 13 cm transmit mixer requires a higher 2 m driving signal power than the 23 cm mixer necessitating the alteration of a few resistances. The values for 13 cm working are given in brackets (fig. 17). The attenuator is so dimensioned, that 1 W of transmit power at 144 MHz drives the transverter to full power output and up to 3 W is dissipated in the 4 x 270 Ω/0.5 W attenuator.

As the VOX switching circuitry is not able to fore-cast when the 2 m transceiver will be switched to transmit, a protection circuit, in the form of a power limiter, is necessary in order to protect the receive mixer from burn-out during the initial switch from receive to transmit. The function of this limiter is made more efficacious by the subsequent provision of a DC bias to the two BA 243 diodes.

The IF amplifier using T 1 has the task of compensating for the loss of power, consequent on conversion and the protection measures to the final output power.

For an operational constrol, two LEDs with suitable dropping resistors can be wired between the + 12 VTX and + 12 VRX supply lines and earth.

9. CONSTRUCTION

As mentioned in the introduction, all microwave circuits of both transverters using microstrip techniques are realized using 1.6 mm thick epoxy-glass PCB material designated FR 4. Various circuit board patterns and their related component placing diagrams are shown in **figures 19 through 27** in actual size. The top side only is shown, of course, as the underside consists of a film of unetched copper. It should be observed that there are two different printed circuit boards for the PIN diode antenna change-over switches,

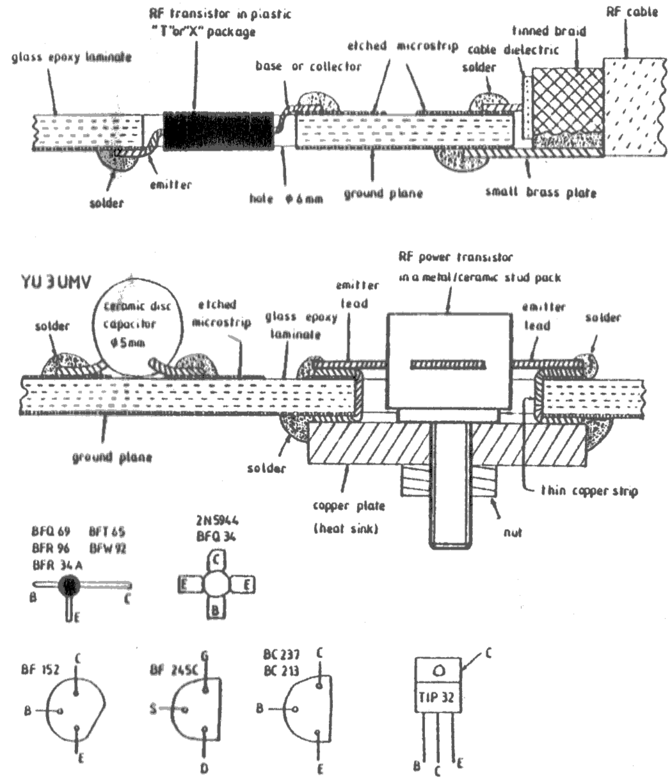


Fig. 18: Mounting instructions for various components

one for 23 cm and the other for the 13 cm band, but the circuit diagrams are identical.

The component layout plans plus diagrams do not, of course, give sufficient information in order that a working replica may be produced and this is normally the case when working at microwave frequencies. Therefore the diagrams of **fig. 18** are given to indicate how various components are physically located upon the board. The plastic packaged transistors are sunk into 6 mm dia holes in the PCB which have been bored at the marked spots. It should be noted that all transistor connections should be as short as possible – especially the emitter. The emitter lead inductance is responsible for most of the transistor's power loss at microwaves.

The power transistors in the metal / ceramic packaging are let into 10 mm diameter holes drilled in the board together with the mounting studs. Prior to this, a strip of sheet copper having the same width as the emitter strip, is soldered around the hole which acts as a very low inductance contact between the two PCB faces. Also a 1 mm copper piece, at least 15 mm × 25 mm, is soldered over the PCB hole with a concentric boring of a suitable diameter to just accept the transistor stud. (**fig. 18**). This serves as the transistor heat-sink. The collector is then identified with a spot of paint put on the ceramic hub and the connection strips cut back to a suitable length. The transistor is then bolted on to its heat-sink complete with a spring washer. The other transistor strip connections can then be soldered onto the PCB.

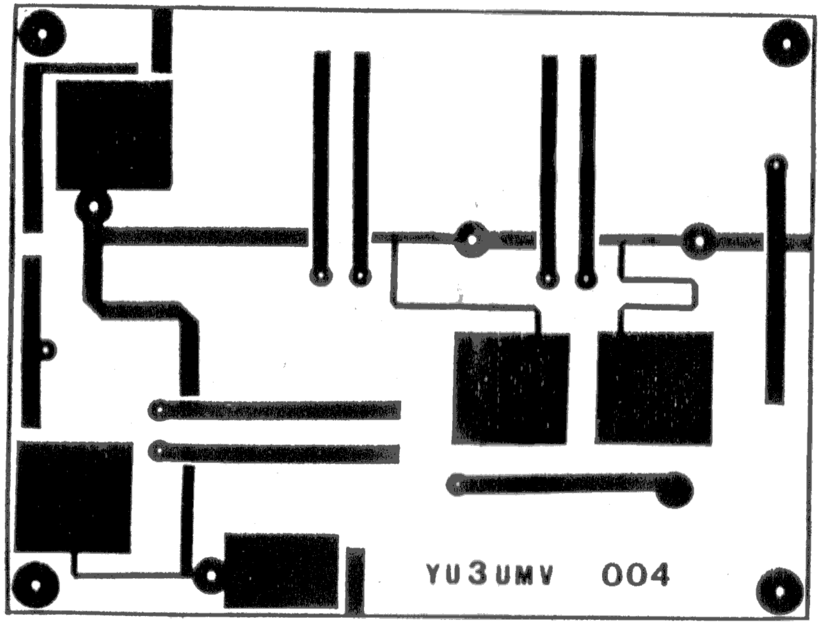
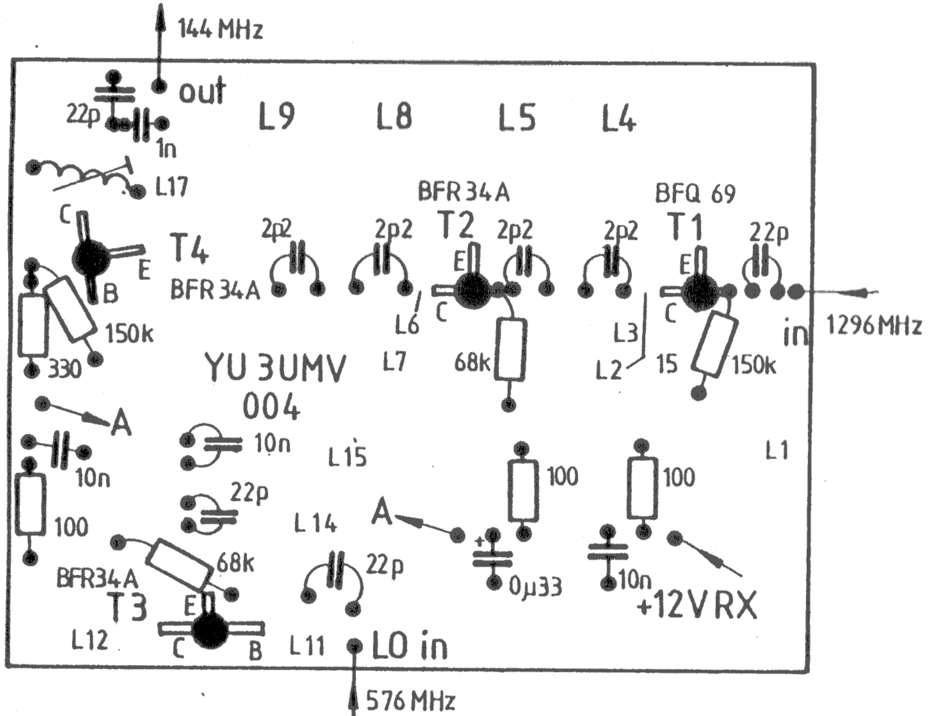


Fig. 19:
PCB YU 3 UMV 004 for 1296 MHz receive converter



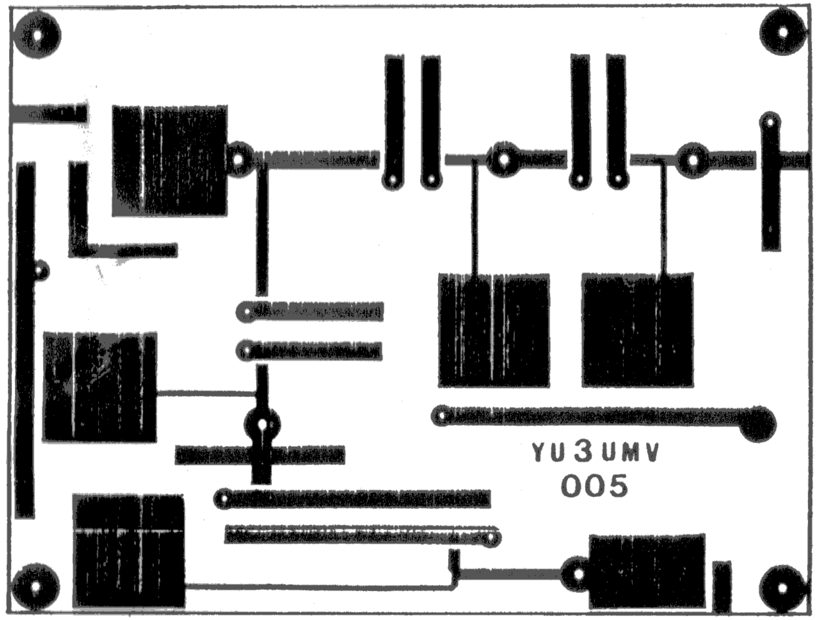
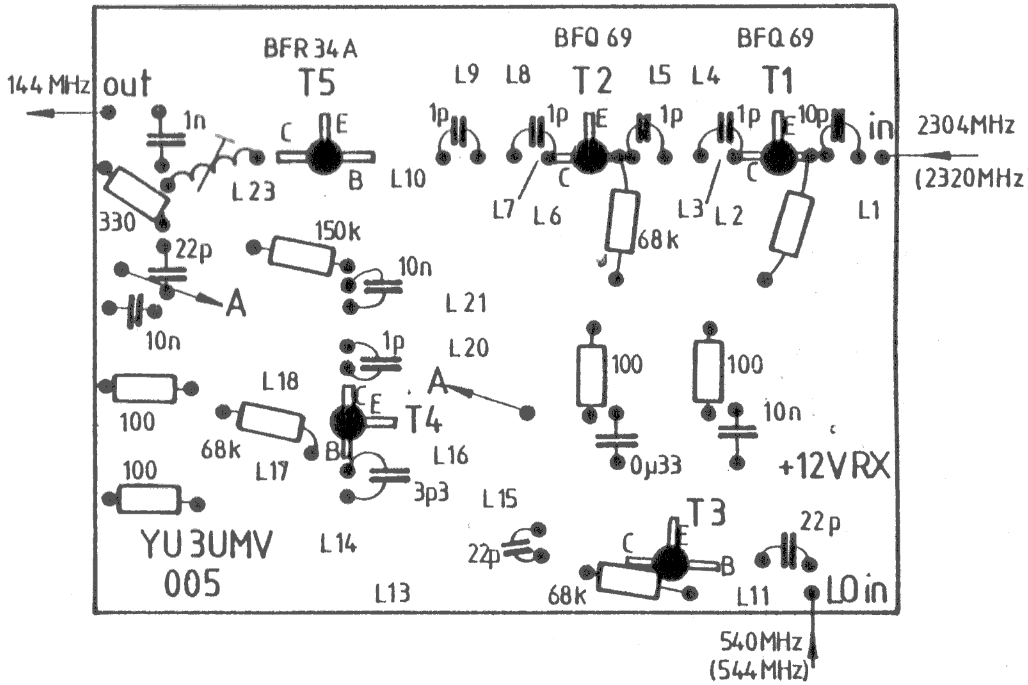


Fig. 20:
PCB YU 3 UMV 005 for 2304 / 2320 MHz receive converter



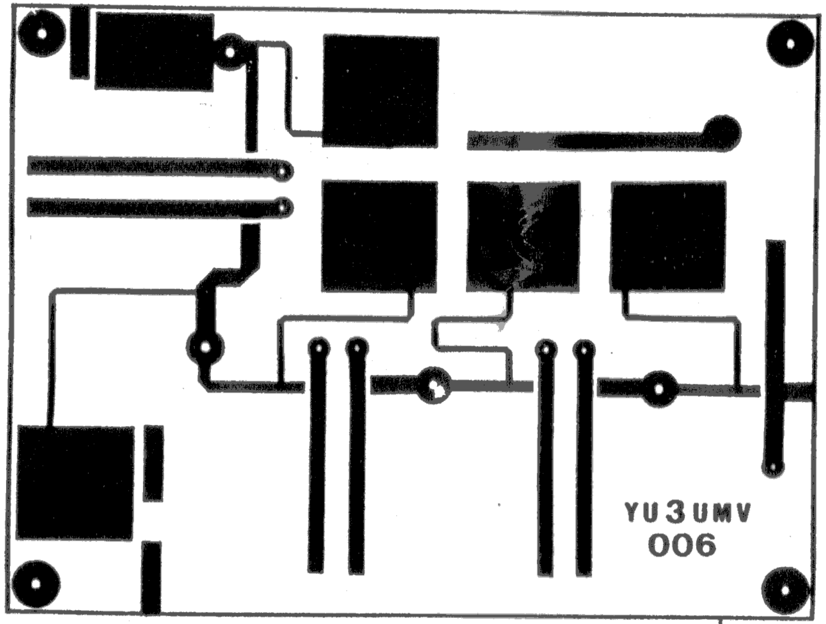
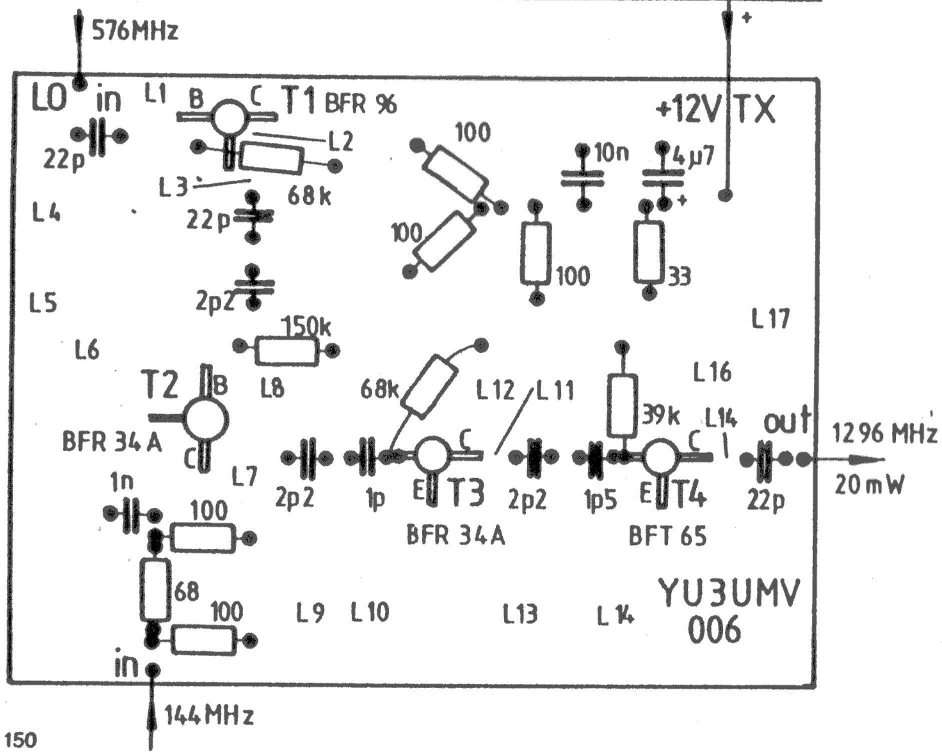
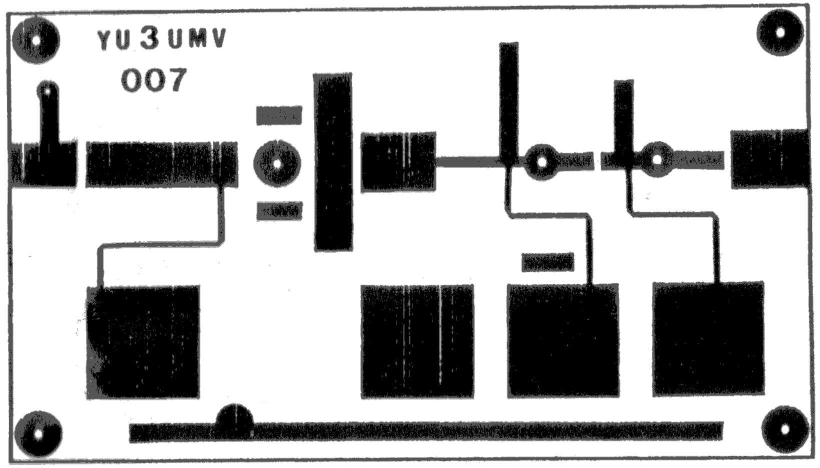


Fig. 21: PCB YU 3 UMV 006 for 1296 / 1270 MHz transmit converter





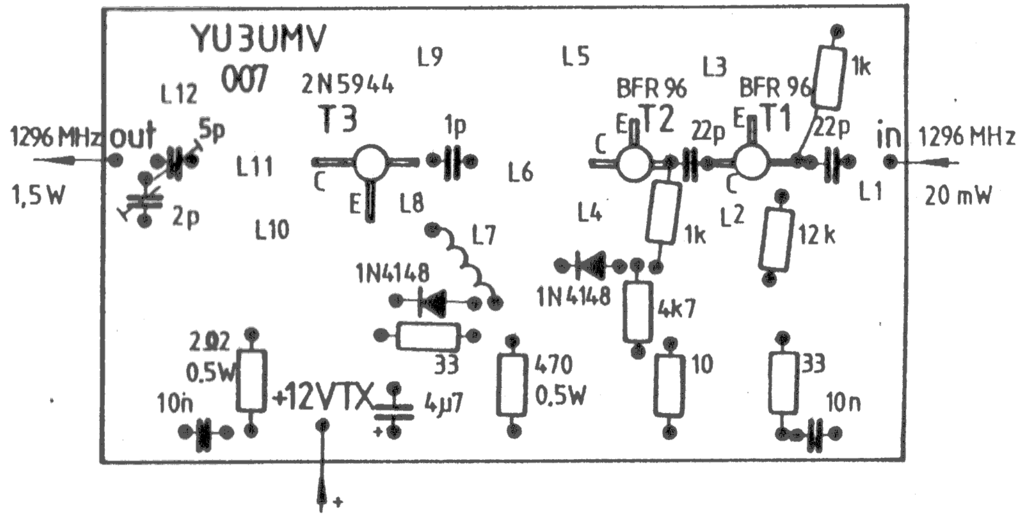


Fig. 22: PCB YU 3 UMV 007 for 1296 / 1270 MHz transmit amplifier

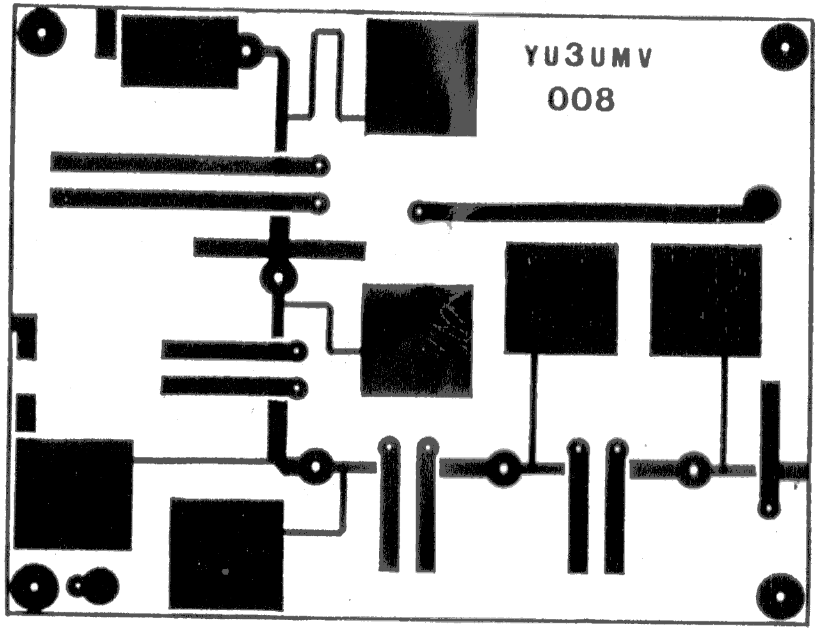
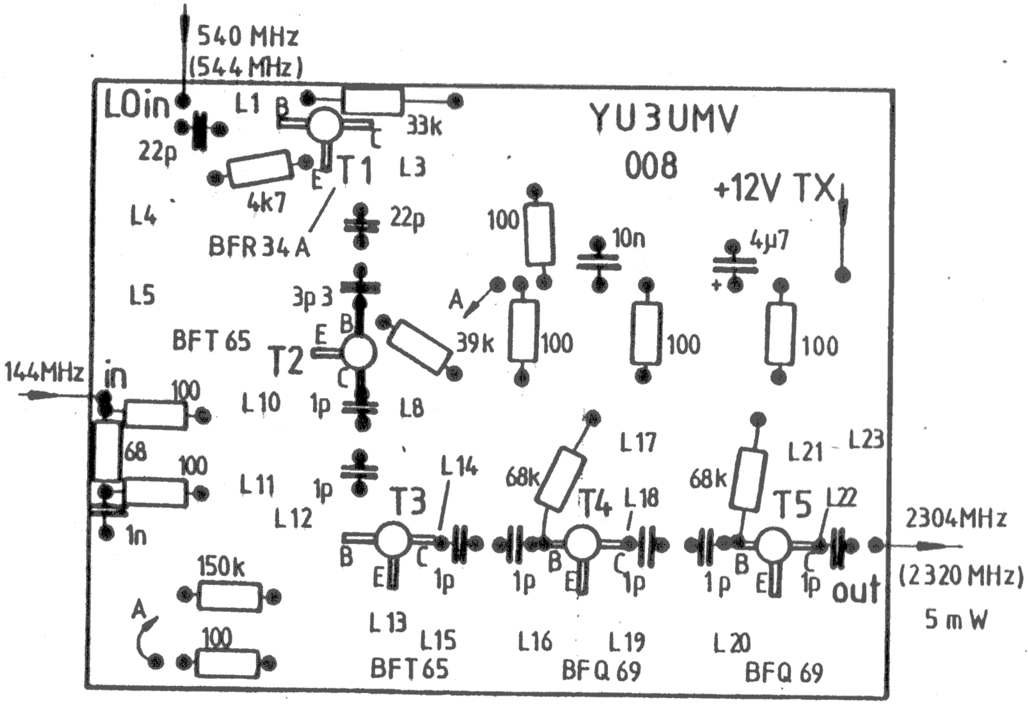
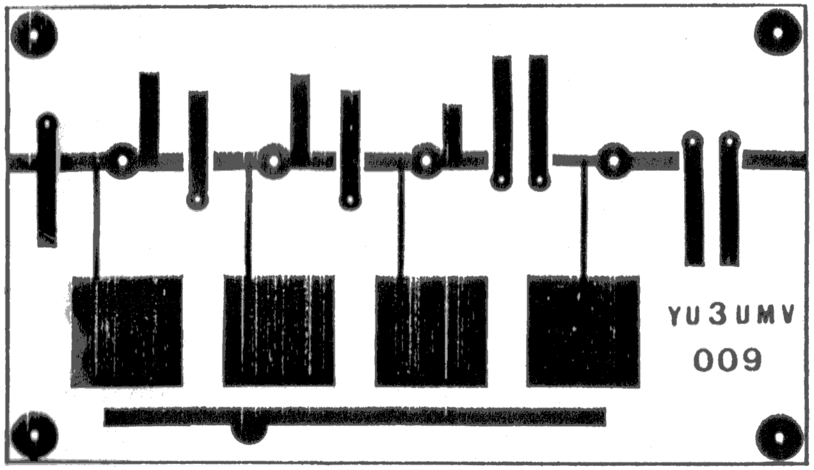


Fig. 23: PCB YU 3 UMV 008 for 2304 / 2320 MHz transmit converter





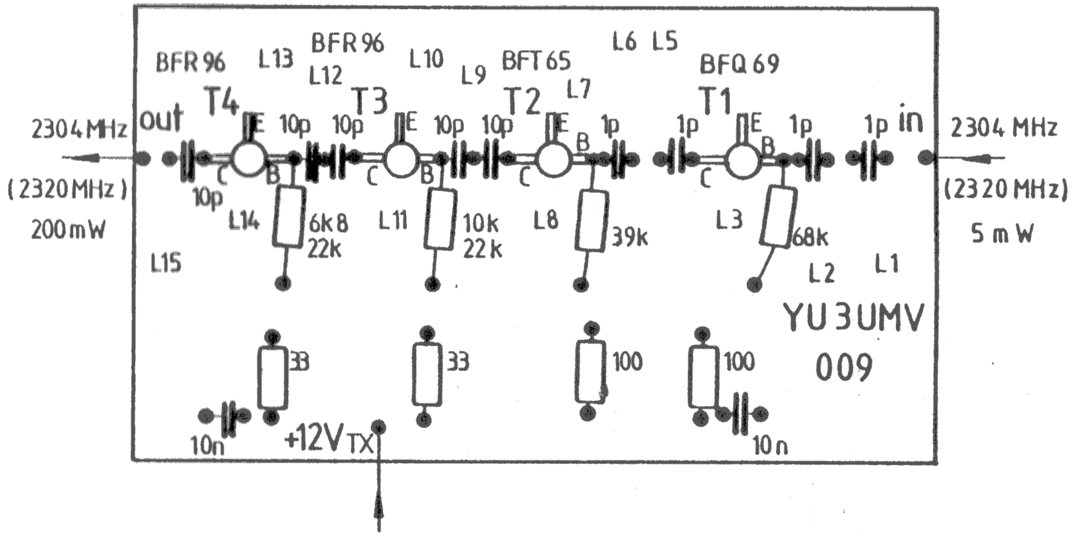


Fig. 24: PCB YU 3 UMV 009 for selective 2304 / 2320 MHz transmit amplifier

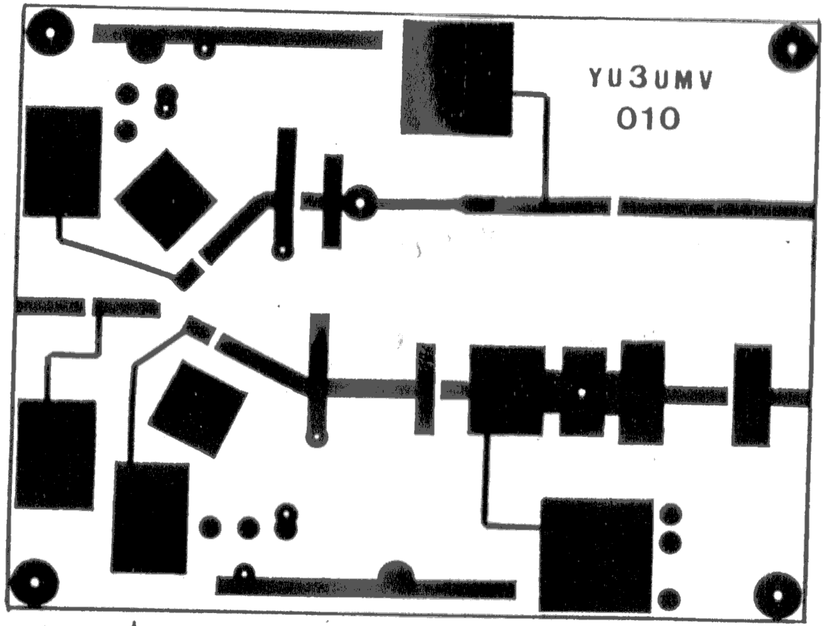
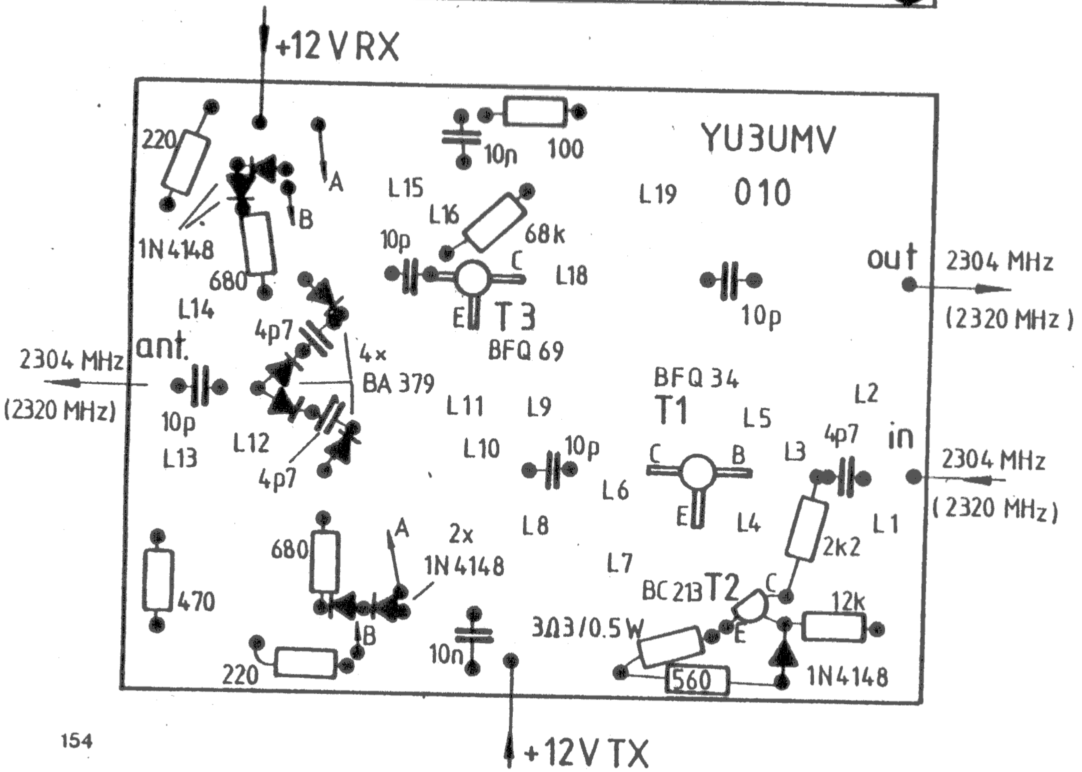


Fig. 25:
PCB YU 3 UMV 010 for transmit amplifier, antenna switch and receive preamplifier for 2304 / 2320 MHz



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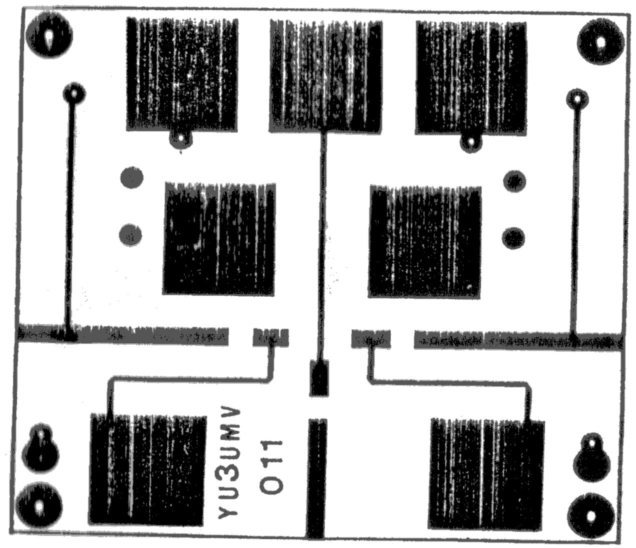
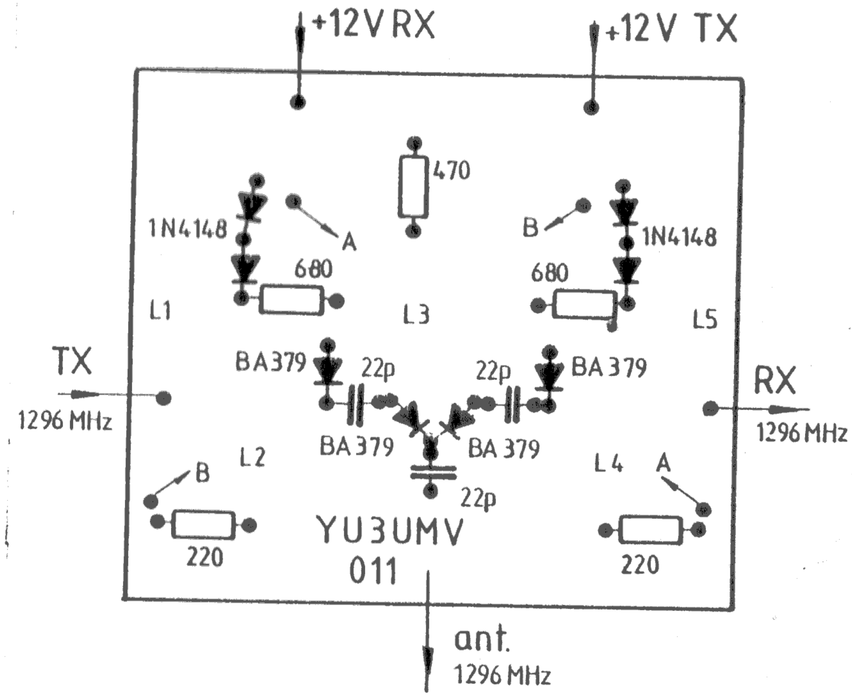


Fig. 26:
PCB YU 3 UMV 011 for 1296 MHz antenna switch



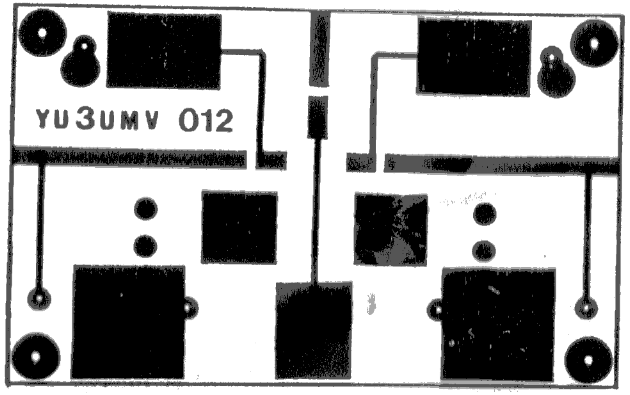
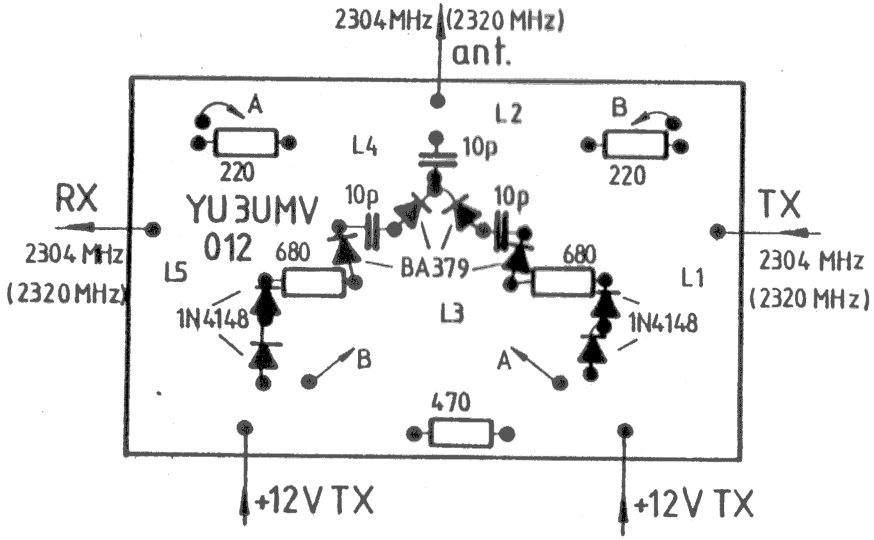


Fig. 27:
PCB YU 3 UMV 012 for 2304 / 2320 MHz antenna switch



The ceramic capacitors must be soldered in with the very shortest connections possible as indicated in **fig. 18**. The leads are, in fact, snipped off so that they protrude only 1 mm from the body of the capacitor and they are then completely covered in solder to effect the connection. A perfectly built transverter should show no trace of SHF capacitor leads.

All low value disc ceramic capacitors are of 3 mm to 5 mm diameter with the exception of the 1 pF capacitors which are pearl types. The higher value ceramics are multilayer types but are also of small format.

The decoupling for the supply voltage is accomplished by ceramic and tantalum pearl capacitors, one leg of which is passed through a 1 mm drilling in the PCB and soldered to the earth plane. Their position is not so critical and their drillings in the layout plan have not been indicated.

Resistors play no part in the microwave portion of the circuit, and where they exist their installation is

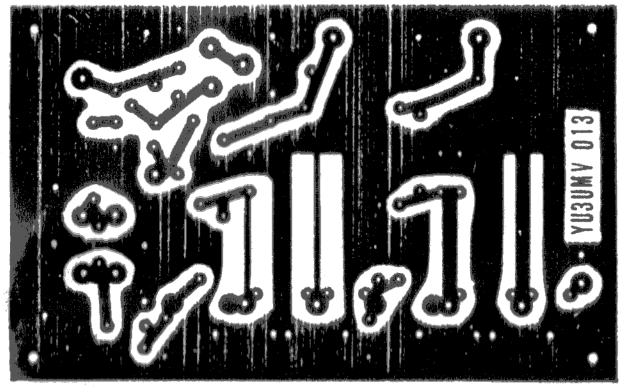
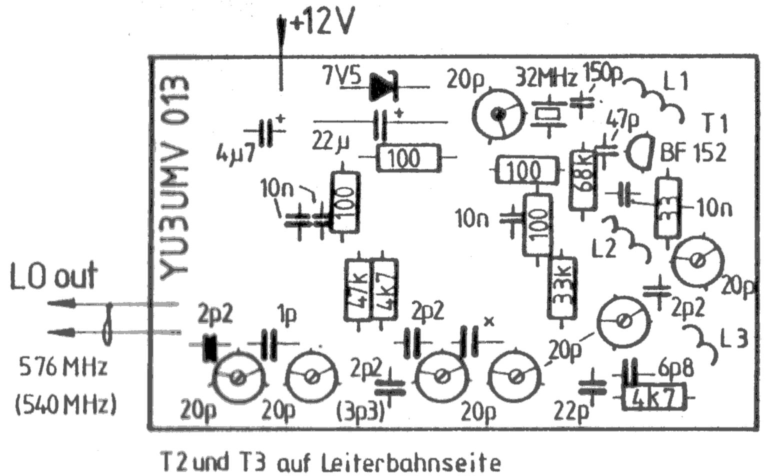


Fig. 28:
PCB YU 3 UMV 013 for oscillator module using 32 (45) MHz crystals



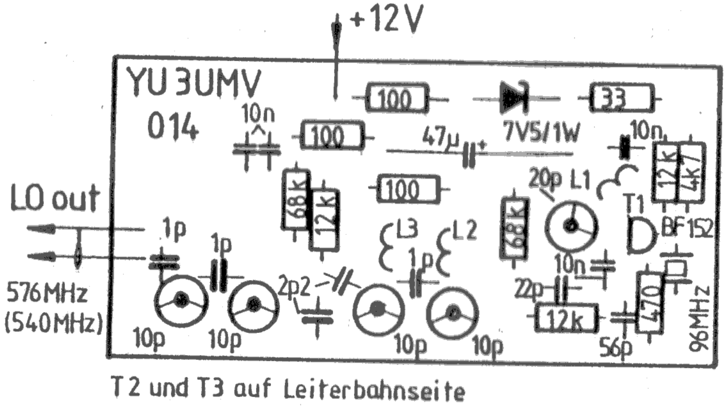
uncritical. The capacitance to earth of connecting leads has, however, to be considered and their length must be minimized to reduce self inductance.

Now comes a most important consideration, the λ / 4 microstrip resonator ground connections. These must be carried out exactly as shown in the prototypes, otherwise not only the resonant frequencies will be off but also the couplings to other active circuit elements will be too.

Always at the indicated spot, a 1.5 mm hole should be drilled. A short piece of 1 mm silvered copper wire is inserted and soldered to both planes. It should be borne in mind that this wire is located exactly in a current anti-node and its parasitic inductance has a marked effect upon the characteristics of the resonator.

The various modules are interconnected by means of short lengths of thin PTFE (teflon) coaxial cable (RG-188). Normal polyethylene coaxial cable (RG-174 / U) can be employed but it is more difficult to work with as the thin dielectric melts very easily with the application of heat. In any case, it is of the utmost importance that the cable ends are terminated at the microstrip circuit exactly as shown in **fig. 18**. Most problems of parasitic resonant effects are caused by inductances formed by the indirect nature of the cable ground

Fig. 29:
PCB YU 3 UMV 014 for oscillator module using 96 (90) MHz crystals



connections. First of all, a suitable length of cable is cut and both ends are prepared as shown, the inner and outer conducters being tinned. A 8 mm × 8 mm piece of tinned thin copper plate (brass will do also) is then soldered to the ground plane of the PCB – a slightly larger size will be required for RG-142 or RG-58 cable – and this is soldered to the coaxial screening mantle of the cable – the inner to the microstrip.

The oscillator modules and also the VOX circuitry can be constructed using ordinary single-sided PCBs. Their etching patterns and component layout diagrams are shown in **figs. 28 to 31** in actual size drawings. All components can be mounted in the usual fashion with the exception of the transistors BFW 92. These are mounted under the PCB in order that the deleterious effects of the lead inductance can be held to a minimum. All trimmers must be of high-quality plastic types – the ceramic trimmers are both mechanically and electrically inferior. The 2 pF to 10 pF trimmers in the 540 / 576 MHz resonator circuits can, in most cases, be replaced with 2 pF to 6 pF types, thereby obtaining a less critical tuning point.

The order of building the modules should follow the order in which the complete assembly would have to be aligned from scratch. First of all the modules which can be independently checked, are completed – the oscillators for example, can be built, checked and aligned first of all. Of course, any order of building is acceptable if the necessary test equipment is readily available but these instructions are intended for the average radio amateur constructor.

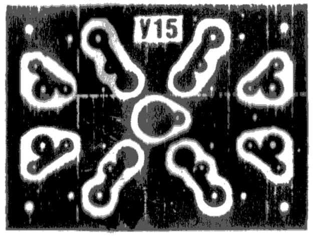
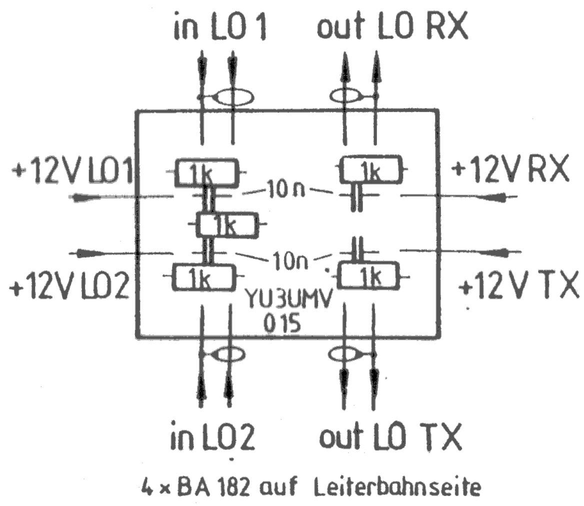


Fig. 30:
PCB YU 3 UMV 015: Change-over switch for oscillator modules



After all the modules have been successfully built, tested and aligned, they can be installed into a suitable metal housing. Internal screening is, in practice, superfluous – quite the contrary in fact, the walls and floors of the cabinet must be kept at least 3 cm away from microstrip circuits in order to inhibit detuning and unwanted coupling effects.

10. ALIGNMENT

The alignment was, in principle, described in the introduction. All the microstrip resonators should lie very close to the specified frequency after an accurate etching process on the board material has been carried out. Only a fine tuning is necessary. Should it be necessary to shorten a resonator because its resonant frequency is too low, then it should only be done by cutting off 0.5 mm at a time, and then testing.

Before testing the resonant frequency of a microstrip circuit, it is as well to bear in mind that the permittivity of glass-epoxy PCB material has a positive temperature coefficient, which means, that as the board is heated, its natural frequency tends to decrease. Therefore, always allow the board to cool to normal ambient temperature before testing (or tuning) following operations with the soldering iron.

The multiplier chain of the heterodyne oscillator is aligned first. All trimmers and / or resonators are tuned to obtain maximum output at the specified frequency. The frequency should be checked by a counter or by a Lecher-line resonant wave meter to ensure that maximum power is being delivered at the correct frequency.

Bipolar transistors, working as mixers and multipliers in a non-linear condition, should be checked following any alteration to the working DC conditions – particularly the base-emitter voltage. Owing to the rectifying action of the base-emitter junction, the applied RF signal develops an additional bias voltage which can be negative (peak to peak) by a few volts. The BE diode of the following transistor behaves then as an RF probe exactly where it is most desired – at the output of the stage which is to be aligned.

The small signal stages of the receive converter should be aligned with the aid of a suitable signal source and a two-meter receiver fitted with a sensitive S-meter. Since the microstrip resonators have already been aligned for the specified frequency a white noise source can be employed as

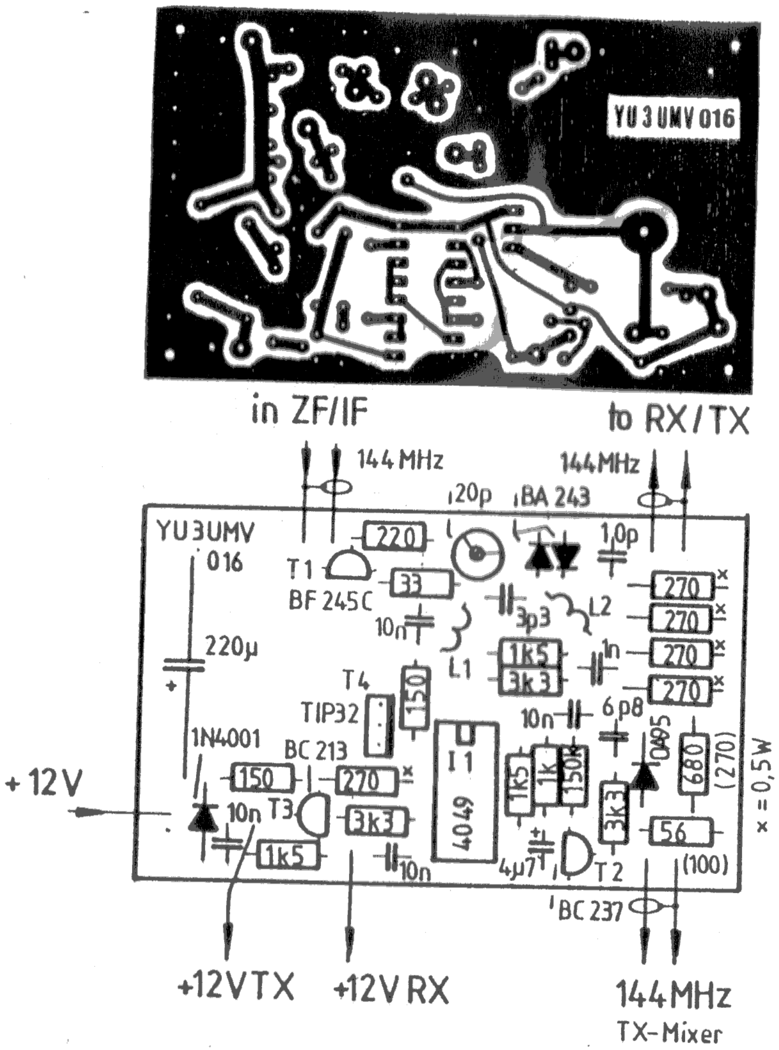


Fig. 31: PCB YU 3 UMV 016: VOX module

a signal generator. All resonators can simply be tuned for maximum possible output as in this transverter the bipolar silicon transistors used at frequencies above 1 GHz show no discernible differencies between noise and signal matching.

To align the transmit converter, an SHF milliwatmeter should be connected to the output. This could take the form of a 50 Ω termination, across which is a Schottky diode detector and suitable indicating meter. As the local-oscillator injection frequency lies very close to the desired mixer output signal – especially in the 13 cm band – and the single ended mixer in no way suppresses it, some care must be taken to ensure that the correct frequency is selected from the mixer out-

put. After the microstrip has been accurately etched to the specified dimensions it means, quite simply, that the resonator must be resonate at the correct frequency with a minimal of fine tuning. The output signal must, of course, immediately disappear when the 144 MHz input signal is removed from the mixer input. The famous Lecher wires are simple to construct and are sufficiently accurate as a frquency measurer (within 1 %).

When the transmit converter has been aligned and tested the output amplifier càn be connected. Again, all stages are tuned to achieve a maximum output power.

During the tuning of the 23 cm power amplifier (fig.8) the 10 Ω resistance in the collector of T 2 should be increased to 47 Ω in order to avoid burning out the BFR 96 under off-tune conditions.

The selective 13 cm output amplifier (fig. 10) should be experimentally adjusted by altering the bias resistors of the last two stages (T 3 and T 4) and penultimate transistors in order to achieve a maximum output power but without exceeding the dissipation limits of these decives.

As already mentioned in the introduction, microstrip circuits, according to theory, require no screening measures. Problems can occur in practice when individual circuits are located in a metal housing and connected by coaxial cable. Every metal enclosure has an unending number of self resonances, some with a high Q, and the situation is made much more complicated by the presence of PCBs wire and coaxial cable of various lengths. Looking at the practical side concerning dimensions, the lowest resonant frequency must fall in low GHz range. For this reason those who work with frequencies below 500 MHz are hardly aware of this problem – quite the contrary to the experience of people working above 1 GHz!

Resonance problems are very difficult to cure by using screening measures as they only serve to create more resonances. Such problems are best tackled by reducing the Q of the unwanted resonance by the use of damping materials and circuits, the introduction of absorber pieces and soldering-in of low inductive resistors between the resonator hot points.

Absorber material is the best method of damping resonances caused by metal enclosures and large surface area microstrips. Of course, care must be taken that the wanted resonance is not damped as well as the unwanted when using this material. If professional damping materials are difficult and / or expensive to obtain, a cheaper alternative solution is recommended: the conductive plastic foam which is used to store MOS semiconductor devices.

Individual resistors instead of absorbing material can also be used. They are used mainly to solve problems arising from the use of physically small components. This applies also to the coaxial cable which connect the various microstrip circuits together. The resonances do not occur on the inner conductor, or between the inner and screening mantle because both ends are terminated in the characteristic impedance of the cable. The screen of the coaxial cable forms, together with metal surfaces, another transmission line. As both ends of the coaxial screen are connected to ground, n λ / 2 resonances are formed which possess quite a good Q. Owing to the continuous parasitic inductance of the screen-ground connection at both ends, these resonances are radiated out of the cable as radio waves. This is made worse, of course, if the ground connections have not been carried out correctly. These resonances can be damped effectively by soldering one or more 100 Ω resistors between screen and ground but the voltage nodal points of the spurious must be avoided.

Finally, there are coaxial cable lengths to consider, the critical ones which carry the signal from the oscillator modules to the multiplier stages are very sensitive to output termination impedances and can be rendered inoperative by an »unfortunate« choice of cable length. The 13 cm power amplifier chain too, perhaps could do with a trimming of cable lengths in order that the output power may be optimized.

11. CONCLUSION

The transverters described in this article have been in operation now for over two years and

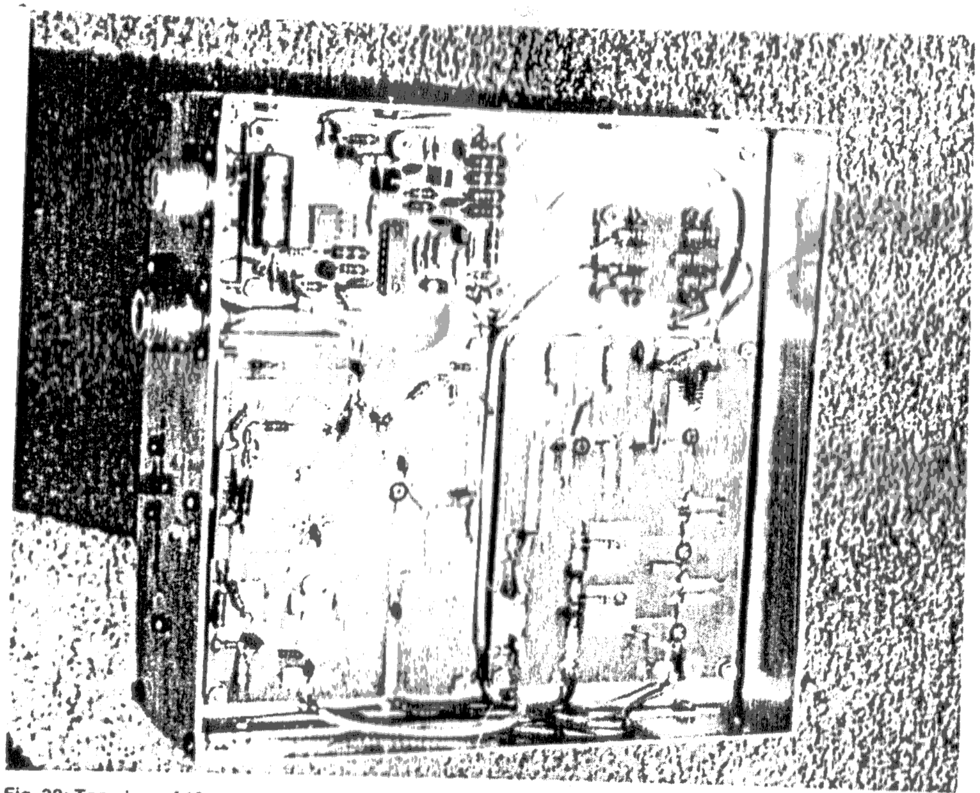


Fig. 32: Top view of 13 cm transverter prototype

several examples have been constructed. Many experiments were carried out in order to weed out potential problem areas in the circuit. The one main problem was caused by the frequency multiplier stages i. e. the transistors which were employed there. Besides the normal causes of transistor failure, over-current, over-voltage etc. another failure mechanism was observed. If relatively large RF signals are applied to the base of a transistor, as normally required for the efficient operation of a frequency multiplier, a parasitic Schottky barrier is gradually formed which lies in parallel to the normal BE barrier. As the breakdown of this pseudo Schottky barrier is only 0.3 V, it is much lower than the 0.7 Volt of the normal barrier. This leads to a progressive deterioration of the transistor current amplification factor until it is zero. This process is gradual and sometimes can last for weeks or months of continous operation before any deterioration is noticed.

It was discovered that differences in the susceptibility of a transistor to this mechanism was exhibi-

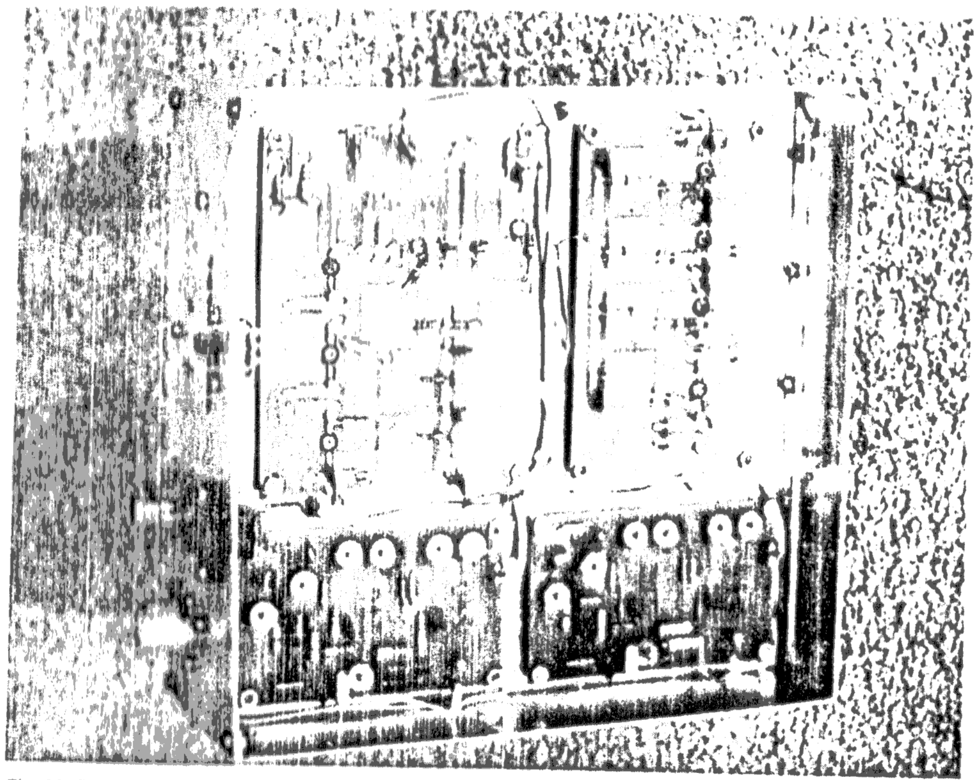


Fig. 33: Bottom view of 13 cm transverter prototype

ted within the same type manufactured by diverse firms. An important factor appears to be, that first grade transistors from reputable firms, such as Valvo or Siemens, seldom exhibit this failing when other specified operating conditions have not been exceeded.

On the other hand, the microwave stages with lower levels can employ a wide variety of transistors as the microstrip band filters allow a very wide tuning range. Microwave transistors in ceramic packaging ("micro-X" or "Cerec") always result in higher gain and noise figure but they are almost an order dearer to buy.

Fortunately, a short time ago, two plastic packaged types with emitter fins have appeared on the market which are both reasonable priced and have data approaching that of the ceramic types.

Both transverters were designed in order that the adjustment procedure was as simple as possible, such items as wave trap resonators, balanced mixers and other complex circuit devices which require expensive test equipment to set up, were deliberately avoided. Of course, these circuits can be used in microstrip technology. Besides the transverters, many other circuits were designed using the same technique: 13 cm and 23 cm transponders, receive down-converters for various satellite bands such as the meteorological satellite at 1.7 GHz, the MARECS downlink band at 1540 MHz and the NAVSTAR frequency 1575.42 MHz. In fact, the first converters built, using the technology described, were tested on METEOSAT signals at 1694.5 MHz.

It was found in practice, that circuits built, using this technology, will work in extreme ambient conditions determined both by temperature and by humidity, they can be hauled up mountains, accepting rough usage such as inadvertent "drop tests" without being detuned.

Josef Grimm, DJ 6 PI

Frequency Modulated Amateur Television (ATV)

Amateur television (ATV) was, until a few years ago, almost exclusively carried out in the 70 cm band. There was then sufficient room for an amplitude modulated colour television signal using commercial transmission standards.

As the 70 cm band grew increasingly busy with FM transponders, direct FM, satellite communication and commercial space safety installations, ATV signals were interfered with more often. Since this inconvenient eventuality cannot now be changed, many ATV amateurs are leaving the 70 cm band and are using frequency modulated signals in the SHF bands.

The advantages of FM will be compared with the previously universally employed amplitude modulation. This report is a compendium of articles published in radio amateur literature and of the author's experience with FM ATV. The latest components from satellite technology, which are employed in ATV, will also be mentioned. In forthcoming issues of VHF Communications there will be articles from various authors which describe send and receive equipments for FM ATV in the GHz range. Some of this equipment will be suitable for the reception of commercial satellite television.

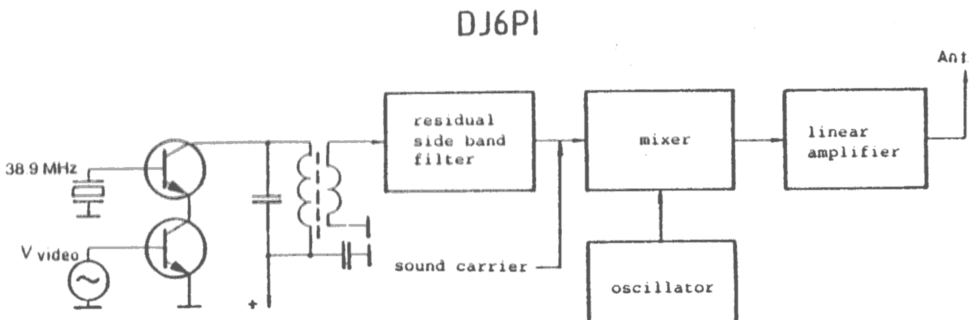


Fig. 1: Principle of an AM TV Transmitter

1296 MhZ Transverter Parts List

Receive Converter Board #004.

Semiconductors

* 1 - BFQ69 (BFQ65)
* 3 - BFR34 (BFR91)

Capacitators

* 4 - 2.2 pfd
* 4 - 8 pfd (etched on PC)
* 4 - 22 pfd
* 1 - 1 n
* 3 - 10 n
  1 - 33 ufd

Resistors

3 - 100 ohm
1 - 330 ohm
2 - 68 Kohm
2 - 150 Kohm

Power Amplifier Board #007

Semiconductors

* 2 - BFR96S
* 1 - 2N5944
  2 - 1N4148

Capacitators

* 1 - 1 pfd (use 1.5pfd)
* 1 - 2 pfd trimmer (3.5 p[illegible]
* 1 - 5 pfd trimmer (3.5pf[illegible]
* 4 - 8 pfd (etched on PC)
* 2 - 22 pfd
* 2 - 10n
  1 - 4.7 ufd

Resistors

1 - 2.2 ohm 1/2W
1 - 10 ohm
2 - 33 ohm
1 - 470 ohm 1/2W
2 - 1 Kohm
2 - 12 Kohm

Switch Board #011

Semiconductors

* 4 - BA379 (HP 5082-3081)
  4 - 1N4148

Capacitators

* 7 - 8 pfd (etched on PC)

Transmit Converter Board #006

Semicoductors

* 2 - BFR34 (BFR91)
* 1 - BFT65 (BFR91)
* 1 - BFR96

Capacitators

* 1 - 1 pfd (use 1.5 pfd)
* 1 - 1.5 pfd
* 3 - 2.2 pfd
* 5 - 8 pfd (etched on PC)
* 3 - 22 pfd
* 1 - 1 n
* 1 - 10n
  1 - 4.7 ufd

Resistors

1 - 33 ohm
1 - 68 ohm
5 - 100 ohm
1 - 39 Kohm
2 - 68 Kohm
1 - 150 Kohm

Oscillator Board #014

Semiconductors

* 1 - [illegible] (BF199)
* 1 - [illegible]
* 1 - 7.5 v zener

* 1 - 96 MhZ Xtal

Capacitators

* 3 - 1 pfd (use 1.5 pfd)
* 2 - 2.2 pfd
* 2 - 5 pfd trimmers
* 2 - 10 pfd trimmers
* 1 - 20 pfd trimmer
* 1 - 22 pfd
* 1 - 56 pfd
* 4 - 10n
  1 - 4.7 ufd
  1 - 47 ufd

Resistors

1 - 33 ohm
2 - 100 ohm
1 - 100 ohm 1/2W
1 - 470 ohm
1 - 4.7 Kohm

Alan Board
Flat 12 No 19 LANE COVE RD
RYDE
Sydney 2112

* 3 - 22 pfd

Resistors

- 2 - 220 ohm
- 1 - 470 ohm
- 2 - 680 ohm

Notes

1) Trimmer Colour Codes

3.5 pfd	red dot
5.0 pfd	black dot
10.0 pfd	yellow body
20.0 pfd	green body

2) Clean off tinned copper track off outside edge of board

3) transistor in brackets are supplied as alternatives to those specified in circuit. BFR91 will work in all positions.

- 3 - 12 Kohm
- 2 - 68 Kohm

VOX Board #016

Semiconductors

- * 1 - BF245
- * 1 - BC237 (BC547)
- * 1 - BC213 (BC557)
- 1 - TIP32
- 2 - BA243 use IN914 or IN4148
- 1 - OA95
- 1 - IN4001
- 1 - 4049

Capacitors

- * 1 - 3.3 pfd
- * 1 - 6.8 pfd
- * 1 - 10 pfd
- * 1 - 20 pfd trimmer
- * 1 - 1 n
- * 4 - 10 n
- 1 - 4.7 ufd
- 1 - 220 ufd

Resistors

- 1 - 33 ohm
- 1 - 56 ohm 1/2W
- 2 - 150 ohm
- 1 - 220 ohm
- 5 - 270 ohm 1/2W
- 1 - 680 ohm 1/2W
- 1 - 1 Kohm
- 2 - 1.5 Kohm
- 3 - 3.3 Kohm
- 1 - [illegible] Kohm

COMPONENTS MARKED WITH AN * ARE SUPPLIED