



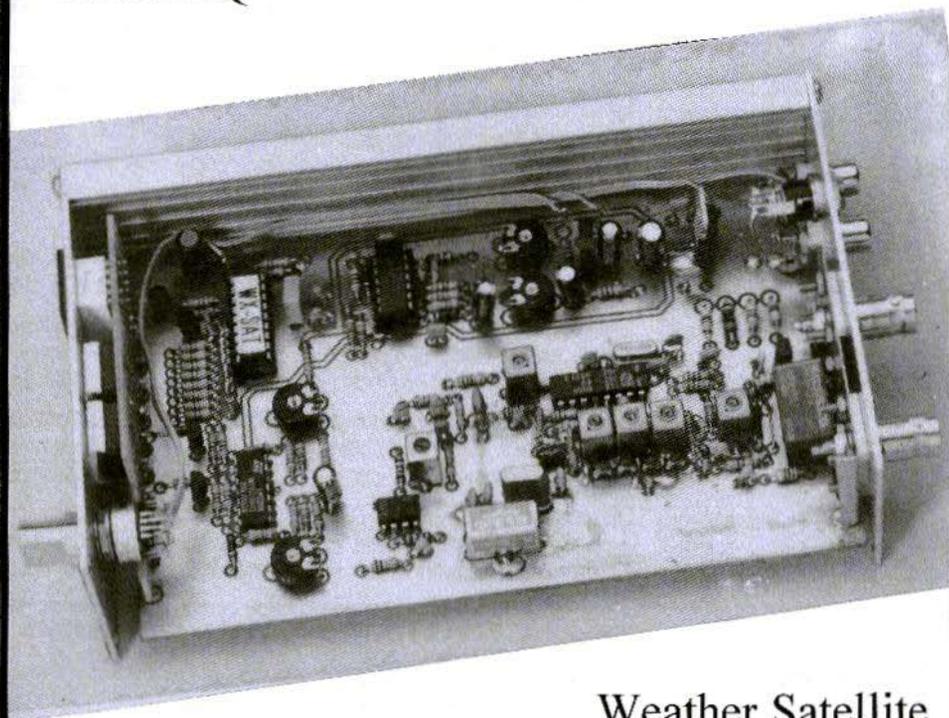
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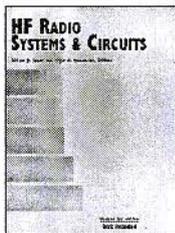
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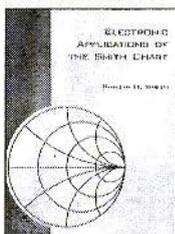
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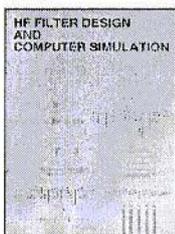
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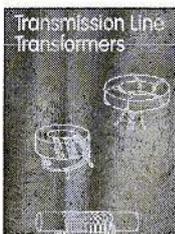
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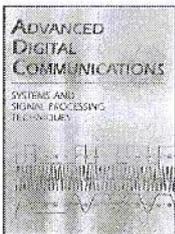
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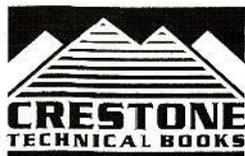
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Matjaz Vidmar, S53MV

Tracking Generator for the Spectrum Analyser 100 kHz to 1750 MHz

1. TRACKING GENERATOR PRINCIPLES

A basic and very useful addition to a spectrum analyser is a tracking generator. A tracking generator produces a RF signal on the exact frequency where the spectrum analyser is receiving at the same time. A tracking generator therefore allows testing many passive and active RF circuits that do not produce any RF signals on their own, like filters, amplifiers etc. With an additional directional coupler or bridge, reflection measurements can be performed in the whole frequency range covered by the spectrum analyser.

The simplest tracking generator is a wideband noise source, since part of the noise power always falls in the instantaneous receiving bandwidth of the spectrum analyser. Since the noise figure of typical spectrum analysers is rather high, in the range around 15-20dB, the required ENR (excess noise ratio) of the noise source is very high. Further, the

dynamic range of noise measurements is quite limited, since the device under test and the input of the spectrum analyser are both loaded with much larger power levels than are actually being measured. Finally, noise measurements always require some averaging (video filtering) due to the random nature of noise signals.

To avoid the drawbacks of noise measurements, almost all manufacturers offer tracking generators that produce a coherent sinusoidal signal on the receiving frequency of the spectrum analyser, either built-in or as separate units. A tracking generator usually does not include very expensive components. Nevertheless, commercial tracking generators may be quite expensive, up to one quarter of the cost of the spectrum analyser. Therefore it makes sense to build companion tracking generators even for commercial spectrum analysers.

A tracking generator designed to operate with the spectrum analyser described in [1] or [2] will be described in this article. Since most commercial spectrum analysers use similar frequencies for the



first IF (usually in the range 2...3 GHz), many of the described circuits are also useful to build a tracking generator for a commercial spectrum analyser. Of course, the exact circuit diagram of a tracking generator should match the frequency conversions in the corresponding spectrum analyser.

A tracking generator obtains its output signal exactly in the opposite way the different frequency conversions are made inside the spectrum analyser. A tracking generator therefore requires the signals from all variable oscillators (VCOs) inside the spectrum analyser. Most spectrum analysers therefore have the signals of the first, second and sometimes even third local oscillator accessible on suitable RF connectors on the front panel or sometimes back panel.

The spectrum analyser, described in [1] or [2], includes two variable oscillators (VCOs) for the first and second conversion. A suitable tracking generator requires the same conversion processes in the reverse direction, as shown on the block diagram in Fig.1. First, the tracking generator should add the frequency of the second IF (around 70 MHz) to the second VCO at 2.03 GHz (+/-10 MHz). The sum is a signal at the first IF around 2.1 GHz. The latter is afterwards subtracted from the frequency of the first VCO (2.1...3.85 GHz). An automatic-gain-control (AGC) loop is used to stabilise the output-signal amplitude over the whole frequency range 100 kHz...1750 MHz.

Frequency summation and subtraction can be performed in different ways: using mixers and filters or with phase-locked loops (PLLs). All signals inside

the tracking generator have relatively high levels. Unwanted mixing products should be monitored carefully. On the other hand, thermal noise is not very important if compared to the design of a spectrum analyser or other receivers. Finally, suitable shielding should be provided to avoid unwanted signal paths as well as using buffer amplifiers to provide the required separation wherever necessary.

In the described tracking generator, the first frequency summation 70 MHz and 2.03 GHz is performed by a phase-locked loop. The second conversion (subtraction) of the 2.1 GHz first IF from the first LO 2.1...3.85 GHz is performed by a balanced mixer followed by a lowpass filter on the output. Two identical buffer amplifiers are used for both VCO signals to prevent any interference from the tracking generator back into the spectrum analyser.

The AGC sets the output-signal level to about 1mW (+0dBm). The latter can be further attenuated down to -40dBm when required on the main tracking-generator output. An auxiliary, fixed -10dBm output is provided for a frequency counter.

The described tracking generator includes nine shielded RF modules and a power supply. Some modules are identical to those in the spectrum analyser: the PLL mixer is identical to the second mixer in the spectrum analyser while the output mixer is identical to the first mixer in the spectrum analyser. The output attenuator with four 10dB steps is identical to the input attenuator in the spectrum analyser. Of course, the description of all identical modules will not be repeated.

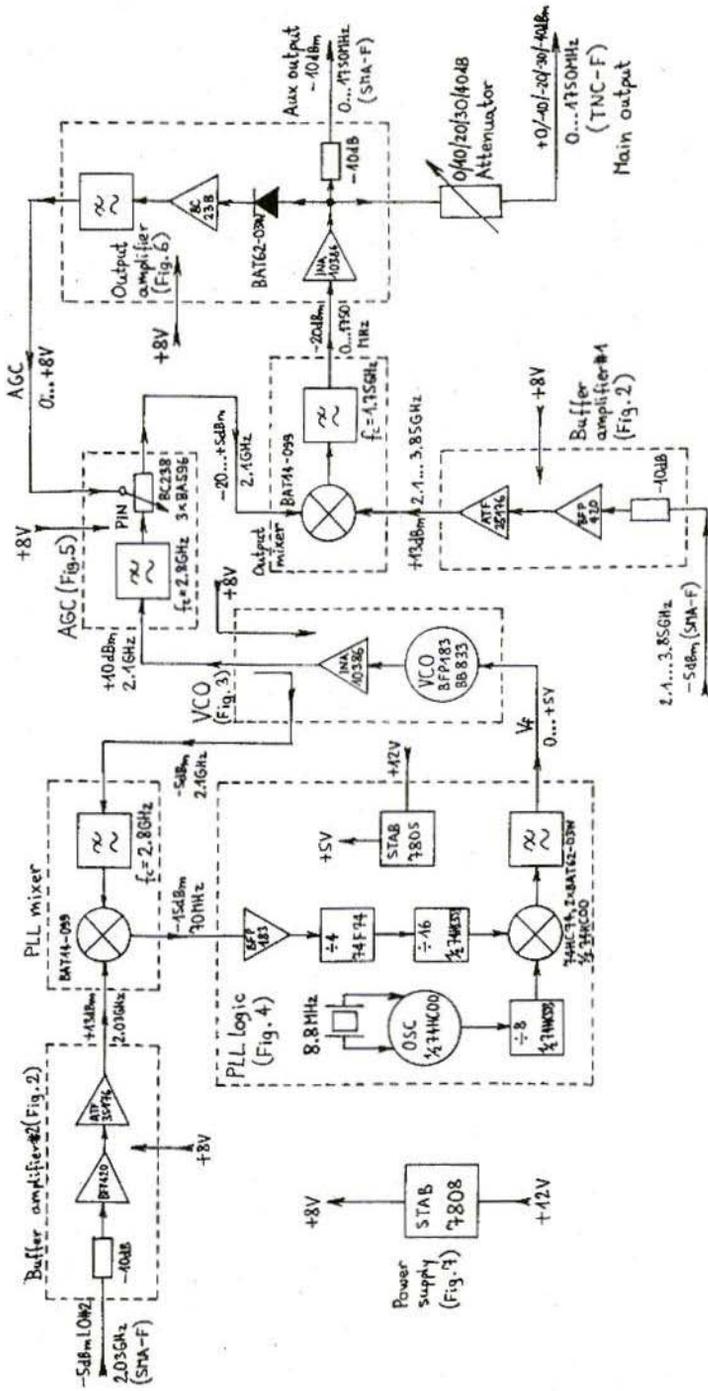


Fig.1: Tracking Generator Block Diagram



2. BUFFER AMPLIFIERS

The tracking generator includes two identical buffer amplifiers for the signals of both VCOs in the spectrum analyser. The buffer amplifiers have two main functions: amplify the signals of both VCOs to about +13dBm to drive both mixers and prevent any signals from the tracking generator returning back in the VCOs and mixers of the spectrum analyser.

The circuit diagram of (one) buffer amplifier is shown in Fig.2. The buffer amplifier includes a -10dB attenuator and two amplification stages. The input attenuator allows a good input impedance match even with the power turned off. Even more important, the input impedance does not change much after

turn-on, so that the frequency pulling of both VCOs inside the spectrum analyser is kept small enough.

The input attenuator is followed by a simple amplifier stage with a BFP420 transistor. The high-frequency gain decay of the latter is partially compensated by the 0.68pF capacitor in parallel to the 68 resistor in the attenuator. The output stage uses an ATF35176 HEMT to obtain the required output power. The output of the buffer amplifier is connected through a short piece of UT-085 semi-rigid cable, used at the same time as a balun inside the balanced mixer.

Both amplifier stages are supplied with +8V through suitable resistors and feedthrough capacitors. The operating point of the output stage may change due to the I_{dss} tolerances of the ATF35176 HEMT, but these changes

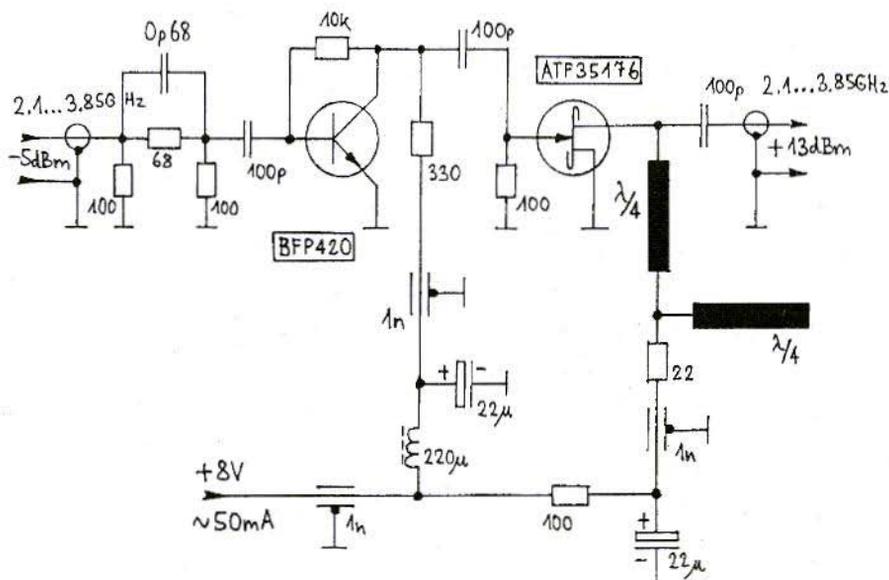


Fig.2 Buffer Amplifier

have little effect on the performance of the amplifier. The drain voltage (U_{ds}) may be anywhere in the range 2...3.5V.

3. PLL FOR 2.1 GHz

The first signal-processing step inside the tracking generator is the addition of the second VCO frequency and the second IF. Of course, the tracking generator should contain its own oscillator to generate a signal on the second IF around 70 MHz. The two frequencies could be simply added in a mixer. Unfortunately, the filtering of the output signal would require a complex cavity bandpass filter to remove the image at 1.96 GHz as well as the second VCO leakage at 2.03 GHz. Further, an AGC circuit would be required to avoid overdriving the mixer.

The same task can be performed by a phase-locked loop with its own VCO operating at the output frequency of 2.1

GHz. A small fraction of the latter is mixed with the second VCO signal at 2.03 GHz coming from the spectrum analyser. The difference of the two frequencies is compared with the nominal second IF value and the result of this comparison is used to correct the frequency of the 2.1 GHz VCO. The 2.1 GHz signal does not require any further filtering. However, the PLL should be designed to achieve lock reliably and to track the 2.03 GHz signal even when the frequency of the second VCO in the spectrum analyser is swept.

The VCO for 2.1 GHz (shown in Fig.3) is very similar to the second VCO in the spectrum analyser operating at 2.03 GHz. Since only a relatively narrow frequency range around 2.1 GHz has to be covered, a single tuning Varactor BB833 is sufficient in the central finger of the interdigital bandpass.

A BFP183 transistor is used as the active device inside the VCO, while an INA10386 MMIC is used as the output buffer. A directional coupler takes a small part of the output signal (around -5dBm) for the PLL mixer.

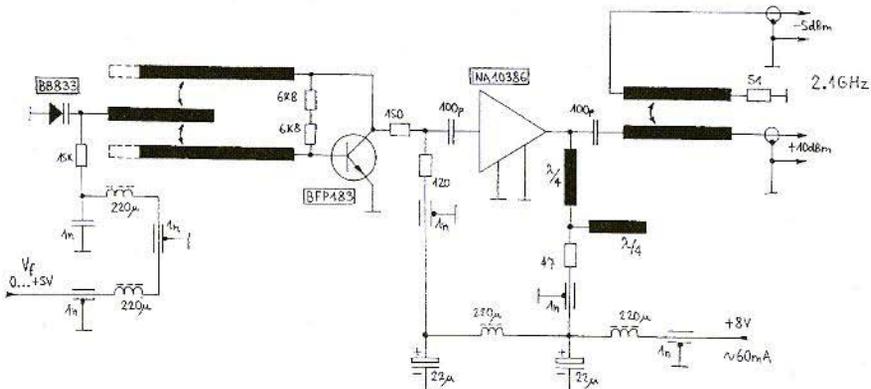


Fig.3 VCO for 2.1 GHz

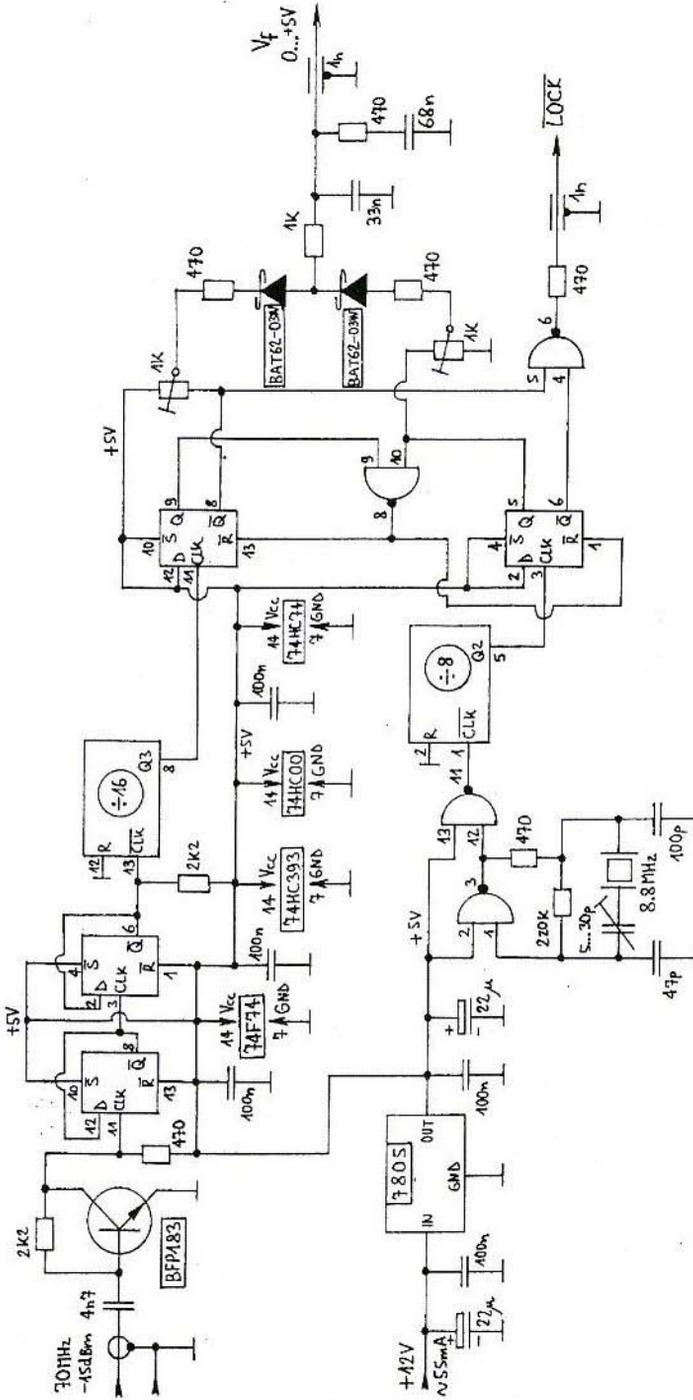


Fig.4: PLL Logic

The VCO for 2.1 GHz is built on an identical printed-circuit board as the narrowband VCO in the spectrum analyser. Besides a different Varactor, both the collector and base microstrips are shortened by about 2mm at their open ends to achieve a higher operating frequency. The control-voltage lowpass network has to be designed carefully to allow a fast response of the control loop.

The PLL mixer is identical to the second mixer in the spectrum analyser. The -5dBm VCO output is connected to the 2.8 GHz lowpass inside the mixer module, while the buffer amplifier is connected to the semi-rigid balun. The frequency difference (around 70 MHz) is taken from the 800 MHz lowpass.

The comparison between the nominal value 70 MHz and the actual frequency difference is performed by the PLL logic shown in Fig.4. The PLL logic includes a reference crystal oscillator, two dividers for the reference frequency and the actual frequency difference and a charge-pump frequency/phase comparator.

While the PLL is unlocked, the actual frequency difference may deviate substantially from the nominal value around 70 MHz. An upper limit of about 140 MHz is set by the input divider 74F74 with the suggested BFP183 driver. The frequency difference is divided by 64, while the 8.8 MHz reference is divided by 8. The 1.1 MHz comparison frequency allows fast tracking of the narrowband VCO in the spectrum analyser even when the frequency of latter is swept.

The reference crystal-oscillator fre-

quency is selected to 1/8 of the second IF of the spectrum analyser. Of course, the exact value depends on the components used in the spectrum analyser, in particular the crystal used in the third conversion (usually 60.000 MHz) and crystal filter (usually 10.700 MHz). For a nominal second IF value of 70.700 MHz a reference crystal oscillator at 8837.5 kHz is required.

The output-voltage (V_f) range of the charge-pump frequency/phase comparator (max 0...+5V) usually has to be further reduced by the two 1kohm trimmers. The frequency difference should never become too large (140 MHz limit imposed by the 74F74 divider) nor should the PLL be allowed to lock onto the image response of the mixer. The charge pump with the two Schottky diodes BAT62-03W is followed by a RC lowpass network, defining the settling time and stability of the feedback loop.

4. OUTPUT MIXER, AMPLIFIER AND AGC

The second signal-processing step is the subtraction of the 2.1 GHz frequency (first IF of the spectrum analyser) from the first LO frequency 2.1...3.85 GHz. All unwanted spurious conversion products from a correctly-designed mixer can simply be removed by a lowpass filter on the output, leaving only the desired signal in the frequency range 0...1750 MHz. On the other hand, it is very

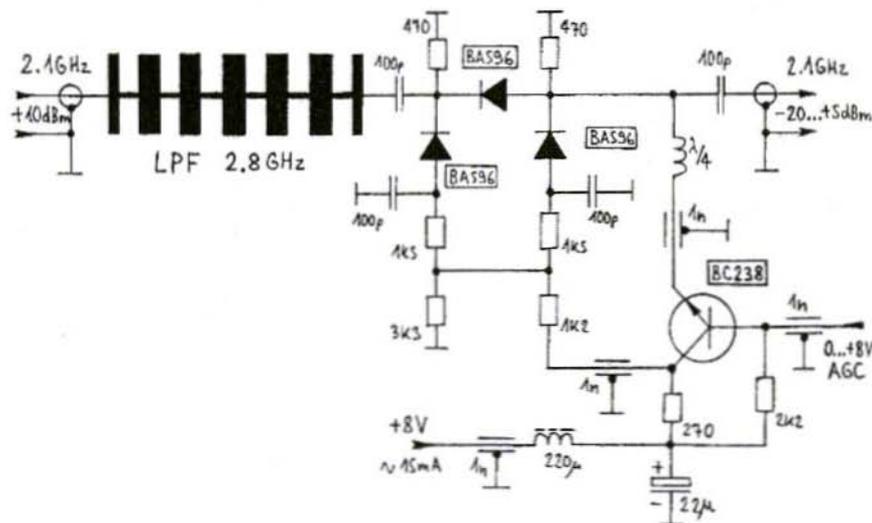


Fig.5: AGC

difficult to build a PLL covering such a wide frequency range.

The output mixer is identical to the first mixer in the spectrum analyser. The latter module includes a 1.75 GHz lowpass. The wideband-VCO signal (2.1...3.85 GHz) is fed through a buffer amplifier directly to the semi-rigid balun of the output mixer. No further processing is required, since the harmonics of the first spectrum-analyser LO do not cause any harmful spurious frequencies in the tracking generator.

On the other hand, the tracking generator is sensitive to the harmonics of the 2.1 GHz first-IF signal. The latter may cause spurious mixing products in the output-frequency range 0...1750 MHz. Therefore, any harmonics of the 2.1 GHz signal have to be removed. Further, the 2.1 GHz signal level has to be controlled carefully to avoid mixer non-linearities.

The 2.1 GHz harmonics are removed by a 2.8 GHz lowpass filter, followed by a PIN-diode attenuator to set the signal level for the mixer. Both circuits are included in the AGC module shown in Fig.5. The PIN attenuator is followed by another 2.8 GHz lowpass in the output mixer module.

The PIN attenuator includes three PIN diodes BA596 connected in a PI network. The BC238 transistor feeds all three PIN diodes with suitable DC currents to keep the input and output impedances stable while adjusting the attenuation. The AGC control voltage spans from zero to +8V. Minimum attenuation is achieved at +8V while zero provides maximum attenuation.

Since the output mixer operates in the linear region, the output signal is relatively weak, around 10uW or -20dBm. To increase the output signal level and drive the AGC detector, an output

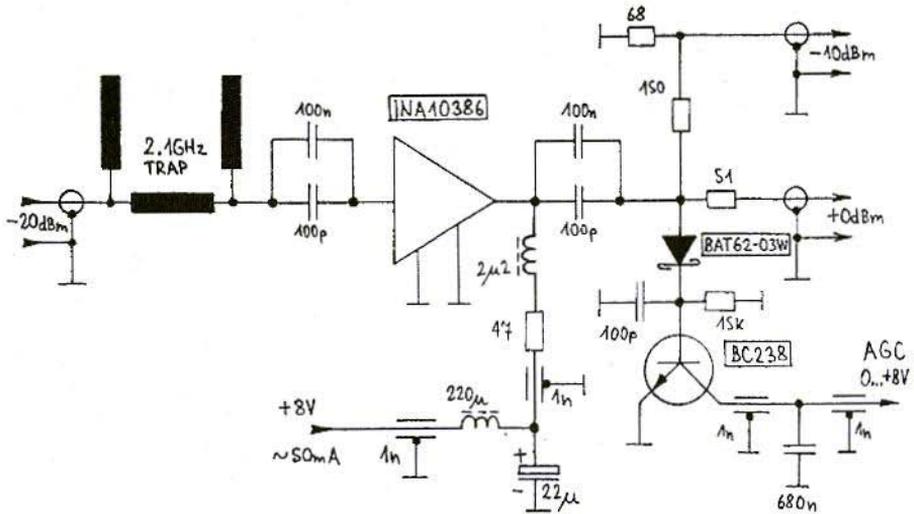


Fig.6: Output Amplifier

amplifier is required as shown in Fig.6. The latter is built with an INA10386 MMIC that provides 26dB gain and an output power of more than +10dBm in the whole frequency range up to 1750 MHz.

The INA10386 MMIC amplifier requires DC decoupling capacitors on both the input and output. The latter limit the lower end of the frequency range of the tracking generator to about 100 kHz. Each coupling capacitor includes a parallel connection of two SMD capacitors: first a low-loss, 100pF NP0 0805 capacitor is soldered on the printed-circuit board and afterwards a 100nF, lossy Z5U 1206 capacitor is soldered across the 100pF capacitor.

The AGC detector (Schottky diode BAT62-03W) is connected directly to the output of the INA10386 amplifier. The BC238 transistor operates as the AGC feedback amplifier, while the time

constant of the feedback is set by the 680nF capacitor. Thanks to the AGC action, the output amplifier behaves as a voltage source. The main output (+0dBm) therefore requires a 51Ω series resistor. The 150ohm and 68Ω resistors allow an auxiliary -10dBm output and provide a DC path to ground for the AGC detector.

The main output of the tracking generator includes a step attenuator to decrease the output signal level down to -40dBm in four 10dB steps. The design of this attenuator is identical to the attenuator used on the input of the spectrum analyser. Due to the non-ideal frequency response of the latter, better measurement accuracy can be achieved with the auxiliary -10dBm output, especially for wideband measurements.

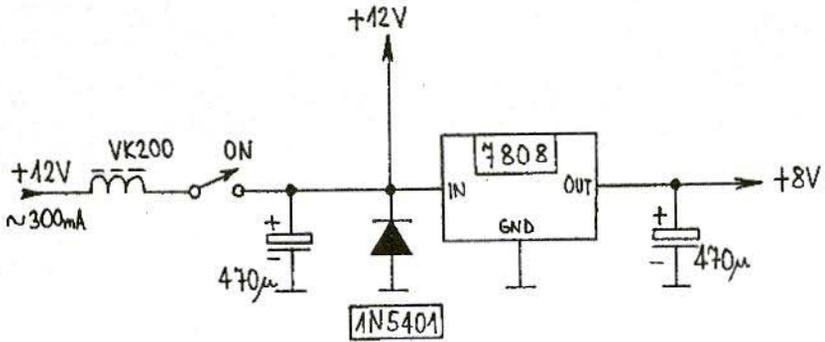
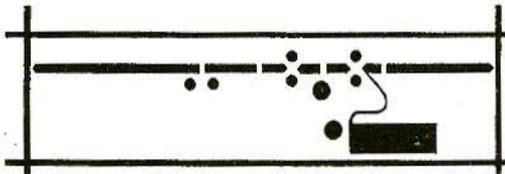
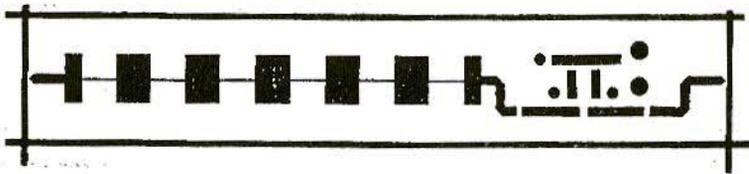


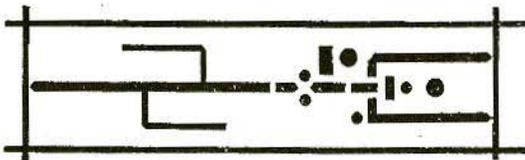
Fig.7 Power Supply



Buffer amplifier
(two identical 20x80)



AGC
(20x120)



Output amplifier
(20x80)

Fig.8: Microstrip Printed Circuit Board Layouts (dimensions in mm)

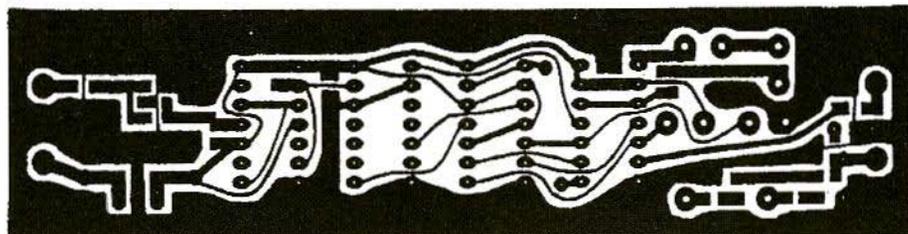


Fig.9: PLL Logic PCB Layout

5. CONSTRUCTION TIPS

The tracking generator is built in a similar way as the corresponding spectrum analyser, described in [1] or [2]. All modules are installed in shielded boxes made from 0.5mm thick brass sheet. Most modules require a +8V supply voltage obtained from the 7808 regulator shown in Fig.7. The PLL logic has its own 7805 regulator inside the module and requires a +12V external supply. The tracking generator is equipped with an ON/OFF switch on the front panel to allow an immediate check of the signals shown on the spectrum-analyser display.

Except for the PLL logic, all other modules use microstrip circuit boards. Some of the latter were already described in the spectrum-analyser article, while the new ones are shown in Fig.8. All microstrip boards are etched on one side of a 0.8mm thick FR4 Glass-fibre-epoxy laminate, while the other side is not etched to act as a ground-plane. The PLL logic is built on a single-sided board shown in Fig.9 and etched on 0.8mm thick FR4 laminate.

Microwave absorber foam is built under the cover of two modules: buffer ampli-

fier for the wideband (2.1...3.85 GHz) VCO and output mixer. Other modules usually do not require any absorber foam inside. Microwave absorber is also not required in the space among the shielded boxes.

The tracking-generator module location is shown in Fig.10. The tracking generator has the same depth (240mm) and width (220mm) as the spectrum analyser, so that the two boxes can be stacked easily one over another. The height of the tracking generator is only 32mm, since all modules are located in a single plane. The bottom of the box is simply a piece of 1mm thick aluminium sheet, bent in the form of an "U". The cover is a similar "U" made from 0.6mm thick aluminium sheet.

6. ALIGNMENT AND CHECK-OUT

The tracking generator is a much simpler piece of test equipment than the spectrum analyser. Correspondingly the alignment and checkout should be much simpler too.

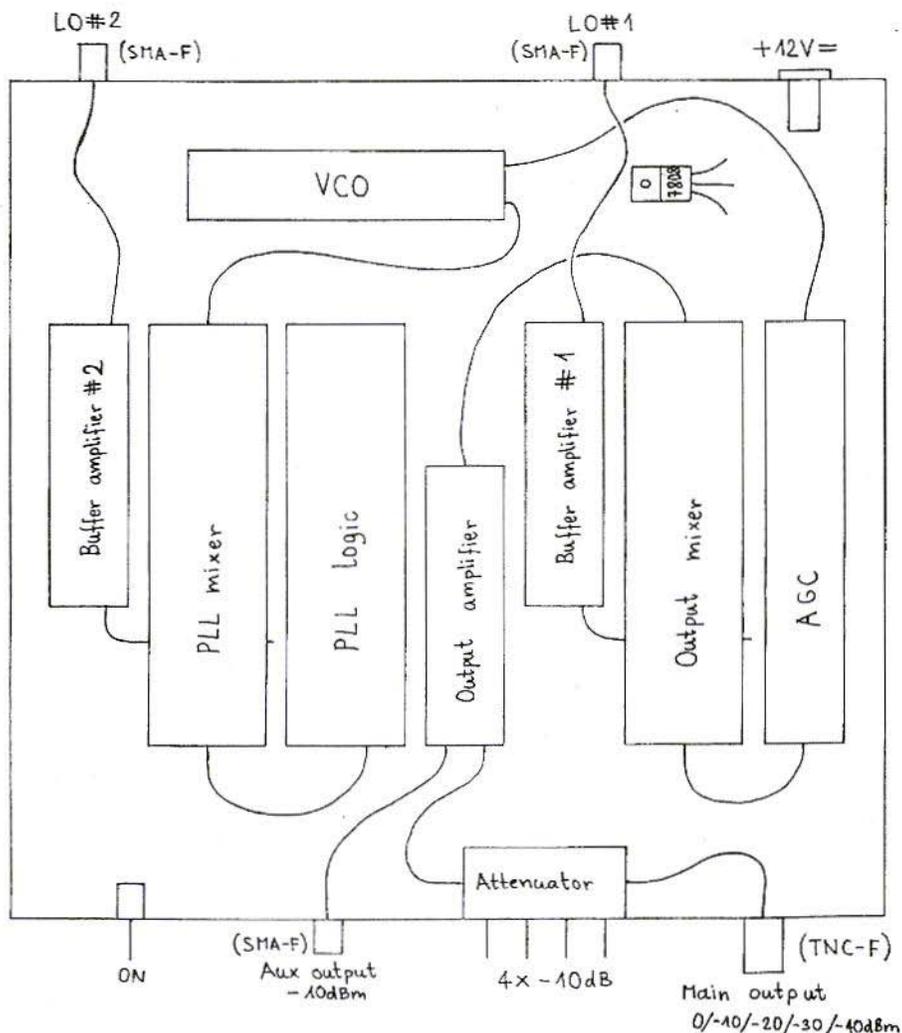


Fig.10: Tracking Generator Module Location.

Of course, the spectrum analyser should be available at all times to supply the signals of both VCOs. Before connecting all tracking-generator modules together, it makes sense to make a few simple checks on each module separately. In all modules it makes sense to check the DC bias points of all semicon-

ductors. Most simple errors can be detected in this way. Next some RF checks can be performed. Both buffer amplifiers should be checked for the output RF power level when driven by the corresponding VCO in the spectrum analyser.

The VCO needs fine tuning of the frequency coverage, since the microstrip board was originally designed for a slightly lower frequency in the spectrum analyser. Of course, the operating frequency of the VCO should be monitored while carefully shortening the base and collector strips. The VCO should achieve the nominal operating frequency of 2.1 GHz at a tuning voltage of 3...3.5V. Further, the VCO should operate without dropouts through the whole range of tuning voltages 0...+5V. Trimming of the centre finger of the BB833 Varactor may also be required.

In the PLL logic, the operation of the crystal oscillator should be checked first. Both 1kohm trimmers should be set initially with the sliders on the respective "hot" ends, to allow the widest span of the output voltage V_f . Without any input signal V_f should reach almost +5V. If a frequency above 70.7 MHz is fed to the input (for example a grid-dip meter coupled through a small wire loop), the V_f output should drop to zero.

Using the already tested modules, the 2.1 GHz PLL can be assembled and tested. While the PLL is locked, the DC voltage on the LOCK test point should drop to less than 0.2V. Both 1kohm trimmers in the PLL logic are then set for a frequency coverage of +/-60 MHz around the nominal frequency of 2100 MHz. While testing the PLL, the main output of the VCO should be terminated properly on a matched load.

After the alignment of the PLL is completed, the remaining modules of the tracking generator can be wired together. The latter do not include any

alignment points, but their operation should be verified. The spectrum analyser should be set to scan the whole frequency band 0...1750 MHz with the widest (4 MHz) IF filter. While connecting the output of the tracking generator to the input of the spectrum analyser, the whole trace should raise almost to the full scale. The trace will not be perfectly straight. The required fine adjustments to flatten the response will be described later. The next step is to test the tracking generator with narrower IF filters. There will probably be no change with a 700 kHz IF bandwidth. However, the response will probably drop with narrower IF filters. The response should be brought back to its original value by adjusting the crystal oscillator in the PLL logic. If the range of the trimmer is insufficient, a parallel capacitor may be added or the trimmer may be replaced with an adjustable inductor. In the worst case, the crystal in the PLL logic and/or the 60 MHz crystal in the spectrum analyser may have to be replaced.

In order to check the operation of the AGC, the Y input of the oscilloscope (used as the display of the spectrum analyser) is temporarily connected to the AGC line. Although the AGC voltage may span 0...+8V, it should remain in the range +1...+3V in a correctly-operating tracking generator. A too high or too low AGC voltage may suggest what is actually wrong with the tracking generator. While testing the AGC voltage, the main output of the tracking generator should be terminated to a matched load, otherwise the AGC voltage may move out of the required +1...+3V range.



The main checkout of the tracking generator is completed at this point. However, several minor adjustments may be necessary both in the tracking generator and in the spectrum analyser to optimise the operation. In particular, the frequency response of the spectrum analyser should be flattened as much as possible. Without any adjustments, the response of the tracking generator connected directly to the spectrum analyser may deviate by as much as ± 5 dB.

Amplitude variations of the frequency response of the spectrum analyser are mainly caused by the first mixer and its termination impedances. Additional impedance matching may flatten the overall response and/or move dips to compensate upward bumps in the response. Some dips and bumps move quickly while changing the length of the cable connecting the first mixer to the cavity filter.

Small pieces of copper foil, soldered to the output microstrip of the wideband VCO module, are usually very efficient in improving the impedance matching of the LO mixer port.

A substantial mismatch, especially above 1 GHz, is also caused by the two step attenuators with toggle switches. Above 1 GHz it is therefore recommended to use the aux -10 dBm output to avoid at least one of the two step attenuators. A better solution is to connect external, fixed 10 dB microwave attenuators (with SMA connectors) to both the tracking-generator output and the spectrum-analyser input.

After all described fine adjustments, the frequency response of the tracking generator and spectrum analyser connected together should be within ± 2 dB from 100 kHz to 1.6 GHz. The decay above 1.6 GHz is caused by the lowpass on the input of the spectrum analyser. An accuracy of ± 2 dB is reasonable even for much more expensive professional spectrum analysers.

Of course, since the response of the spectrum analyser and tracking generator can be measured easily, the resulting error can simply be subtracted from the real measurement result. Most manufacturers therefore offer an electronic storage/normaliser unit to be connected between the spectrum analyser and the oscilloscope display. An electronic storage/normaliser (using cheap integrated circuits) was also developed and built for the described spectrum analyser and tracking generator.

7.

LITERATURE

- [1] Matjaz Vidmar: 'Spectrum Analyser 0 to 1750 MHz, VHF Communications 1/99, pp 2-30
- [2] Matjaz Vidmar: 'Spektrum-analyser 0...1750 MHz', to be published in AMSAT-DL Journal.



Gunthard Kraus, DG8GB

Design and Realisation of Microwave Circuits

Part-7

10. MICROWAVE MIXER STAGES

10.1. Principles

The word "mixing" is a rather unfortunate way of describing the function of this stage. All mixer stages are actually amplitude modulators.

A "tone" signal, which we shall henceforth refer to as the high-frequency signal (RF signal), is amplitude-modu-

lated through a "carrier signal", which is usually generated by a carrier oscillator hence the description "LO signal" in the literature. In the ideal mixer, thanks to this amplitude modulation, only the total and difference frequencies are generated, whilst all other frequencies are completely suppressed (Fig.65).

Both these signals are of interest here, for we can mix upwards (filtering out the total frequency) or mix downwards to form a lower intermediate frequency, as usually occurs in a superhet receiver.

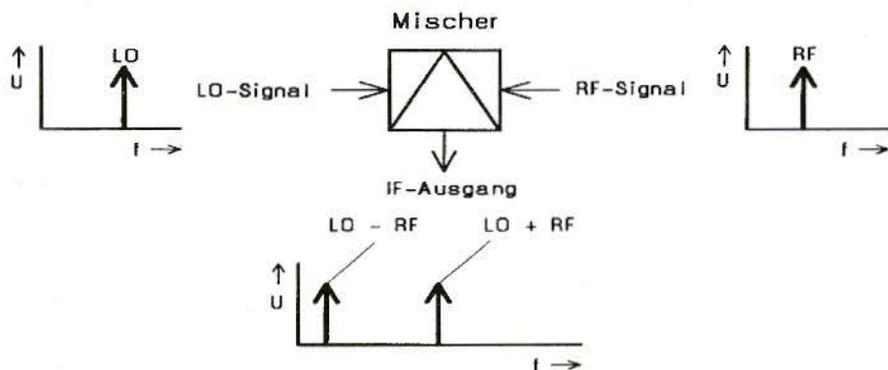


Fig.65: Principle and Function of a Mixer
Mischer = Mixer, *IF-Ausgang* = IF output

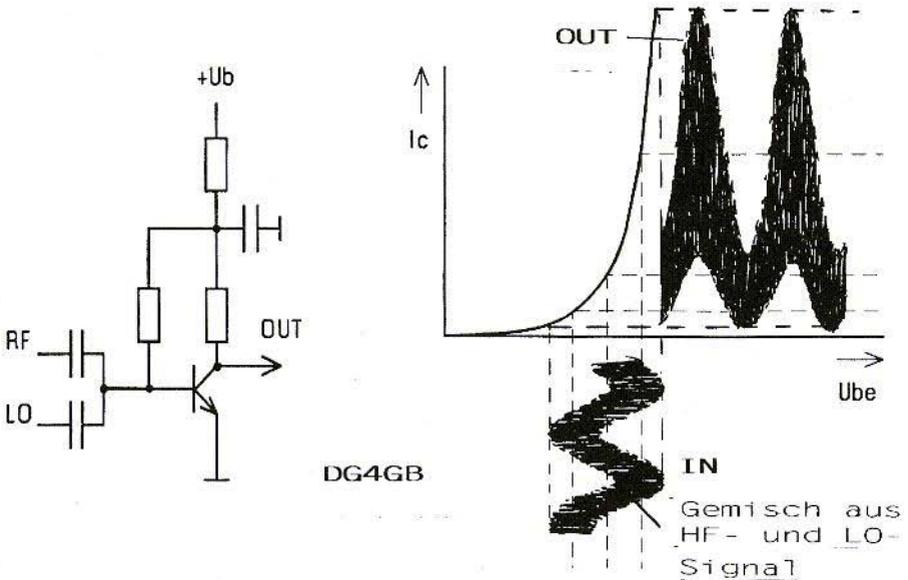


Fig.66: Additive Mixing on a Curved Characteristic
Gemisch aus HF- und LO-Signal = Mixture of HF and LO signals

10.2. Additive Mixer Stages

They probably represent the oldest form of mixer, for all that is needed is any component with a curved characteristic or, to put it more precisely, with a non-linear relationship between the current and the voltage or between the input signal and the output signal.

It makes no initial difference whether this is a diode, a transistor, a field effect transistor, a voltage-dependent resistor or a filament bulb.

If both voltages are simultaneously applied to the component selected as mixer the LO signal usually has a higher amplitude this LO signal now continuously pushes the operating point to and fro along the characteristic. So the characteristic gradient for the smaller high-frequency signal is permanently

changing and the amplitude modulation required for the high-frequency signal fraction is already present at the output!

Of course, no amplification can be expected from the simple mixing circuits with diodes. However, even today this method is still used for mixing in many Astra LNC's with Schottky diodes. Diode mixers in the dwarf wavelength range will be with us for some time yet.

If, on the other hand, the basis emitter segment of a bipolar transistor or the gate source segment of an FET is used as the modulator element, things are already looking much better with regard to amplification.

Since 1930, millions and millions of superhet receivers have demonstrated that this technology works well (see Fig.66).

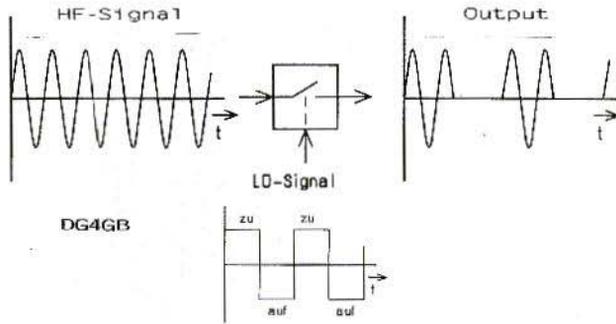


Fig.67:
Simple Multiplicative
Mixer

However, if we now look at the precise gradient of the output signal, we can immediately recognise the main disadvantage of this very simple principle the curved control characteristic does not yield textbook AM but something which is decidedly distorted. So in addition to the total frequency and difference frequency required, this signal gradient also contains both the input signals fed in (LO and HF), together with innumerable harmonics and their combination frequencies.

The only way to obtain acceptable inter-modulation ratios is consequently to work with low signal levels and good filtration.

A pure diode mixer stage unfortunately also has two additional secondary effects: Firstly, the attenuation already referred to must be compensated by additional amplification. Secondly, attenuation always leads to an increase in the noise factor of this assembly by the amount of attenuation.

Thus we need a very good, low-noise pre-amplifier with still higher amplification and a very conscientious check calculation to determine the total noise factor of the system.

10.3. Multiplicative Mixer

It would be better to identify this model through its most important characteristic. We are actually talking about the relay mixer group. In the most simple type of multiplicative mixer, the simple modulator, the high-frequency signal is "chopped up" to fit the rhythm of the LO signal (Fig.67). A square-wave modulation is generated in this way, in which the undesirable signal fractions and interference products in the output signal are reduced by 20 to 30dB, as against additive mixing.

Of course, we now need a considerably higher LO voltage, which must usually have a peak-to-peak value of at least 1 to 1.5 V, which is equivalent to a minimum level of +7dBm. A genuine square-wave voltage of this amplitude, with flanks as steep as possible, is even better than the usual sinusoidal form. The electronic switch must actually be opened in a flash to fit the rhythm of the carrier frequency and must be closed again just as fast. Only in this way can we arrive in the initial control characteristic segment with the strongest curvature, which also produces the most distortions, as quickly as possible. This naturally has a more favourable effect overall on the distortions arising.

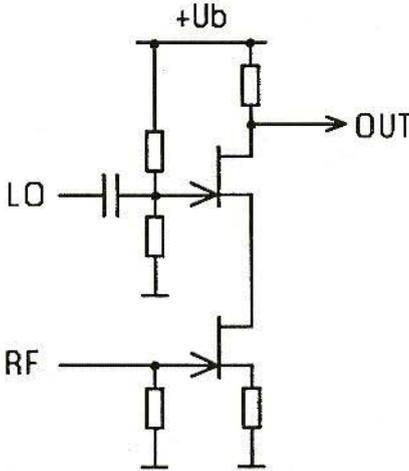


Fig.68: Modern Multiplicative GaAsFET Mixer

But there is one disadvantage we should not hide. Neither the HF signal nor the LO signal is suppressed in this circuit. They emerge at the output and must be eliminated with filters. By using active components such as transistors or FET's, mixing amplification can be obtained with this circuit as well.

A simplified version of the inner workings of a modern, integrated mixer module constructed according to this operating principle can be seen in Fig.68. It is easy to recognise the two GaAsFETs, the lower one of which operates as an amplifier for the high-frequency signal. The other one, by contrast, acts as an "on/off switch, activated by the LO signal".

The circuit will certainly look familiar to one or two people. It is a "well-known cascade connection".

This circuit is at present replacing the Astra LNCs at breakneck speed i.e. all

kinds of passive diode mixers at 12 GHz, since they provide considerable amplification. At the same time, the FET acting as "chopper up" serves as the conversion oscillator. This means additional components can be saved on and the costs can be reduced.

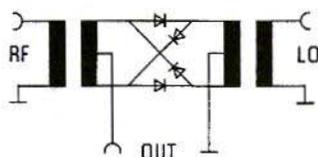
However, anyone wanting to come as close as possible to the ideal mixer must resort to the balanced mixer. This is well known as the ring modulator or double balanced mixer, with 2 transmitters and 4 diodes. It not only chops signals up, but switches poles through 180° as well (Fig.69). This balanced principle not only reduces the harmonic content further but also reduces the suppression of the two input signals. If we now also take it into account that the LO signal acting as the turn-on voltage should, as far as possible, be 20dB above the high-frequency level, there is not much left to be desired in a mixer.

Naturally, the double balanced mixer can also be created using active components. A distinction should certainly be made between two methods:

a) The diodes in the circuit are replaced, for example, by power MOSFET's, to make use of their superior characteristics as almost perfect electronic high-current switches. This naturally increases the saturation resistance and the IP3, but we now need quite a few Volts for the turn-on voltage amplitude to switch these components on and off rapidly enough. Unfortunately, this circuit also has the passive mixer's attenuation at least 6dB.

b) In contrast, if we do not wish to dispense with mixing amplification, because we are actually using active

Ringmodulator = Double Balanced Mixer



ARBEITSWEISE

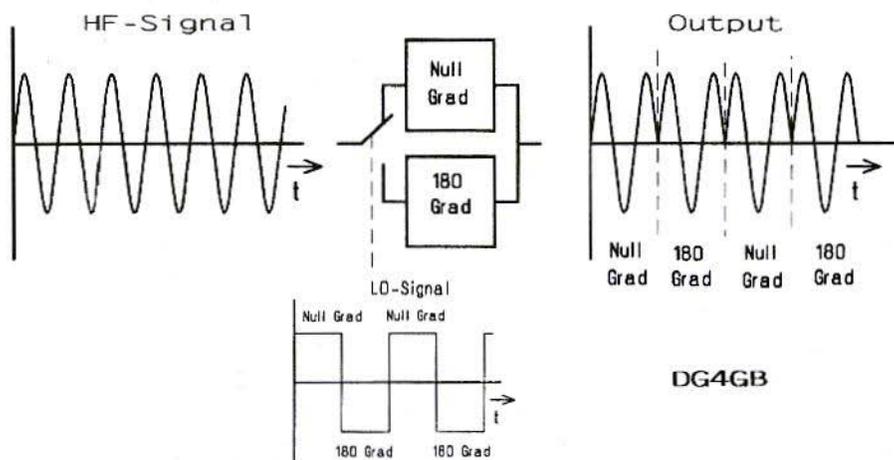


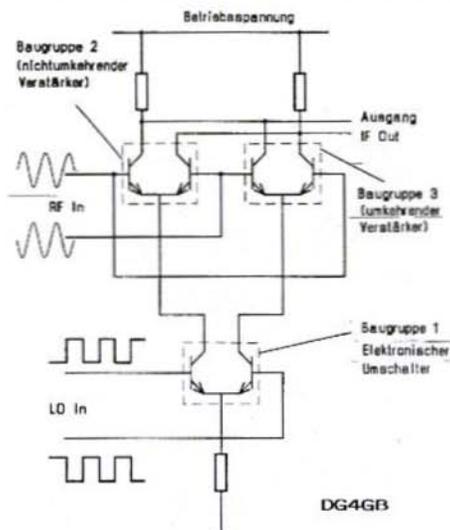
Fig.69: Principle and Functioning of a Ring Modulator

Arbeitsweise = Functioning, Null = Zero, Grad = Degrees

components, then we will need three differential amplifiers in all. The way they interact can easily be seen from Fig.70. For assembly 1, we actually need only switch the power supply voltage over to fit in with the rhythm of the LO signal, thus putting either amplifier 2 (without phase reversal) or amplifier 3 (with 180° phase rotation) into operation alternatively. These two amplifier stages are therefore connected up in parallel at the input and output for the high-frequency signal path.

In the English-language literature, this layout is called a Gilbert cell. It has recently become an established constituent part of, for example, all radio phones and mobile phones, in a frequency range going far beyond 2 GHz.

The only disadvantage it has proved to have so far is the maximum level control, characterised by the IP3. Even for the best designs (with suitably high levels of zero signal current), at app. 32dBm this is still below the best brands, which are set, through the



N.B.: All assemblies for setting operating point have been omitted!

Fig.70:

Principle of the Gilbert Cell

Baugruppe 2 (nichtumkehrender Verstärker) = Assembly 2 (non-reversing amplifier),

Baugruppe 3 (umkehrender Verstärker) = Assembly 3 (reversing amplifier),

Baugruppe 1 = Assembly 1, Elektronischer Umschalter =

Electronic change-over switch, Betriebsspannung = Operational voltage, Ausgang = Output

pedigree passive ring modulators referred to, with diodes or power MOS-FET switches (giving $IP_3 = +40$ to 45dBm).

To sum up, we can say that the multiplicative mixer has made the breakthrough in terms of professional applications where the spectrum purity of the output signal has to be high.

The industry has recently become able to supply a very wide range of suitable circuits and finished modules, the SMD versions in particular being more advantageously priced (for the same quality). There are also numerous data manuals and application manuals, with excellent articles explaining the principles involved and notes on applications [1], [2] and [3].

10.4. Examination of Diode Ring Mixer Used

Here, once again, we begin with a short list of the advantages of this type of mixer:

- High level control, characterised by a high IP_3 value.
- Universal module, since it is suitable for use as a mixer, modulator or de-modulator for AM or SSB, a phase detector or a controllable attenuator.
- No additional current supply required, since this is a passive component.
- Enormously wide bands can be covered for all inputs (high-frequency and LO inputs run through small magnetic core transmitters, which means there is a lower limiting frequency of app. 20 MHz there, whilst the upper limiting frequency of this transmitter is app. 2 GHz).
- High-frequency and intermediate-frequency outputs may be transposed, which now leaves only high-frequency signals from DC up to 2 GHz to be high-mixed.
- Automatic matching in 50Ω systems.

But there are disadvantages as well:

a) As a passive component, the ring modulator displays at least 6 to 7dB attenuation with diodes.

b) This not only means we need additional amplification in the system, it means the noise factor in this assembly also increases by this value.

c) The turn-on voltage required (LO signal) must have a very high minimum amplitude. Depending on the type of mixer, this means an LO level of between + 7 and + 25dBm.

d) The matching, expressed, for example, in terms of the VSWR ratio, is not especially good, for in the connecting stretch between the high-frequency input and the intermediate-frequency output the transmitter must always be taken into account, together with the dynamic resistances of the diodes. The diodes are switched, and so the resistances in the signal path are continually changing and change again if the LO level fed in is increased or reduced!

e) Meticulous attention must be paid to obtaining a broad-band 50Ω adapted load, as near perfectly as possible, at the intermediate-frequency output, since otherwise for example, if a quartz filter is directly connected all signals located outside the filter's transmission band are reflected back into the mixer. They "wander around" there like ghosts, with unknown phase positions. In certain circumstances, they re-mix, and thus almost always worsen the hard-won high attenuation values for the undesirable signal fractions!

f) In professional systems, connecting up broad-band attenuators (usually 3dB)

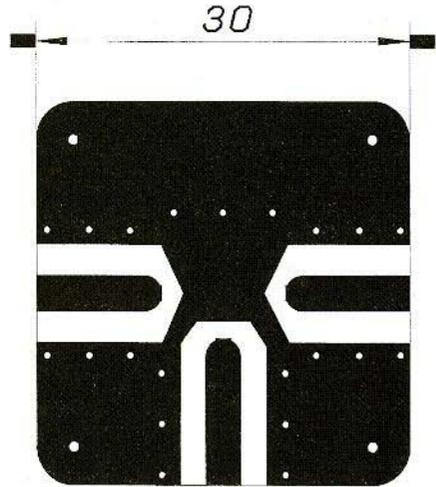


Fig.71: Board Layout for Mixer Test Structure

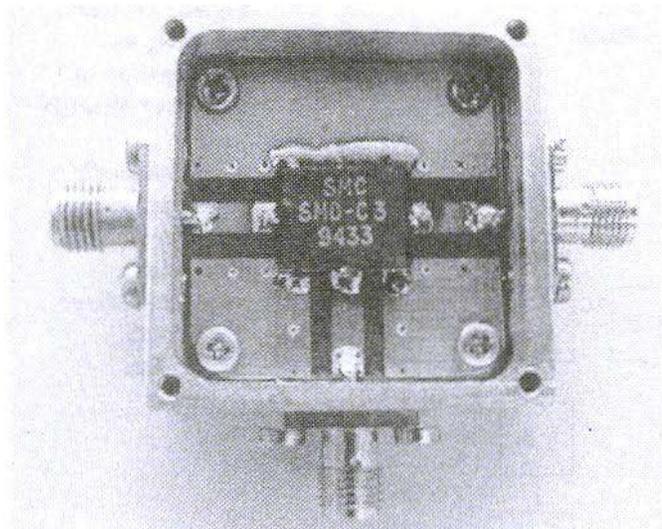
The recommended application range extends beyond 2 GHz; + 7dBm should be allowed for as the LO level.

Inside this cast resin cube are two ring cores for the necessary transmitters and a quartet of Schottky diodes. These components are mounted on a little printed circuit board, the three connection points of which are connected with the exterior. 50Ω microstrip lines are now connected up to these on the board.

With the layouts used so far, this has not proved a problem. The strip line width is 2.92mm for FR4 material with a thickness of 1.5mm at a frequency of 1.7 GHz. Naturally, the required feedthroughs from the continuous earth surface to the individual "earth islands" of the layout should once again be created using silver-plated compression rivets ($d = 0.8$ mm.). The finished board is fastened into a milled aluminium



Fig.72:
Photograph of
Prototype



housing, with external dimensions of 35 x 35 x 15 mm, with four nickel-plated M 2.5 x 6 cylindrical screws.

The SMA sockets are precisely positioned in such a way that their neutral conductors lie just on the 50Ω microstrip lines, and can be soldered on with a little solder containing silver, without any problem. In order to avoid irregularities, all neutral conductors are chamfered below 45° with a fine file before being soldered on.

Nor should we forget to drill a “matching” hole for the neutral conductor through the housing wall. With a conductor diameter of 1.27 mm., and for 50Ω, this gives a hole diameter of 2.9mm.

Fig.71 shows the board layout. Fig.72 gives us a look at the prototype, ready to operate, but with the lid off.

Now to the electrical properties. The following characteristics are of interest to users. Most serious manufacturers include them in the data sheet:

- a) Extent and frequency range of conversion loss or amplification
- b) 1dB compression point
- c) IP3
- d) SSB noise factor
- e) Insulation of various ports from one another, i.e. a dimension for the reciprocal crosstalk
- f) VSWR or reflection factors at individual ports

Naturally, a very extensive and expensive array of gauging equipment is required to capture all these mixer data, and not everyone has this available.

However, tests have been carried out on mixers, the most important being the spectrum analyser with a small signal frequency response. The gauging rig shown in Fig.73 was somewhat modified as required.

The two standard signal generators represent the high-frequency and LO signal sources. The output level must be

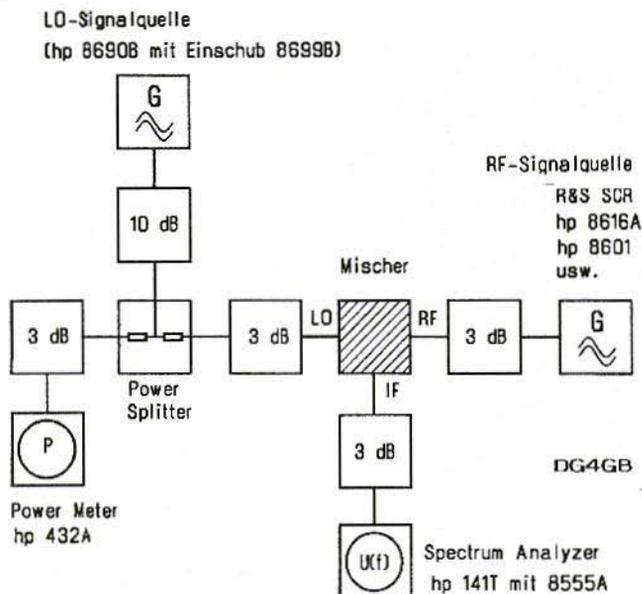


Fig.73: Measurement Rig for Investigation of Mixer Module

*Signalquelle = Signal source,
Mit Einschub = With insert,
Mischer = Mixer,
Mit = With,
Usw. = Etc.*

stabilised, and it must also be possible to alter it within precise limits. The attenuators at all the mixer ports ensure correct broad-band tuning.

10.4.1. Conversion Loss

As is well-known, when the high-frequency signal is converted in a mixer, the energy is divided into a lower and an upper sideband component frequency, namely into the total frequency and the difference frequency. Since both these fractions are of equal size, each contains half of the high-frequency energy. Thus with an ideal mixer the conversion loss is 6dB.

In practice, because of unavoidable losses in transmitters and diodes, the value is higher. Moreover, there is another big problem in determining data

sheet specifications, as the mixer manufacturers are apparently not fully in agreement among themselves:

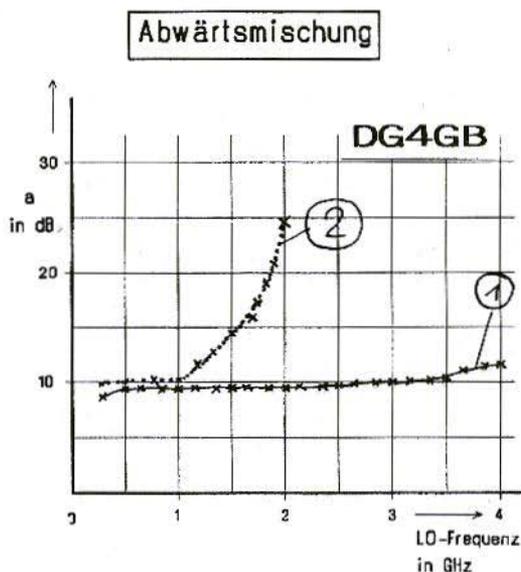
a) Some work at a constant intermediate frequency (in point-to-point radio systems, for example, this is 70 or 140 MHz). The high frequency and the LO frequency are increased until the amplitude of the intermediate frequency output signal falls by 3dB, or the conversion loss increases by 3dB. The frequency interval of the two signal generators must naturally correspond to the intermediate frequency and is kept constant.

This represents a feasible, practically-orientated method of designing a receiver with a low intermediate frequency or a reception converter, and it corresponds to the operating case of the downward mixer.

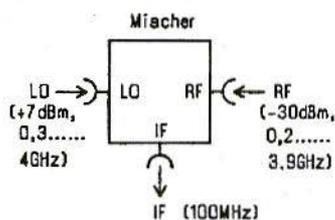
b) In the second option, the high frequency is kept constant (for example,



Conversion loss = f (Frequenz)



Meßschaltung für Kurve 1



Meßschaltung für Kurve 2

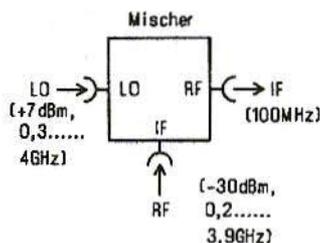


Fig.74: Downwards Mixing; *Abwärtsmischung* = Downwards mixing, *Frequenz* = Frequency, *Meßschaltung für Kurve* = Measurement circuit for curve, *Mischer* = Mixer

HF = 100 MHz), and the LO frequency is increased until the amplitude of the total or difference frequency arising decreases by 3dB or the conversion loss increases again by 3dB. This corresponds to operation as an upward mixer, as is done in SSB transmitter frequency synthesising, or in receivers with a high first intermediate frequency. In order to be clear on the behaviour of this circuit, both options are systematically examined below.

A) Operation as downward mixer:

Here a constant intermediate frequency of 100 MHz was selected. So there must always be a frequency spacing of 100 MHz between LO and high-fre-

quency signals, and they are increased together until the conversion loss becomes unacceptably great. The high frequency is lower than the LO frequency here.

The precise operational values here, based on the mixer inputs, are:

LO signal: 300 MHz to 4 GHz, with a level of +7dBm at the LO port

HF signal: 200 MHz to 3.9 GHz, with a level of -30dBm at the HF port

The intermediate frequency observed was 100 MHz. Fig.74 shows the conversion loss frequency response. The curve is so flat that this low-cost component can actually be used at up to 4 GHz!

As part of the trial, the intermediate-frequency and high-frequency ports were

Conversion loss = f (Frequenz)

Aufwärtsmischung

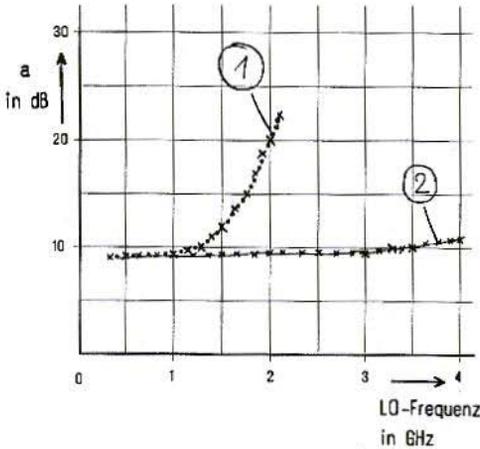


Fig.75: Upwards Mixing

Aufwärtsmischung = Upwards mixing, *Frequenz* = Frequency,
Schaltung für Kurve = Circuit for curve, *Mischer* = Mixer

also swapped round once, and the measurement was repeated (see also Fig.74). The result is quite shattering, for now the conversion loss is already 10dB greater at 1.5 GHz, and is more than 20dB greater at less than 2 GHz.

B) Operation as upward mixer:

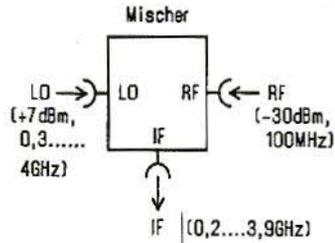
Now a constant high frequency of 100 MHz is fed into the mixer, and the LO frequency starts at 300 MHz and is increased in 100 MHz steps.

The following operating conditions were maintained:

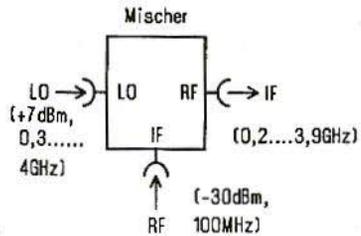
LO signal: 300 MHz to 4 GHz, with a level of +7dBm at the LO port

HF signal: 100 MHz, with a level of -30dBm at the HF port

Schaltung für Kurve 1



Schaltung für Kurve 2



The lower sideband component frequency (thus LO-HF) observed went from 200 MHz to 3.9 GHz.

As our curiosity had been aroused by the result of the first trial, we swapped the high-frequency and intermediate-frequency ports round here as well after the first pass, and then we repeated the measurements. The results can be seen in Fig.75.

There was a big surprise here too.

A circuit which can be used up to 4 GHz, with constant and relatively small conversion loss, is obtained only if the "high-mixing" HF signal of 100 MHz is fed into the intermediate-frequency port. However, if the above downwards mixer circuit is used by mistake, the mixer can be used at up to just 1.5 GHz. If the



Conversion loss = f (LO - Pegel)

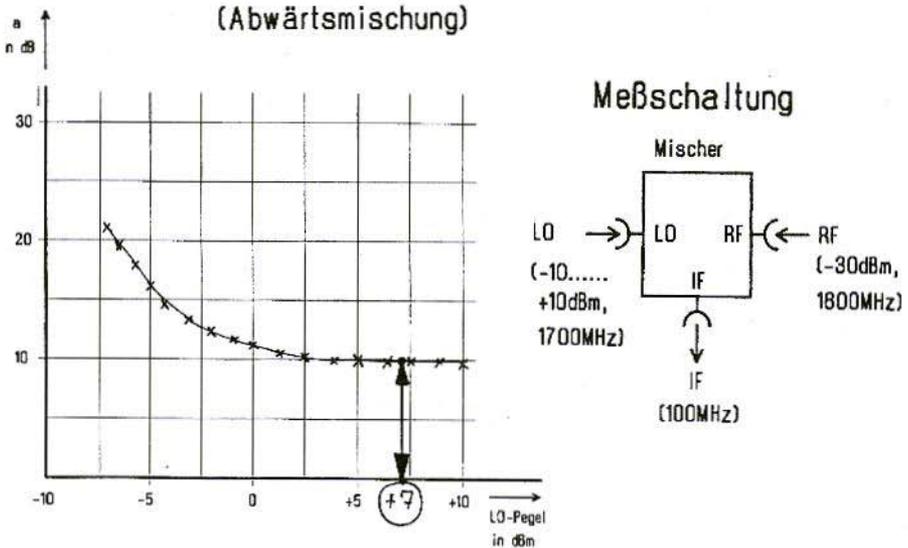


Fig.76: Relationship Measured between Conversion Loss and LO Level with Constant HF Signal; *Aufwärtsmischung* = Upwards mixing, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer

frequency is increased any further, the conversion loss immediately takes on astronomical proportions.

10.4.2. Relationship Between Conversion Loss and LO Level

It was something that was being hinted at back in the chapter on principles. The faster and the more abruptly the diodes are switched through the LO signal, the better the mixer will be. In this context, this means better with respect to the conversion loss and the undesirable mixing products generated.

Fig.76 now demonstrates the relationship measured between the conversion

loss and the LO level for a constant HF level.

The curve was obtained under the following conditions:

LO signal: 1,700 MHz, with a level of -10 to +10dBm at the LO port

HF signal: 1,800 MHz, with a level of -30dBm at the HF port

The difference frequency observed was 100 MHz. It can easily be recognised from this that the manufacturer's note "Please select the LO level at + 7dBm" is certainly absolutely right. If the LO level is increased any higher, this brings no further improvements, at least with this HF amplitude and with respect to the conversion loss.

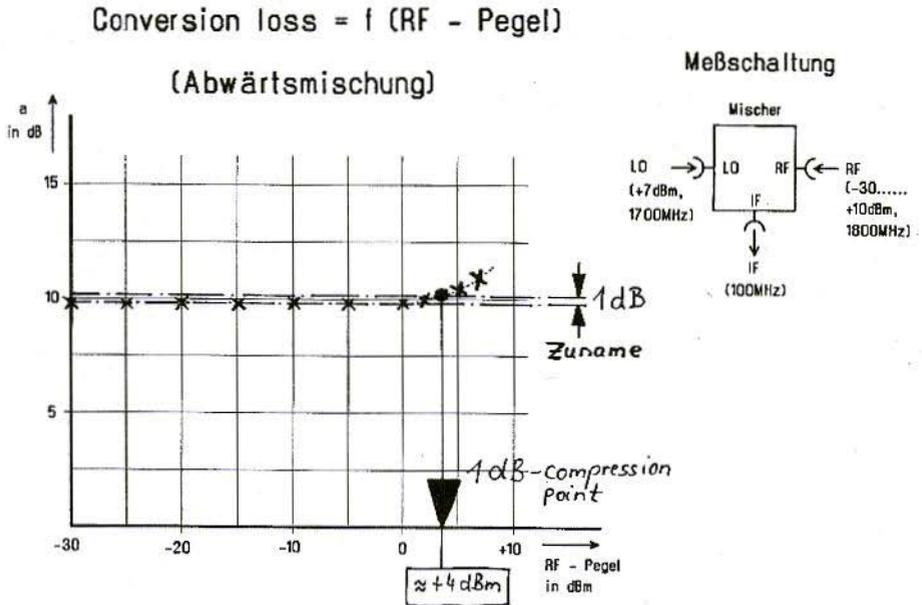


Fig.77: Measurement Curve with 1dB Compression Point: Curve shows Conversion Loss plotted against HF Level
Abwärtsmischung = Downwards mixing, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Pegel* = Level

10.4.3. The 1dB Compression Point

If we assume that the LO level of +7dBm has been correctly selected, we can see the consequences of changing the HF level.

Fig.77 shows the measurement results for the following operating case:

LO signal: 1,700 MHz, with a level of +7dBm at the LO port

HF signal: 1,800 MHz, with a level of -30 to +10dBm at the HF port

The difference frequency of 100 MHz was observed. The marked point at which the conversion loss increased by 1dB can (I hope) be clearly seen. This "1dB compression point" is important for three reasons:

a) This is where the saturation or limitation begins. From this point onwards, there is no point in increasing the HF level any further, because the intermediate-frequency output level no longer alters in line with the change, but very quickly becomes almost constant again.

b) The inter-modulation signals generated rise rapidly from this point onwards, and so clearly visibly that the spectrum analyser connected at the intermediate-frequency output suddenly begins to show a "lattice" of interference lines.

c) However, if we look more closely at the design manuals already mentioned, in [5], for example, we can find this:

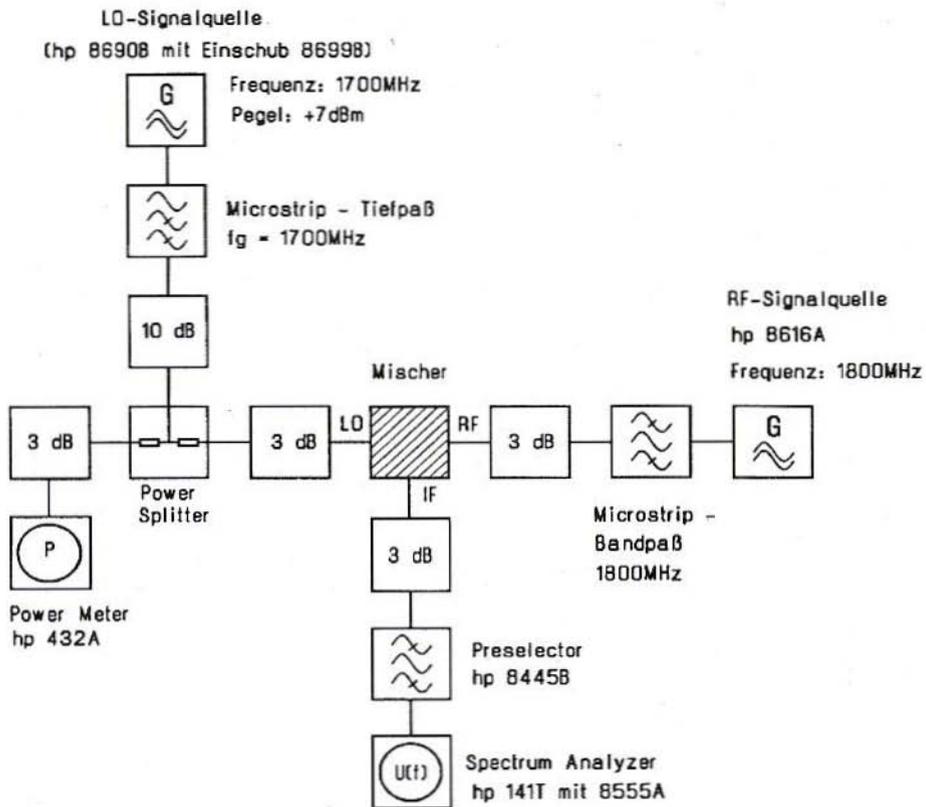


Fig.78: Extended Measurement Circuit for Determining IP

Signalquelle = Signal source, *Mit Einschub* = With insert,
Frequenz = Frequency, *Pegel* = Level, *Tiefpaß* = Low pass,
Mischer = Mixer, *Bandpaß* = Band pass

"If about 15dB is added at the 1dB compression point at the lower band end, this gives the approximate value of the third order intercept point, IP₃. The value to be added to the upper band end should then be reduced to 10dB."

10.4.4. The Intercept Point

The biggest problem here is probably the measurement rig needed to determine the values.

The intercept point itself is actually a fine thing. It supplies very accurate information concerning the distance between the undesirable interference signals arising and the desirable intermediate-frequency signal at a specific HF level. The formulation often heard of in connection with receivers, "inter-modulation free dynamic range" is relevant here as well.

Here is a short summary of the theoretical and physical basic principles:

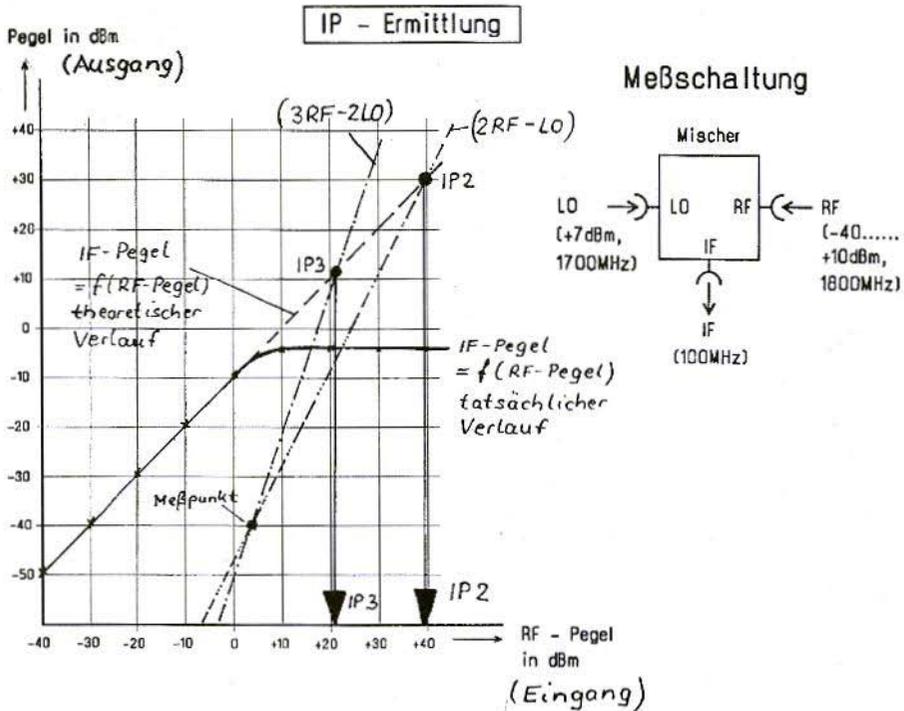


Fig.79: Graphic Determination of IP's

Pegel = Level, *Ausgang* = Output, *Theoretischer Verlauf* = Theoretical gradient, *Tatsächlicher Verlauf* = Actual gradient, *Meßpunkt* = Measurement point, *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Eingang* = Input

As soon as the HF level fed into the mixer is increased, the desirable intermediate-frequency output level also increases, up to saturation. Unfortunately, harmonic mixing on non-linear characteristics leads to many new combination frequencies, namely the notorious intermodulation signals. And this is where things get really unpleasant. Although their amplitude is very small at first, they increase much faster than the intermediate-frequency level when the HF input level rises. At some point (if no saturation takes place), this undesirable fraction becomes as large as

the amplitude of the desirable total frequency or difference frequency, and this level intersection point is the famous IP!

It should also be pointed out that the rise becomes steeper and steeper as the higher-order harmonics are mixed in.

The following conditions apply:

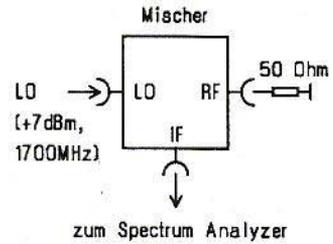
If the HF level is increased by 10dB, then:

- a) Second-order signals increase by 20dB. The annoying frequency here is usually $(2 \times \text{HF} - \text{LO})$.



Meßschaltungen

Kurve 1



Kurve 2

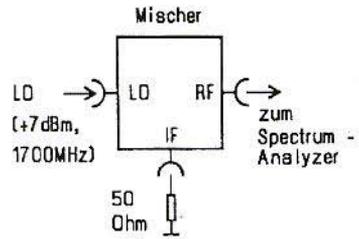


Fig.80: Measurement of Isolation of LO Signal

Frequenz = Frequency, Meßschaltungen = Measurement circuits, Kurve = Curve, Mischer = Mixer, Zum = To

b) Third-order signals increase their amplitude in this case, but by 30dB. The frequency ($3 \times HF - 2 \times LO$) is particularly unpleasant here.

Signals of still higher orders normally cause no interference, since their amplitudes usually remain insignificantly low, as long as we keep far enough away from mixer saturation.

But be careful. The individual level intersection points, IP2 and IP3, are not identical!

There are particular problems here with IP3, for the third-order characteristic rise is unfortunately three times as great as for the intermediate-frequency signal.

As a result, the associated inter-modulation products increase to a correspondingly great extent.

This is particularly unfavourable in practice if the mixer for example, in a receiver has too high an input level, and the HF signal consists of a fully occupied amateur band!

In addition, of all places for combination frequencies of this kind to be, they are often just at the place where the desirable intermediate frequency also emerges. For can a human being or a filter keep wanted signals and interference signals apart?

So now to the measurements.

Mixer manufacturers operate with a "two-tone HF signal" here, consisting of two closely adjacent measurement frequencies. This group of two is now converted with the LO signal, and the result is examined on the spectrum analyser. Again, the total signal and the difference signal naturally consist of these groups of two. But in addition, we discover, in all their glory, the harmonics produced in the mixer and also converted. Their attenuation can be easily read off against the desirable intermediate frequencies. The overall name for this is "dual tone third-order inter-modulation testing", and it is a feature of the specifications of commercially available mixers.

Of course, this method of measurement is very costly, since every one of the three separate signal sources must also be freed from inherent distortion. For this is the only way to determine the fractions which arise from the mixing process. The measurement rig used here is still very extensive, and is based on the following observations (Fig.78):

- a) The HF is selected at 1,800 MHz, and the LO signal at 1,700 MHz.
- b) The LO level is set precisely to +7dBm at the LO port and is kept constant.
- c) Both signals are freed from inherent distortion through the filters discussed in the earlier chapters (microstrip band-pass and microstrip low-pass).
- d) The signal ($2 \times \text{HF} - \text{LO}$) is observed to determine the IP2. It is located at $(2 \times 1,800 \text{ MHz} - 1,700 \text{ MHz}) = 1,900 \text{ MHz}$.
- e) The IP3 is determined by observing the combination frequency ($3 \times \text{HF} - 2 \times$

LO). It has the value $(3 \times 1,800 \text{ MHz} - 2 \times 1,700 \text{ MHz}) = 2,000 \text{ MHz}$.

f) The HF level is now raised higher and higher until both these new frequencies can clearly be recognised on the screen and emerge from the internal noise level of the analyser. The HF level was then advanced until the amplitudes of the two inter-modulation products were exactly the same. It only remained to set the three levels at the HF level arriving at the HF port, together with the two levels ($2 \times \text{HF} - \text{LO}$), or ($3 \times \text{HF} - 2 \times \text{LO}$), as accurately as possible.

The results showed that with an HF level of +4dBm at the HF input of the mixer, the two signals ($2 \times \text{HF} - \text{LO}$) and ($3 \times \text{HF} - 2 \times \text{LO}$) are exactly -40dBm.

Now to the diagram, into which we first plot the relationship determined in the previous trial between the HF input level and the intermediate frequency amplitude arising. Because of the saturation, the curve naturally goes over into a horizontal line at the end. But the ideal theoretical course, which forms a straight line, is extended until we arrive at the limits of the diagram.

Two straight lines are now drawn through the measurement result (-40dBm). Their gradients are two or three times as acute as that for the relationship already plotted between the HF signal and the intermediate frequency signal, and they intersect this curve at two points (Fig.79). These two intersections correspond to IP2 and IP3!

As conscientious researchers, we also had to determine the amplitude of the intermediate-frequency signal arising



and plot it in the diagram. It is immediately apparent how far into saturation we already are.

We thus know that:

IP2 is at about +39dBm. IP3, on the other hand, is at about +21dBm.

Anyone who wants to can check the opinion repeated in the previous chapter on the relationship between the 1dB compression point and the IP3.

With the help of this diagram, it is now no longer a problem to determine the inter-modulation distance. We need only draw a vertical line through this level (on the horizontal axis), mark its intersections with the three straight lines (gradients 1, 2, 3), read off the distances of the inter-modulation points from the intermediate frequency point, and express the result in dB.

Anyone who prefers to work with formulae, on the other hand, could easily put this geometrical relationship into a corresponding expression and use it to calculate the values desired. A still easier way is to follow the method also referred to in [6]:

a) Determine the value of IP3 either from the data sheet or by adding 15dB (lower band end) or 10dB (upper band end) to the 1dB compression point.

b) Then determine the difference in dB between the HF input level applied and IP3.

c) Multiply this difference by the order of the harmonic (here: 3).

d) Subtract this result and the conversion loss from IP3. This gives us the level of the third order inter-modulation interference signal.

If this method is applied to the HF level used for all measurements, -30dBm, this gives us:

a) Measured value: IP3 = +21dBm

(The rule of thumb would have been somewhat less accurate: 1dB compression point of +4dBm +15dB = 19dBm!)

b) IP3 - HF = +21dBm - (-30dBm) = 21dBm + 30dBm = +51dBm

c) 51dB x 3 = 153dB

d) IP3 - 153dB - 10dB conversion loss = +21dBm - 153dB - 10dB = -142dBm

In this operational case, the level for inter-modulation signals is thus -142dBm, which is equivalent to 112dB below the HF level. But the difference in level at the desired intermediate frequency is much more important for the user, for there the mixer's conversion loss (app. 10dB) must also be taken into account.

Thus the final value should be 102dB below the intermediate frequency signal, and this is precisely what can be seen on the spectrum analyser.

If this seems an unbelievably good result, just imagine an HF level 30dB higher. This rise unfortunately brings about an increase of 90dB in the inter-modulation products (!).

If the intermediate frequency is now -10dBm, because of the app. 10dB conversion loss, the unwanted IM3 signal unfortunately bounces up to -14dBm +90dB = -52dBm. And so suddenly the distance between the intermediate frequency signal and the interference product is only 42dB. That could just cause some difficulties.

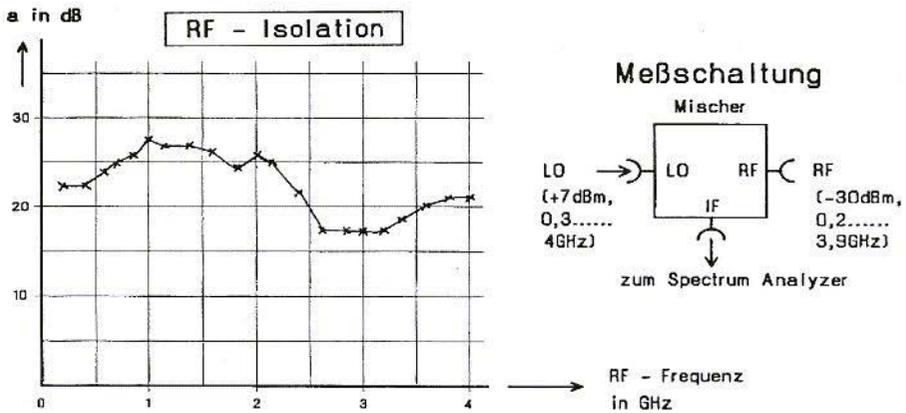


Fig.81: Measurement of Isolation of HF Signal; *Meßschaltung* = Measurement circuit, *Mischer* = Mixer, *Zum* = To, *Frequenz* = Frequency

10.4.5. SSB Noise Factor

Luckily, this is something which can be settled very quickly. As already mentioned, the intermediate frequency amplitude is reduced by the conversion loss with respect to the HF signal fed in. Thus the noise factor must rise by this amount. According to the mixer manufacturers, the SSB noise factor is calculated by adding a mere 0.5dB to the conversion loss. This gives us the noise factor we need.

10.4.6. Isolation of Ports

Nothing is perfect, and so here too we find remains of the signals fed into the mixer arriving at the intermediate frequency output port or at the opposite signal port. The difference in level between the signal fed in and the measurable residue at another port is known as isolation. If this value deteriorates too much, we can get some nasty surprises. For example, imagine a direct conversion receiver with an LO signal

which suddenly starts to wander around on the antenna. The isolation behaviour should therefore also be entered in the mixer data sheet.

Fig.80 shows the isolation values of the LO signal for the frequency range 0.2 to 4 GHz, with a level of + 7dBm at the LO mixer input. At the intermediate frequency output the spectrum analyser is hanging, with the help of which we can look for the LO residues. The HF port, which is not required, is blanked off with 50Ω.

The measurement was also repeated with the HF and intermediate frequency ports swapped round. The striking features in the result are the very high conversion losses at low frequencies and their deterioration when the frequency increases. This is very unpleasant for many applications.

Fig.81 shows how high the HF signal at the intermediate frequency port can still be found to be with "correct downwards mixer operation".

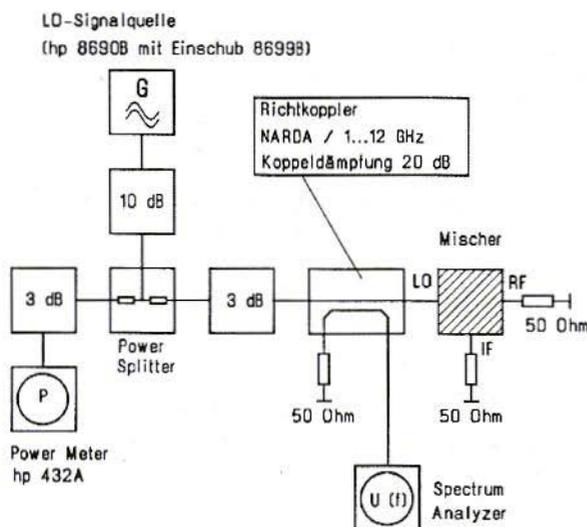


Fig.82: Experimental Set-Up for Determination of Reflection Attenuation at LO Port

Signalquelle = Signal source,
Mit Einschub = With insert,
Richtkoppler = Directional coupler,
Koppeldämpfung = Coupled attenuation,
Mischer = Mixer

Operational values:

LO signal: 0.3 to 4 GHz, with a level of +7dBm at the LO port

HF signal: 0.2 to 3.9 GHz, with a level of -30dBm at the HF port

Intermediate frequency: 100 MHz

As against Fig.80 (LO isolation), these

readings show markedly fewer fluctuations and are less dependent on the frequency. But unfortunately, the downside is that they are worse results.

But be careful: the conversion loss for the intermediate frequency signal also has to be taken into account if we are trying to determine the distance between

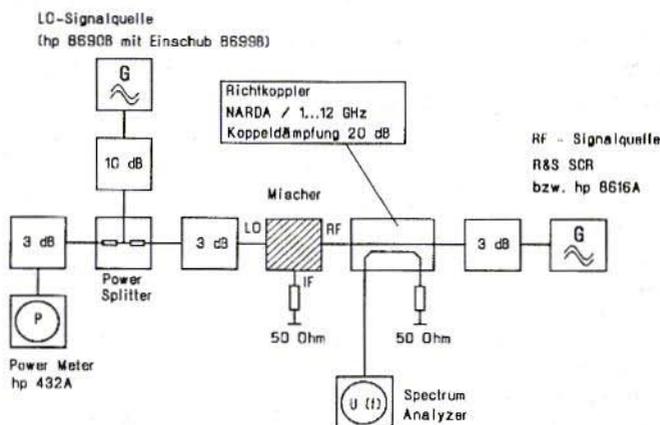


Fig.83: Determination of Reflection Attenuation at HF Port

Signalquelle = Signal source, *Mit Einschub* = With insert,
Richtkoppler = Directional coupler, *Koppeldämpfung* = Coupled attenuation,
Mischer = Mixer, *Bzw.* = And / or

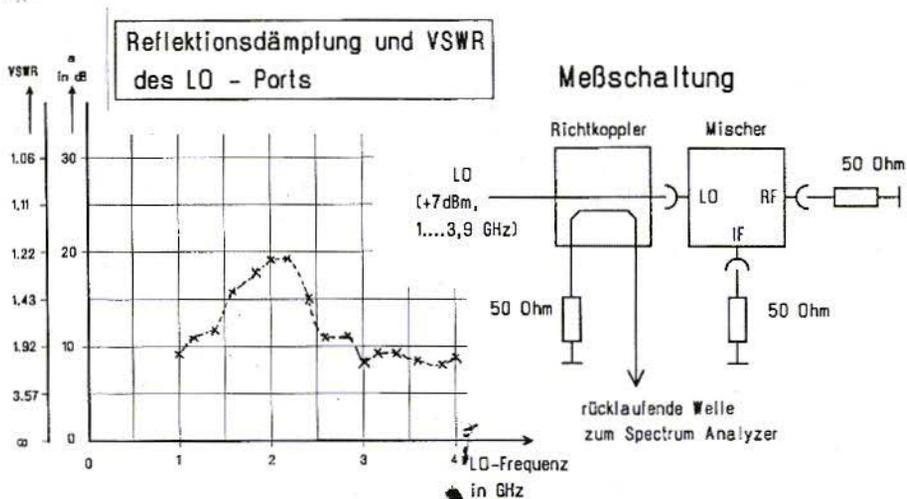


Fig.84: Measured Behaviour of Reflection Attenuation at LO Port
Reflektionsdämpfung und VSWR des LO-Ports = Reflection attenuation and VSWR of LO port, Frequenz = Frequency, Meßschaltung = Measurement circuit, Richtkoppler = Directional coupler, Mischer = Mixer, Rücklaufende Welle zum = Return wave to

the desirable intermediate frequency signal and the HF residue at the intermediate frequency output, for this makes the level differences approximately 10dB worse. This is important for the layout of the subsequent intermediate frequency band pass, with reference to the inhibition attenuation required to suppress undesirable signals!

10.4.7. VSWR or Reflection Attenuation Levels for Individual Ports

Some manufacturers give only meagre information about this, or none at all, but others give precise details. However, when we look at these more closely, they provide very little encouragement, for the VSWR reaches, and sometimes exceeds, a value of 4. And no wonder, for not only the transmitter in the mixer

and the diode grades, but also the LO level at which the diodes are switched, have a role to play here, as has the adapted load error at the other ports. It is therefore inevitable that attenuators or supplementary circuits should be mounted around this mixer. The mixer used here is itself not perfect, which means it imposes even higher requirements on its environment.

The measurement is somewhat costly, but the manufacturers can use the most modern network analysers for it. With these appliances, the level fed into the test object can be precisely adjusted and monitored.

This is exactly what is important, for example, at the LO port, for the correct switching through of the diodes in the mixer. It does not become really tricky unless measurements have to be made at

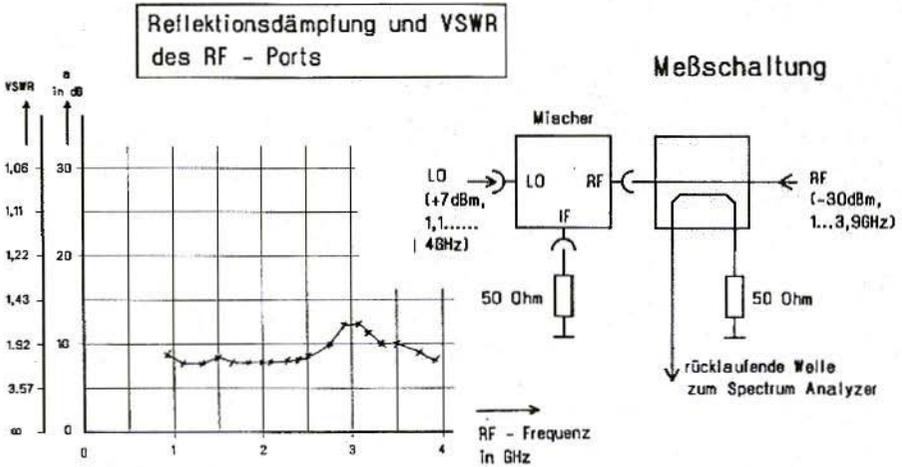


Fig.85: Measured Behaviour of Reflection Attenuation at HF Port
Reflektionsdämpfung und VSWR des RF-Ports = Reflection attenuation and VSWR of RF port, Frequenz = Frequency, Meßschaltung = Measurement circuit, Mischer = Mixer, Rücklaufende Welle zum = Return wave to

the HF port, since then the LO port must also be selected, and both frequencies must be moved in synchronisation with one another for downwards mixer operation. This can not be done on older appliances, so we have to fall back on older methods using directional couplers (see Figs. 82 and 83). Fig.84 therefore shows this type of measurement only for the reflection attenuation behaviour at the LO port for the frequency range between 1 GHz and 4 GHz.

Fig.85 provides the reflection values required for the HF port in "downwards mixer operation", measured in the frequency range between 1 GHz and 4 GHz, with an intermediate frequency of 100 MHz. As conjectured, the results are still just usable.

10.5. Development Trends

It has already been pointed out that the integrated, multiplicative GaAs mixers are making great advances, and have driven all older designs, including diode ring modulators, blow-through mixers, etc., out of the market (and not just for mobile telephones).

The SAT receivers behind the Astra mirrors also contain only complete single-chip receivers, including PLL synthesisers for frequency ranges going up to 3 GHz. The LNC's for the frequency range from 12 GHz now contain almost exclusively the cascade stages referred to, acting as combined mixers and oscillators.

So this article will undoubtedly not be the last word on the subject of microwave mixers.

Moreover, the newest specimen IC's are already waiting on my workbench (manufacturer's specifications give the frequency range as 0.1 to 8 GHz), to be inserted into a suitable "Puff design" board. I shall be reporting on this right here at the appropriate time.

11. Literature

- | | |
|--|--|
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App. 700 pages, DIN A4</p> <p>[2] Communication Components / GaAs & Silicon Products Designer Catalogue from HP/Avantek</p> | <p>[3] RF- and Microwave Semiconductors (Discrete, Monolithic and Multifunction) Handbook from M/A-COM</p> <p>[4] Berberich, Eugen, DL8ZX Diplexers for Ring Mixers VHF Communications 2/96, pp. 75-81</p> <p>[5] RF/IF Designers Handbook Mini-Circuits Chapter on "Two-tone, third-order distortion effects" P. 1/6</p> <p>[6] RF/IF Designers Handbook Mini-Circuits Chapter on "Two-tone, third-order distortion effects" P. 1/9</p> |
|--|--|

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Hermann Hagn, DK8CI

Hot-Cold Noise Temperature Measurement in the Laboratory and in Nature

Based on a paper read at VHF-UHF 1996 in Munich

Measurement methods are described which make it possible to measure the noise temperature of receiver installations in the range between 1.3 GHz and 47 GHz. The noise numbers to be measured here lie in a range from about 0.4dB at 1.3 GHz up to 10dB at 47 GHz.

For noise numbers above and below these limits, these methods become inaccurate, because a level variation which is either greater or smaller must be measured very precisely.

1. PRINCIPLES

The laboratory measurements are based on the discoveries of Johnson and Nyquist [1], and those in nature on

Planck's radiation formula. Since we are carrying out our measurements at lower frequencies than those corresponding to the maximum on the Planck radiation curve, the Rayleigh-Jeans approximation applies. These inter-relationships are described in detail in [2]. In the laboratory, we use a resistance, $R (= Z)$, in nature a matched, loss-free antenna (which "sees" the temperature, T , over the entire apex angle), with the radiation resistance, R , at the input of our receiver installation (Fig.1). The input impedance is not changed, and we have $R = Z$ (impedance level).

In both cases, we measure the output:

$$P = k \cdot T \cdot B \cdot V$$

P in W, B in Hz, T in K

Boltzmann constant $k = 1.38 \cdot 10^{-23}$ J/K

V is the power amplification factor for our installation.

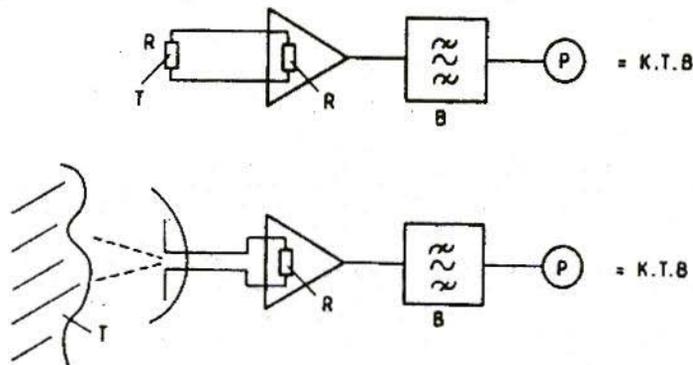


Fig.1: Comparison of Load Resistor with matched, polarised Antenna during Line Adaptation ($R_i = R_o$) leads to identical Noise Power

2. MEASUREMENTS IN THE LABORATORY

The resistor, $R (= Z)$ at the input of our installation is brought to two defined temperatures: e.g. 78K with liquid nitrogen and room temperature, 293K, and the precise values are measured using a thermometer.

Now, if we had an ideal amplifier, so we would measure a power ratio at the output which corresponds precisely to the ratio of the two temperatures. But since all amplifiers contribute some noise, the power ratio measured will be smaller than the temperature ratio. To the reference temperatures, we must also add the noise temperature fraction, T_E , of the installation to be measured. From the two outputs measured, or else from the measured power ratio, we can calculate the noise contributed by the installation (H = hot, K = cold):

$$P_H = k \cdot (T_H + T_E) \cdot B \cdot V$$

$$P_K = k \cdot (T_K + T_E) \cdot B \cdot V$$

Here, B and V must be very accurately known, so that it is simpler and more precise to use the power ratio for the calculation:

$$\frac{P_H}{P_K} = \frac{T_H + T_E}{T_K + T_E} = y$$

$$T_E = \frac{T_H - yT_K}{y - 1}$$

The noise figure, F (dB), can then be calculated from T_E , based on the standardised temperature of 290 K:

$$F_{(dB)} = 10 \cdot \log \frac{290 + T_E}{290}$$

(F in dB, T_E in K)

Likewise, the following relationship applies:

$$F_{(dB)} = 10 \cdot \log k T_0$$

F in dB, $k = 1.38 \cdot 10^{-23}$ J/K,
 $T_0 = 290$ K



If we wish to determine the noise temperature or the noise number of the input amplifier from the overall values measured, then we have to measure the noise temperature of the subsequent electronics and determine the amplification factor of the input amplifier.

According to H.T.Friis [3], the following relationship applies:

$$F_{(\text{ges})} = F_1 + \frac{F_2 - 1}{V_1}$$

$F_{(\text{overall})}$, F_1 and F_2 in $k T_0$

V_1 = output amplification factor of pre-amplifier

3. MEASUREMENTS IN OPEN AIR

We now need two radiating media, the temperatures of which are known. The warm radiator is the ground, and here it is best to measure downwards at an angle into thick bushes or into a wood.

At low frequencies (below 10 GHz), the ground temperature measured should be multiplied by a correction factor [5]. Our own measurements gave us a value of 0.9 at 1.7 GHz. At these frequencies, the ground is still no "black radiator", and there are also still reflections from the cold sky to be added. A clear difference is noted, depending whether we "look" down at an angle onto a meadow or measure into a wood. This correction factor becomes 1 from about 10 GHz onwards, so that interpolations

can be made with good accuracy between 1 and 10 GHz.

The cold radiator is the cloudless sky (50 to 80° elevation to the North; for very sensitive installations, you need to pay attention to the CAS A radio source). The ground temperature is measured using a thermometer, and 35 K (T_A) can be taken as a good value for the cold sky in our range widths [4, 5]. This value for the cold sky is made up of: the background radiation [6], unresolved far distant radio sources, atmospheric radiation, ohmic losses from the antenna and ground radiation through side lobes of the antenna. For this measurement it is of no importance what the antenna's gain is, provided the antenna is completely "illuminated" by the radiator in question (warm or cold) over its apex angle.

Example:

Antenna 1.75 m. diameter
Frequency 1.69 GHz
Apex angle app. 7.5° (measured)

Measured power ratio, ground / cold sky, 3.8dB; 3.8dB = 2.4 fold, $y = 2.4$;
 $T_H = 290\text{K}$; $C = 0.9$ correction factor at 1.7 GHz; $T_K = 35\text{K}$

$$T_E = \frac{C \cdot 290 - 2.4 \cdot 35}{2.4 - 1} = 126 \text{ K}$$

$$F_{(\text{dB})} = 10 \log \frac{290 + 126}{290} = 1.57 \text{ dB}$$

$$T_{\text{system}} = T_E + T_A = 126 + 35 = 161 \text{ K}$$

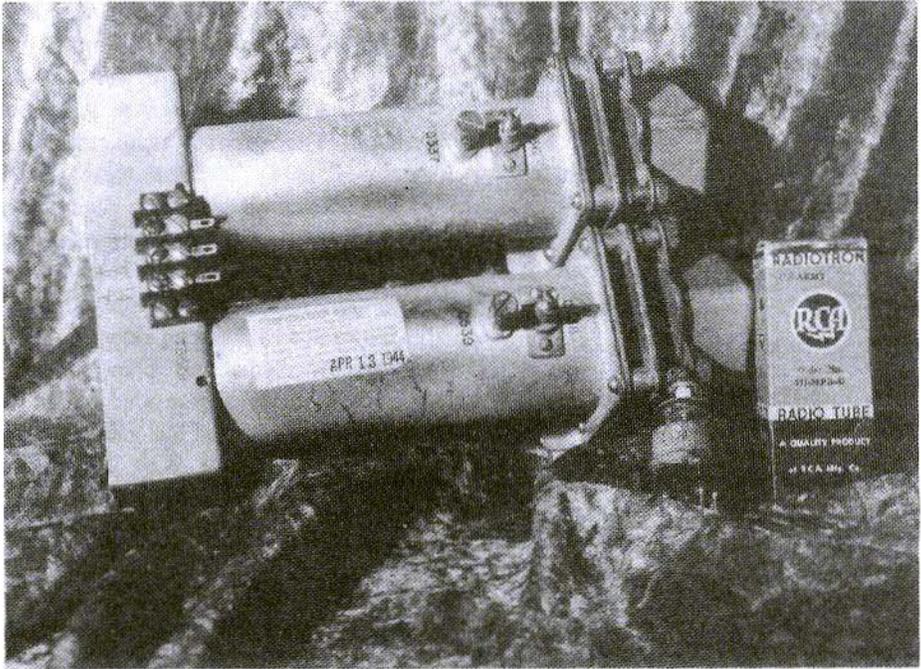


Fig.2: American UHF Tube Amplifier from 1944 (Photo: Kerstin Hang)

In many installations, the noise power at the output changes markedly if we go over from the resistor at the input to the antenna. This effect depends on how good or bad the SWR is between the input amplifier and the antenna.

It is very strongly recommended that you should use a low-attenuation (< 0.3 dB) circulator / insulator at the amplifier input. Naturally, another option is to calibrate to the maximum output divergence between the ground and the sky and then make no further changes on the pre-amplifier.

4. MEASUREMENT APPARATUS

The band width up to the AM demodulator should be 2 MHz and above, to avoid excessive fluctuations in the readings displayed.

Readability of level at 0.1dB. An attenuation circuit which has 0.1dB steps before the demodulator is better (USVD from R & S, already a few years old, but affordable). It is also possible to proceed as described in [2].

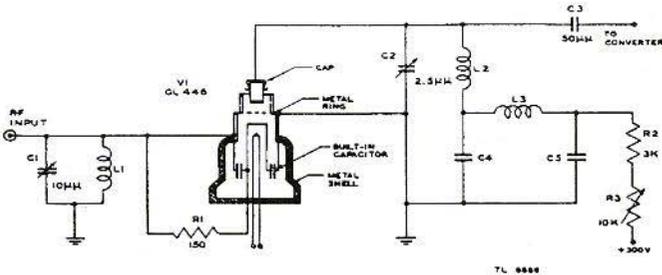


Figure 202. Grounded-grid r-f amplifier.

TL 8884

Fig.3:
Circuit for a Stage
with GL 446

In measurements at 10 GHz and above, a well-loaded horn antenna should be used at the start of a measurement, as home-made antennas with relatively large areas (= relatively high gain) often have thick side lobes and mismatching. This can falsify the measurements.

These measurements naturally do not replace a hot-cold measurement carried out very carefully in the laboratory, or using a high-accuracy noise measurement rig, but they give a good idea of the points in an installation at which improvements could be made - e.g. the pre-amplifier or the antenna.

5.

HISTORICAL EXAMPLE

The UHF pre-amplifier shown on the photo (Fig.2) is one of the first commercially manufactured amplifiers for this purpose. It was used in an American radar appliance in 1944.

Data:

Frequency range: 400 - 435 MHz

Amplification: 26dB (at 432 MHz)

Noise figure F: 8.3dB (hot-cold measurement)

Power consumption: 150V @ 0.03A;

6.3V @ 1.4A

Tubes: 2 x 446 A (forefather of all "lighthouse" tubes)

Very strongly-fixed high-level signals

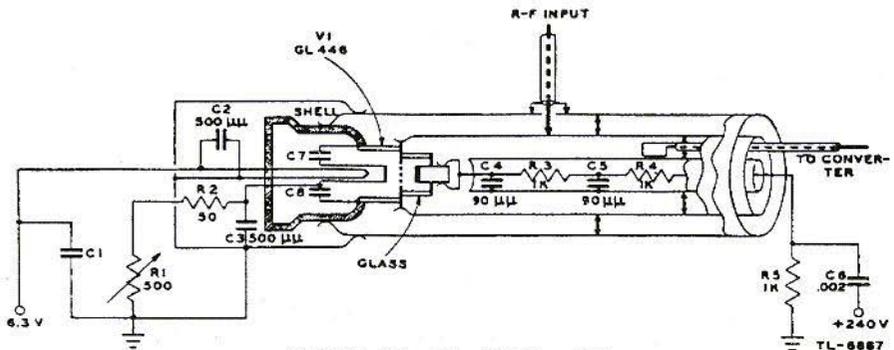


Figure 203. R-f amplifier with lighthouse tube.

Fig.4: Wiring Diagram of HF Amplifier with "Lighthouse" Tube (Voltage Feed +240V and Earth are probably the wrong way round!)



(important in a radar appliance), output power of tubes app. 1W. Selective, owing to four matched oscillation circuits. After this pre-amplifier came a mixing stage, with a 955 wired up as diode (noise number app. 23dB = 200 kT₀). A semi-diode from this period with a better noise number would not have been able to "get above" the primary radar pulse (Fig's.3 and 4).

Overall noise figure:

$$F_{\text{ges}} = F_1 + \frac{F_2 - 1}{V_1}$$

$F_{\text{overall}} = 7.26 \text{ kT}_0 = 8.6\text{dB}$
with $F_1 = 8.3\text{dB} = 6.76 \text{ kT}_0$;
 $F_2 = 200 \text{ kT}_0$ and $V_1 = 400$.

6.

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D.Eckart Schmitzer, DJ4BG

Designing Chebyshev Filters for Practical Operation

Part-2

6. SPECIMEN CALCULATIONS

6.1. Specimen Calculation for a Low-Pass Filter

We can now calculate the components of a filter, based on the clarifications in the previous chapters. As an example, let us design a fifth-grade low-pass filter, with a limiting frequency of 10 MHz and an impedance of 50Ω , which is to have around 0.2dB ripple.

The reference values for the inductance and capacity can be obtained using formula (2) above (similarly converted to the more practical values H and pF):

$$L_b = 7.95775 / f_b = 7.95775 / 10 \text{ (MHz)} \\ = 0.795775 \mu\text{H}$$

$$C_b = 3183.1 / f_b = 3183.1 / 10 \text{ (MHz)} \\ = 318.310 \text{ pF}$$

If we now multiply these reference values by the coefficients from the table

(here, for example, for 0.2dB ripple), we immediately obtain the component values sought. The values used in the example are underlined in Table 1.

The calculation process is shown again below in diagrammatic form:

- Calculate reference values L_b and C_b from impedance and desired limiting frequency.
- Multiply reference values by coefficients from table:

Component 1 = Reference value (here C_b) · coefficient 1

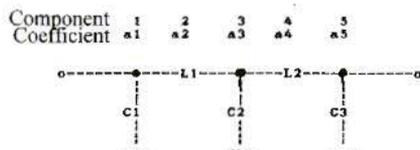
Component 2 = Reference value (here L_b) · coefficient 2, etc.

Coefficients from table for 0.2dB ripple:

$$a_1 = 1.300426, a_2 = 1.345877$$

$$a_3 = 2.127107, a_4 = 1.345877$$

$$a_5 = 1.300426$$



$$C1 = C_b \cdot a1 = 318,310 \text{ pF} \cdot 1,300426 = 413,9386 \approx 414 \text{ pF}$$

$$L1 = L_b \cdot a2 = 0,795775 \text{ } \mu\text{H} \cdot 1,345877 = 1,071015 \approx 1,071 \text{ } \mu\text{H}$$

$$C2 = C_b \cdot a3 = 318,310 \text{ pF} \cdot 2,127107 = 677,0794 \approx 677 \text{ pF}$$

$$L2 = L_b \cdot a4 = 0,795775 \text{ } \mu\text{H} \cdot 1,345877 = \text{wie } L1$$

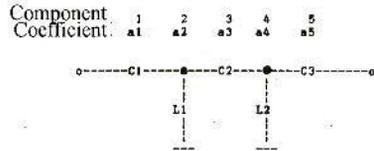
$$C3 = C_b \cdot a5 = 318,310 \text{ pF} \cdot 1,300426 = \text{wie } C1$$

In all cases it is recommended that component values and coefficients should be entered into the calculation as precisely as possible, and that the component values should not be rounded off until the end of the calculation. It will usually not be possible to avoid some calibration work to achieve the right inductance values, while the capacitances must be made as close as possible to the rated value by wiring up standard values in parallel: for example, by combining 390 and 24pF to give 414pF.

Parasitic elements, such as the self-capacitances of the coils and the external inductance of the capacitors, have not been taken into consideration yet! These were discussed in greater detail in [2].

6.2. Specimen Calculation for a High-Pass Filter

A fifth-grade high-pass should be laid out with the same basic data as the low-pass referred to above. So it should have a limiting frequency of 10 MHz and an impedance of 50Ω, together with app. 0.2dB ripple. The reference values determined are the same as those in Section 6.1., which gives us:



$$C1 = C_b \cdot a1 = 318,310 \text{ pF} : 1,300426 = 244,7736 \approx 244,8 \text{ pF}$$

$$L1 = L_b \cdot a2 = 0,795775 \text{ } \mu\text{H} : 1,345877 = 0,591269 \approx 0,591 \text{ } \mu\text{H}$$

$$C2 = C_b \cdot a3 = 318,310 \text{ pF} : 2,127107 = 149,6445 \approx 149,6 \text{ pF}$$

$$L2 = L_b \cdot a4 = 0,795775 \text{ } \mu\text{H} : 1,345877 = \text{wie } L1$$

$$C3 = C_b \cdot a5 = 318,310 \text{ pF} : 1,300426 = \text{wie } C1$$

6.3. Simulation Results for Above Specimens

The rounded-off component values for the two examples were entered into PSPICE as a node list, and displayed over a range of from 1 to 100 MHz.

For the low-pass, Fig.3, curve a gives the result for the attenuation range up to 100dB, in order to show the long-range attenuation. Fig.4, curve a gives the attenuation up to 3dB, to demonstrate the ripple in the transmission band. In Fig.3, both cursors are exactly on the limiting frequency of 10.0 MHz, while in Fig.4 cursor 1 is on the limiting frequency and cursor 2 is at the left-hand edge of the diagram, i.e. on 1 MHz.

In Fig's. 3 and 4, the "b" curves show the results for the high-pass. It can easily be seen from these diagrams that the low-pass and the high-pass produce gradients which are exactly symmetrical



Date/Time run: 10/17/96 14:37:57

Temperature: 27.0

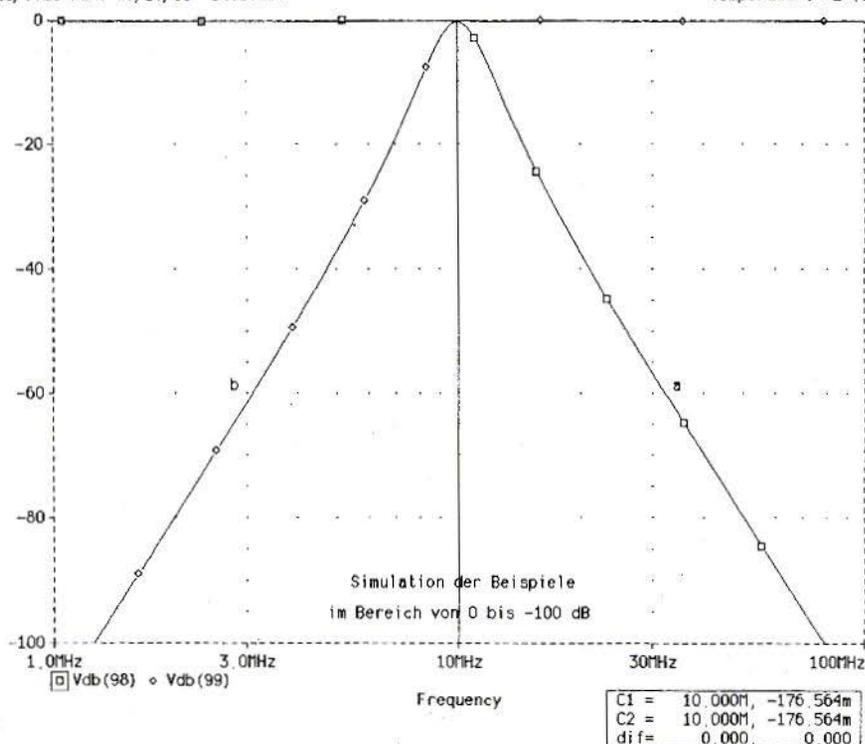


Fig.3: Simulation of Specimens in Range between 0 and -100dB

to the selected limiting frequency. It is also clear that, as stated initially, the ripple values are rounded off. What we are actually seeing here in the display field is a simulation of the basic ripple value of about 0.176dB (cursor C1 = 10,000 M, - 176.564 m., i.e. 176,564 milli-dB)!

It should be pointed out once again here that in Chebyshev filters the limiting frequency is not the 3dB point (!) but the point of the curve at which the attenuation exceeds the maximum desired ripple value. So for a filter with 0.2dB ripple, the limiting frequency is the point after which the attenuation just exceeds the value of 0.2dB!

7.

FILTER TABLES

In Part 1, filter types with varying ripple values were selected. The components for these filters, and those for the specimens in Section 6, were calculated for a limiting frequency of 10 MHz and for 50Ω, and were listed in tables. So the values for these types of filter can be converted by means of simple scaling up or down (see Section 8) for any limiting frequency or impedance required.

Date/Time run: 10/17/96 14:37:57

Temperature: 27.0

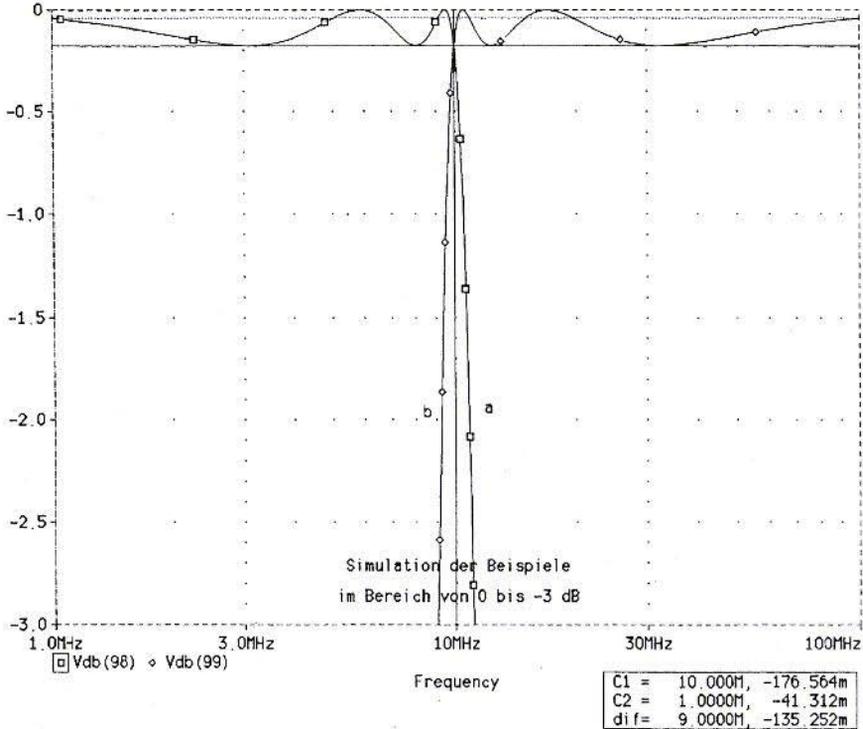


Fig.3: Simulation of Specimens in Range between 0 and -3dB

7.1. Low-Passes

The components are equally suitable for an impedance of 50Ω with a limiting frequency of 10.0 MHz.

The reference values involved have already been calculated in Section 6.1. with:

$$L_b = 0.795775 \mu\text{H} \text{ and } C_b = 318.310 \text{ pF}$$

A delta circuit is used (not so many coils). Thus we obtain the component values in pF and H shown in Table 3.

The numbering differs from the usual sequence in that the component numbers coincide with the coefficients, as can also be seen from the circuit diagrams in

Fig's.2a to 2c in Part 1.

The values calculated were precise, and were not rounded up or down until later to give practical values (with sufficient accuracy).

7.2. High-Passes

The components are also suitable for an impedance of 50Ω with a limiting frequency of 10.0 MHz. The reference values calculated in Section 6.1. also apply here, with:

$$L_b = 0.795775 \mu\text{H} \text{ and } C_b = 318.310 \text{ pF.}$$

A star circuit is used here (not so many coils). Thus we obtain the component



Echodämpfung 10 dB (Welligkeit rund 0,5 dB)

Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	494,3	0,8800	494,3	-	-	-	-
5	529,2	0,9894	794,4	0,9894	529,2	-	-
7	539,2	1,0133	825,7	1,0844	825,7	1,0133	539,2

Echodämpfung 14 dB (Welligkeit rund 0,2 dB)

Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	378,1	0,9185	378,1	-	-	-	-
5	413,9	1,0710	677,1	1,0710	413,9	-	-
7	424,5	1,1053	712,6	1,2064	712,6	1,1053	424,5

Echodämpfung 17 dB (Welligkeit rund 0,1 dB)

Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	316,5	0,9094	316,5	-	-	-	-
5	355,0	1,0934	618,8	1,0934	355,0	-	-
7	366,1	1,1361	658,3	1,2603	658,3	1,1361	366,1

Butterworth-Verlauf (Welligkeit 0 dB)

Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	318,3	1,5915	318,3	-	-	-	-
5	196,7	1,2876	636,6	1,2876	196,7	-	-
7	141,7	1,0082	573,6	1,5916	537,6	1,0082	141,7

Table 3: Component Values for Low-Passes at 10.0 MHz and 50 Ω
Echodämpfung = Reflection loss, *Welligkeit* = Ripple, *Rund* =
 Approximately, *Grad* = Degree, *Bauteilwerte* = Component values,
Und = And, *Butterworth-Verlauf* = Butterworth gradient

Echodämpfung 10 dB (Welligkeit rund 0,5 dB)Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	205,0	0,7196	205,0	-	-	-	-
5	191,5	0,6399	127,5	0,6399	191,5	-	-
7	187,9	0,6250	122,7	0,5841	122,7	0,6250	187,9

Echodämpfung 14 dB (Welligkeit rund 0,2 dB)Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	267,9	0,6894	267,9	-	-	-	-
5	244,8	0,5913	149,6	0,5913	244,8	-	-
7	238,7	0,5730	142,2	0,5249	142,2	0,5730	238,7

Echodämpfung 17 dB (Welligkeit rund 0,1 dB)Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	318,5	0,6964	318,5	-	-	-	-
5	285,4	0,5792	163,7	0,5792	285,4	-	-
7	276,8	0,5574	153,9	0,5025	153,9	0,5574	276,8

Butterworth-Verlauf (Welligkeit 0 dB)Grad: Bauteilwerte in pF und μH :

	C1	L2	C3	L4	C5	L6	C7
3	318,3	0,3979	318,3	-	-	-	-
5	515,0	0,4918	159,2	0,4918	515,0	-	-
7	715,2	0,6281	176,7	0,3979	176,7	0,6281	715,2

Table 4: Component Values for High-Passes at 10.0 MHz and 50 Ω
Echodämpfung = Reflection loss, *Welligkeit* = Ripple, *Rund* =
 Approximately, *Grad* = Degree, *Bauteilwerte* = Component values,
Und = And, *Butterworth-Verlauf* = Butterworth gradient



values in pF and H listed in Table 4. The numbering differs from the usual sequence in that the component numbers coincide with the coefficients, as can also be seen from the circuit diagrams in Figs.2d to 2f in Part-1.

The values calculated were precise, and were not rounded up or down until later to give practical values (with sufficient accuracy).

8. SCALING UP OR DOWN

8.1. Conversion to Another Limiting Frequency

In order to convert the filters above for use with other limiting frequencies, the capacities and inductances merely need to be multiplied by the ratio of the frequencies:

$$C'' = C' \cdot (f_1 / f_2)$$

$$L'' = L' \cdot (f_1 / f_2)$$

where f_1 for our limiting frequency of 10 MHz, used in the tables, is:

$$C'' = C' \cdot (10 \text{ MHz} / f_2)$$

$$L'' = L' \cdot (10 \text{ MHz} / f_2)$$

Example:

The dimensions of a filter are to be changed to give it a limiting frequency of 14.5 MHz. This gives us:

$$C'' = C' \cdot (10 \text{ MHz} / 14.5 \text{ MHz})$$

$$L'' = L' \cdot (10 \text{ MHz} / 14.5 \text{ MHz})$$

i.e.

$$C'' = C' \cdot 0.68966$$

and

$$L'' = L' \cdot 0.68966$$

where C' and L' are the original values for a capacity or inductance at 10 MHz and C'' and L'' are the desired values for the new frequency.

Check: Higher frequencies, smaller capacities and inductances; lower frequencies, bigger capacities and inductances!

8.2. Conversion to Another Impedance

To convert to another impedance, the capacities are multiplied by the ratio of the impedances (Z_1/Z_2), while the inductances are divided by the same ratio, i.e.

$$C'' = C' \cdot (Z_1 / Z_2)$$

$$L'' = L' / (Z_1 / Z_2)$$

Starting from the impedance of 50Ω for Z_1 selected here, this gives:

$$C'' = C' \cdot (50 / Z_2)$$

$$L'' = L' / (50 / Z_2)$$

Example:

A filter is to be converted to an impedance of 75Ω. This gives us:

$$C'' = C' \cdot (50 / 75)$$

$$L'' = L' / (50 / 75)$$

i.e.

$$C'' = C' \cdot 0.6666$$

$$L'' = L' / 0.6666$$

where C' and L' are the original values at 50Ω and C'' and L'' are the desired values at the new impedance.



Check: Higher impedance, smaller capacities and bigger inductances; lower impedance, bigger capacities and smaller inductances!

But it must be pointed out once again that parasitic elements are not taken into account here see [2]. Thus in practise it will be necessary to use slightly lower component values to obtain the precise limiting frequencies desired.

9.

CHECK

The follow-up article by D.Eckart Schmitzer, DJ4BG on this subject will be appearing in a futures issue of VHF Communications:

All the component values shown in the above tables were entered into PSPICE and the curves obtained were checked.

LC Filters with Attenuation Peaks Based on Chebyshev Filters

Editor

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Gerald Dillberg, DL8DAQ

Microprocessor-Controlled FM Receiver for Weather Satellites

Part-1

A modern FM receiver is introduced, with a station-finding function for the 137 MHz range. The central elements of the unit are an IC receiver, a PLL module and the PIC processor.

A remote power supply for masthead pre-amplifiers is also provided, together with a second input for a 1.7 GHz Meteosat converter.

1. INTRODUCTION

The PLL receiver described below is designed to receive signals from weather satellites in polar orbit, e.g. NOAA or Meteor on 137 MHz. Six channels are provided for this purpose, with a station-finding function; these six channels are: 137.3 MHz, 137.4 MHz, 137.5 MHz, 137.62 MHz, 137.8 MHz and 137.85 MHz.

Channels seven (137.5 MHz) and eight (141 MHz) are prepared to receive

signals from geo-stationary weather satellites, e.g. Meteosat on 1.7 GHz. An external converter is required for this, with an LO frequency of 1,553.5 MHz.

If these channels (7 and 8) are selected, the receiver input is switched to the second antenna input through a relay.

DC is applied to both antenna inputs to provide a remote power supply for amplifiers or converters.

Thanks to the IC's used, the entire structure is compact and easy to copy.

2. DESCRIPTION OF CIRCUITRY

Fig.1 shows the circuitry for the high-frequency stage.

Two separate high-frequency inputs are provided, to give a choice of reception, through a 137 MHz antenna or through a Meteosat converter.

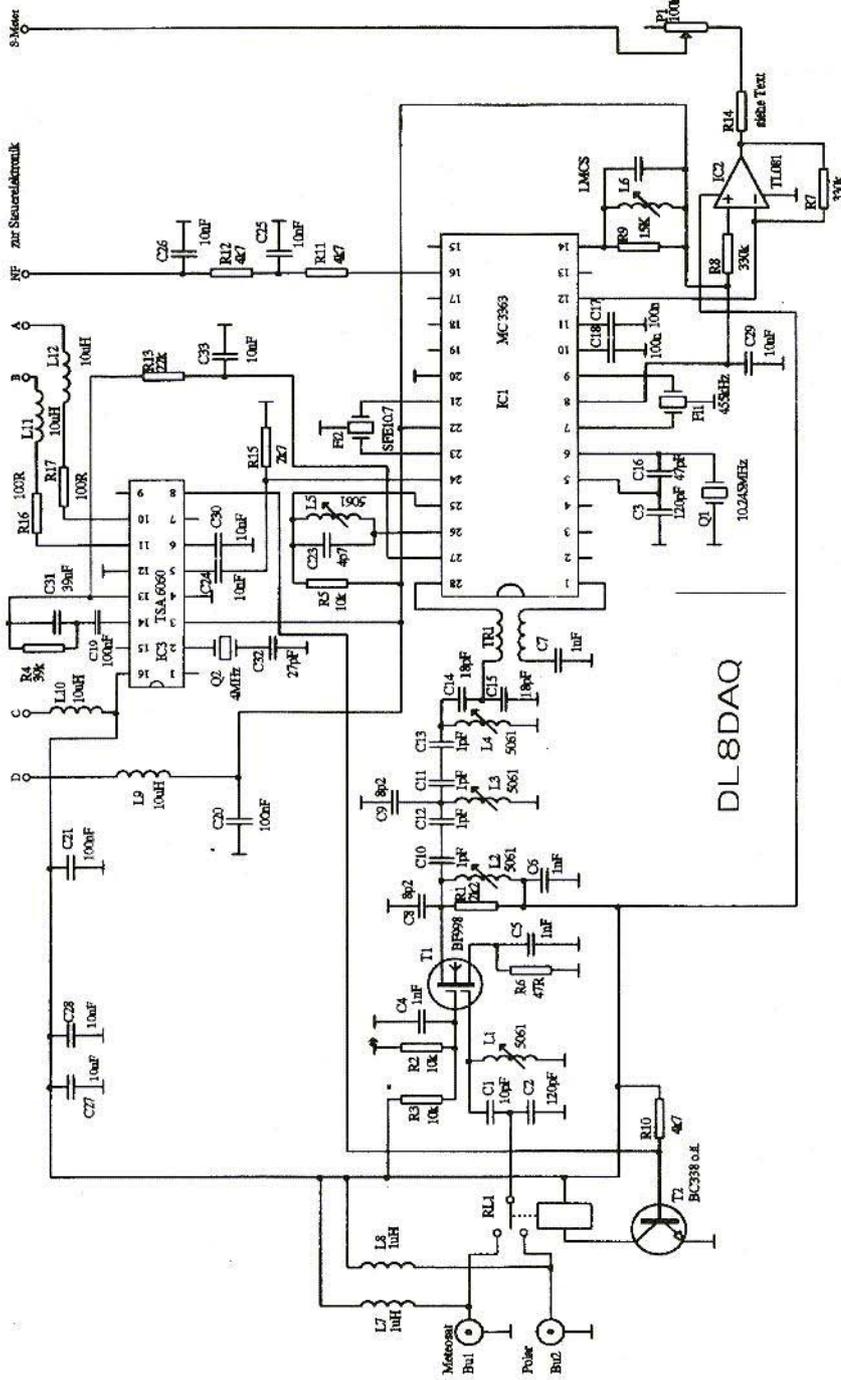


Fig.1: Wiring Diagram of High-Frequency Section of the Weather Satellite Receiver



Fig.2:
Front View of
Prototype showing
small number of
Controls

The input signal is transmitted through a relay onto the capacitive tap of the single-circuit preliminary filter. The high point of this circuit goes directly onto the dual-gate MOSFET BF 998.

At its output is a triple-circuit filter for image frequency suppression.

The signal is then symmetrically uncoupled through a transformer into the MC 3363 receiver IC. This is the core of the receiver, which, with very little external circuitry, forms a complete double super receiver, with a de-modulator and an S-meter output.

The first intermediate frequency of 10.7 MHz is filtered with a ceramic filter, with a band width of app. 150 kHz. The 35 kHz wide 455 kHz ceramic filter is responsible for the adjacent channel sensitivity itself.

This band width is needed to receive signals from weather satellites, since these transmit with a wider dispersion than is otherwise usual in narrow-band FM.

The low-frequency signal is then already available at pin 16 of the receiver IC.

This signal initially passes through a simple low-pass, consisting of an R11,

R12, C25 and C26, and is then split between three paths:

The first path runs through an operational amplifier circuit (IC4C) with an amplification of app. 26dB.

This signal is the desired low-frequency output signal; the loudspeaker signal and the output signal for the subsequent weather fax decoder can be adjusted independently of one another through two trimmers.

The current amplification for the loudspeaker output uses a BD 135 as an emitter follower. This simple circuit for low-frequency output amplification is fully adequate here, since only a monitoring function is involved, and the loudspeaker usually remains silent during image reception.

The second path (IC5C) is used to evaluate the DC voltage fraction of the de-modulator for the AFC. This is particularly necessary for the reception of polar satellites, in order to balance out the Doppler effect. First, the low-frequency fractions are removed from the output voltage with a passive low-pass filter (R29 / C48) and the resultant DC voltage is amplified to approximately 6 Volts. In two subsequent

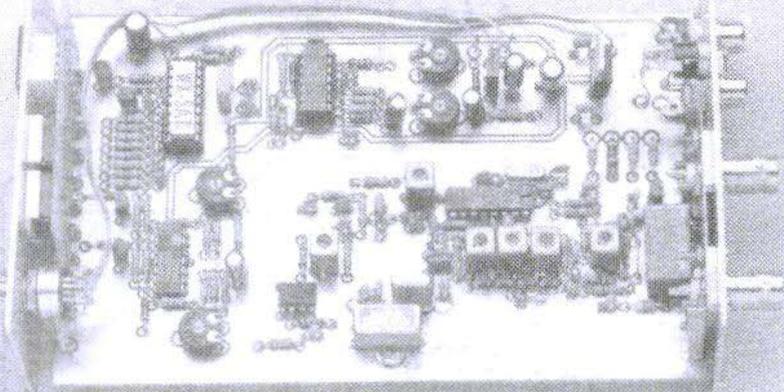


Fig.4: View of Components Side with wired-up Components

comparator circuits, it is compared with the rated voltage, which is set using P3.

The rated voltage corresponds to the DC voltage of the de-modulator, with tuning at the frequency mid-point.

One comparator (IC5B) generates a "1" for too high and a "0" for too low a frequency at the processor port.

The other comparator has a feedback system which brings about a hysteresis. This ensures that the reception frequency is not continuously oscillating around the rated value, as in a two-point regulator. The feedback system is adjusted in such a way that the hysteresis corresponds to app. 1 kHz of the maximum possible tuning error.

Fig.3 shows the control electronics for the complete receiver.

The frequency correction itself is calculated by the processor and is transmitted to the PLL through the I²C bus.

Finally, the third path (IC4D) has an active band-pass filter for 2,400 Hz with a downstream comparator, which goes to the timer input of the processor through a voltage divider, R23 / R 24.

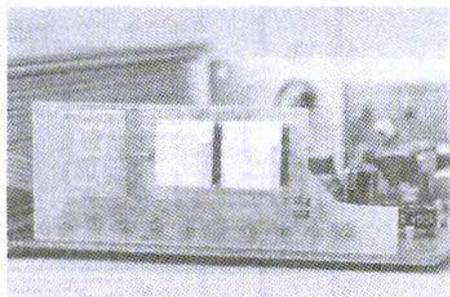


Fig.5: Operational Controls on Display Board

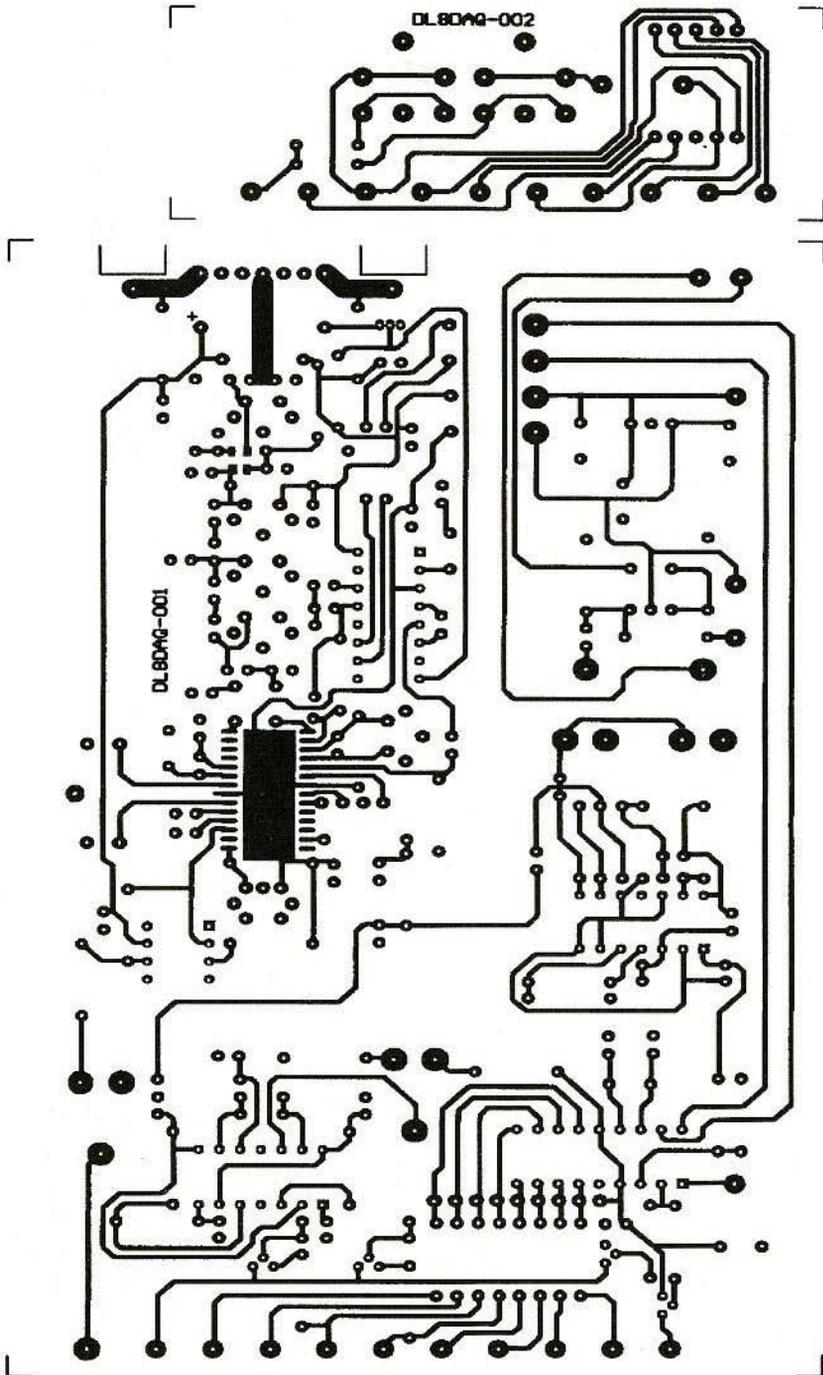


Fig.6: Component Side of the WX Receiver & Display PCB

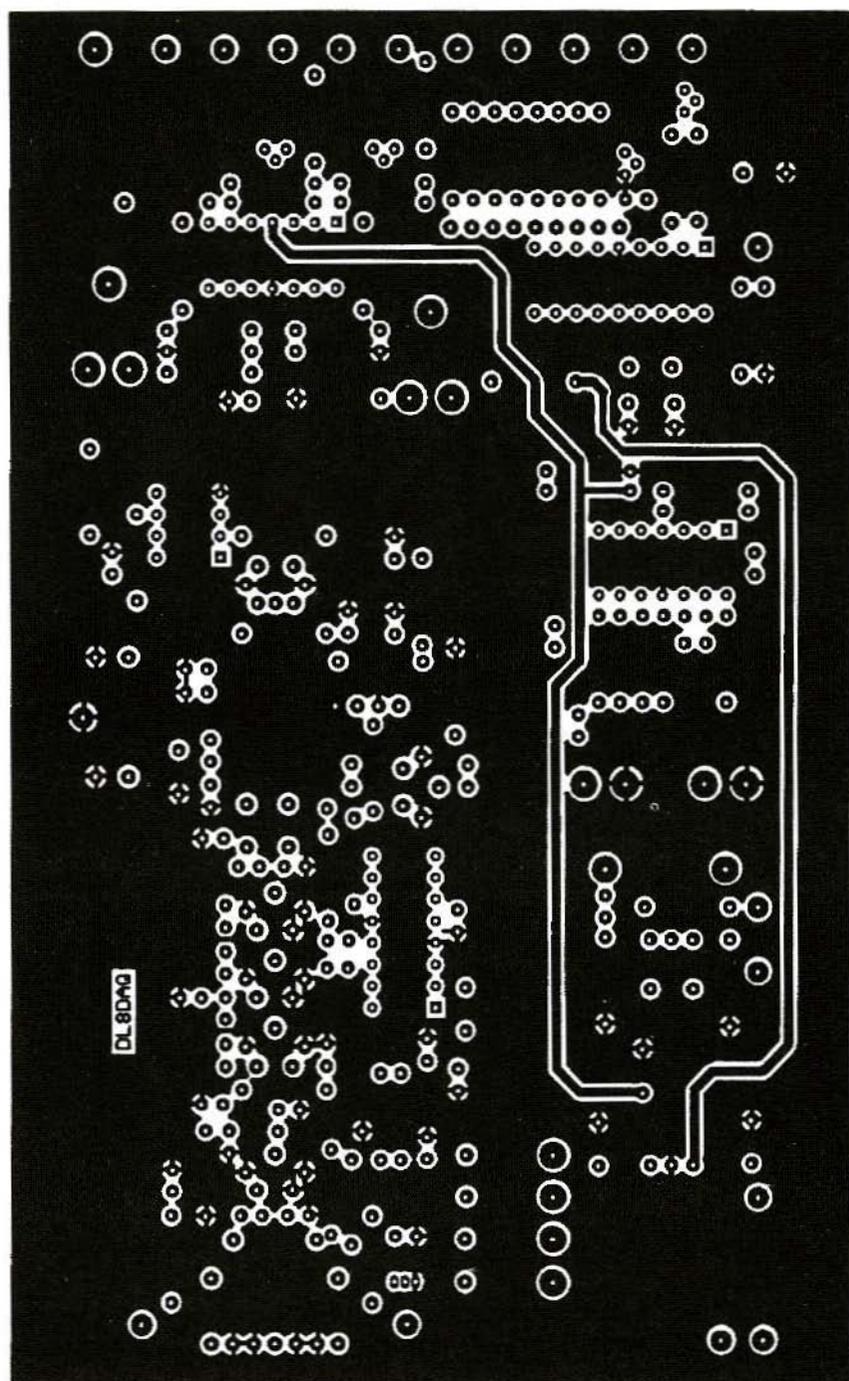


Fig.6: Earth Side of the WX Receiver & Display PCB with Two Tracks

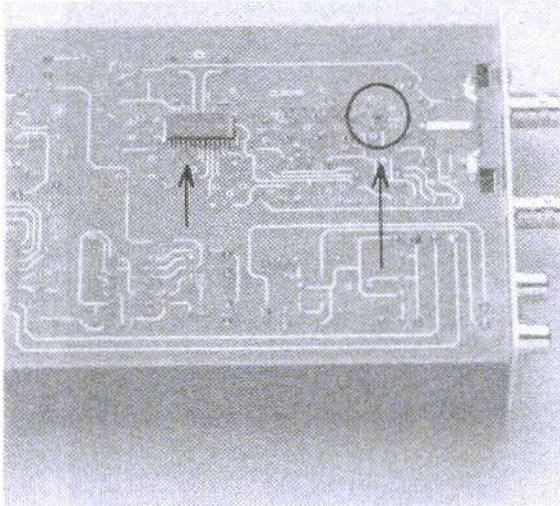


Fig.8:
Positioning of two
SMD Components
on Underside

Here the APT tone of the satellite signal is evaluated. When a satellite appears, the station finding function is terminated.

This method is more reliable than conventional noise inhibition, because it does not respond to noise impulses. Moreover, this method does not require equalization, because an appropriate active filter band width has been selected.

The internal LO in the MC 3363 is used to mix the reception frequency down to the first intermediate frequency. This LO oscillates at approximately 127 MHz, and is locked into the required frequency with the PLL-IC TSA 6060. The PLL module also required very little external wiring.

The LO signal is fed to pin 5 through a C coupler. The tuning voltage is available at pin 13. The loop filters are formed by C19, C31 and R4. Any remains of the reference frequency are filtered out by the subsequent RC filter R 13 / C 33.

In the PLL module too, the reference frequency is generated in combination with an external 4.0-MHz crystal.

To adjust the frequency, the divider ratio of the internal frequency divider of the processor is adjusted accordingly through the I²C bus.

The TSA 6060 also has a band switch output (pin 8), which is used here to switch over the antenna relay. The switchover data also come from the processor through the I²C bus.

The receiver has an S-meter output, with a display dynamic of app. 60dB. Depending on the signal strength, the output voltage lies between 6.5 and 9 Volts. If an appropriate instrument has to be connected up then, should the situation arise, the R14 resistance (high-frequency stage) should be adapted. No S-meter was included in the specimen unit for reasons of space. However, it can be useful, especially for adjusting the Meteosat antenna.

To be continued



COMPLETE KITS & PCB's

KIT	DESCRIPTION	ISSUE	No.	PRICE
DG4RBF-003	HF Synthesiser 5 to 1450 MHz, TCXO	2&3/98	-----	£135.00
DG4RBF-004	IIF Synthesiser 5 to 1450 MHz, Processor	2&3/98	-----	£125.00
DG4RBF-005	HF Synthesiser 5 to 1450 MHz, Bus Driver	2&3/98	-----	£ 12.00
DG4RBF-006	HF Synthesiser 5 to 1450 MHz, Regulator for VCO	2&3/98	-----	£ 12.00
HF-WOBB	Software for PC: HF-WOBB	2&3/98	-----	£ 55.00
KOMBI	EPROM HF-SYNTH + Software HF-WOBB	2&3/98	-----	£ 85.00
DC8UG-007	5W PA for 13cm	3/94	06938	£286.00
DJ8ES-019	Transverter 144/28 MHz	4/93	06385	£143.00
DJ8ES-019mod	Transverter 50/28 MHz	2/95	06392	£143.00
DJ8ES-020	Hybrid Amplifier for 144 MHz	1/94	06387	£179.00
DJ8ES-021	13cm FM ATV Exciter	2/94	06388	£ 67.00
DJ8ES-022	28/432 MHz Transverter Oscillator	2/95	06395	£ 59.00
DJ8ES-023	28/432 MHz Transverter Converter	2/95	06396	£114.00
DJ8ES-031mod	Mixer / LO	2/98	-----	£ 52.50
HF-MESS	PC Software for the DJ8ES projects	TBA	03999	£ 32.50
OE2TLZ 001	PreMixer for 23 and 13cm	1/97	-----	£115.00
YT3MV	Low Noise L-Band Aerial Amplifier	2/92	06358	£ 29.80
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DJ8ES-023	28/432 MHz Transverter Converter	2/95	-----	£ 16.50
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DJ8ES-030	VCO with PLL	3/97	-----	£ 15.00
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DJ8ES-032	Pin Diode Controller	3/97	-----	£ 15.00
DJ8ES-033	A/D Converter with Detector	4/97	-----	£ 15.00
DJ8ES-034	Direct Mixer with Logarithmic Display	4/97	-----	£ 15.00
HF-MESS	Software for the DJ8ES PC projects	TBA	-----	£ 32.50
OE2TZL-001	PreMixer for 23 and 13cm	1/97	-----	£ 15.00

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THE G8OZP RANGE OF MICROWAVE KITS FOR AMATEUR TELEVISION

ASSEMBLED GUNNMOD2 3cm ATV TRANSMITTER PCB & KIT

Gunnmod2 PCB forms the heart of a compact, fully featured FM. Gunn diode PAL, SECAM, or NTSC Amateur Television Transmitter with 5.5, 6.0 or 6.5 MHz intercarrier sound. The PCB is suitable for most types of 3cm and other frequency Gunn oscillators up to 25mW (up to 300mW Gunn oscillators may be used with the addition of one extra transistor). The single 3 by 2 inch quality PCB incorporates a stable, adjustable Gunn diode modulator with video pre-emphasis and adjustable deviation, together with a stable intercarrier sound generator with selectable pre-emphasis, audio limiting, input level control and output filtering. The PCB requires a single 10 - 15V DC unregulated supply and consumes only 20mA, excluding the Gunn diode. This makes it ideal for portable or mast head use.

PCB with full instructions - you supply components, Gunn Osc, case, connectors, etc..... £ 7.50
Kit with full instructions - you supply Gunn Osc, case, connectors, etc (Note 1)..... £ 25.00
Kit with Gunn oscillator - you supply case, connectors, etc. (Notes 1 & 2) £ 35.00

Note 1: Please state 5.5 MHz, 6.0 MHz or 6.5 MHz audio sub carrier frequency. (6 MHz standard) *Note 2:* At present Gunn oscillators are good tested surplus units. Gunn oscillators are WG16 square flange mounting, 8 - 12mW output and pre-tuned to 10.340 GHz, other frequencies at request. 8 - 12mW oscillators, when equipped with a suitable antenna and a low noise RX are capable of providing line-of-site transmissions well in excess of 100Km. With minor modifications this TX may be used to transmit data.

3cm LNB's

Low Noise Block receive converters have been the main contributive factor that has opened up the 3cm band to ATV and other modes. With the low receive noise figures now available, transmissions are no longer restricted to clear line-of-sight paths. Test transmissions have conclusively shown that over the horizon transmissions by various propagation modes are now possible. None line-of-sight transmission by means of scatter from rain clouds has also been achieved. Equipped with suitable antenna systems, etc., operation well in excess of 100km is readily achievable. Integral feed horn types will fit directly on to standard offset satellite dishes with a 38 - 40mm mount.

Brand new with integral feed horn 0.7dB noise figure £ 50.00
Brand new with integral feed horn 1.0 - 1.2dB noise figure..... £ 45.00
Brand new with integral 20dB antenna 0.8db noise fig. £ 60.00
Brand new 22mm circular input 0.7db noise figure £ 55.00
Brand new 22mm circular input 1.0db - 1.2db noise figure £ 50.00

All LNB's. are fully tested with the local oscillator at 9.0GHz $\pm 0.01\%$ (this is often better than original). Other local oscillator frequencies available by request. Horizontal or vertical polarisation is selected by applying either approximately 13 or 1BV DC.



ASSEMBLED DOVE ATV TUNEABLE IF / RECEIVER KIT

The Dove ATV Tuneable IF/RX forms the ideal heart of an ATV receiving system for either 24cm (see note) or the higher bands when used in conjunction with a suitable Low-Noise-Block Converter. It is designed as a compact easy to construct unit that requires no adjustment, setting up, or alignment. The kit contains all board mounted components and full instructions. The only other requirements are a case, power and output connectors, four potentiometer controls (all 10k lin) and a switch (SPST). The unit has been designed on a flexible modular concept. Plug-in modules are being developed which include: a fully tuneable sound demodulator, plus single and multichannel synthesisers. Unprocessed DC-coupled baseband output is provided for expansion into data reception or multiple subcarrier demodulation. An AGC output is provided for signal strength, etc. A divided by 256 local oscillator signal is also available.

Size:	105mm wide x 97mm deep x 50mm high
Power:	11 - 14v DC @ 400mA
Frequency Range:	925 - 1800 MHz
Sensitivity:	-57dBm (300µV)
Audio Subcarrier:	6 MHz
Video Bandwidth:	25 Hz to 3-5.5 MHz Variable
Video Outputs:	1V P-P Composite Unprocessed Baseband
Audio Output:	1.5 W into 4Ω minimum
Controls:	Main Tuning, Volume, Video Gain, Video Bandwidth, LNB Power ON/OFF
LNB / Preamp Powe:	12V DC Switched
Connections:	RF Input F-type, all others 0.1" pitch (supplied)
Mounting:	Fixing pilot hole in each corner

NOTE: When used as a 24 cm Receiver unless the signal is a strong local one a preamp will be required.

DOVE RECEIVER KIT £ 57.00

24cm HEMT GaAsFET ATV PREAMP

Satellite receivers are basically tuneable IF units which require, for satisfactory operation, input signal levels of many millivolts. In a satellite installation most of the RF gain is provided by the dish mounted LNB. This means that if used on the 24cm band, unless they are used only for local working, they will require a high-gain pre-amp at the front end to compensate for the gain normally provided by the LNB. This 24cm low-noise, high-gain pre-amp is designed specifically for ATV use, but may be used on any modes if required. Micro striplines plus Helical band pass filter make alignment very simple and greatly reduces the possibility of interference from out of band signals.

1dB noise figure HEMT GaAsFET front end, 40db gain.

I2 - 20V DC supply via coax. (from satellite RX etc.) or separate feed if required. Sturdy tin plate enclosure. High quality BNC connectors. Very simple to align.

Note: This kit contains some surface mount components. TX/RX switching is not included

Complete Kit £ 77.50
Fully assembled, aligned & tested £ 90.00

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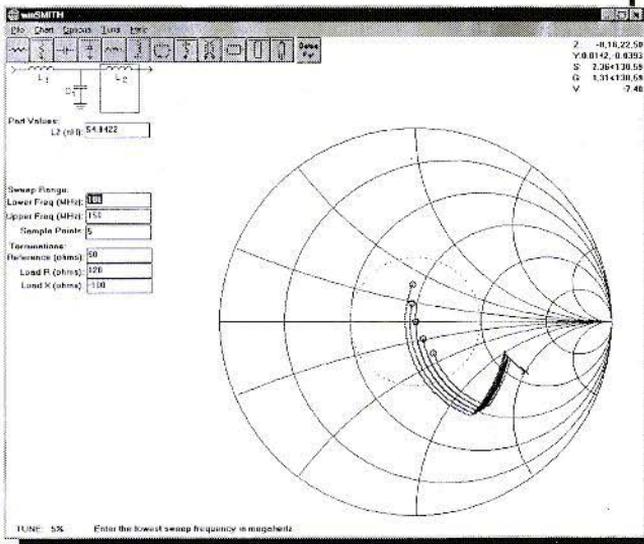
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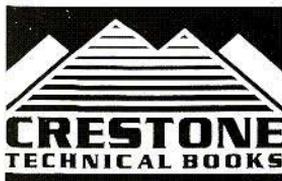


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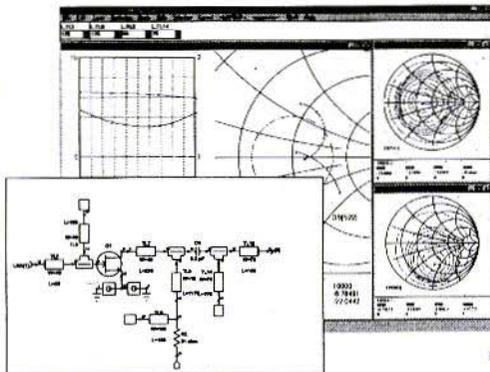
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