

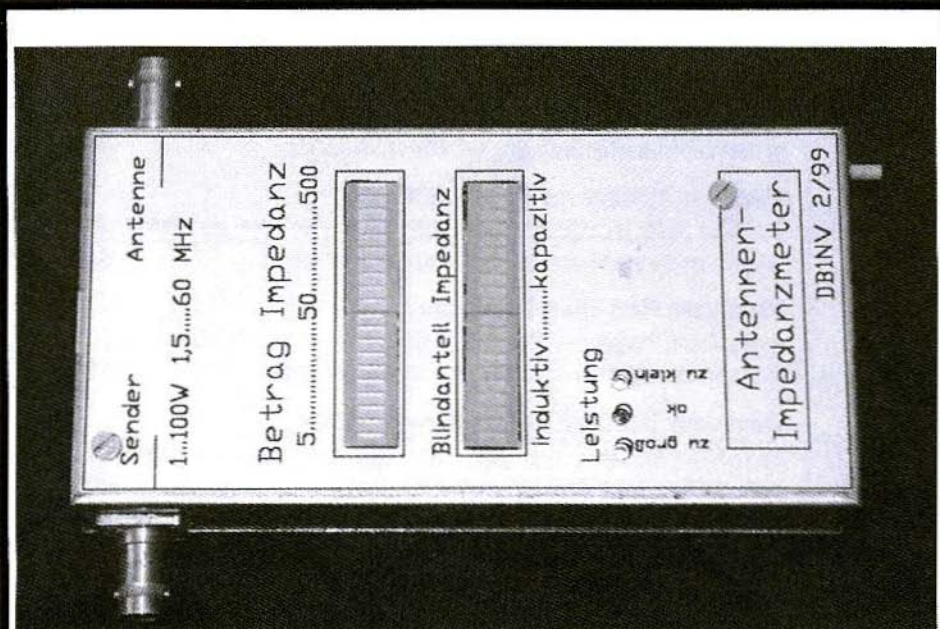


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## Aerial Impedance Meter

*Dr. - Ing. Jochen Jirmann*

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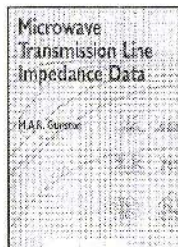
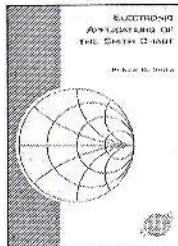
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*This is my second issue as editor. The articles this time are quite technical but there is a lot of useful and interesting information.*

*I had planned to put more articles in this issue but there was not enough space to squeeze another article in this time. I will be looking at different DTP packages over the summer to see if it is possible to increase the contents without decreasing the quality.*

*Thank you to the readers who have sent comment on my first issue (mostly good !), I am always pleased receive these, it helps me to know what you want. See page 125 for an update on plans for the 2000. - Andy*



**KM Publications**, 63 Ringwood Road, Luton, LU2 7BG, UK

Telephone: 01582 581051; INT: +44 1582 581051; FAX: +441582 581051

Email: [andy@vhfcomm.co.uk](mailto:andy@vhfcomm.co.uk)

WWW: <http://www.vhfcomm.co.uk>



*Dr.-Ing. Jochen Jirmann, DB1 NV*

# A VECTORIAL AERIAL IMPEDANCE METER FOR THE SHORT-WAVE RANGE

Aerial impedance measuring bridges have been described for years in relevant aerial literature such as the Rothammel [2], and yet most radio amateurs know nothing about them. In contrast to the standing wave meter, with a bridge information can be obtained concerning the true value of the aerials resistive and reactive components.

Anyone who has ever worked with an aerial measuring bridge will come to think of the standing wave meter method as something out of the Stone Age. So why is it that the bridge method is so unpopular, in spite of its advantages? Probably because a two-stage low-power balancing procedure has to be carried out for each measurement procedure and the bridge can not remain permanently in the feeder.

Both these disadvantages can be eliminated with modern electronics:

The aerial impedance meter described below represents a (relatively) simple way of creating a vectorial network

analyser, and is intended as an aid for radio amateurs who do a lot of experimenting with short-wave aerials or who are active on field days. It makes it possible to measure the impedance of an aerial in terms of magnitude and phase, and to do this it is looped into the feeder like a standing wave meter. Using this information, balancing aerials is made considerably simpler.

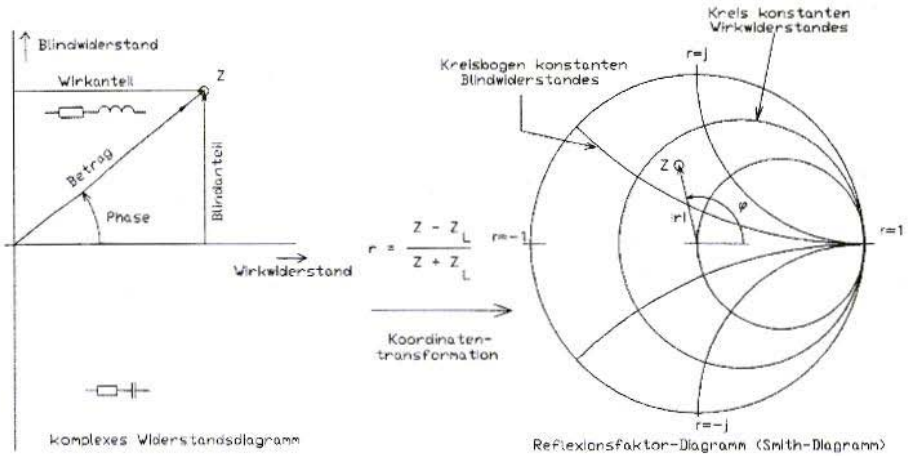
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## 1. PRINCIPLES OF ALTER- NATING CURRENT MEASUREMENT METHODS

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This section is intended for those readers who have no electrical engineering training. Basic terms such as impedance, real resistance and reactive impedance, level and phase are explained here. A more comprehensive picture can be obtained, for example, from Chapter 32 of [2].

Direct current technology involves only



**Fig. 1: Representation of Complex Resistances in Resistance Diagram**  
 (Blindwiderstand = Reactive impedance, Wirkanteil = Active component, Betrag = Amplitude, Blindanteil = Reactive component, Wirkwiderstand = Effective resistance, Komplexes Widerstandsdiagramm = Complex resistance diagram, Koordinatentransformation = Co-ordination transformation, Kreisbogen konstanten Blindwiderstandes = Constant reactive impedance arcs, Kreis konstanten Wirkwiderstandes = Constant effective resistance circle, Reflexionsfaktor-Diagramm (Smith-Diagramm))

effective resistances which use the power supplied and convert it into other forms of energy. Alternating current technology also involves components which store energy in the form of electrical fields (capacitors) or magnetic fields (coils). A capacitor also picks up current from an alternating current source almost without heating itself up. If the voltage and current at the capacitor are measured, e.g. with a multimeter, it will nevertheless have a resistance in accordance with  $R = V/I$ . What we actually did here was to carry out a scalar (relating to level) impedance measurement, which does not allow for

any differentiation between a resistive load and a reactive impedance (capacitor, coil). Transferred to aerial technology, this means that we do not know whether energy is being radiated in the form of an electro-magnetic wave or merely generating reactive power. This fine distinction is not noticed until the phase position between the voltage and the current is evaluated, i.e. a vectorial impedance measurement is carried out. With a pure effective resistance, the voltage and current are in phase. For a pure reactive impedance, they are phase-displaced through  $90^\circ$ . With a coil, the voltage leads  $90^\circ$  in front of the current,



and with a capacitor it is  $90^\circ$  behind it. In practice, aerials have a complex impedance, which has an effective fraction and also a certain reactive impedance.

To represent complex resistances, we use the resistance diagram (Buschbeck diagram), in which the effective fraction is plotted towards the right, positive reactive impedances (inductances) are plotted upwards, and negative reactive impedances (capacitances) are plotted downwards, as shown in Fig. 1. Resistances can be represented in the upper half-plane as a series circuit of a resistance with an inductance, and in the lower half-plane as a series circuit made up of a resistance and a capacitor. A complex resistance (impedance) can be described either in terms of an active component and a reactive impedance or in terms of the level (the vector length from the origin to the impedance point) and the phase angle between the level indicator and the X axis.

The representation as a Smith diagram (reflection factor diagram) which is obtained through a transformation (conformal representation) in accordance with the equation

$$r = (Z - Z_L)/(Z + Z_L)$$

should be more familiar to VHF-UHF amateurs. By this means, the impedance diagram is immediately standardised to the system impedance level (e.g. 50 Ohms or 75 Ohms). The vertical lines of the constant effective resistance are formed into closed circles, while the horizontal lines of the constant reactive impedance form arcs. The tuning point lies in the centre of the diagram and the reflection factor (in accordance with the

level and phase) is the indicator from the diagram centre to the location of the impedance,  $Z$ . The voltage standing wave ratio, again, is derived from the reflection factor through the equation

$$SWR = s = (1 + |r|)/(1 - |r|)$$

If we are investigating an aerial with various measuring instruments, an SWR meter sees only the impedance variation from the 50-Ohms point. Thus we know only that the impedance lies on a circle around the phasing point.

A tuning measuring bridge with a variable multiplier (e.g. the Matchmaker in [2]) continuously delivers the active component of the aerial impedance correctly. A reactive impedance is recognised indirectly from the fact that the balancing of the measuring bridge supplies no sharp zero but only a fuzzy minimum. Only a correct alternating current measuring bridge and measurement of the reactive impedance supplies a complete impedance measurement and thus makes it possible to carry out purposeful tuning measures.

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## 2. PROCESSES FOR MEASURING AERIAL IMPEDANCE

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For most radio amateurs, the concept of aerial impedance measurement brings to mind only the phrase standing wave meter. Balancing an aerial is thus somewhat like an attempt to repair a modern transceiver using a Dupol or an LED voltage tester.

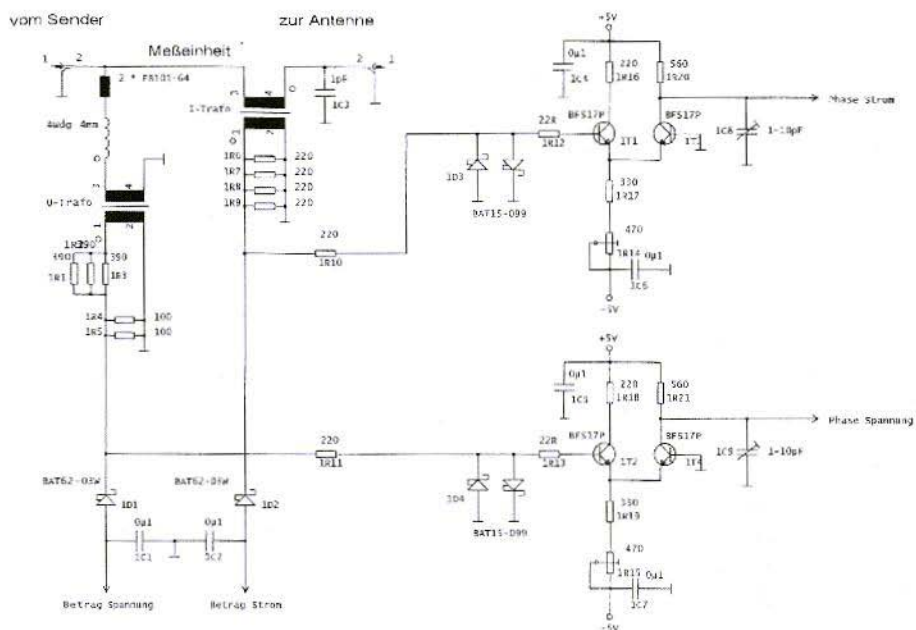


Fig. 2: Measurement Unit Wiring Diagram

Vom Sender = From transmitter, Meßeinheit = Measurement unit, Zur Antenne = To Aerial, Zweig = Branch, U-Trafo = Transmission transformer, Betrag Spannung = Voltage level, Betrag Strom = Current level, Phase Strom = Current phase, Phase Spannung = Voltage phase

Even an ideal standing wave meter merely supplies information concerning how far away from the tuning point we are. It does not say whether the Aerial impedance is too low or too high, or whether an inductive or capacitive reactive impedance is causing tuning to deteriorate.

In reality, things look even grimmer, for anyone who has not been able to obtain a Bird Wattmeter at a good price, or

who still has a Drake W4 or a WV4, has to rely on the questionable accuracy of so-called amateur SWR meters. The sad truth is that accuracy of measurement is not merely a matter of price. Investigations by DH1 NAW have shown that many cheap SWR meters measure more accurately than expensive equipment. Moreover, nobody knows how long a model will actually continue to be produced without any changes. Often a



different core material for the current transformer or a new type of diode is enough to bring about alterations.

For years, the best solution has lain in looking in the Rothammel [2], where various aerial measuring bridges are described. The much-hyped noise bridges, in the authors opinion, are at present pointless if one has a transceiver capable of continuous transmission, which can be reset to a low power level of less than 10 W. With a fixed 10-dB attenuator, one has an ideal test transmitter for aerial trials. Receiver / test transmitters have too low an output for this purpose. Since wireless transmitters already generate total power levels exceeding 1 mW on a large short-wave aerial, a test transmitter power of 100 mW is the minimum. The obtainable, and much-used, 3200B power oscillator from Hewlett-Packard from the Valve Age is extremely suitable.

More remarkably, the advantages of aerial measuring bridges are scarcely known about, or they are reputed to be very laborious to operate. It is therefore time to consider a modern procedure which automatically generates the information content of a manual impedance bridge.

The specification was defined as follows:

Measurement in a coaxial 50-Ohms system

Frequency range 1.8 to 30 MHz or, better, to 60 MHz (6-m. band)

– Displaying of aerial impedance level and phase

- Operation with standard transceiver power levels of max. 100 W
- More reliable displaying from 0.5 to 1 W for QRP transmitters
- Battery operation
- Format comparable with normal SWR meter
- Operation as automatic as possible

A difficult problem lay in finding the optimal forms of display, for here the desirable representations differ a great deal. The ideas range from simple indicator instruments through luminous spot displays and digital displays, all the way to small LCD screens with a Smith diagram representation. For this reason, the reading generation and the display have been separated, so that we can all find our own solution, up to the interface or the notebook.

The prototype was equipped with 20-stage luminous spot displays. The accuracy of reading is adequate, and the solution is cheaper, more compact and more robust than using two analogue instruments! To make it simpler to set the 50-Ohm point, the level scale in this range is electronically expanded.

Incidentally, impedance measuring is very simple with the current/voltage method. A current transformer measures the aerial current, a voltage transformer measures the aerial voltage. The quotient of the levels is the aerial impedance level and the phase position between the voltage and the current supplies the impedance phase angle.

All the user has to do during the aerial





balancing is to bring the two luminous spots for the level and the phase to the centre of the scale! With a little practice, one can recognise the difference between the resonance point of an aerial (the real input resistance, but not necessarily at 50 Ohms) and the result of conventional balancing, in which the input resistance is complex, but by chance lies closer to the 50-Ohm point. Moreover, it can be seen at once whether a dipole is too short (capacitive reactive impedance) or too long (inductive reactive impedance).

---

### 3.

## THE EQUIPMENT CONCEPT OF A VECTORIAL AERIAL IMPEDANCE METER

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In the earlier sections, it was assumed that we were talking about an independent solution with a built-in high-frequency source. This will not be taken any further, for two reasons.

Firstly, a high degree of accuracy in targeting frequencies is necessary to set resonances reliably for narrow-band aeriels, one which requires at least a built-in counter, or better a synthesiser source.

Secondly, for a reliable measurement, uninfluenced by strong wireless transmitters, a high-frequency output of 100 mW or more is required, which unnecessarily increases the current consumption.

Since many transceivers with continuous transmission ranges have recently become available, with power levels which

can be regulated down to under 1 W, there is no longer any reason to have a built-in high-frequency source.

So the measurement circuit was conceived for high-frequency outputs of between 1 and 100 Watts. A level monitoring system included which signals if this power is outside the acceptable range. Moreover, it is even possible to leave the impedance meter in the aerial line, since the maximum power take-off is 5% of the throughline power.

A further point worth considering is the power supply. As is well known, accumulators are always empty when you need them, and external power supplies are difficult on a field day. For this reason, a normal 9-V battery (alkali/manganese) is used as the power source which, in spite of the circuits very high current consumption, allows for a good hours operating time.

A control circuit activates the power supply only if high-frequency energy is being applied or if a key is pressed. So operation is bound to be intermittent, which means the battery could last out for an entire portable season. The internal current supply circuit is designed in such a way that it will operate with any DC voltage from 3.5 to 20 Volts, so that nothing stands in the way of operating with an external supply.

As already mentioned, two 20-stage luminous spot displays are used in the prototype. If so desired, they can be replaced by two indicator instruments with the zero point in the middle. A solution using a PIC or 80C51 microcontroller was also taken into consideration for the all-digital people.

## 4. REALISATION

The aerial impedance meter consists of the following functional blocks:

Reading measurement with voltage and current transformer

Level measurement with quotient output and display scaling

Phase measurement

Luminous spot displays with driver circuit

Power supply

### 4.1. Reading measurement

The readings for the voltage and the current in the feeder are taken by two conversion transformers. It is simplest if the current transformer takes the form of a superstructure. The internal conductor of the coax cable leads through a ferrite ring core (FT 50-43 in the prototype) and forms a primary winding for the transformer. The secondary coil has 7 turns made from 0.5 mm. enamelled copper wire (uncritical) and feeds a 50-Ohm resistance which, for reasons of high-frequency capability and current-carrying capacity, consists of four parallel-wired 0.6 W-metal layer resistors, type 0207.

In Fig 7 - a photograph of the specimen apparatus, the transformer can be seen made from  $\frac{1}{4}$  semi-rigid cable with a ring core transformer sunk in a slot in

the printed circuit board, it has a compensation capacitor of 1 pF at the aerial output. A band width of exactly 1 MHz to 1 GHz was obtained. The inherent reflection remains below 20 dB, which should suffice.

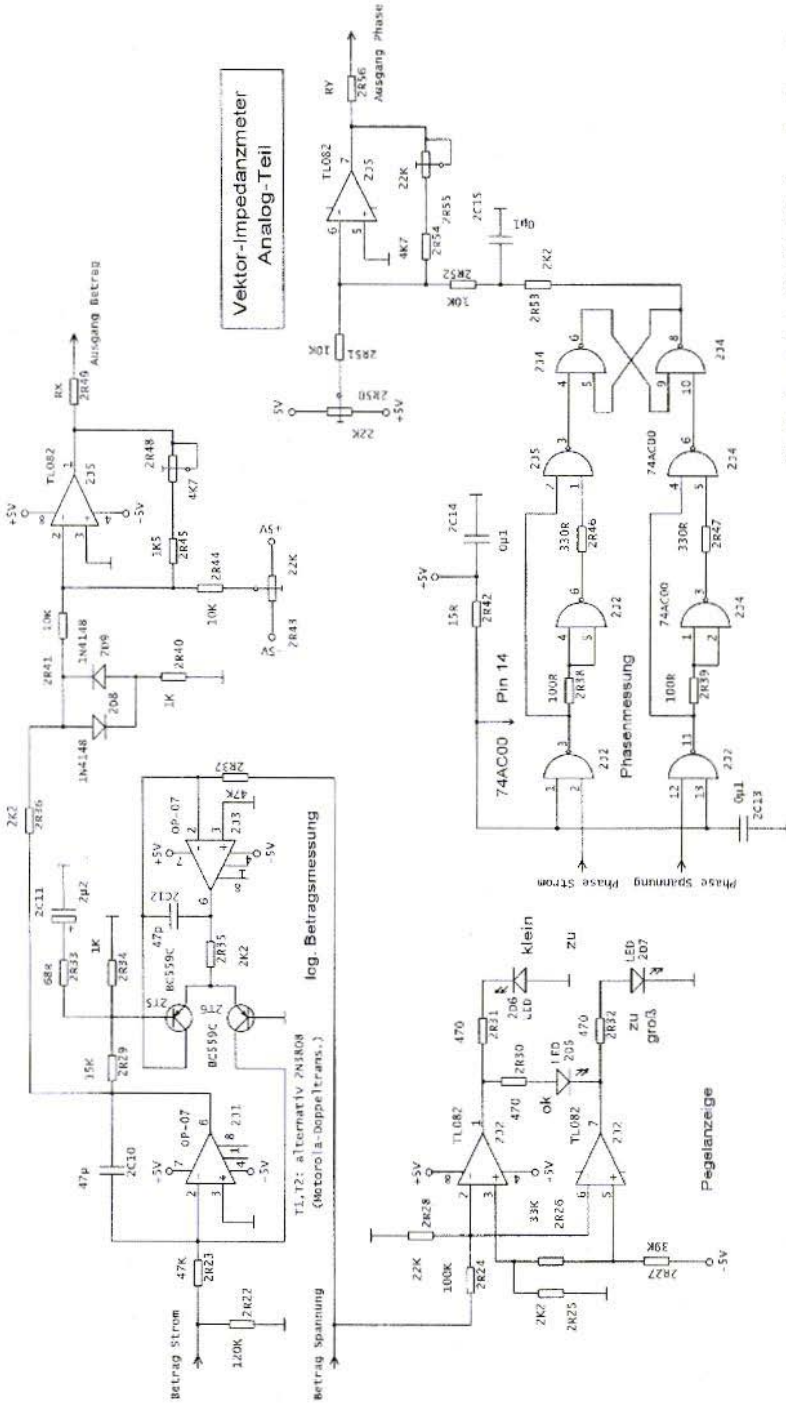
The voltage transformer was initially wound as a simple 7:1 transformer but, in spite of the lower frequency, that was a failure. Depending on the type of winding, either the stray capacitance or the stray inductance was too high. One way out was a 3:1 trifilar transformer (3 x 13 turns of Teflon-insulated wire-wrap wire on FT 50-43) followed by a voltage divider made from metal layer resistors (due to the current-carrying capacity). With two ferrite beads and a small longitudinal coil (they attenuate control resonances in the VHF range) at the internal conductor tap, a frequency response error was attained of better than  $\pm 0.15$  dB from 1 to 100 MHz. The amplitude error for the two measuring channels remains less than 10% in the range from 1 to 60 MHz!

How did the winding ratio of 7:1 for the measuring transformer come about? With 100 Watts and tuning, a high-frequency voltage of 70.7 V is obtained on the line. After transformation, that is 10 V in the voltage path. With standard high-frequency diodes with a blocking voltage of 40 V, this voltage can be reliably rectified. If the high-frequency power is reduced to 1 W, high-frequency voltages of 1 V are obtained, and we are still in the linear range of the rectifier characteristic.

Even if most OMs think only of germanium diodes for the rectification



Log. Betragmessung = Logarithmic level measurement, Ausgang Betrag = Level output, Ausgang Phase = Phase output  
 Betrag Strom = Current level, Betrag Spannung = Voltage level, Doppeltrans. = Double transistor



Vektor-Impedanzmeter Analog-Teil = Vectorial impedance meter analog section  
 Phase Spannung = Voltage phase, Phasenmessung = Phase measurement  
 Pegelanzeige = Level display, Zu groß = Too high, Zu klein = Too low, Phase Strom = Current phase

Fig. 3: Analogue Section of Vectorial Impedance Meter

of the higher H-F voltages, there are more modern solutions. Siemens (now Infineon) have developed the BAT 62 Schottky diode as a replacement for germanium detectors. It has data corresponding somewhat to the AA 119 or AA 143 demodulator diodes known of old, the blocking voltage being somewhat higher (40 V) and the capacitance considerably lower. Among other things, Fig. 3 shows the reading measurement circuit section.

At the output of the rectifier, two direct voltages are available which represent the readings for the voltage level and the current level at the aerial connection.

#### 4.2. Level measurement

To measure the impedance level, the output voltages of the voltage transformer and current transformer are rectified at high resistance. The two DC voltages actually had to be divided by one another which, in analogue circuit technology, requires a bit of effort. A digital divider circuit, in the form of a micro-controller, is certainly simpler, but you then also need two analogue/digital converters with 10 to 12 bit resolution and corresponding display technology. So that idea was ruled out from the start.

If it is desired to represent readings over a wide range of values, then a logarithmic scale suggests itself. The better analogue logarithmic circuits form the logarithm from the ratio of two voltages. Such circuits have been put forward many times in VHF Reports. A look at Tietze-Schenk [1] will provide further

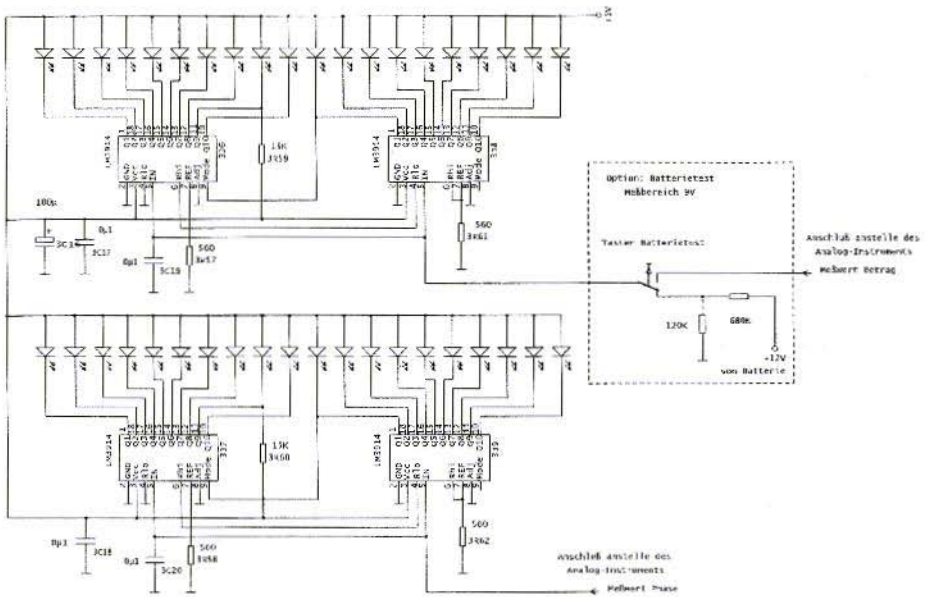
assistance.

As Fig. 3 shows, the logarithmic circuit consists of two precision OPs (OP 07 or MAX 400) and a p-n-p double transistor, or two selected BC 559 individual transistors. It forms the logarithm of the quotient of the two voltage values. See [1] for more details on this method.

At the output of J1 the logarithmic level signal appears, and a voltmeter with a central zero setting could be directly connected here. Since such display instruments are rare, there follows an amplifier stage with a TL 082 (J5), which provides adjustable amplification and voltage offset. The output voltage is scaled for the LM 3914 luminous spot modules used in the specimen, which require a voltage of 1.25 V for full display.

A non-linear voltage divider (R 36, R 40, D 8, D 9) is incorporated between the logarithmic circuit and the display amplifier. This addition was DJ 3 VY's idea and ensures that the scale is extended by the factor 3 around the 50-Ohm point, thus making it easier to adjust for precise tuning.

The window comparator J 2 is a constituent part of the level measurement equipment and, depending on the voltage measured, it switches on one of the three LEDs Level Too Low (< 1 W), Level OK (1 to 100 W), or Level Too High (> 100 W). Reliable measurements can be obtained even with power levels



**Fig. 4: The LED Display Module**

**Batterietest = Battery test, Meßbereich = Measurement range, Taster Batterietest = Battery test key, Von Batterie = From battery, Anschluß anstelle des Analog-Instruments = Connection instead of analogue instrument, Meßwert Phase = Phase reading**

of under 500 mW!

**4.3. Phase measurement**

For phase measurement, the output voltages of the measurement transformer are first converted into square-wave voltages. Two gates running as monostables generate spike pulses, which set and reset an RS flip-flop. The mean value of the flip-flop output voltage is then the phase measurement value.

To start with, the high frequency is fed

directly to the corresponding series-connected 74 AC gate. As long as RF is present, everything is fine. But when no RF is fed in, the gates go into linear drive, and their current consumption rises giving the risk of overheating. For this reason, pulse formers are series-wired with differential amplifiers.

As Fig. 2 shows, a limiter with two Schottky diodes which are connected prior to the inputs. The operating point of the amplifiers is set in such a way that the open-circuit output voltage lies at around 2 V, i.e. below the critical

linear range. The two trimmer capacitors at the outputs are used for tuning at high frequencies.

The differential amplifiers were initially assembled using the correct high-frequency transistors, but the circuit seemed to have ideas of its own. It was not clear what was happening until a sampling oscilloscope was used with a 10 GHz band width. On the pulse tops of the square-wave voltage there were self-oscillations bouncing up to 2 to 5 GHz! A telephone conversation with DF 8 CA brought the solution. The transistors were too good. What was needed were lame high-frequency transistors. BFS 17 is a derivative of the ancient BFY 90 type, with a transit frequency of app. 1 GHz.

Following the differential amplifiers, the two signals are square-wave to some extent. Two 74 AC 00 gates provide for the final square-wave formation, and two monostables, which utilise the gate running times and the RC time constants of the inputs, generate negative spike pulses with a width of 3 to 5 nsec, which set and reset a flip-flop which is also formed using 74 AC 00 gates.

The current signal is phase-displaced by  $180^\circ$  with regard to the voltage signal, which is set to have the polarity of the current transformer. With a resistive load ( $0^\circ$  phase between voltage and current), a rectangular voltage with 50% mark-space ratio appears at the output of the flip-flop. Depending on the phase angle of the aerial impedance, the mark-space ratio varies from 25% to 75%, or the mean voltage following the low-pass filtration varies from 1.25 to 3.75 Volts. As in the level channel, a

T1. 082 (second part of J 5) provides for the tuning of the level and the open-circuit voltage. Here too, a voltmeter can be connected up directly.

#### 4.4. The display module

Anyone with lots of space will be well served with two voltmeters or milliammeters as displays, especially as the operational amplifier supplies enough voltage ( $\pm 3$  V) and current ( $\pm 20$  mA) to drive almost any instrument.

A 20-place luminous spot display with LED bar modules saves space and is precise enough for the application. Triggering can easily be provided using two LM 3914 cascaded LED luminous spot drivers from National Semiconductor [3]. They are wired up for a display range from 0 to 1.25 V, because that is the simplest external circuit. The circuit used is practically the application circuit.

There are no constraints for the display assembly. If we start with a micro-controller with an LCD module as a display assembly, then the following options can be obtained:

Combined display as numerical value and bar diagram

Graphic-option LCD display with representation as Smith diagram

The signal frequency which is anyway present in the CMOS level is divided down using a fixed divider and counted by the micro-controller.

The reactive impedances can then be converted into inductances or capaci-



tances.

The micro-controller supplies control signals for an automatic aerial tuner

The last concept, in particular, is fascinating, since the aerial tuners present in amateur transceivers, or externally, can function well only if hardly any tuning is actually necessary. They are still a long way from what can be bought second-hand from Marconi or Rockwell-Collins.

#### 4.5. Power supply

The measurement electronics require two stable distribution voltages of  $\pm 5$  V, the power consumption on the positive side being much higher. Initially, we thought of two accumulators and two fixed voltage controllers, but then it would have been necessary to monitor the charging condition of two accumulators. The alkali/manganese batteries already occasionally used are more advantageous, for they have a long-term storage capability and an energy content which is about three times as big as that of an NiMH accumulator column. Against that, the voltage drops upon discharge to around half the rated voltage. Thus to utilise the battery capacity only a switching controller power supply with a wide input voltage range is advantageous.

In the present case, a 9-V power unit is sufficient in terms of capacitance, even if this type of battery, in contrast to, for example, DIN 40863 round cells, is not optimised for high loads. Now things get difficult. The initial voltage of the

batteries is far above the output voltage, while the end voltage of the batteries is a little below it!

This is not a problem with a converter transformer, but the expense still scares people off. Enlightenment came from some pages in the 1996 Data Book of Linear Technology [4]. There we find a smart transformer circuit called a Sepic transformer, which can transform upwards and downwards and, in addition, needs only a bi-filar wound storage choke and an AC voltage fixed electrolytic capacitor. So, in brief, a test circuit was soldered together with a standard IC for upwards transformers (LT 1172 from Linear Technology) and things went well over an input voltage range from 3 to 20 V. Above that, things became critical for the integrated power transistor and the output-side Schottky diode from the point of view of voltage. With the help of an auxiliary turn and a linear secondary controller, even the negative supply can be obtained. Fig. 5 shows the result. This is virtually the cheapest way to do it.

But there still remained one standard problem for all battery-driven equipment. It is put away switched on and forgotten. Then one can only hoped that the battery really is leak proof and will not pour its electrolyte contents into the equipment. But something could be done here using the stand-by mode of the LT 1172. If pin 2 is earthed at low impedance, the switching controller goes into stand-by and needs 50  $\mu$ A max.! A timer, triggered by pressing a key, switches the equipment off after a short time, but an LMC 555 (CMOS version of the NE 555) needs about 200  $\mu$ A, far

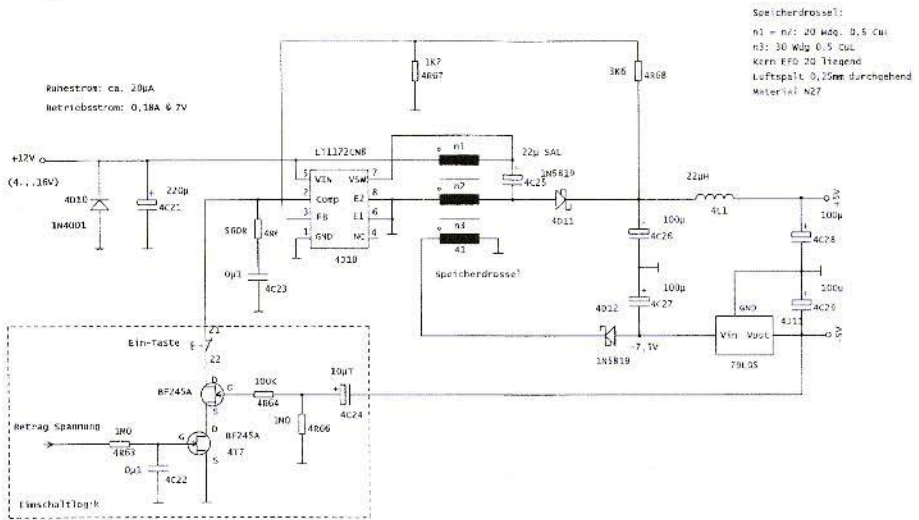


Fig. 5: Current Supply with Switch-On Control

**Ruhestrom = Open-circuit current, Betriebsstrom = Operating current, Betrag Spannung = Voltage level, Einschalttoggle = Switch-on toggle, Ein-Taste = On key, Speicherdrossel = Storage choke, Kern = Core, Liegend = Horizontal, Luftspalt = Air gap, Durchgehend = Continuous**

too much.

After one day, the timer circuit shown in Fig. 4 came about, with two BF 254 As. The key (the break contact!) starts the transformer IC. The negative voltage blocks the upper FET through the R 66 time function element, C 24, for approximately 10 seconds. Then the transformer switches off again. The lower FET triggers switching on if high frequency is applied. If the voltage of the voltage rectifier is sufficient to block the FET (app. 1 V on the BF 245 A), then the equipment switches on and remains

on as long as the high frequency is present, but for a minimum of 10 seconds. Now only continuous high frequency or a mechanically blocked key can suck the battery dry. In intermittent drive (10 minutes continuous transmission, 10 minutes pause), a measurement time of 70 minutes was obtained anyway, which should be sufficient for a field day season.

Naturally, there is no obstacle to an external power source at 4 to 16 V either. However, since the transformer power supply has no voltage isolation, caution is advised so as to avoid





short-circuits through the coax cable external conductor. With external power supplies, in particular, the battery must be powered in a practical way.

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## 5. INSTEAD OF ASSEMBLY INSTRUCTIONS

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From discussing the matter with a few radio amateurs, we found that literally each one of them would have preferred a slightly different solution. For this reason, the author does not want to try to lay down full assembly instructions, but rather to stimulate people to produce their own custom-built solutions using the components described above, in several assembly stages. Since the circuit contains some SMD components, it can anyway be recommended only for experienced solderers!

### **Assembly stage 1: Reading measurement and cross-pointer instrument**

Anyone who has had the opportunity to inspect old medium-wave transmitters will have been struck by the large Buschbeck-type cross-pointer wattmeters in the control panels. Anyone who has a sensitive cross-pointer instrument available can just assemble the reading measurement system as per Fig. 2, with voltage and current transformers and the two rectifiers, for some first trial steps. You will then have a combination of relative output display and impedance level display. As long as the voltage and

current readings are the same, the two pointers will form an equilateral triangle, which inclines to the left or right when impedance discrepancies occur.

### **Assembly stage 2: Level measurement**

In Phase 2, the logarithmic apparatus and, if necessary, the level display, must be added. A sensitive measurement apparatus with the zero in the centre can be directly connected, with a matching voltage multiplier, to the output of the operational amplifier, J1.

### **Assembly stage 3: Level and phase measurement**

For universal use, the circuit as per Fig. 3 is fully assembled. Then any display instruments desired can be universally connected. If necessary, the wiring of the output operational amplifier (J5) can be modified.

### **Supplement 1: Luminous spot display**

For a compact apparatus, the luminous spot display module as per Fig. 4 is the optimal solution. A format of 150 x 75 x 30 mm. can easily be realised using a two-level construction (display printed circuit board piggyback on the measurement printed circuit board).

### **Supplement 2: Power supply**

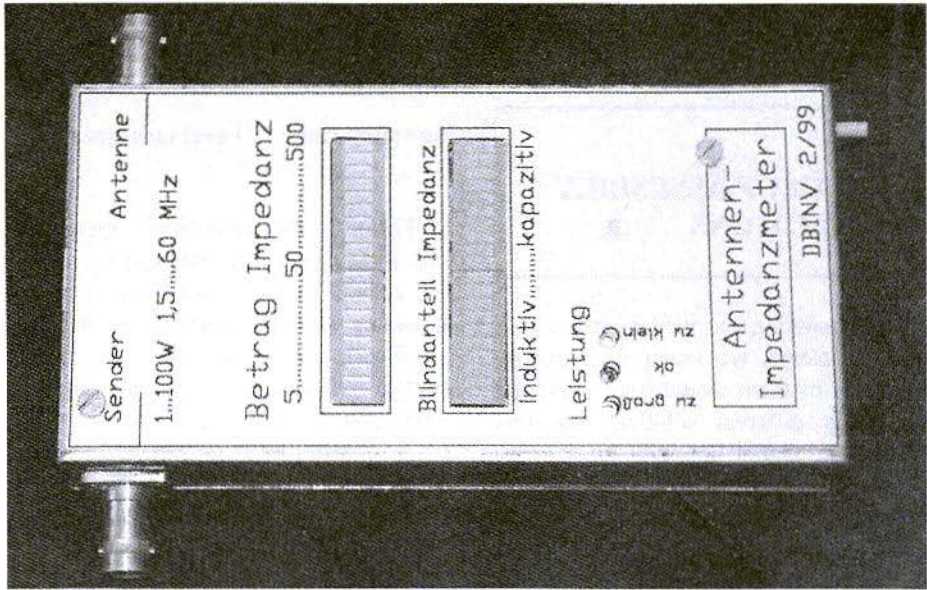


Fig. 6: Specimen Apparatus with LED Bar Display

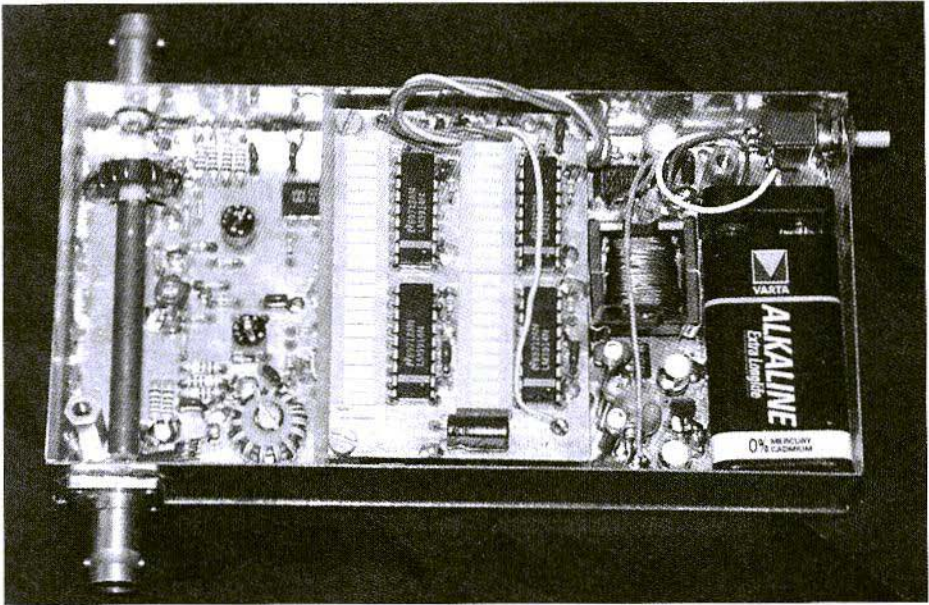
Supplement 2 involves a power supply using the Sepic transformer which is used for the battery operation of the apparatus. Anyone wanting only to carry out measurements at home can use any properly stabilised power supply which can supply = 5 V at max. 300 mA.

Something else which could come into question is a micro-controlled evaluation circuit which, in addition to the level and phase display, can, for example, measure the power or count the transmission frequency, in order to convert reactive impedances into component values.

## 6. BALANCING INSTRUCTIONS

For balancing and operational checks, in addition to a transceiver or a high-power signal generator, we need a (good) 50-Ohms loading resistor and an oscilloscope. A power attenuator is also useful. You can construct a few test standards yourself at little expense:

- BNC connectors with short cable ends (max. 2 to 3 cm.) have the following adapted test loads:



**Fig. 7: View inside Specimen of Vectorial Aerial Impedance Meter for Short Wave**

- 2 x 10-Ohms 0.6 W metal layer parallel, yields 5 Ohms for lower range limit
- 2 x 1 kOhm 0.6 W metal layer parallel, yields 500 Ohms for upper range limit
- Capacitor with 320 pF (or standard value 330 pF) yields capacitive reactive impedance of 50 Ohms at 10 MHz
- Coil with 800 nH yields inductive reactive impedance of 50 Ohms at 10 MHz

The measurements are carried out with a

power level of a few Watts. Those who have no faith in the behaviour of their transceivers in the event of mismatch should connect up a power attenuator to the output of the transceiver, so that 1 to 2 Watts of power will come out.

First, feed in approximately 1 W and connect the load at the output. There will then be two voltages of about the same size to measure at the Current Level and Voltage Level measurement points (discrepancy < 10%), they being a little more than 1 V. This step indicates the functioning of the measurement transformer voltage and current, and any coil faults existing are discovered.



With an open output the current reading must drop to almost zero, while with a short-circuited output it will be the voltage reading! The first series of tests is completed by a test at the lower band limit (1.8 MHz) and the upper band limit (30 MHz).

A tip: many multimeters, especially in the cheap range, indicate house numbers for measurements in high-frequency circuits. If the results are not clear, first check the high-frequency operation of the multimeter.

For the next steps, the Sepic transformer must be running, or else two laboratory mains supply circuits are used to generate the  $\pm 5$  V. There is nothing to balance in the transformer power supply, for if R 67 and R 68 in Fig. 4 are correctly chosen as 1% metal film resistors the 5-V supply is correct!

Only the input voltage range can be tested. With a load amounting to 200 mA at 5 V and 50 mA at 5V, the voltages must be correct over an input voltage range of 3 to 20 V. If the transformer is not working properly, check whether the switch-on key is open and whether the windings in the storage choke are wrongly connected!

In a third step, the operating points of the differential amplifier are adjusted with no RF present - using R 15 and R 16 set the outputs between 1.8 and 2 V at the Voltage Phase and Current Phase measurement points. If high frequency is present, two square wave voltages with the input frequency, which are in phase opposition, are to be measured at the measurement points. If they are not in phase opposition, reverse the polarity of the secondary coil of the current trans-

former. If the voltages are not square wave, there is probably a problem with the oscilloscope. It should have a bandwidth amounting to 100 MHz, and the probes must have very short connections to the assembly ( $l < 2$  cm.), especially the earth line! The two phase trimmers remain turned right down (minimum capacity!).

The fourth step sets the measurement range of the level and phase displays. 10 MHz and 1 to 3 Watts are again used for the test signal.

With R43 in Fig. 2, we bring the luminous spot indicating the level with a 50-Ohm dummy load to the centre of the scale. With R 48, we set the maximum amplitude in such a way that for a test resistance of 5 Ohms the left-hand scale end is reached, while the right-hand scale end is reached with 500 Ohms. The two balancing points are interdependent, so several passes are needed.

The same steps are taken for the phase channel with R 50 (scale centre) and R 55 (full scale deflection), but the coil and the capacitor act as the test standard.

Finally, the phase compensation is set: dummy load at output, test signal 30 MHz or (if possible) 54 MHz. The phase display will no longer be in the centre here. By experimentally turning in the trimmer capacitors in the differential amplifiers, the phase display is centred. If a trimmer is operating in the wrong direction, it comes back to the minimum position and we go to the other trimmer.

This completes the balancing. It is



recommended that you test the function-  
ing again at regular intervals.

If we now connect up the switch-on key, the equipment must switch on when there is high frequency present and switch off when the high frequency switches off, or after 10 seconds. If the switch-on control is not working, it can be guaranteed that this will be due to an unsuitable BF 245. You must use a BF 245A!

## 7.

### TRUTHS AND RUMOURS ABOUT AERIALS

There is scarcely any area in electrical engineering (with the possible exception of the Hi Fi high-end people) which is so alive with assumptions and rumours as aerial engineering. Thus, for example, a poor VSWR is made responsible for standing waves and TVI, and new miracle aerials regularly draw attention to themselves.

In order to interpret the measurement values of the aerial impedance meter correctly, a little theoretical background is necessary. For this reason, we should concern ourselves for a moment with the objectives of aerial optimisation. Instead of making a stab in the dark, it is recommended that, intellectually at least, you move forward in several steps, which have nothing to do with one another, but which are regularly mixed up:

a) Adjust current distribution on the aerial conductor which is suitable for the

radiation

b) Compensate for the reactive impedance of the input resistance at the feed point

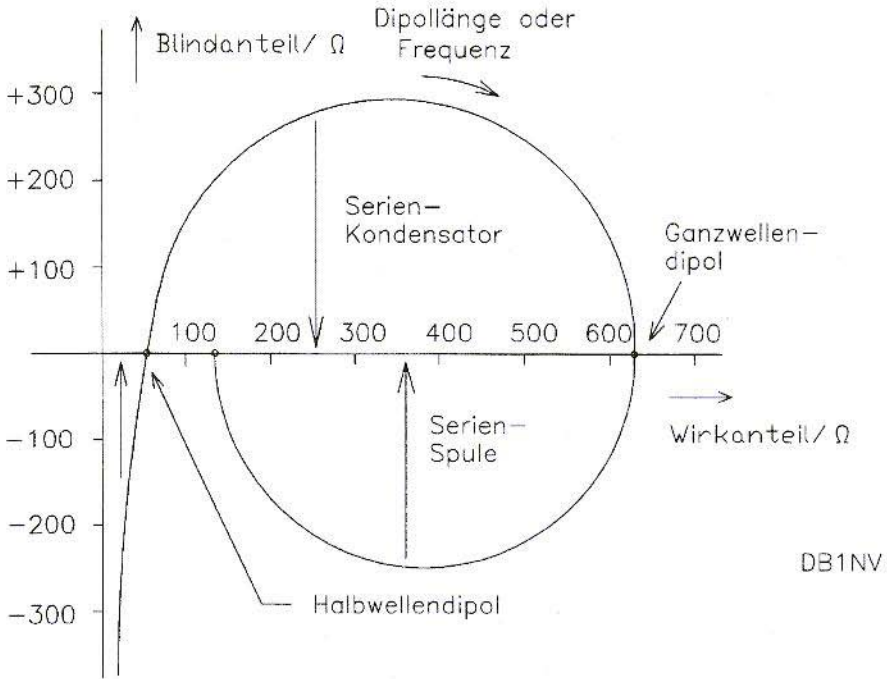
c) Transformation of the now real input resistance to the characteristic impedance

d) Balancing transformation for symmetrical coaxially powered aerials

#### 7.1. The radiator element, less critical than people think

Despite some peoples views to the contrary, the length of the radiator element makes little difference, provided it is under  $5/8$  of the wavelength for a monopole or under one wavelength for a dipole. Thus a shortened dipole (Hertzian dipole) supplies only about 0.4 dB less gain, as against the half-wave dipole, and the full-wave dipole just supplies a gain of exactly 1.7 dB.

For aerial lengths in this range, one current distribution automatically suggests itself which generates a beautiful, dipole-like directional diagram in the well-known eight form. For a  $5/8$  lambda aerial, we also see the first side lobes in the diagram. Longer radiator elements display very fanned-out diagrams which are of little use in practice. This is due to the fact that the current fractions in phase opposition on the aerial conductor generate fields in several directions in space, which compen-



**Fig. 8: Impedance Diagram of Dipole**

**Blindanteil = Reactive component, Dipollänge oder Frequenz = Dipole length or frequency, Serienkondensator = Series capacitor, Serienspule = Series coil, Halbwellendipol = Half-wave dipole, Ganzwellen-dipol = Full wave dipole, Wirkanteil = Effective component**

sate for each other. One remedy is to build in phasing lines for phase balancing, which is known about from stacked radiator elements. Another method reduces the destructive interference of the partial currents by avoiding the reflection of the waves running along the aerial conductor at the end of the wire. We find this on aerials with moving loads such as Beverage aerials, rhombic aerials or attenuated broad-band dipoles.

The reason why half-wave dipoles or

quarter-wave monopoles are so popular lies elsewhere. They have a real input resistance of approximately 70 Ohms for a half-wave dipole and 500 to 1,000 Ohms for a full-wave dipole. Short dipoles have a capacitive reactive component and a low effective resistance, while a dipole length between a half and a complete wavelength has an inductive reactive impedance. For lengths exceeding lambda, the reactive impedance again becomes capacitive.

Fig. 8 gives a simplified version of the



typical loop curve of the dipole input impedance. The active component is plotted towards the right and the reactive impedance towards the top. In order to arrive at the real axis, for short dipoles an inductance (extension coil) with a positive reactive impedance must be wired up in series. For a dipole length of more than  $\lambda/2$ , a capacitor (shortening capacitor) with a negative reactive impedance should be wired up in series. If the aerial is longer than  $\lambda$ , then a coil is required at the power supply point again, as we know from the base coil of the  $5/8$   $\lambda$  aerial.

The same considerations also apply to monopole aerials (rod aerials) over an earth which conducts well.

### 7.2. Impedance transformation

If a real aerial input resistance has been attained, then it should be transformed into a common cable characteristic impedance (50 or 75 Ohms). True, everything can be tuned using a reactance transformation (variable disc capacitor and rolling coil or Pi filter), but the higher the transformation ratio the narrower the band is, and the more reactive components are involved. This is also the reason why broad-band high-level stages can scarcely be constructed with valves. The only solution, the cascade amplifier, has to be ruled out for amateur applications, but power levels of a few kilowatts above a frequency range of 100 Hz to 200 MHz can be achieved with it!

Broad-band transformers can be created

for frequency ratios of 1: 10 or more, provided we are working in the resistance range of 5 to 500 Ohms. At very low resistances, the stray inductance interferes, and at very high resistances the stray capacity does.

Even greater band widths can be achieved using line transformers, but this commits us to whole-number turn ratios and thus to resistance transformations of 1, 4, 9, 16, etc..

### 7.3. The balun, the unknown component

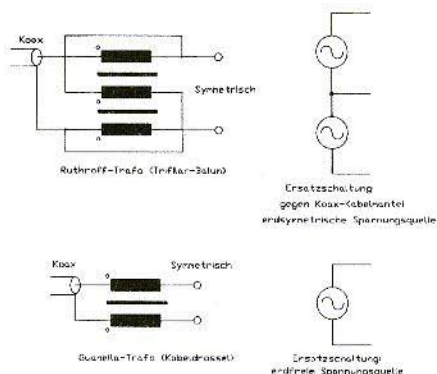
The final part of the task is balancing the circuits, i.e. the transition from the coax cable onto a symmetrical dipole. This procedure, simple in itself, is an ideal breeding ground for seductive half-truths and beliefs arrived at from reading tea leaves.

Some people essentially consider balancing network as superfluous and simply clamp coax cores and casings to the dipole. This even frequently works without any relatively large amount of high frequency being carried into the house. The deformed directional diagrams are in any case not checked.

Others curse baluns and Guanella transformers because they so often go wrong. However, they are usually overloaded.

A third group believes only in specified, well marketed miracle brands of balancing circuits.

So let's shed some light on the darkness. The balancing network has two functions, namely to prevent high-frequency



**Fig. 9a: Cable Choke and Guanella Transformer**

**Fig. 9b: Trifilar Balun**

**Symmetrisch = Symmetrical,**  
**Ruthroff-Trafo = Ruthroff transformer,**  
**Trifilar-Balun = Trifilar balun,**  
**Ersatzschaltung = Alternate circuit,**  
**Gegen Koax-Kabelemente erdsymmetrische Störungsquelle = Voltage sources balanced to coax cable,**  
**Guanella-Trafo (Kabeldrossel) = Guanella transformer (cable choke),**  
**Erdfreie Spannungsquelle = Earth-free voltage source**

flux on the external conductor of the coax cable and simultaneously to drive the aerial with balanced voltage or with balanced current. It is the last point where the fine difference resides. People tend to overlook this, for the normal baluns or cable chokes fulfil only one of the two functions.

An ideal dipole is a balanced formation. Here any balancing circuit brings the desired success. We are most likely to find this condition in multi-element Yagis in the VHF/UHF range or in larger shortwave beams, if they are mounted at a distance of several wavelengths from interfering ambient components. In addition, a definite earth reference point exists here, in the form of the boom in the plane of symmetry of the aerial.

With dipoles for lower frequencies, one is frequently not in a position to maintain sufficient distance from structures, trees, power lines or other aeriels. The two dipole branches then have different stray capacitances to earth.

The cable choke or Guanella transformer as per Fig. 9a suppresses high-frequency current conduction on the cable casing through its direct cycle inductance, while the useful load flowing in the cable is unaffected by it, for the fields of the internal and external conductor currents cancel each other out in the external direction. To put it another way, the casing of the coax cable is impenetrable for the useful load. Thus the power cable remains free of high frequency externally, provided it is fed away at right angles from the power feed point in the plane of symmetry of the dipole.

The Guanella transformer ensures that the current being fed into the dipole is symmetrical. The Guanella transformer acts like an isolation transformer and simulates an earth-free voltage source at the power feed point. Due to ambient influences, asymmetrical voltages to earth build up easily on the aerial





halves, but this has little effect.

There is one effect which even an ideal balun can not prevent the radiation effect of the feeder cable. If the field of the aerial is undisturbed, then a plane of symmetry of the fields, standing vertical to the dipole, exists at the power feed point.

Provided the feeder cable is fed away vertical to the dipole in the plane of symmetry, it consumes no power from the high-frequency field of the aerial. Otherwise, there are electrical field components parallel to the cable, and magnetic field lines which intersect the cable. The cable acts as an aerial.

In spite of correct balancing at the power feed point, high H-F voltages arise on the external conductor of the station-side cable end, especially if the power line has a length which is resonant for the operating frequency. A second cable choke, directly on the point where the cable enters the house, is of assistance here!

Things are different for a standard trifilar balun as per Fig. 9b. It generates two voltages to its earth connection, in phase opposition, of exactly the same size. If the earth connection is linked via a Yagi at the power feed point to the boom, then the problem is solved here.

But what about the earth connection for a freely suspended dipole? The only connection to earth is the external conductor of the coax cable. If the dipole is now unbalanced, as a result of the ambience, then the reactive currents flow to the earth connection of the balun and from there onto the external conductor of the cable.

In spite of the balun, we have standing waves on the cable and high-frequency irradiation onto other equipment in the house. Anyone who does not understand this merely comes to the conclusion that the balun has not achieved anything with regard to the TVI and the irradiation interference, so the balun is superfluous!

A second cable choke, directly on the point where the cable enters the house, is of assistance here as well!

There are even more subtle differences between the Guanella transformer / cable choke and the trifilar balun. With the Guanella transformer or the cable choke, the useful power runs inside the coax cable and thus does not load the transformer. The trifilar balun works more like a normal transformer, in which the entire power is carried over the windings and the core. Details regarding the dimensioning of broadband transformers for tuning and balancing can be found in [5].

With this knowledge, it is a simple matter to interpret the displays of the aerial impedance meter correctly.

If we measure directly at the power feed point of an aerial, the reactive component is the indicator for the adjustment of the resonance length.

If the half-wave dipole is too short, we measure a capacitive reactive component. If it is too long, we measure an inductive reactive component.

If the reactive component,  $0$ , of the active component is a long way from 50 Ohms, then with Yagis this can be traced back to the influence of the parasitic elements. In general, the parasitic elements reduce the input resist-



ance, so that now and again a loop dipole has just 50 to 60 Ohms. As can be seen from [2], it is scarcely possible to optimise the development objectives gain, good front-to-back ratio, band width and base impedance simultaneously. If the base impedance can not be brought to a point close to 50 Ohms, then transformation circuits must be provided for at the power feed point.

If we measure at the end of the feeder cable, then the aerial input resistance over the line is transformed. The reflection factor level and thus the standing wave ratio are still maintained. Only the phase is rotated. With electrical line lengths of half a wavelength (the mechanical length is reduced by the shortening factor) the precise Aerial base impedance is present at the start of the cable. For experiments, a half-wave feed is the solution. For a test on standing waves, i.e. inadequate balancing, the feeder cable should be coiled four or five times through a large ferrite ring core and the result should be compared with the measurement without the ring core. If there has been no change in the impedance, the cable is largely free of standing waves.

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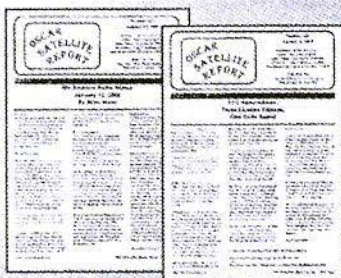
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# TEFLON OR EPOXY BASE MATERIAL?

**Two identically constructed low-noise pre-amplifiers with source counter-coupling for 1,200 to 1,400 MHz in a comparison test.**

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## 1. INTRODUCTION

---

Ultra-low noise HEMTs have recently attained such low noise figures that the noise figure which is practically attainable for frequencies of 1 to 2 GHz is clearly specified through the circuit loss of the input voltage.

A type ATF 36077 transistor (HP) or even a type MGF 4919 (Mitsubishi) has a theoretical internal noise of 0.1 dB ( $< 10$  K) in this range. Thus at room temperature and with low circuit losses (typically 10 - 20 K) at 23 cm., noise temperatures of 20 - 30 K (0.3 - 0.5 dB) are obtained with transistors of this type. The minimum noise figure here is determined by optimal matching of the transistor input (noise matching), in combination with the lowest input circuit losses. The optimal generator im-

pedance is very high-ohmic in the frequency range selected (typically,  $R_{opt} = 91$  Ohms,  $X_{opt} = j 350$  Ohms), which means good broad-band input matching can be attained only with great difficulty. Moreover, these extremely low-noise transistors are not absolutely stable except from approximately 5 GHz upwards ( $K > 1$ ), so that their use at low frequencies is always linked to a tendency to self-excitation. Stability and noise matching over a wide frequency range with almost matched input ratios (good return loss) without the use of insulators sounds like a contradiction.

The amplifiers presented here work by the method of inductive source counter-coupling, which was described by A. Anastassius and M. Strutt in 1974 [1] and put into practise by Nevin and Wong [2] and by D.R. Williams, W. Lum and S. Wienreb [3] in 1980.

This type of circuit had already been investigated by the author at the beginning of the eighties, using GaAs-FETs of types MGF 1400 and CFY 13, successfully tested and described in a publication by G. Schwarzbeck [4]. A

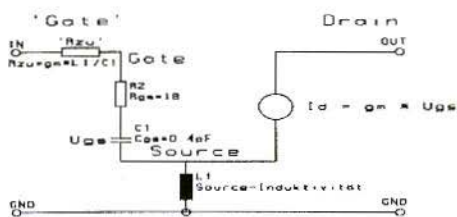


Fig. 1: Influence of source inductance on input impedance

further circuit description of this kind, using the ATF 10135 / 10136, can be found in [5].

An attempt is made below to clarify the effect of inductive source counter-coupling on two identically-constructed amplifiers with modern semi-conductors. These were assembled using different base materials (Teflon and epoxy), so

that the influence of the circuit quality on the noise figure and the amplification could be practically investigated simultaneously.

## 2. EFFECT OF INDUCTIVE SOURCE COUNTER-COUPING

This effect was already being investigated in low-noise triodes at the end of the forties [6], and this interest was transferred to GaAs-FETs in the seventies which, like the high-transconductance triodes, required high-Ohmic generator resistances for noise matching [7], [8], [9]. A precisely measured inductance is incorporated in the source-base circuit here, in series with the source connection. This increases the real fraction of the input resistance almost

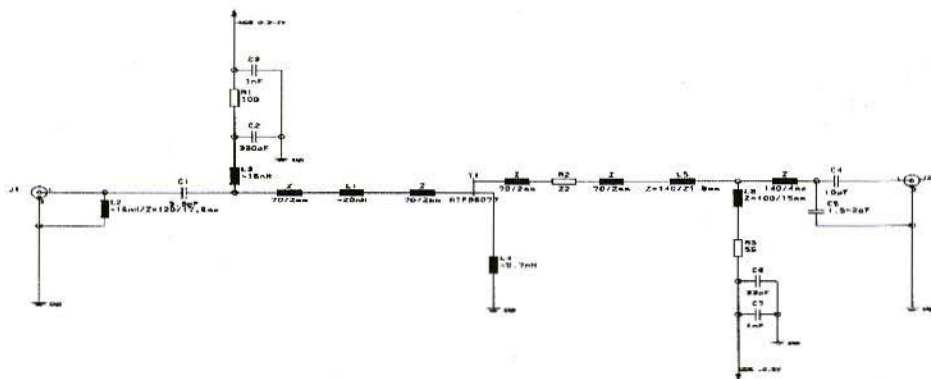


Fig. 2: Amplifier circuit without voltage supply

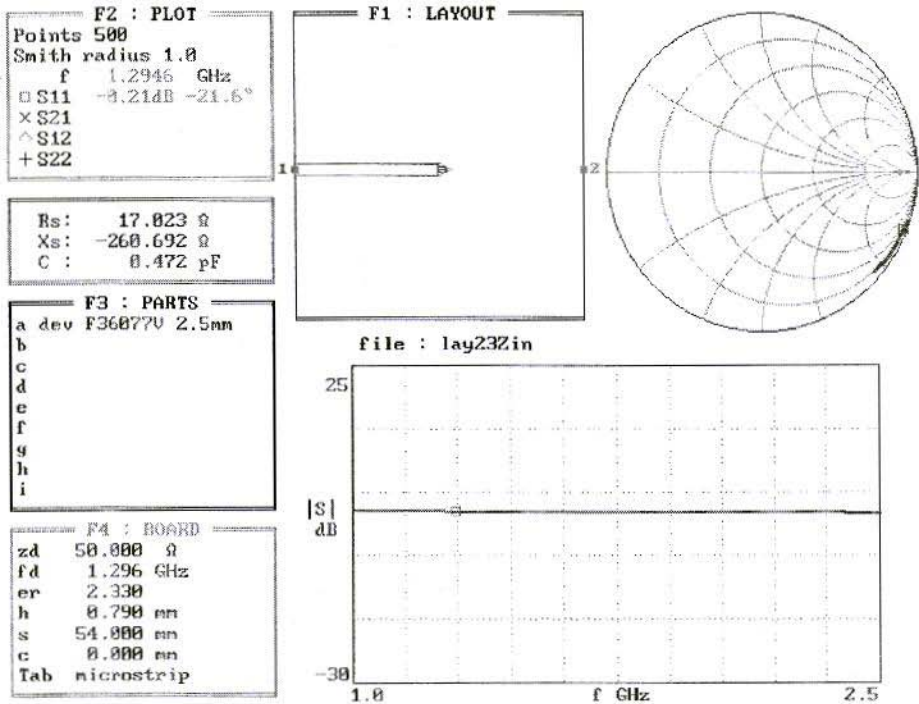


Fig. 3: Input impedance of FET

independently of the frequency, because the input quality of the FET falls by an amount which depends on the gain of the transistor, the input capacitance and the level of additional source inductance (Fig. 1).

The physical relationship of the noise parameters, in relation to the material, component size, voltage and current values, is rich in complexity and is comprehensively presented in [7].

In formal terms, the increase in the input resistance can be reduced to the following expression:

$$R_{zu} = g_m \times L/C$$

where

$R_{zu}$  = additional real fraction of input resistance through counter-coupling

$g_m$  = transconductance in V/mA

$L$  = inductance in source feed in nH

$C$  = gate-source capacitance, pF

Typical values for modern HEMTs:

$g_m = 50\text{-}75 \text{ mA/V}$

$C = 0.2\text{-}0.5 \text{ pF}$



For the noise matching of a type ATF 36077 transistor at 1.3 GHz, we need an optimal generator resistance of  $Z_{opt} = 91 + j350$  Ohms, which must be formed by loss-free transformation [14].

Counter-coupling in the source brings about real fraction transformation through the increasing of the input resistance by  $R_{zu}$ , in this case by 41 Ohms. The imaginary fraction can then be matched by means of a simple series inductance in the gate connection, so that the entire transformation circuit consists of two inductances.

An additional effect arising is that the input matching (return loss  $S_{11}$ ) is broad-band in this kind of transformation, and also still displays decidedly good values. The increase in the real fraction in the input is almost independent of the frequency, because the input capacitance changes only insignificantly, or  $C$  remains  $< gm$ .

### 3.

## CIRCUIT ASSEMBLY

Fig. 2 shows the amplifier circuit. The input has a high-pass filter with a limiting frequency of 550 MHz. The series inductance,  $L_1$ , neutralises the internal capacitances and thus determines the frequency at which the lowest noise figure is present. Through the input capacitance and the transistor gradient, the source inductance,  $L_2$ , determines the size of the additional Ohmic resistance before the gate connection and thus makes broad-band noise matching possible.

The output has a 1:4 attenuator, this is broad band which matches the FET. The size of 25s capacitance makes it possible to fine tune the output impedance to 50 Ohms at the desired frequency. The amplifier circuit pictured corresponds to the circuit principle as per [3], and has been investigated using the well-known Puff software, using not the S-parameters but the output matching parameters for noise matching.

This is because the effect of source inductance cannot be simulated with the above-mentioned software, it only allows transformation using scatter parameters.

The *noise matching* is essentially determined through *the real fraction*, so that the circuit can be laid out to optimise output matching and acquires the noise matching through the correctly measured source inductance. The optimal generator resistance for the *output matching* is 1.3 GHz at  $Z_{opt} = 17 + j280$  Ohms, i.e. with a low-Ohmic real fraction, as shown in Fig. 3, based on the calculated input impedance of the transistor.

For *noise matching*, the generator resistance must be at  $Z_{opt} = 91 + j350$  Ohms, i.e. with a high-Ohmic real fraction.

Fig. 4 shows the graph for the reflection factor,  $opt.$ , for  $U_{ds} = 1.5$  Volts and  $I_d = 10$  mA, in accordance with the technical data of HEMT ATF 36077 from Hewlett-Packard [14].

Therefore, for output matching the real fraction in relation to 50 Ohms must be transformed downwards, whilst it must be increased for noise matching. The imaginary fractions for both types of

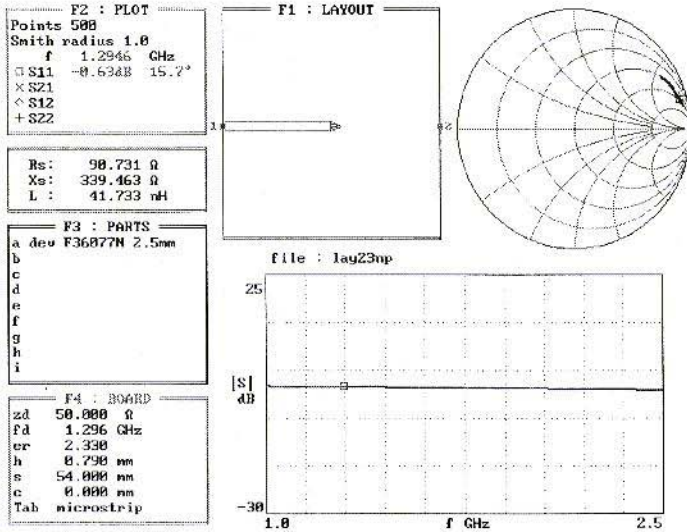


Fig. 4: Optimal generator resistance for noise matching

matching are not far apart. It is thus obvious to lay out the circuit in the simulation initially in the input matching direction. The actual sizes of the source inductance and the series inductance are

empirically determined in order to arrive at the necessary broad-band noise matching.

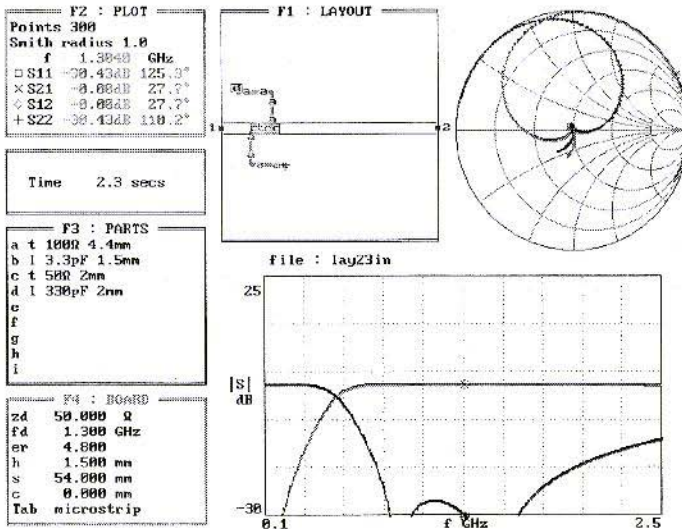


Fig. 5: High-pass filter



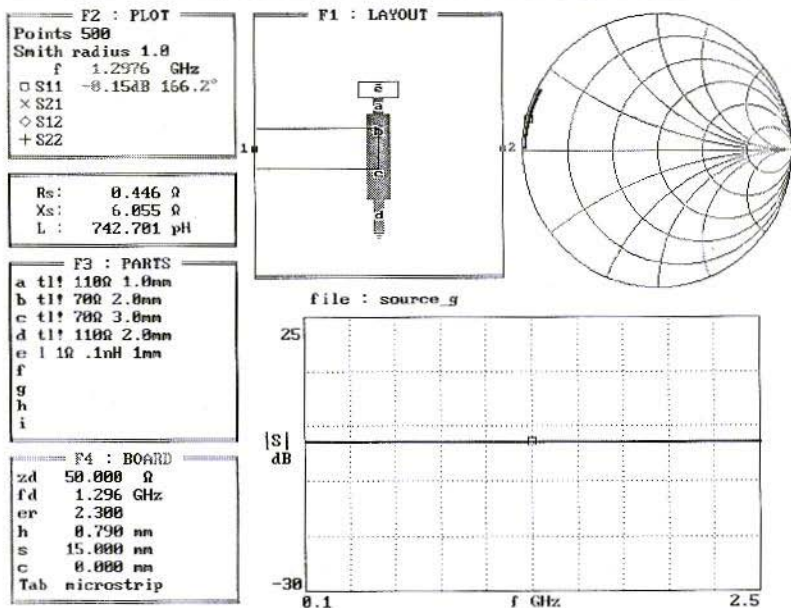


Fig. 6: Simulation of L4 source inductance

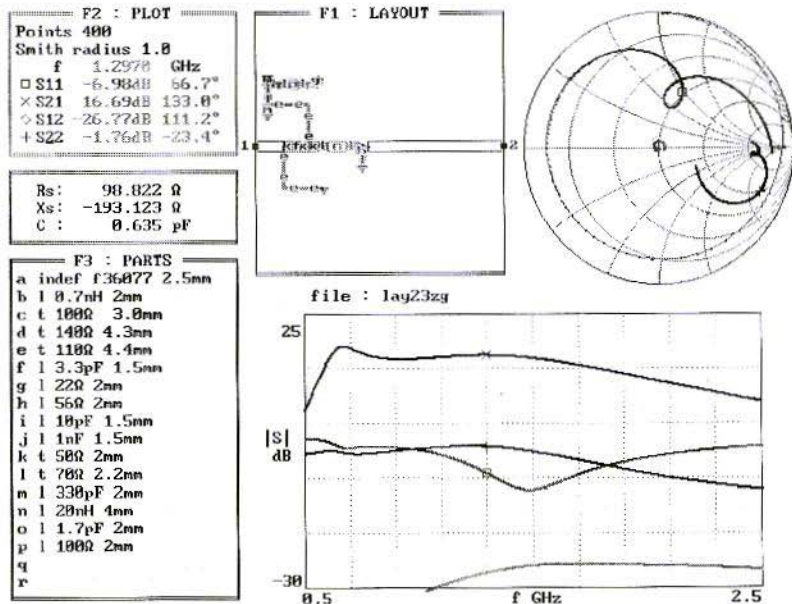


Fig. 7: Influence of source inductance on drain circuit impedance

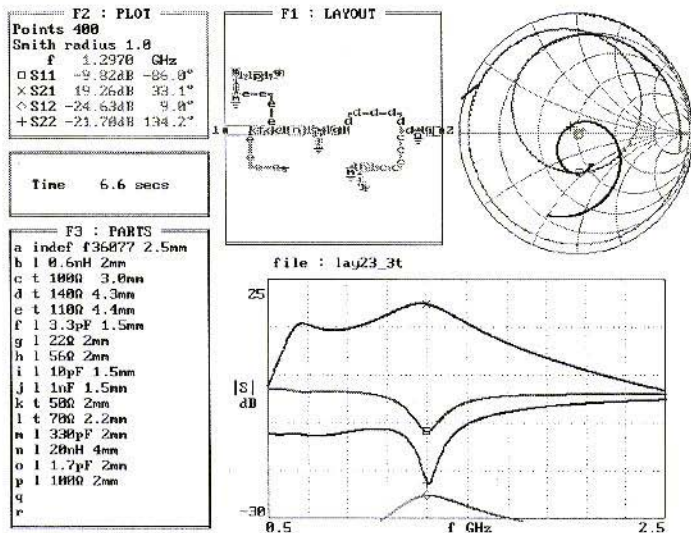


Fig. 8a: Simulation of amplifier circuit on Teflon material

## 4. SIMULATION RESULTS WITH PUFF

### 4.1. Input circuit with high-pass filter

The high-pass filter is initially developed in the input circuit. Its reciprocal frequency lies at 550 MHz and the impedance level is at 50 Ohms. Fig. 5 shows the simulated filter on epoxy material (FR4). It is relatively loss-free and is intended to reduce the strong signals below 500 MHz. The inductances, L2 and L3, of approximately 16 nH, are formed by two striplines with  $Z = 100$  Ohms. In combination with the series inductance, L1, in the input, this gives a quasi-matched procedure for a value of 20 to 25 nH.

Quasi-matched because the influence of the source inductance and the connec-

tion inductances on the FET is not yet being taken into consideration here. This means that the circuits real and imaginary fractions are still too small.

### 4.2. Source counter-coupling

The designing of the source inductance is of decisive importance for the functioning of this type of circuit. By using Puff, a size estimate can be obtained for the realisation of an inductance of approximately 0.7 nH. The source connections are formed using stepped, short microstrip lines, and one side is earthed through a 0-Ohm or 1-Ohm chip resistance.

Fig. 6 shows the result, in which the required circuit lengths and wave resistances are specified. The calculated lengths must be increased by approximately 10% in the actual layout, so that

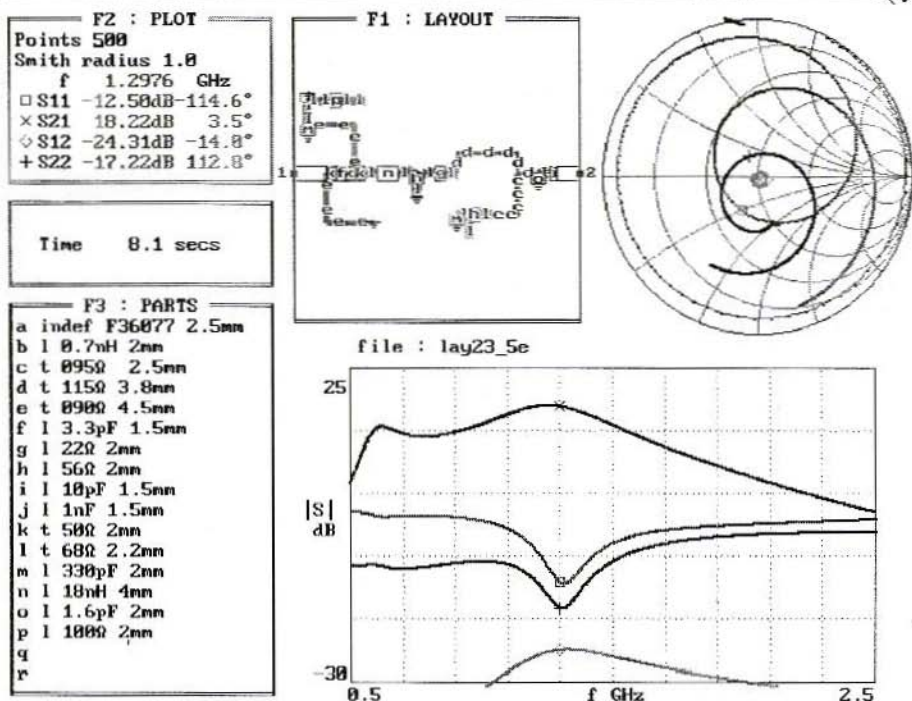


Fig. 8b: Simulation of amplifier circuit on epoxy material

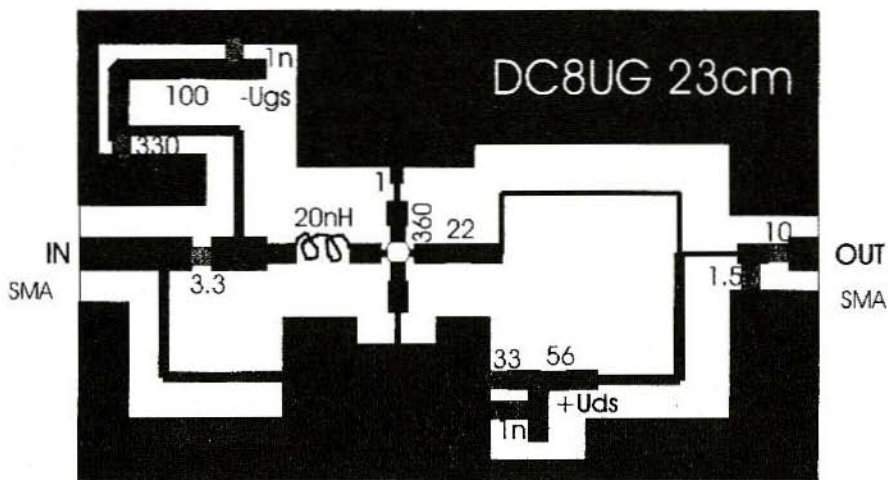


Fig. 9: Layout of amplifier circuit with components

Printed circuit board 54 x 35 mm, Teflon substrate 0.79 mm

L3 = 2 nH - 3 turns, 0.3 mm CuAg, 2 mm Diameter, 2 mm long, 1.5 mm leads

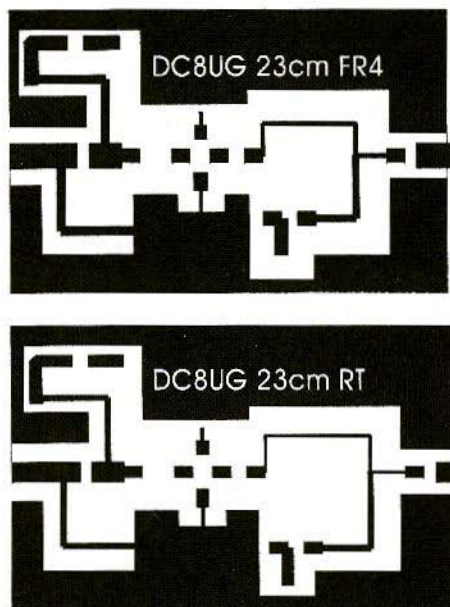


Fig. 10: Layouts for the two materials, RT-duroid and epoxy-FR4

the dispersion at the transition from  $Z = 120$  Ohms to  $Z = 0$  Ohms (KS) is balanced out. This is then brought about by the unshortened source connections.

#### 4.3. Output circuit

The source inductance leads to a change in the generator resistance of the FET in the drain circuit. In accordance with the scatter parameter,  $S_{22}$ , the output resistance is approximately  $Z_{out} = 150 \text{ j}80$  Ohms for 1,300 MHz. The additional source inductance alters the value to approximately  $Z_{out} = 100 \text{ j}190$  Ohms, because it brings about a reduction in the series output capacitance from ap-

proximately 1.5 pF to 0.6 pF, with a reduction in the effective resistance from 150 to 100 Ohms. This change is strongly dependent on the size of the source inductance. Fig. 7 shows this change in the results window,

#### 4.4. Total circuit

Figs. 8a and 8b show the simulations of the developed circuits for Teflon and epoxy base materials. The amplification expected is 18 dB and the return loss in the output approximately 17 dB. A comparison of the two simulation results leads one to believe that the epoxy layout is rather more broad-band, due to the greater dielectric losses ( $\tan$  is greater by a factor of 10) i.e. the attenuation is somewhat reduced, but then the return loss is better.

The input matching, which is simulated here, can not correspond to the actual situation, because the influence of the counter-coupling inductance can not be represented using Puff. So here we must fall back on other software products, e.g. Pspice 9.0..

Fig. 9 now shows an enlarged layout assembly for the Teflon version. Fig. 10 shows the layouts for RT duroid and FR 4.

The amplifier needs a current supply to operate which generates max. + 3 Volts drain-source voltage and 3 Volts gate-source voltage. A corresponding current supply is described, for example, in [12].

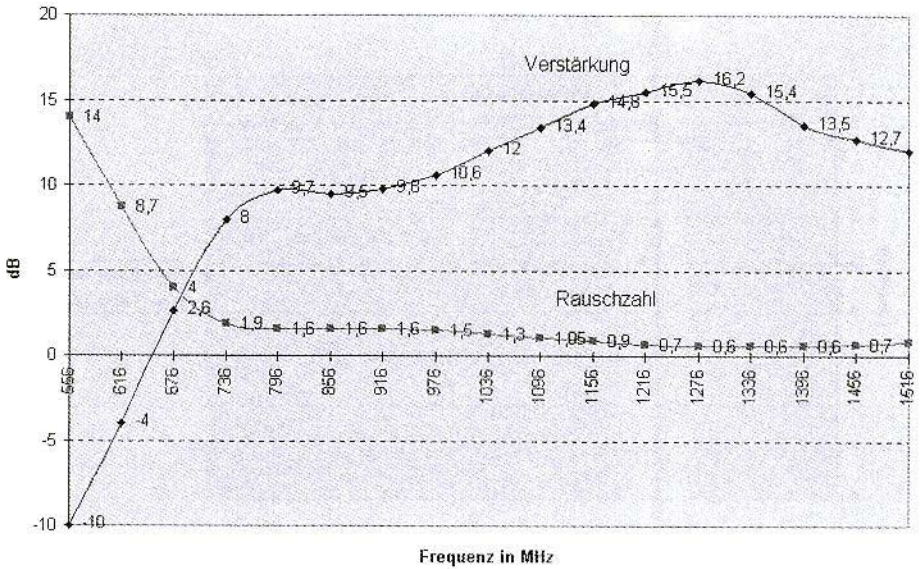


Fig. 11: Measured noise figure and amplification of Teflon assembly  
 Verstärkung = Amplification, Rauschzahl = Noise figure,

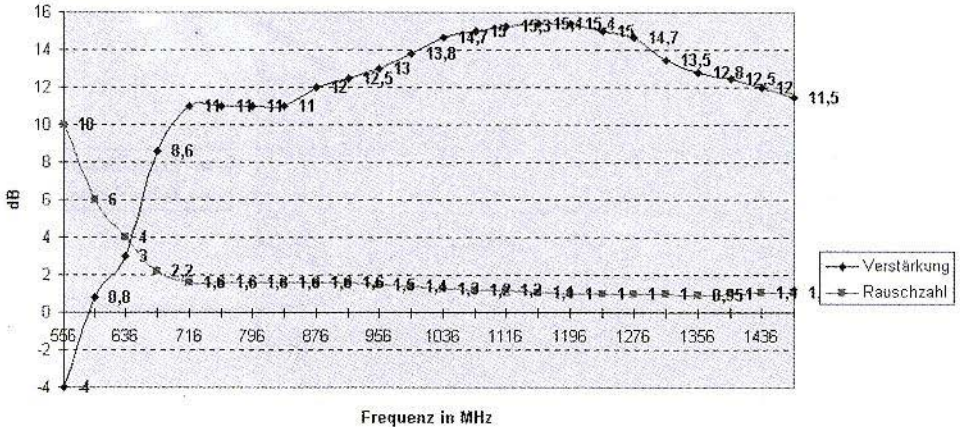


Fig. 12: Measured noise figure and amplification of epoxy assembly

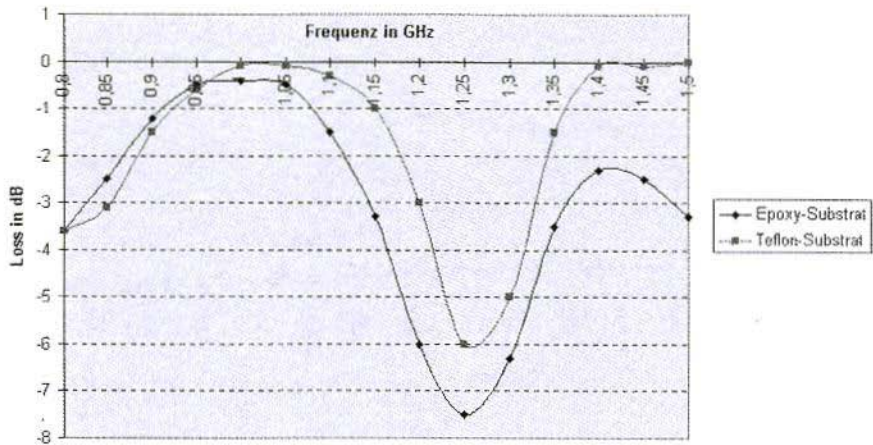


Fig. 13: Comparison of return loss levels obtained in input

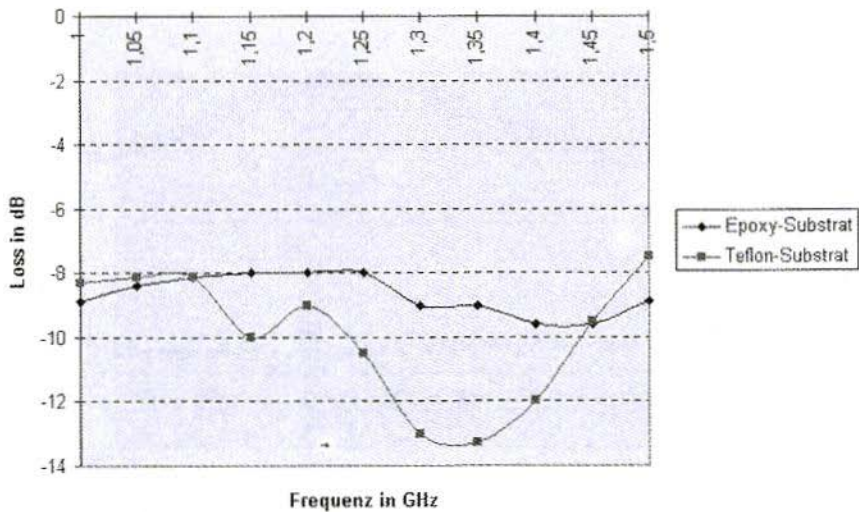
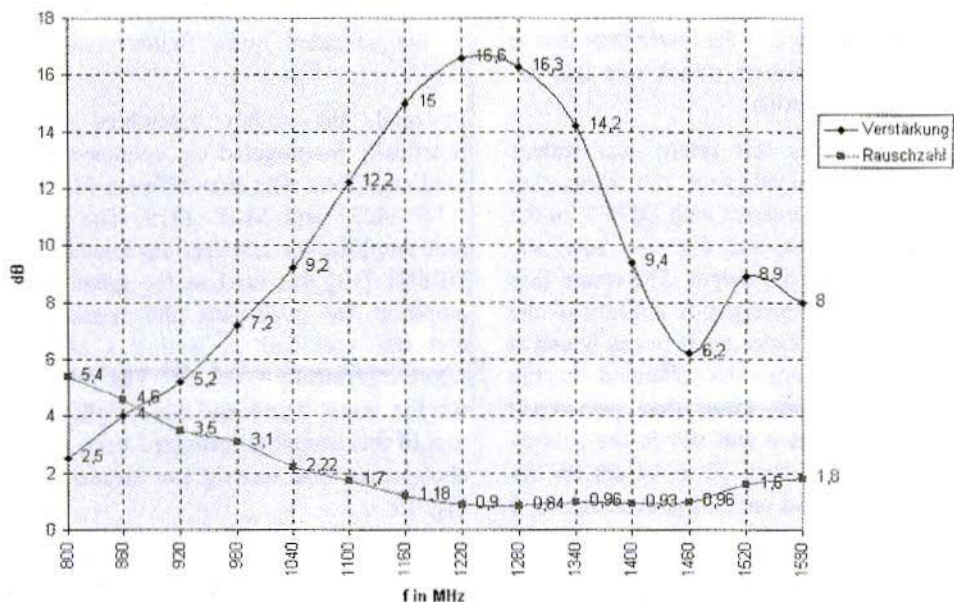


Fig. 14: Comparison of return loss levels obtained in output



**Fig. 15: 1.3 GHz amplifier according to DB6NT**

**Verstärkung = Amplification, Rauschzahl = Noise figure**

## 5. READINGS

In electronics development, the circuit simulation is the basis for short development times and also, with high-end software, a guarantee of a believable verification of the switching function.

Even if nowadays it is scarcely necessary to manufacture expensive prototypes, metrological investigation of what is initially only a simulated circuit still remains the touchstone for the success of the design.

Here it is now definitively clear whether what we have simulated also coincides with reality, for the well-known Mur-

phys law tells us that the slightest alteration can be the trigger for big changes.

Fig. 11 plots the noise figure and the amplification against the frequency for the amplifier on the Teflon base material (Rogers 0.79 mm.) and Fig. 12 shows the measured results for the same amplifier on epoxy material (FR4 1.5 mm.).

The measurements were carried out using an automatic noise test rig (HP 8970) and a noise source with a high return loss (HP 346A, 5 dB ENR). The accuracy of measurement of this layout is  $\pm 0.15$  dB for the values measured.

The noise values for the circuit assembled on Teflon are 0.3 dB lower than those for the epoxy assembly. The



amplification obtained is slightly greater but, as expected, 3 dB lower than that in the Puff simulation, which was laid out for line adaptation.

Fig. 13 plots the return loss values actually obtained with the measuring bridge in accordance with DJ7VY in the amplifier input, and Fig. 14 shows the return loss in the output. The return loss in the input corresponds entirely to the values to be expected between 6 and 8 dB. The return loss obtained in the output is rather lower than previously simulated. But it still lies in the acceptable range of from 10 to 14 dB for the assembly based on Teflon material, or 8 to 10 dB for the layout on the FR4 substrate. The capacitor, C5, in the output for fine tuning is omitted here, in order to test the broad-bandedness of the 1:4 transformation.

---

## 6. SUMMARY

---

Two amplifiers with source counter-coupling were assembled on different base materials. These were metrologically investigated, with regard to the attainable noise figure, the amplification and the return loss in the input and output.

If we compare the measurement results in this frequency range with other commercially available amplifiers, the Teflon-based assembly with its 0.6 dB noise figure is in the upper range. The assembly on epoxy material is more likely to be found in the lower range of the GaAs-FET amplifier. The reference

values used here were those determined in the so-called noise figure contests [15].

In total, the author assembled and practically investigated six counter-coupled amplifiers with two different FETs, ATF 36077 and MGF 4919. The no-tune amplifier for 1.3 GHz developed by DB6NT [13] was used as the reference amplifier for evaluating the readings, and was specified as having a noise figure of typically = 0.7 dB. The values for the noise figure and the amplification of this assembly measured using the above-mentioned test rig can be seen in Fig. 15.

The question still has to be clarified why the theoretical noise figure of 0.3 dB was missed by three-tenths or 100%.

For one thing, the quality of the input matching circuit is an essential factor for the attainable noise figure, with the relationship between the basic Q and the loaded quality giving the deflection.

The source counter-coupling determines the loaded quality by means of an increase in the effective series resistance in the input. The microstrip lines, with a discrete inductance, L3, including the source connection, are responsible for the basic Q.

With the epoxy assembly, as one might accept, the basic Q of the input circuit is already rather lower than that for the Teflon assembly, which leads in this case to a noise figure which is 0.3 dB higher. As regards the Teflon assembly, it remains to be stated that here too the circuits basic Q e.g. through inherent losses and radiation losses is still rather too low.





A further point is the tolerance of the measurement set-up. The measurement tolerance of  $\pm 0.15$  dB gives a range of 0.45 dB to 0.75 dB for a reading of 0.6 dB.

If we deduct an attenuation of approximately 0.1 dB as the transmission loss of the high-pass, then the actual noise figure of the test object is in the range between 0.35 and 0.65 dB.

If pre-amplifiers are used, in radio astronomy or in EME installations, only an assembly which operates with the lowest losses in the input is tolerable e.g. with cavity resonators. For other applications, though, good return loss in the input and good stability should always be preferred for a noise figure one or two tenths of a decibel better. There is a very good description in [11] of how the unavoidable mismatching loss on the aerial impedance leads to an increase in the system noise figure.

To sum up, we can say that an assembly on epoxy material, in spite of the noticeably lower basic Q due to the use of almost noiseless semi-conductors, can lead to amplifier noise figures of up to 0.8 dB. A few years ago, this was restricted to only special assemblies.

The low-loss amateur assemblies (e.g. with cavity input) in the L-band nowadays demonstrably attain noise figures of 0.3 0.4 dB [15].

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## 7.

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*Dipl.-Ing. J.v.Parpart*

# THE I/Q MODULATOR DESCRIBED

Digitalised image and sound signals have become an everyday occurrence in the last decade. Polished algorithms for data reduction make it possible to transfer more and more information in a restricted frequency band, with or without a wire. Many digital modulation processes require a so-called I/Q modulator as key element. What this element does is to control a carrier frequency as regards phase and, if necessary, amplitude, with each carrier condition or each condition change represented by a bit sequence.

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## 1. PRELIMINARY REMARKS

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In principle, the I/Q modulator is fully digitalisable. Its output signal can be predicted, depending on the data to be transferred, and can be generated through D/A conversion. Its all a question of technology.

However, the I/Q modulator presented

here stems from a time in which components were still wired. It operates as an analogue or a linear unit. The circuit, laid out using purely passive techniques, allows for an insight into what is going on. Various applications are discussed, starting from a basic circuit (Fig. 1), which can be easily built on the board shown in Figs. 2 and 3.

---

## 2. FROM DSB MODULATOR TO I/Q MODULATOR

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By multiplying an oscillator signal (frequency:  $f_0$ ) with a modulation signal (frequency:  $f_m$ ), we obtain two new signals with the frequencies  $f_0 - f_m$  and  $f_0 + f_m$ . If a complete base band signal is multiplied by an oscillator signal - the modulation process two side bands are obtained (hence DSB, for double side band). If a DC voltage is applied to the multiplier, this means that the carrier (otherwise suppressed) appears in the spectrum.

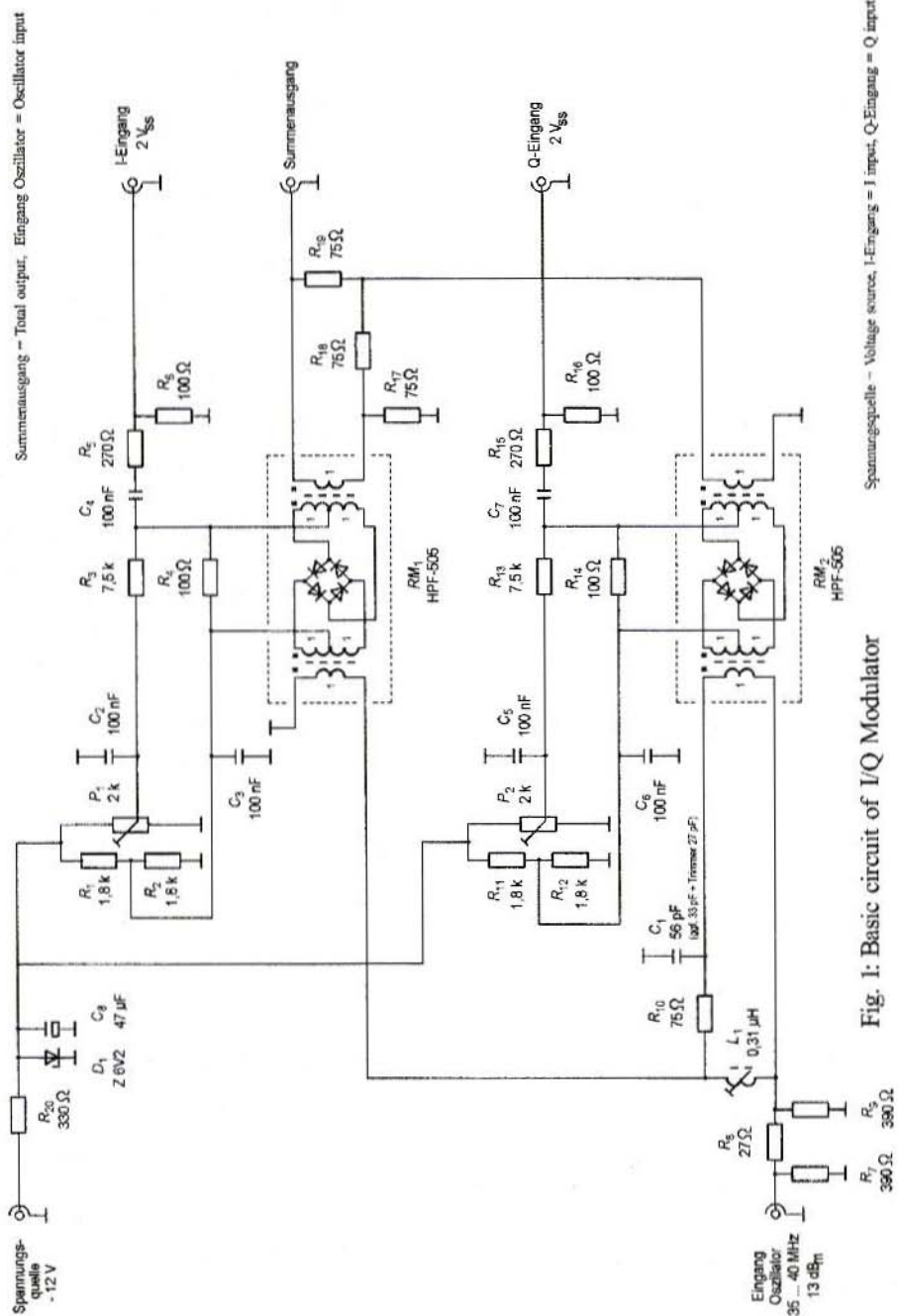


Fig. 1: Basic circuit of I/Q Modulator

In practise, multiplication is often carried out with the help of an electronic commutator switch (e.g. a diode ring mixer). The polarity of the modulation signal thus changes within the oscillator frequency cycle. Additional products are generated by means of this process, as well as the modulation products referred to above. They are grouped around the odd-numbered harmonics of the oscillator frequency and must be eliminated through filtration.

In comparison to the base band, the DSB signal takes up double the band width. But since frequencies are always scarce, it would be nice if we could double the use of the doubled band width. It is actually possible to separate two DSB signals again on the reception side, insofar as they have been generated with the same oscillator frequency and with a defined oscillator phase. In the ideal case, this gives us two transmission channels independent of each other.

From the mathematical point of view, the I/Q modulator consists of two multipliers which contain the modulation signals ( $I(t)$  and  $Q(t)$ ). There is  $90^\circ$  difference between the oscillator signals. By convention the leading oscillator signal is applied to the multiplier in the  $Q$  path. If we look at the  $I$  oscillator signal as a reference, then the  $Q$  oscillator signal is displaced by a quarter-period to the left on the time axis. In the vector representation, is rotated  $90^\circ$  anti-clockwise

The output signals from the multipliers are added and then transmitted. Fig. 4 shows the interaction of modulator and demodulator, on the assumption that the

multipliers have orthogonal phases in the  $I$  and  $Q$  channels, and that the transmission signal causes no signal delay (direct connection). It can be seen from the drawing that if the receiver oscillator is running synchronously with the transmitter oscillator ( $\phi = 0$ ), we get back  $I(t)$  and  $Q(t)$  through synchronous rectification and subsequent low-pass filtration (provided with a system-dependent proportionality factor).

There are four conditions altogether free from cross-talk, namely  $\phi = 0^\circ, 90^\circ, 180^\circ$  and  $270^\circ$ . So at the demodulator outputs there is either the original, a transposition with  $I$  inversion, a two-sided inversion, or a transposition with  $Q$  inversion. Mind you, the phase position must be kept to very precisely with respect to channel separation (e.g.  $\pm 1^\circ$  already causes a 35 dB cross-talk). Moreover, any deviation from the required orthogonality brings problems with it. Modulator errors and demodulator errors can compensate for each other or be cumulative, which makes this area very undefined. If we assume that we have an orthogonal demodulator, a  $1^\circ$  modulator error causes a channel cross-talk of 35 to 41 dB.

---

### 3. COMPLEX MODULATION

---

We are now making the assumption that the two modulation signals,  $I(t)$  and  $Q(t)$ , receive discrete amplitudes ( $n$  different values, multi-valent data transfer), and that the change from one value to the other takes place at fixed times. It is best if we represent this in vector



Fig. 2: Layout of I/Q modulator experimental printed circuit board

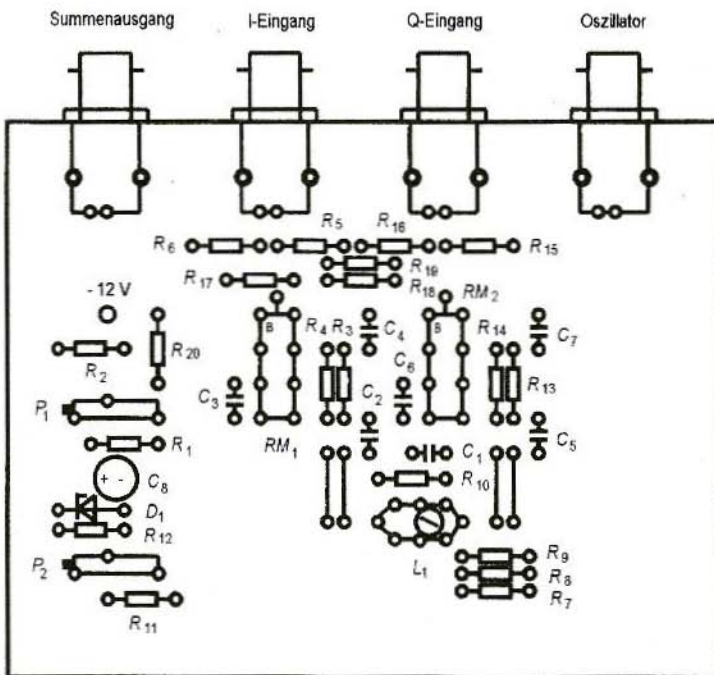
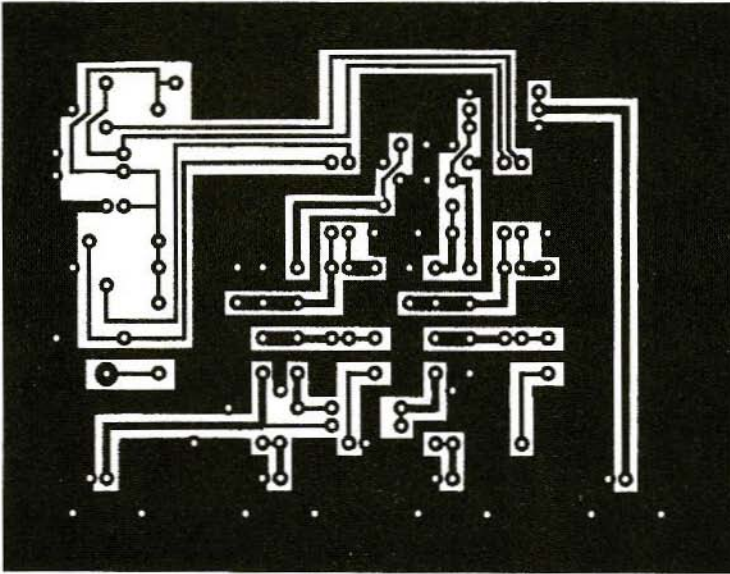


Fig. 3: Component plan of experimental printed circuit board

Summenausgang = Total output,  
 I-Eingang = I input,  
 Q-Eingang = Q input,  
 Oszillator = Oscillator

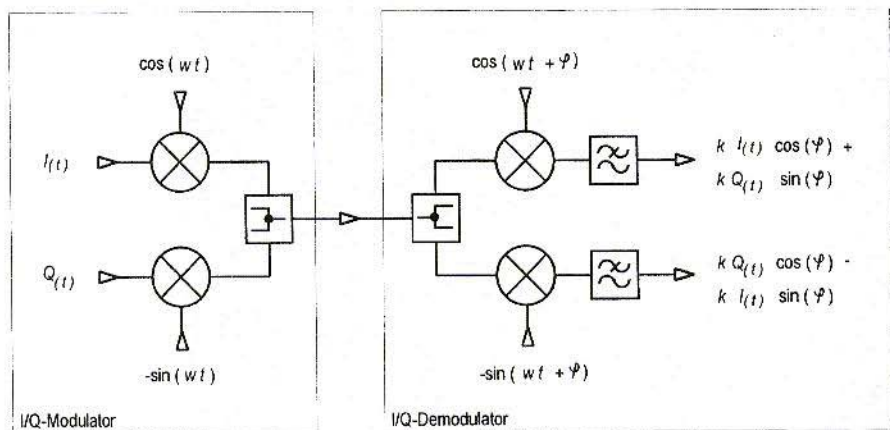


Fig. 4: Interaction of I/Q modulator and I/Q demodulator

terms. The  $I$  signal is plotted on the x axis and the  $Q$  signal on the y axis. Each  $I$ - $Q$  combination corresponds to a reading with a defined amplitude and phase (an oscilloscope can make that very clear in x-y representation).

One known representative of the complex modulation procedures is, for example, QAM (quadrature amplitude modulation). QAM is distinguished by equidistant amplitude values in the  $I$  and  $Q$  channels. All possible pairings are also allowed. With four different values per channel (e.g. 1 V, -0.33 V, +0.33 V, +1 V), we have 16-QAM, for there are 16 different combinations.

Also frequently encountered is PSK (phase shift keying). PSK sets the amplitude values in the  $I$  and  $Q$  channels in such a way that the total vectors always lie on a circle. With five different values per channel (e.g. 1 V, -

0.71 V, 0 V, +0.71 V, +1 V), we are dealing with 8-PSK, for we can then represent eight phase-differentiated readings with the same length (only specified  $I$ - $Q$  combinations are permitted here).

If any reading ( $Z_s = I(t) + jQ(t)$ ) is sent over the transmission system, in accordance with Fig. 4, the corresponding combination of the demodulator outputs gives

$$ZE = kZ_s e^{-j\phi}$$

In plain English, the reading received corresponds to the reading transmitted, but its phase angle is rotated through  $\phi$  in a clockwise direction. If  $\phi$  is sufficiently stable, the information transmission could take place in such a form that readings following each other in a





temporal sequence were compared with one another. It is advantageous if the carrier recovery circuit provides for  $\phi = 0^\circ, 90^\circ, 180^\circ$  or  $270^\circ$  (a Costas loop can do this), for then the amplitudes at the demodulator outputs take on only the pre-specified discrete values (no channel cross-talk), and can consequently be recognised using window comparator circuits without interference. It would be even better if there were only one scanning point, but the Costas loop does not cater for this. The uncertainty remains as to which quadrant we are in. Codewords can be agreed, or the difference can be evaluated for phase angles following one another.

Understandably, people try to transmit as much information as possible per unit of time. The finer the  $I$ - $Q$  amplitude sub-division is, the more data can be transmitted for a given bandwidth. But then a good signal/noise ratio is required, with linear behaviour from the components taking part in the transmission. We can allow ourselves 64-QAM in the cable television systems. In satellite technology, we do not wish to differentiate amplitudes from one another. We rely on robust 4-QAM = 4-PSK, through which we can obtain a high degree of efficiency for the transmitter high-level stage (constant level control).

#### 4.

### AN I/Q MODULATOR IN A PRACTICAL FORMAT

Following this theoretical introduction, let us now consider an I/Q modulator in

practise. The circuit can easily be fitted onto half a European standard size pc board. This is certainly very large by today's industrial standards, but against this we can get by with standard commercial components. Fig. 2 shows the layout (solder side) and Fig. 3 the components plan for the I/Q modulator experimental printed circuit board. The top face of the 1.5-mm. epoxy printed circuit board is not etched so that all earth connections can be through-hole plated. The remaining bores are counter-sunk. BNC angle sockets form the interfaces to the outside world.

The heart of the circuit consists of two 7-dBm Schottky diode ring mixers ( $RM1, RM2$ ). It is important that all mixer connections are accessible and not connected to earth at the manufacturing stage. Each mixer port is separated from the external inputs and outputs by means of additional attenuation. The system characteristic impedance is 75 Ohms.

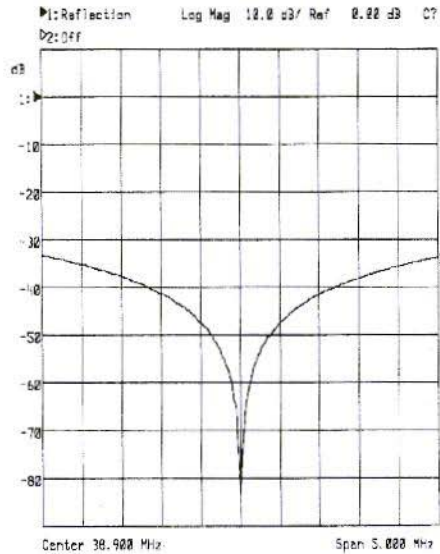
The oscillator frequency should lie in the range between 35 and 40 MHz. An external generator supplies the oscillator signal (13 dBm). A 3-dB attenuator ( $R7, R8, R9$ ) is followed by a 3-dB  $90^\circ$  distributor, de-coupled at the outputs, in a bridge circuit ( $R10, L1, C1$ ). One of the distributor outputs must be tapped off potential-free, which is provided for by the transformer integrated in  $RM2$ . The ring mixers thus each have an oscillator level of 7 dBm ( $RM2$  receives a signal running  $90^\circ$  ahead of  $RM1$ ), so that they see a broad-band matching, and are de-coupled from one another.

The reactive impedances of  $L1$  and  $C1$ , together with the resistance  $R10$ , must correspond to the system characteristic

impedance. Moreover, the source (oscillator) must be broad-band matched.  $L1$  takes the form of a balcable inductance (11 turns of enamelled copper wire on a 4-mm. rod, ferrite core).  $C1$  should likewise be balcable, if applicable (trimmer). The criterion for the correct values of  $L1$  and  $C1$  is a sharp form of reflection attenuation at the distributor input at the desired oscillator frequency (Fig. 5), with the minimum being of no interest as an actual reading (because of the final directivity of the measurement system) but being of interest only for reasons of reproducibility.

The mixers output signals are combined by a bridge circuit ( $R17$ ,  $R18$ ,  $R19$ ). In this connection, an attenuation of 6 dB arises in each case. One of the coupler inputs must be switched potential-free, which is provided for by the transformer integrated into  $RM1$ . The condition for the de-coupling is that  $R17$ ,  $R18$  and  $R19$  correspond to the system characteristic impedance, and in addition the sink (total output) must be broad-band matched.

A stabilised DC circuit of approximately 6 V is obtained from the voltage fed in from outside, using  $R20$ ,  $D1$  and  $C8$ . A DC voltage adjustable within a range of 0 mV to  $\pm 20$  mV is applied at the mixer inputs through the network consisting of  $R1$ ,  $R2$ ,  $R3$  and  $P1$  (or  $R11$ ,  $R12$ ,  $R13$  and  $P2$ ).  $P1$  and  $P2$  take the form of 10-turn helical potentiometers. We can thus generate components of the carrier frequency orthogonal to one another at the total output, each with up to approximately 30 dBm. Without DC voltage compensation (0 mV at the mixer inputs), the carrier frequency



**Fig. 5: Reflection attenuation at input of 3-dB 90° distributor**

appears at the total output at approximately 60 dBm, conditioned by the couplings and asymmetries of the mixer and the peripheral circuit. In accordance with the laws of vector addition, we can make the vector to disappear completely, by a suitable adjustment of  $P1$  and  $P2$  (more and more fascinating!).

The modulation signals,  $I(t)$  and  $Q(t)$ , (up to approximately 10 MHz) find their way to the mixer inputs through 18-dB attenuators ( $R4$ ,  $R5$ ,  $R6$  or  $R14$ ,  $R15$ ,  $R16$ ). They are AC coupled through  $C3$ ,  $C4$  or  $C6$ ,  $C7$ . The lower limiting frequency is approximately 20 kHz. If we now apply 2 V<sub>ss</sub> (sinusoidal) either to the  $I$  input or the  $Q$  input, we obtain two sidebands at the total output, each with approximately 20 dBm.

If we feed the modulation signal to both inputs simultaneously, and if we use

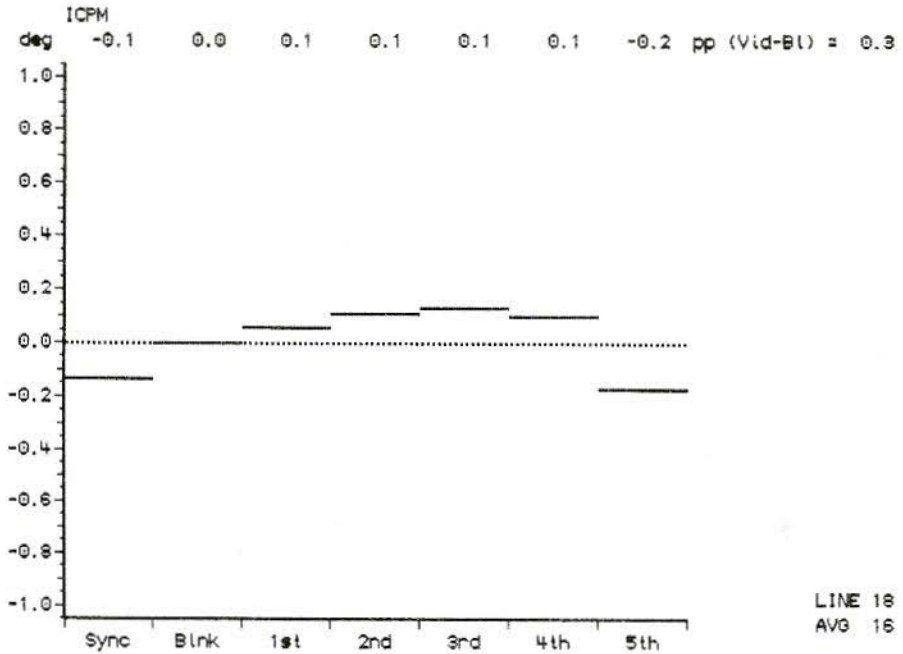


Fig. 6: I/Q modulator as ICPM-free image modulator

orthogonal signals for this, then one sideband can be strongly attenuated (theoretically obliterated). The remaining sideband is increased by 6 dB. Since the circuit is passive, the process can also be reversed. A signal ( $f_s$ ) fed into the total output generates two orthogonal signals with  $f_0 - f_s$  at the  $I$  and  $Q$  inputs (after subsequent low-pass filtration). If  $f_0 > f_s$ , then the  $Q$  signal will take precedence. If  $f_s > f_0$ , it is the other way round. Single-sideband technology has been making use of the options arising from this (including the combi-

nation possibilities) for decades.

As regards data transmission, since the modulator operates in a very linear manner, the  $I$  and  $Q$  modulation signals can be reliably divided into ten or more equal stages in the range between  $-1$  V and  $+1$  V (whether we can also actually make use of this relatively fine quantisation depends on other conditions see above). The filtration of the modulation signals required for band limitation can take place in the basis band, on the basis of the given linearity. Initially, no static voltages are transmitted (no DC cou-



pling). We shall come back to this.

### Parts list:

1 x 27 Ohms, 1%, *R8*  
4 x 75 Ohms, 1%, *R10, R17, R18, R19*  
4 x 100 Ohms, 1%, *R4, R6, R14, R16*  
2 x 270 Ohms, 1%, *R5, R15*  
1 x 330 Ohms, 5%, 1/3 W, *R20*  
2 x 390 Ohms, 1%, *R7, R9*  
4 x 1.8 kOhms, 5%, *R1, R2, R11, R12*  
2 x 7.5 kOhms, 5%, *R3, R13*  
2 x 2 kOhms, spindle trimmer, *P1, P2*  
2 x HPF 505 (mini-circuits), *RM1, RM2*  
Or similar, VHF mixer  
1 x 56 pF (if applicable, trimmer), *C1*  
6 x 100 nF, ceramic, *C2 to C7*  
1 x 47  $\mu$ F, electrolytic capacitor, *C8*  
1 x 0.3  $\mu$ H, 11 winding on 4-mm. core, *L1*  
Vogt kit with balancing core  
1 x ZPD 6V2, Zener diode, *D1*

---

## 5. AN EXCURSION INTO VIDEO TECHNOLOGY

---

Video signals can be converted into the intermediate-frequency plane with the I/Q modulator shown here. The following circuit change is required for negative modulation (e.g. standard B/G). *R3* changes from 7.5 kOhms to 1.5 kOhms,

*C3* and *C4* remain short-circuited, *R1* is removed.

The 90° coupler (*L1, C1*) is tuned to 38.9 MHz. The video signal must be aquired at the black level (0 V), and comes to the *I* input. With *P1*, the residual carrier can be set as desired. With a 10% residual carrier, the synchronous level at the total output is approximately 14 dBm.

What are the *Q* channel and *P2* for? Well, because of inter-carrier sound demodulation, the modulator may give off only a flawless *I* signal. Any *Q* remains (and there are always some) cause a phase modulation of the image carrier. The I/Q modulator compensates for this, by generating an inverse *Q* carrier (DC voltage at *RM2*).

Professional television test demodulators can measure the image carrier phase modification (ICPM). Fig. 6 shows the precision of measurement to be described as perfect. It should be pointed out that the tuning of the 90° coupler is not critical in this application case. The compensation is outstandingly temperature-stable. We just have to pay careful attention to the DC voltages.

---

## 6. SIMULATION OF A MULTI-VALENT DATA TRANSMISSION

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In this experiment, the *I* and *Q* channels are to be loaded with data. For simulation purposes, non-synchronous, stepped video signals are fed into the modulator

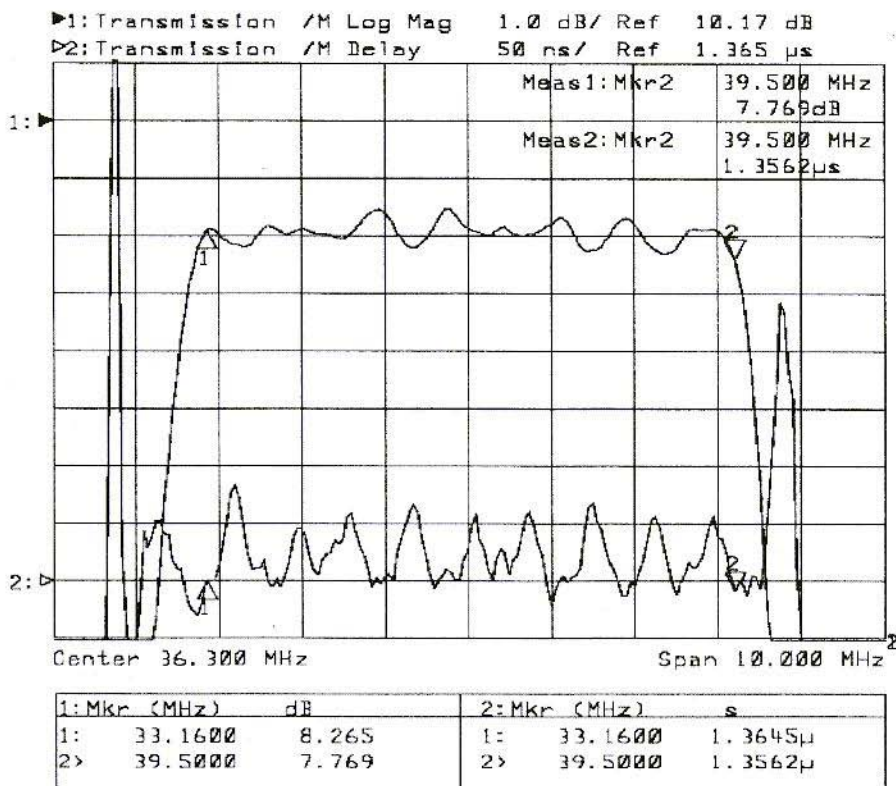


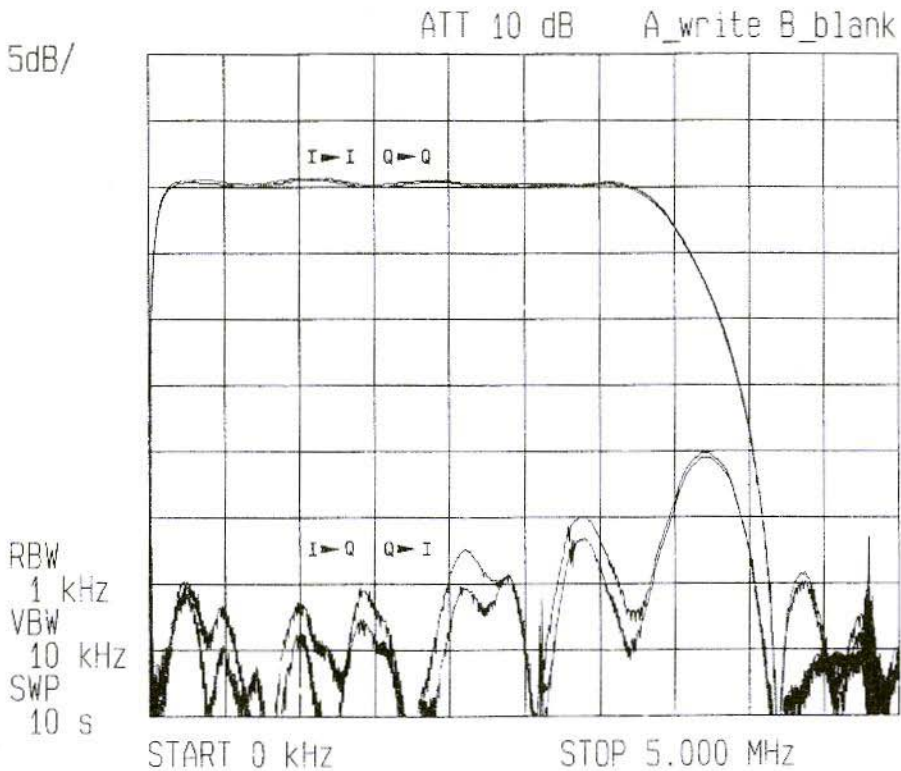
Fig. 7: Transmission curve and group delay of filter

inputs (DC coupling). The stages correspond to the voltage values 0.3 V, 0 V, + 0.14 V, + 0.28 V, + 0.42 V, + 0.56 V and + 0.7 V. Considered vectorially, this gives at least 49 possible combinations, of which only the first quadrant is completely occupied.

The I/Q modulator is driven by DC coupling i.e. C3, C4, C6 and C7 are short-circuited. An external voltage source is dispensed with (if the position of the zero point in the receiver were critical, a bipolar adjustable DC voltage would have to be fed to the mixers through R3 and R13).

An advantageously priced residual side-band filter, with pre-amplification and post-amplification, forms the transmission channel. Its transmission curve (approximately 0.7 dBpp) and group delay (approximately 50 nspp) are logged in Fig. 7. So that the modulator signal matches into the filter, the oscillator frequency is pushed into the centre of the band (here: 36.3 MHz). L1 and C1 are carefully balanced.

Now we still need a suitable demodulator. The author used a circuit with the same structure as that in the modulator (Fig. 4 suggests this). The oscillator

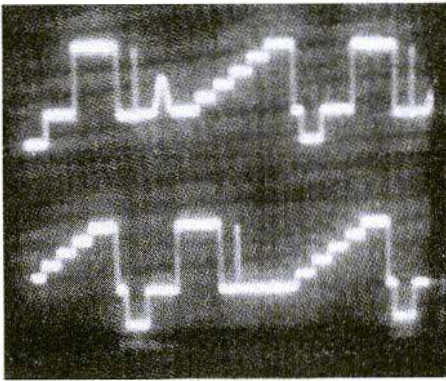


**Fig. 8:** Transmission curves  $IS \rightarrow IE$  and / or  $QS \rightarrow QE$  and cross-talk  $IS \rightarrow QE \rightarrow IE$

signals for the modulator and demodulator came from the same source. One of the branches was phase-adjustable, using a variable-length line (a so-called trombone).

The amplitude cycle of the total system consisting of the modulator, the filter and the demodulator was tested first (Fig. 8). The transmission curves  $IS \rightarrow IE$  and  $QS \rightarrow QE$  are quasi-identical the transmission range goes up to approximately 3.5 MHz. The cross-talk attenuation levels,  $IS \rightarrow QE$  and  $QS \rightarrow IE$  are extremely similar (more than 25 dB to approximately 3 MHz).

The demodulated signals can be looked at with the oscilloscope after low-pass filtration with  $f_g \approx 20$  MHz. Fig. 9 shows both channels (the photo quality is unfortunately not optimal). The DC voltage values are correctly transmitted, i.e. the black value remains at 0 V. However, the channel cross-talk expands the original clear contours. An x-y representation is very informative. Since the modulation signals, in terms of time, move over each other, new  $I$ - $Q$  combinations are arising all the time. Some of these random constellations are shown in Fig. 10. From the purely qualitative



**Fig. 9: Oscilloscope trace of demodulated signals**

point of view, it can be stated that there should be no great difficulty in finding a reliable answer to the question of exactly what complex numerical value is being transmitted at any moment.

## 7.

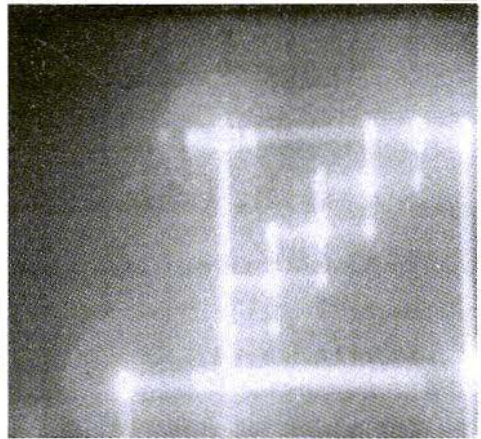
### CONCLUSION

The transmission of digital data in the high-frequency plane is a new challenge for high-frequency technology. The I/Q modulator often forms the interface between digital and analogue technology. This article is intended to awake an understanding of the problems involved and to provide stimulation for further observations and experiments.

## 8.

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**Fig. 10: X-y representation of demodulated signals**

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Wolfgang Schneider, DJ 8 ES

# LOGARITHMIC AMPLIFIER UP TO 500 MHZ WITH AD8307

As a supplement to a good stock of measuring equipment, a logarithmic amplifier is described below with a useful dynamic range of 90 dB. Suitable outputs are provided so that the reading can be displayed, for example, on an oscilloscope. The signal to be measured can likewise be fed out over the serial 3-conductor bus by means of an optional A/D converter for further processing using a micro-controller or a PC.

---

## 1.

### GENERAL

---

Wobble measurements, for example, using the short-wave synthesiser for the frequency range 1 to 65 MHz [1] need a logarithmic display of the reading with as extensive a dynamic range as possible. Here the AD 8307 logarithmic amplifier from Analog Devices represents an almost ideal component, as a detector with dB linear output. A glance at the specimen circuit shows how few components are required (Fig. 1).

---

## 2.

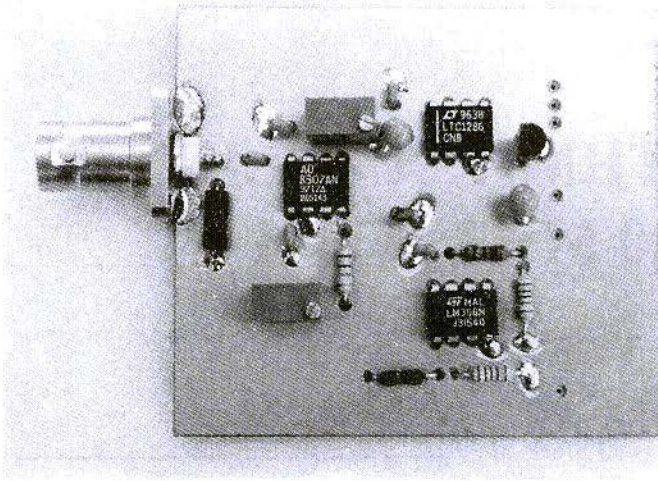
### CIRCUIT DESCRIPTION

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The core of the circuit (Fig. 2) is the AD 8307 logarithmic amplifier from Analog Devices. In the data sheet, the IC is specified as up to 500 MHz, with a dynamic range of approximately 90 dB. This is valid for input levels of +17 dBm to 75 dBm. The AD8307 can even be used (with restricted accuracy) up to 900 MHz and beyond.

As the frequency increases, the display voltage decreases slightly. The associated characteristics are shown as the frequency response of the AD8307 in Fig. 3. In the interesting frequency range between 1 and 65 MHz, the deviation is negligently slight. Compensation might be necessary, if applicable, with correspondingly broad-band measurement applications.

The dB linear output (pin 4) has a gradient of 20 mV/dB in the given circuit and is high impedance. A type



**Fig. 1:**  
**Prototype of**  
**logarithmic**  
**amplifier up to**  
**500 MHz**

LM358 double OP is downstream as a buffer stage. Thanks to the two different feedback resistances, this makes loadable outputs available with gradients of

50 mV/dB or 100 mV/dB.

The linearity and the offset (in relation to the input power) are set with two 50-Ohm spindle trimmers (see also Section 4: Putting into practise). Table 1 shows the output voltage, plotted against the input power following the balancing.

**Ausgangsspannung des**  
**logarithmischen Detektors**

	dbm	20mV/dB	50mV/dB	100mV/dB
E	10	2,00	5,01	10,01
i	0	1,80	4,50	9,00
n	-10	1,59	3,99	7,97
g	-20	1,39	3,48	6,96
a	-30	1,19	2,99	5,97
n	-40	0,99	2,48	4,96
g	-50	0,79	1,98	3,95
	-60	0,59	1,49	2,97
	-70	0,40	1,00	2,00
	-80	0,24	0,60	1,20

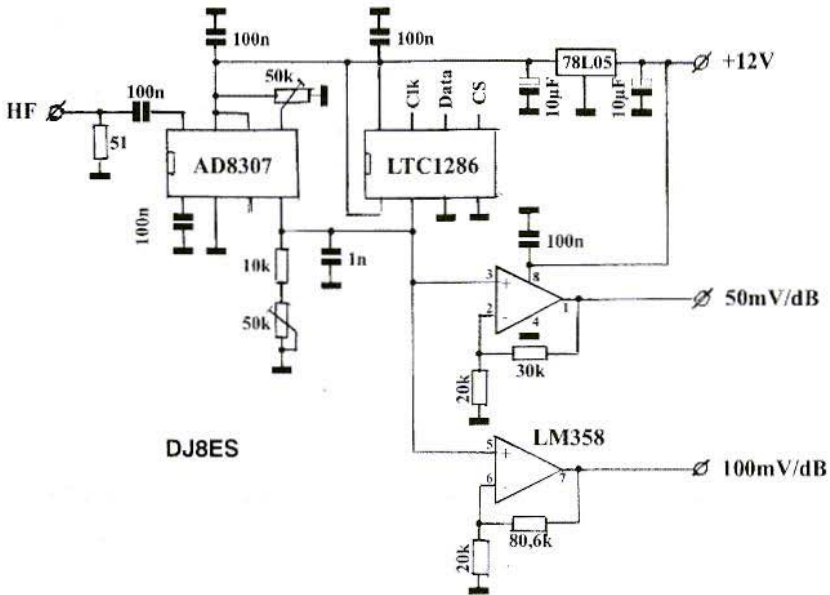
**Table 1: Output voltage plotted**  
**against input power**

**Ausgangsspannung des logarith-**  
**mischen Detektors = Output voltage of**  
**logarithmic detector, Eingang = Input**

As an extension, space was provided on the printed circuit board for a 12-bit type LTC 1286 A/D converter manufactured by Linear Technologies. This makes the signal to be measured available (through a serial 3-conductor bus) for further processing using a micro-controller or PC.

### 3. ASSEMBLY INSTRUCTIONS

The circuit for the logarithmic amplifier for up to 500 MHz (DJ8ES047) is assembled on a double-sided copper-coated epoxy printed circuit board (Fig.



**Fig. 2: Circuit for logarithmic amplifier with AD8307**

4) with dimensions of 54 mm. x 72 mm.. It should be housed in a suitable timplate housing.

A BNC socket is provided for the measurement input. The voltage of + 12 V is fed through a 1-nF feedthrough capacitor. Teflon bushings are used for all outputs.

First, all the resistors and capacitors and the voltage regulator are inserted, in

accordance with the component plan (Fig. 5). Before the ICs AD 8307 and, if applicable, LTC 1286 are soldered in, their supply voltage of + 5 V should be tested at the output of the 78L05 fixed voltage regulator.

Because of the high maximum operating frequency, which exceeds 500 MHz, the AD8307 had to be soldered directly onto the printed circuit board i.e. without a socket.



### 3.1. Component list

1 x DJ8ES 047 printed circuit board  
1 x tinplate housing 55.5 mm. x 74 mm.  
x 30 mm.  
1 x AD8307 logarithmic amplifier, DIL  
8-pole  
1 x LM 358 operational amplifier, DIL  
8-pole  
1 x LTC 1286 A/D converter, DIL  
(optional)  
1 x 78L05 fixed voltage regulator  
2 x 10  $\mu$ F/35 V, tantalum electrolytic  
capacitor  
2 x 50-kOhm vertical spindle trimmer

1 x 1-nF Df capacitor, solderable  
1 x 4 mm. soldering lug, turned down  
5 x Teflon bushings  
5 x 100 nF, RM 2.5 mm., ceramic  
1 x 1 nF, RM 2.5 mm., ceramic  
**¼ W resistances, RM 10 mm.:**  
1 x 51 Ohms  
1 x 10 kOhms  
2 x 20 kOhms  
1 x 30 kOhms  
1 x 80.6 kOhms

## AD8307 - logarithmischer Detektor

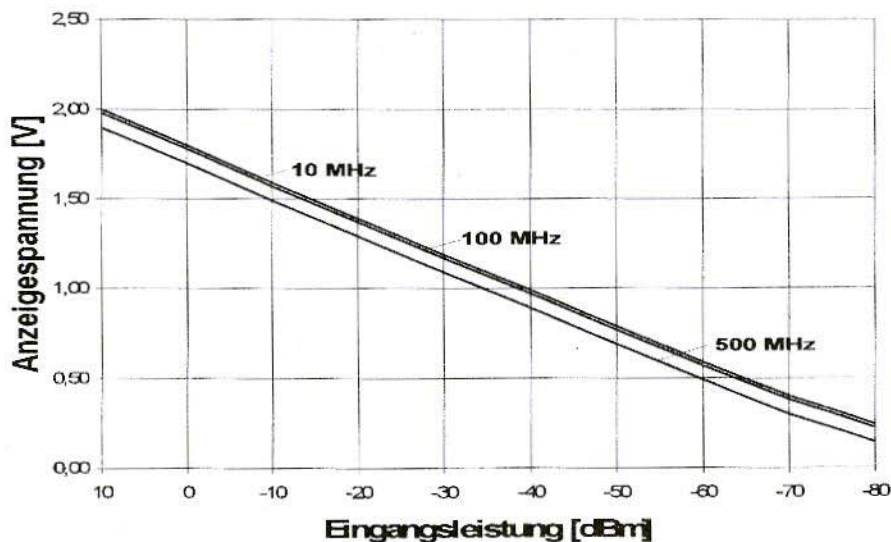


Fig. 3: Frequency response for AD8307

Anzeigespannung = Display voltage, Eingangsleistung = Input power

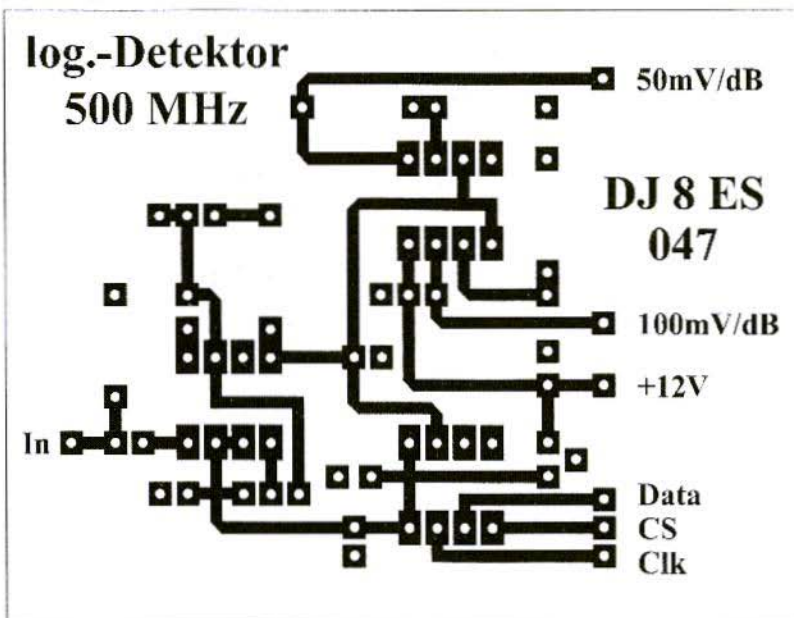


Fig. 4: DJ8ES 047 printed circuit board layout

#### 4. PUTTING INTO OPERATION

Feed in the + 12 V DC power supply and an input signal of e.g. 0 dBm (1 mW). You should already be able to obtain a reading from a measuring instrument connected to the two outputs.

Use the 50-Ohm trimmer (through 10 kOhms to pin 4 of AD8307) to set the linearity of the entire system. The voltage differential between any test signal and a signal weakened by 10 dB must be exactly 1 V.

The simplest way of balancing this is with a switchable attenuator. During the balancing procedure, the absolute value

of the measurement voltage displayed is of no importance. The only thing relevant is the differential of 1 V to be established between the two measurement levels.

Finally, adjust the input signal offset. To do this, feed the test signal into the input at a level of 0 dBm and use the 50-kOhm trimmer at pin 5 of AD8307 to set 100 mV/dB output to precisely 9.0 V.

That completes the balancing. The logarithmic detector now measures outputs between + 10 dBm and 80 dBm in accordance with Table 1 at the beginning of the article. A reading of + 10 dBm (10 mW) thus corresponds to a voltage of 10 V at the 100 mV/dB output or, to put it another way: 1 V per 10 dB.

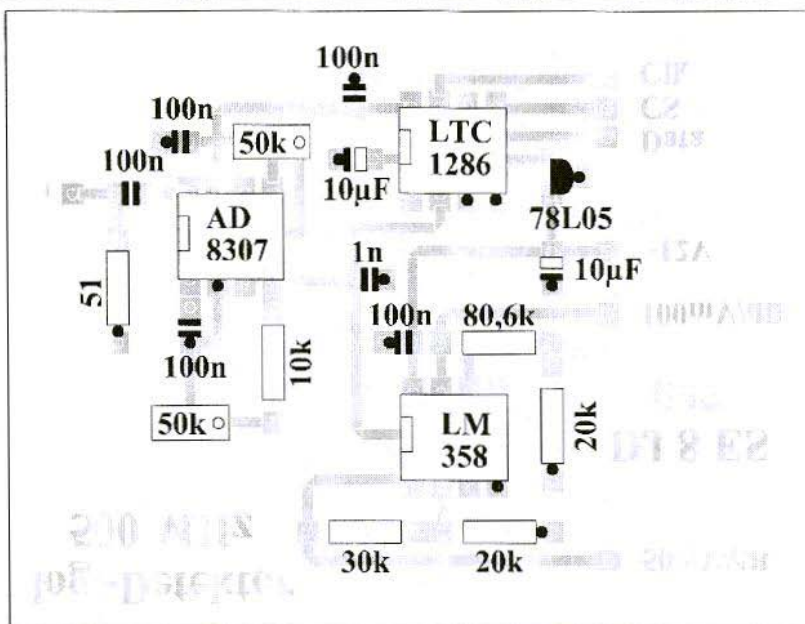


Fig. 5: Component plan for DJ8ES 047 assembly

In wobble measurements with the short-wave synthesiser, the Y input of an oscilloscope is connected up at the same point. Set the oscilloscope to 1 V/cm. for the purpose. The synthesiser supplies the X deflecting voltage. The settings and measurement structure required have already been discussed in [1].

## 5.

### LITERATURE REFERENCES

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1-65 MHz short-wave synthesiser with DDS

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Looking forward to next year I am hopeful that the price of the magazine will stay the same. This depends quite a lot on the circulation, so encourage your friends to subscribe. Two things that will happen is are a review of the price charged for postage, particularly to reduce the air mail rate, and a change to a single price for credit card and cash.

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Andy - G8ATD

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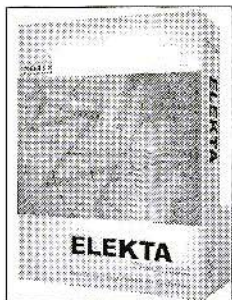
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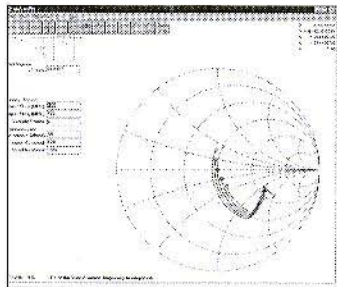
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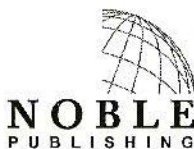
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