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Contents

Wolfgang Schneider DJ8ES	Voltage Converters - 12 / 24V or Above	66 - 71
Dipl.-Ing. Detlef Burchard	Active Reception Antennas: Observations, Calculations and Experiments	72 - 89
Richard A. Formato K1POO	Improved Feed for the Off-Centre Fed Dipole	90 - 93
Denys Roussel F61WF	An Ultra Low-Cost HF SSB/CW Transceiver with 20W Output, an AGC Meter, S-Meter and Audio Filters	94 - 110
Dr. Hans Sapotta DF8FR	Push-Pull receiver Front End Stages in Common Base Circuits. The Solution to the Inter-Modulation Problem	111 - 121
Norbert Kohns	IF Amplifier with Wide Range of Adjustment	122 - 125



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Wolfgang Schneider, DJ8ES

Voltage Converters - 12 / 24V or Above

Whether as the tuning voltage for capacitance diodes or else for the coax relays available at attractive prices in flea markets, now and then we need voltages higher than the normal 12 Volts.

The following article lists various options for DC Converters operating from 12 to 24 or 30 Volts, or even higher. These are simple and economical circuits for the radio amateur.

1. 12 / 24 V VOLTAGE DOUBLER

In the first example, we're going to take a closer look at a conventional circuit (Fig.2) for a 12 / 24V voltage doubler.

An oscillator and driver stage for the high-level stage are constructed using

complementary metal oxide semiconductor 4069 gates. The square wave voltage thus generated is rectified using two Schottky power diodes (low voltage losses and very low capacitance) and added to the input voltage. The clock frequency is app. 30 kHz. The circuit shown is best suited to, for example, a 24V coax relay.

2. TCA 720 VOLTAGE CONVERTER

The TCA 720 integrated voltage converter has a self-blocking oscillator circuit and a temperature-compensated circuit for constant voltage control. The actual frequency of the self-blocking oscillator is determined by the inductance of the external coil and the supply voltage.

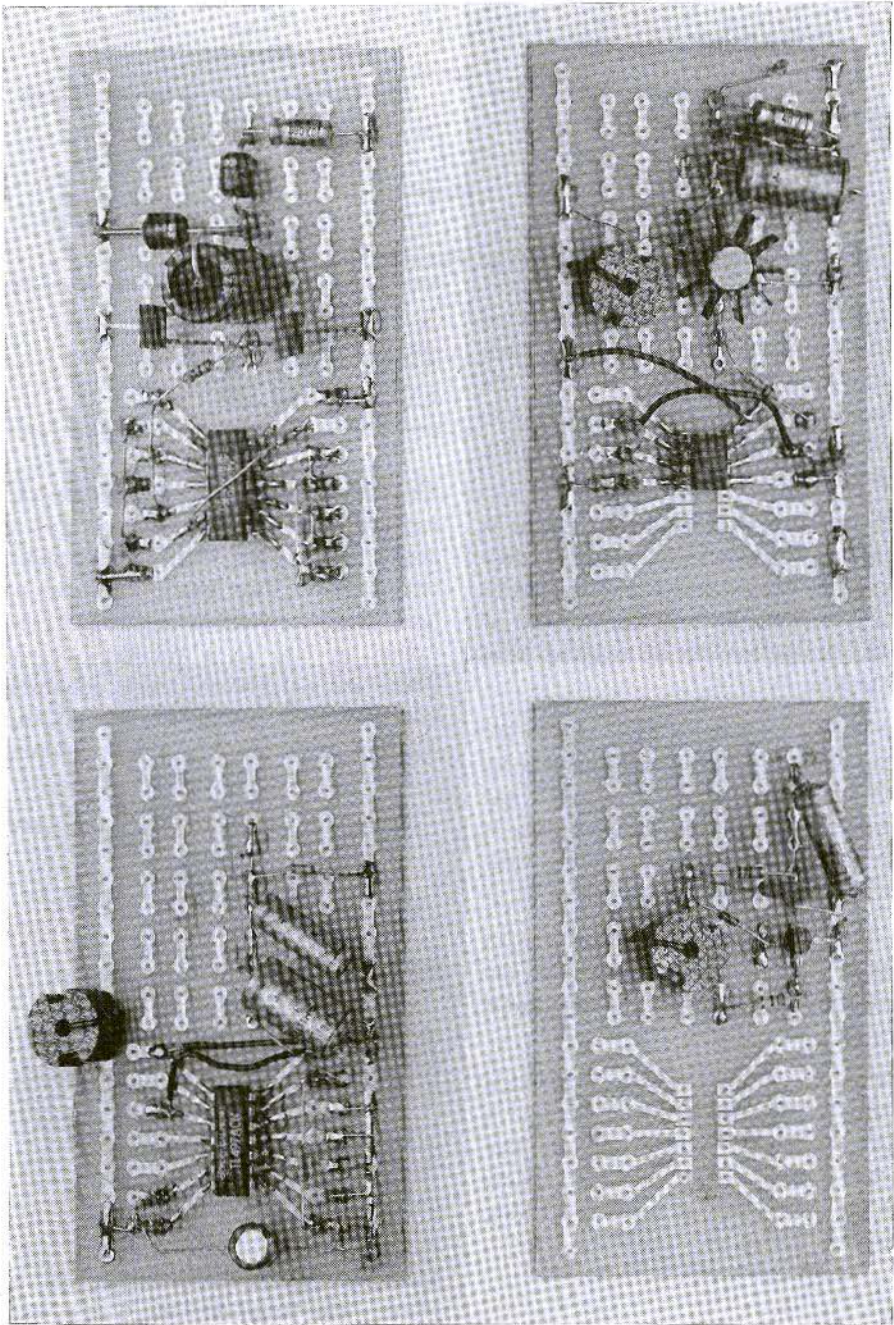


Fig.1: Various Voltage Converters as Test Rigs

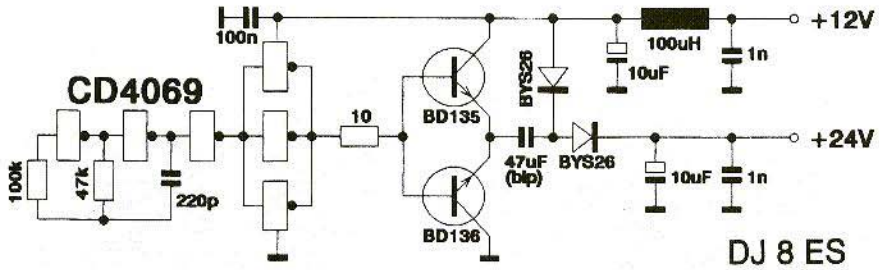


Fig.2: Conventional 12 / 24V Voltage Doubler

It is important to have a high factor of quality for the inductance, L. If a shell core is used with $AL = 250$, the necessary 5mH is obtained with 150 turns of enamelled copper wire (CuL), diameter 0.1mm.

Unfortunately, this integrated circuit is no longer manufactured, although they can still be obtained here and there. And as it is a relatively well known type of circuit, it can't really be omitted from this article.

The wiring diagram in Fig.3 shows the normal circuit for standard applications. With an output voltage of 30 - 35V and a current, $I_A = 1mA$, the assembly is absolutely pre-destined for tuning voltages - e.g. for capacitance diodes.

Fig.4 shows the quality of the output voltage. All residual noise fractions can be suppressed by means of additional filtering.

In open circuit, the current consumption is only 2.5mA. Even under load (1 - 2mA), only app. 10mA is required, due to the comparatively high efficiency, with the output voltage measured falling by a good 1 Volt.

3. TL497A SWITCHING CONTROLLER

The entire circuit for a universal switching controller can be comprised on a single chip. The TL497A integrated circuit from Texas Instruments can be used as an upward or downward controller, or else as a voltage inverter. There are no problems with obtaining this integrated circuit at present. The TL497A requires only a few external

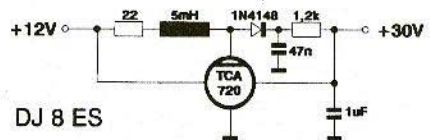


Fig.3: Voltage Converter using a TCA 720

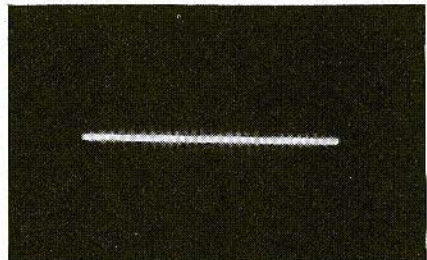
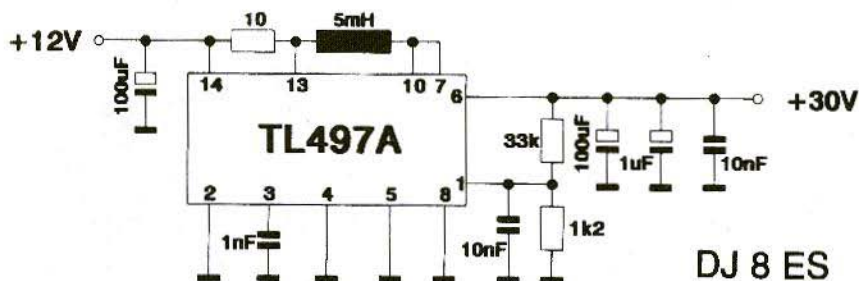


Fig.4: The Residual Noise Fractions can be seen on the Scope



DJ 8 ES

Fig.5: TL 497A Switching Controller with External Components

components, which must be dimensioned for the output voltage desired (Fig.5). The output voltage is stabilised, which effectively limits the current at the same time.

The switching controller integrated circuit supplies 500mA max. output current. An additional switching transistor can be provided for higher current levels. The output voltage level depends on the relationship of the two resistances, R2 to R3. As a first approximation, the 33k resistance, R2, can be expressed by:

$$R2 = (V_{out} - 1.2) \text{ k}\Omega$$

Here the full manufacturing tolerances are naturally taken into account. Measured resistances, or resistances with close tolerances, should be used to give a precise setting here. On the test rig,

using the equipment referred to, the output voltage is app. 34V.

In open circuit, the current consumption of the TL497A amounts to only 12mA. Naturally, when a load is applied there is a corresponding rise in the current level.

4. VOLTAGE TRANSFORMER WITH STORAGE CHOKE

Fig.6 shows a combination of integrated and conventional systems. This voltage transformer with a storage choke has already been used in FM-ATV transmitters for the 23cm band [3].

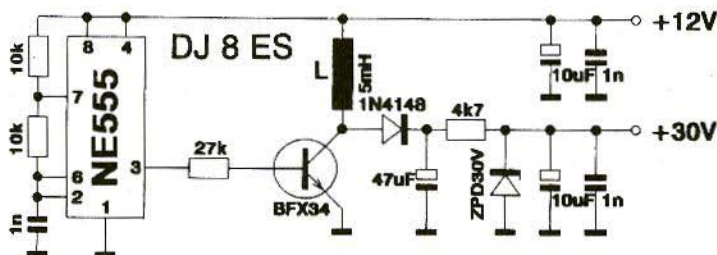


Fig.6: Voltage Converter with Storage Choke



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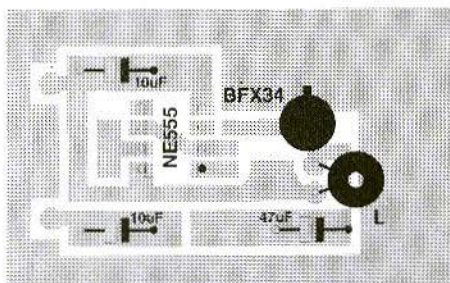
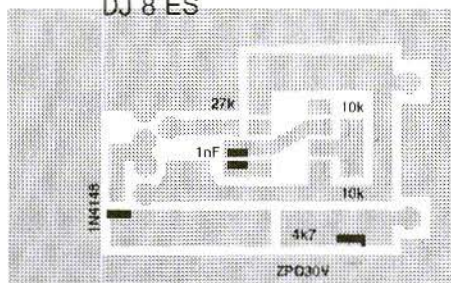


Fig.7: The Components are Mounted on Both Sides

The oscillator with the NE 555 timer integrated circuit oscillates at app. 40 kHz. The frequency in this case is determined by the resistance / capacitance combination consisting of the 1 nF capacitor and the two 10k resistances. The integrated circuit drives the BFX34 switching transistor, in the collector circuit of which the storage choke, L, generates a multiple of the working voltage.

A "high voltage" obtained in this way still has to be rectified and filtered. The subsequent ZPD 30V Zener diode stabilises the voltage and limits it to the desired 30V at the output.

The circuit above has a maximum output voltage of 80V (US measured at the storage choke). The base protective resistor (27k) must be optimally dimensioned for the transistor used here.

The downstream voltage stabilisation supplies the 30V required at the output in this example. For other output voltages, Zener diodes and protective resistors should be selected accordingly. The current consumption in open circuit or under load corresponds to that for the previous circuits referred to.

5. PRACTICAL IMPLEMENTATION

Naturally, all the voltage transformers previously described entirely fulfil their purpose. There are, of course, certain differences, depending on the individual application. In the final analysis, though, the decision on the optimum circuit really comes from a gut feeling: the voltage transformer with storage choke (see Section 4).

The small, double-sided board has dimensions of 34mm x 54mm and so it can be mounted in a standard tinplate housing. The assembly is thus given a screened housing, avoiding any problems with perturbing radiation.

The combination of SMD and conventional circuit technology pays equal attention to how easily components can be obtained and to making assembly as simple as possible. Fig.7 shows the layout of the components on the front and rear faces of the board. Special assembly instructions are not in the least necessary for this clear circuit.



5.1. Component list

- 1 x NE555 (NEC)
- 1 x BFX34 (Siemens)
- 1 x AL = 250 shell core (Siemens)
(for winding data, see text)
- 2 x 10 μ F, 63V tantalum
- 1 x 47 μ F, 100V tantalum
- 2 x 1 μ F feedthrough
- 1 x tinplate housing
(37 x 55.5 x 30 mm)
- 1 x double-sided board, 34 x 54mm

And in SMD format:

- 1 x 1N4148 diode
- 1 x ZPD30V Zener diode
- 1 x 1 μ F ceramic
- 1 x 4.7k
- 2 x 10k
- 1 x 27k

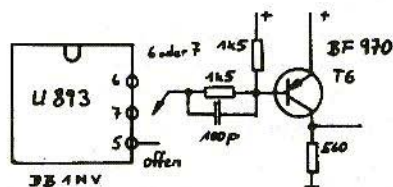
6.

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Extensions / Improvements

THE DB1NV SPECTRUM ANALYSER ISSUE 2, 1989



oder = or; offen = open

A :64 divider, type SDA 4211 or SDA 4212, is mentioned in the local oscillator/phase locked loop assembly, DB1NV 007, in the wiring diagram and the components diagram. This divider module is either difficult or simply impossible to obtain.

The replacement we come across most frequently is a pin-compatible type known as U893. However, for this to work satisfactorily, a small circuit modification is required, as shown in Fig.1.

The essential difference is the other equilibrium level at the divider output, which can be eliminated by using a voltage divider.



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Active Reception Antennas: Observations, Calculations and Experiments

The idea of combining antennas with active components has existed for about 60 years, but it was only 30 years ago that it began to be introduced into technical reality. The capacitive broad-band antenna did not become possible until depletion layer field effect transistors were available, which was at the beginning of the seventies. Since then, extensive antenna equipment has no longer been required for the reception of the extremely long wave, long wave, medium wave and short wave ranges.

There is a second important application for active antennas in measuring techniques. They can be dimensioned in such a way that they have a constant conversion factor between the output voltage and the field strength over a wide frequency range. They are thus ideal for use in field strength measurements.

It was this second application which stimulated me to assemble my first

active antenna in 1974. There were noise field strength measurements to be taken in relation to the preparations for Central Office for Telecommunications Technology inspection tests, which at that time were still carried out in the open air, using adjustable-length dipoles, which were mounted on rotatable masts with adjustable heights. The field strength conversion factor was read off from tables or nomographs, which gave ample opportunity for false readings or errors.

The circuit developed at that time is reproduced almost unchanged here. Even at that time, it was distinguished by the fact that the input capacity was infinitesimally small and the source resistance, which the receiver sees, was pretty accurately 50Ω . Questions of low noise levels and low distortion were in the background in those days. But they have to be tackled if we are to try and obtain the best possible reception, as is expected from an active antenna today.

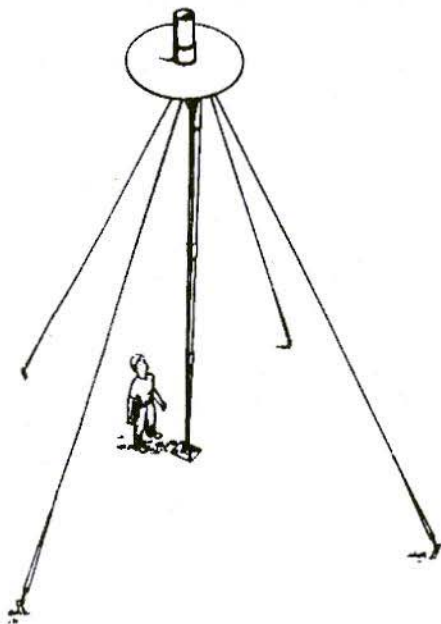


Fig.1: Standard Commercially Available Active Antenna from 1974 on a High Mast
Picture from Brochure (11)

1. INTRODUCTION

If we now return to the same topic 20 years later, it's a good idea to see how far technical progress has come over this period of time:

The theoretical principles behind active antennas had already been developed between 1960 and 1975, and were published in such well-thought of journals as the NTZ and the IEEE Transactions. I would prefer not to go through the mathematical work again, as good summaries are available - e.g. from

Meinke & Gundlach [7] and Rothammel & Krischke [8]. Jirmann [5] has already used Diagram 1 from the first book in his article. I shall follow this example and use other diagrams from Hütte [1] and from Meinke & Gundlach [7].

As the capacitive broad-band active antenna is set up vertically for omnidirectional reception, the question arises of whether this is actually correct - i.e. in other words, whether the polarisation of the incident radiation is really predominantly vertical. Mature and experienced radio amateurs - among whom I do not count myself - can undoubtedly bring great competence to this question. I've read it up in the aforementioned books, and also in the older Meinke & Gundlach publication [6], in Vastenhoud [9], and in Weeks et al. [10]. The practicing radio amateur may still find it worthwhile to hear about what I found there. Weeks et al. [10] contains an outline of the energy transfer at an interface between air, with $E_r = 1$, and water, with $E_r = 81$, explained in a way which is easy to remember, which is why this article is mentioned here.

One of the first commercially available active antennas was the AA300 from Standard Radio & Telefon AB [11]. At that time, its introduction onto the market gave me the incentive to carry out my own experiments. Fig.1 is taken from the prospectus. It has rather a deterrent effect in the perspective selected by the artist, due to the high mast. And the question should therefore be investigated of whether such a mast is actually necessary, or at any rate at least advantageous. The antenna has a base-plate with a considerable diameter,

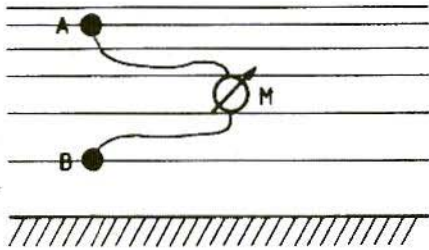


Fig.2: Metallic Conductors A and B receive the Potential from the corresponding Potential Lines. The Difference is displayed high-ohmically by the Measuring Apparatus M

which I simply copied at the time. Today, the question arises of whether it is necessary, harmful, or simply superfluous.

The proposal in [2] to use a CB radio groundplane as a short-wave reception antenna comes from me. In [3], I explained a probe which can be used to measure field strengths. It remains to be tested how far the results obtained then have to be reviewed.

The capacitive broad-band active antenna can be created only because the ambient noise intensity levels rise considerably at low frequencies. In the Hütte [1], I found a diagram which

determines the noise field strength directly, and thus supplies the calculation data required for an antenna which converts the field strength into a voltage.

Internal noise from the amplifier used naturally plays an important role. The selection of the correct transistor for the input stage and the influence of the passive antenna fraction should be represented. For this, we also need a book in which the relationships are shown in a comprehensible manner, which is why the Connor [49] can be found in the reference literature.

Finally, we should not be content with the theory. The antenna should be given a practical test. The groundplane from [2] can act as a comparison, using the short-wave receiver described there.

2. THE PASSIVE ANTENNA FRACTION

The capacitive broad-band active antenna is a receiver which registers potential difference between two points

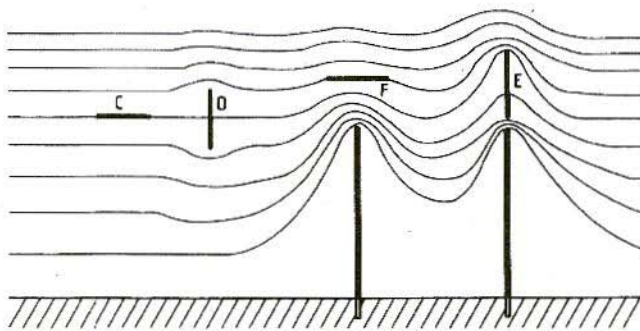


Fig.3: Field Distortion due to Metallic Conductors, which are insulated or connected to Earth

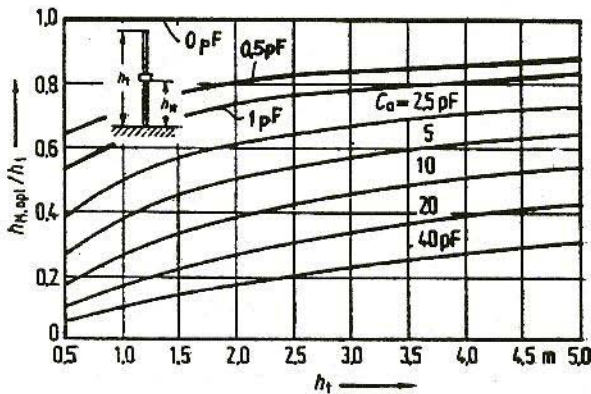


Fig.4: Optimum Height, h_M , optimum for the De-Coupling of the Reception Voltage according to Meinke & Gundlach (7) with newly plotted Curves for 0; 0.5 and 1 pF

in an electrical field. Fig.2 shows equipotential lines parallel to the earth's surface. The intervals between them diminish as the height increases. Two metallic conductors, A and B, which here are spherical in form, receive the potential from the corresponding lines. A high-ohmic measuring apparatus, M, shows the potential difference as a voltage. Direct calibration in field strength units (V/m) is possible if the interval between the conductors remains constant and is always vertical to the potential lines.

The potential difference source impedance can be measured and calculated as the capacity between the two conductors. The loaded impedance of the measuring apparatus must be high in relation to this, if the voltage is to be measured correctly and independently of the frequency. An input capacity in the measuring apparatus means the voltage displayed can be lower, while still remaining independent of the frequency. The highest voltage is undoubtedly obtained with a ball in contact with the ground.

The conductors should on no account be

spherical in shape, as these are not easy to manufacture using machine tools. Fig.3 gives examples of other shapes. A horizontal conductor, C, does not interfere with the field pattern. It can take the form of a rod or a plate. A vertical rod, D, by contrast, distorts the field, because in its immediate environment only the mean potential can prevail, which it then also acquires itself. A mast connected to the ground means considerable field distortion. A feed cable to the upper section E or F, or an earth wire has precisely the same effect. The vertical rod, E, or the horizontal conductor, F, thus obtain a lower potential than would be present for their height with an undisturbed field. The field strength is reduced on the ground in the vicinity of the masts. This shading also takes place in the vicinity of houses and trees, and in even more pronounced form within houses and under bridges.

The feed to the measuring apparatus in Fig.2 goes through several potential lines, and also creates field distortion (not shown here). The effect becomes more and more disadvantageous as the capacity of this feed towards the environment increases. It can thus be of use

to hold the actual receivers with support insulators and to make the feed extremely thin. If the influence of the feed is then discounted, the conversion factor can rapidly be estimated. If the interval between conductors A and B is, for example, a metre, then a field strength of 1 V/m delivers a voltage of 1 Volt. The conversion factor is V/V/m..

The capacities can also be estimated easily, but particularly easily if balls are involved. Older readers will remember that previously the unit of capacity was defined by means of a ball bearing with a radius of 1cm. in space. This ball bearing then had "1cm." capacity, so a ball bearing with a radius of 13cm. had "13cm." capacity. Nature was again very cooperative, and arranged things in such a way that this "1 cm." corresponded fairly precisely with 1pF, so that the 13cm. ball bearing has a capacity of 13pF in space. The vicinity of other conductors (ground, mast, baseplate) can only increase the value determined in this way which, as we shall see later, is advantageous.

The capacity for other forms can be calculated using the following formulae, taken from Meinke & Grundlach (6):

Circular disc in space (C = capacity in pF, s = disc thickness in cm., D = disc diameter in cm.):

$$C = 0.353 D \cdot (1 + 0.637(s/D)) \quad 1$$

Vertical rod at a distance of more than 1/4 from the ground (l = length in cm., D = diameter in cm.):

$$C = 0.24 \cdot l \cdot (1/l_g) \cdot (1/d) \quad 2$$

Horizontal rod above ground (h = height, l = length, D = diameter)

$$C = 0.24 \cdot l \cdot (1/l_g) \cdot (4h/D) \quad 2$$

Some examples may indicate the order of magnitude involved:

For the rod with a diameter of 7mm. and a length of 1 m. which Jirmann [5] uses, the value calculated is 11pF. A thin plate with a diameter of 10cm. gives 3.5pF, and even a thin feed with a diameter of 0.5mm. and a length of 0.5m. already has a value of 4pF.

A layout like Fig.3e can be optimised in accordance with Meinke & Grundlach [7]. This is clarified in Fig.4, which is taken from this book, but which has been amplified by the three curves for the 0, 0.5 and 1pF input capacity of the subsequent amplifier. According to this, with the pre-set mast height, ht, we obtain an optimal height, hM, for the output coupling point, which depends on the amplifier input capacity. If we succeed in making the input capacity zero, then the effective upper part of the antenna can be very short.

The way in which the potential lines are forced back at the tip of the mast indicates that the field strength is greater there than elsewhere. The same is also true above houses and trees. However, all these considerations are valid only as long as the dimensions are small in relation to the wavelength. With an expansion of even 5m. in the upper short-wave range, this no longer applies. The point of having a high mast lies more in increasing the distance from unwanted interference emanating from cars, houses and transmission lines.



And for this reason it can be advantageous, not just to position an active antenna on the roof of a house, but to place it on top of a mast positioned there. Damping a feed can usually be dispensed with until right at the top end of the short-wave range. At a length of 100m., it produces only 2 - 3dB for a measuring frequency of 30MHz with RG-58/U type cable.

3. EXTERNAL NOISE

The thunderstorm activity continuously present on Earth is the origin of the atmospheric noise. It is assumed that 1000 to 2000 thunderstorms are simultaneously active at any given moment, producing about 100 lightning strikes

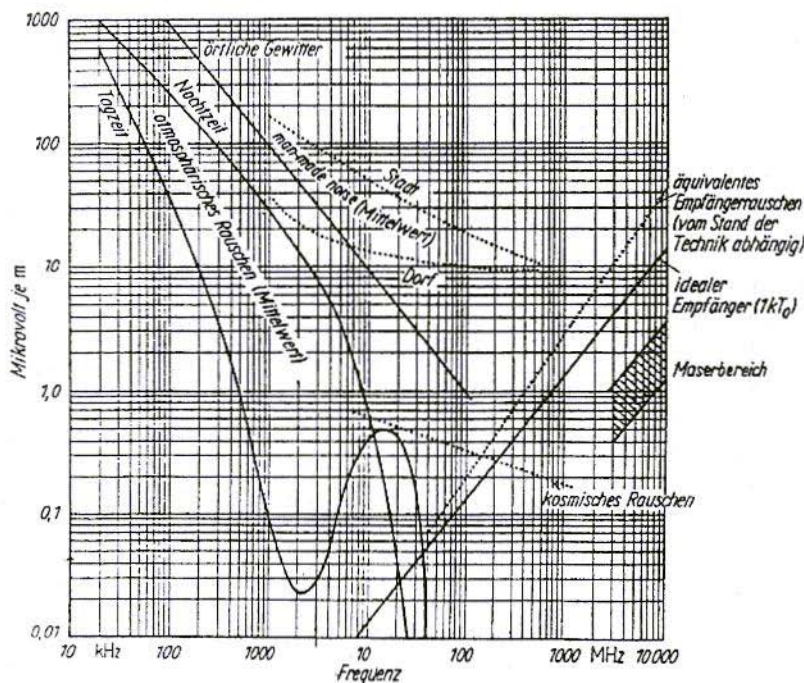


Fig.5:

Noise Field Strength of Various Sources for a Receiver Band Width of 10 kHz according to Hütte (1). A $\lambda/2$ Dipole has been used as the Reception Antenna in the Curve for Equivalent Receiver Noise

Tagzeit = Daytime, *Nachtzeit* = Night time, *Örtliche Gewitter* = Local thunderstorm, *Atmosphärisches Rauschen* = Atmospheric noise, *Mittelwert* = Mean value, *Stadt* = City, *Dorf* = Village, *Kosmisches Rauschen* = Cosmic noise, *Äquivalentes Empfängerrauschen (vom Stand der Technik abhängig)* = Equivalent receiver noise (dependent on state of the art), *Idealer Empfänger* = Ideal receiver, *Maserbereich* =

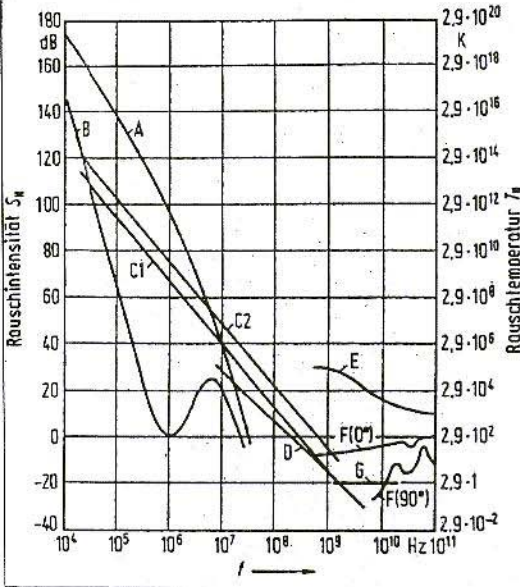


Fig.6:
Interference Power Density for Various Sources according to Meinke & Gundlach (7)
A - Atmospheric Noise (Max)
B - Atmospheric Noise (Min)
C1/C2 - Industrial Noise (Country/Town)
D - Galactic Noise
E - Quiet Sun
F - Noise from O₂ and H₂ at two Angles of Elevation
G - Cosmic Background Radiation 2.7 K

Rauschintensität = Noise intensity, *Rauschtemperatur* = Noise temperature

per second. The propagation of this interference energy is subject to the same laws as those governing electromagnetic waves. The range width is considerably greater by night than by day, the day-night difference being less for long waves. We thus obtain the graph in Fig.5, divided up to show the pattern at night, in the daytime, and during a storm in the area. The diagram is taken from Hütte [1] and the field strengths given are valid for a noise band width of 10kHz, which corresponds to the standardised radio inter-

ference measuring apparatus or to an average AM radio.

A capacitive broad-band active antenna with the conversion factor 1 V/V/m delivers an output voltage which is numerically identical to the field strength. Thus, for example, at 20kHz it emits a noise voltage of app. 1mV at night. In the high-frequency (short-wave) range, the atmospheric, and also the cosmic, noise is usually less than 1µV.

The rising line on the right in Fig.5 represents the noise background for an ideal or actual receiver which is connected to a $\lambda/2$ dipole antenna, converted to the equivalent field strength. We can see that a particularly sensitive receiver is not required in the short-wave range. 20dB worse than 1 kT₀ is still sufficient. The diagram used by Jirmann [5] is shown again here as Fig.6, because it demonstrates an impor-

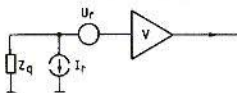


Fig.7: Noise Equivalent Circuit Diagram for an Amplifier. V Noise-Free Amplifier, Zq Source Impedance, Ir Noise Current Source, Ur Noise Voltage Source



tant technical advance, as against Fig.5 - a considerable reduction in interference generated by human activity! In the late fifties, it was realised in the industrial countries that interference had to be reduced, so legislation was brought in imposing interference suppression. The introduction of regulations in stages, depending on the state of knowledge at any one time, has led to a situation today where, in spite of a considerable increase in the amount of electrical equipment, the intensity of interference caused by it has decreased. On the other hand, the country I now live in has no legislation on interference suppression to date.

The ordinates for the two diagrams, Fig's.5 and 6, are measured in different units, and are thus not directly comparable. To convert from intensity, which probably corresponds to the power density, into field strength, we must use the effective antenna area and the field strength output voltage formula for the $\lambda/2$ dipole. I shall save myself the trouble, since for the capacitive broadband active antenna the field strength scale is more practical and easier to read.

4.

INTERNAL NOISE

The amplifier behind the passive antenna section sees a source, the impedance of which corresponds to a few pF. A low radiation resistance of considerably less than 1Ω has actually no part to

play, as the quality remains greater than 100 over the entire range which is of interest.

The internal noise of an amplifier can always be represented in relation to the input - i.e. by a noise voltage source, U_r , and a noise current source, I_r , in Fig.7. The voltage drop generated by I_r at the source impedance, Z_q , combines with U_r to give the total noise voltage, $U_{r \text{ tot}}$. If the noise sources are uncorrelated, the noise voltages are added together geometrically:

$$U_{r \text{ tot}} = \sqrt{(U_r^2 + Z_q^2 \cdot I_r^2)} \quad 4$$

If they are correlated, the phase angle of Z_q then has to be considered. If it is 0° (ohmic), the voltages are added together arithmetically. With anti-correlation, it is even theoretically possible that the total noise voltage will be smaller than the figure you start with. However, according to Murphy this drop is extremely unrealistic. If Z_q , the phase angle, is + or - 90° , as always with capacitive broadband active antennas, then equation (4) applies again. For amplifiers which operate in accordance with a tuning circuit, the dependency on the phase angle is the reason why de-tuning in itself can reduce noise.

The lowest inherent noise value for an amplifier is obtained if it is conditional upon the active element in the first stage alone. This can not always be fully guaranteed, for there is a limit to the amplification of the first stage. However, if it is only 1, then the second amplifier stage contributes just as much as in the input stage. We have given away 3dB. The same is true for amplifi-

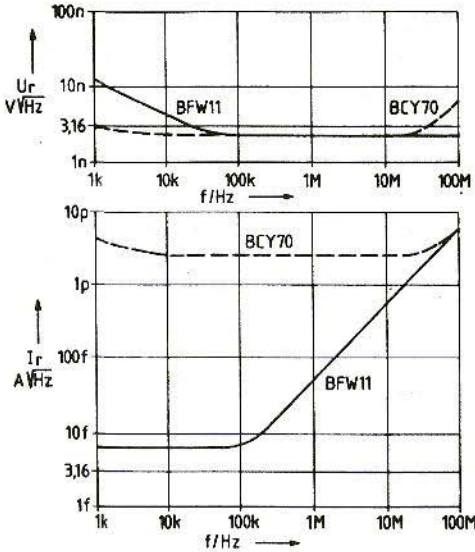


Fig.8:
Noise Voltage and Noise
Current behave very differently
in Bi-Polar and Depletion Field
Effect Transistors, as shown
here through the Examples of
BCY70 (Bi-Polar) and BFW11
(FET)

ers with differential amplifier inputs, because there two transistors are contributing to the noise in the input. An input of this type is necessary with active antennas for horizontal polarisation, but not with those under consideration here. The first transistor can be a field effect or bipolar type. The curves for the U_r and I_r paths, plotted against the frequency, show what has improved. Noise specification is just the area where the manufacturers are very reticent. I found adequate specifications for only a few types. Those in Fig.8 may perhaps be representative. The curves show that, for a low-frequency FET, U_r is somewhat higher than for a bipolar transistor. At high frequencies, the bipolar type performs rather worse. But this may be due to the fact that it is really not intended for high-frequency applications at all. There is no significant difference in the long-wave, medium-wave and short-wave ranges.

I_r behaves in a completely different way. At low frequencies, the noise current from the FET is up to 3 decades less than that for the bipolar type. By contrast, above 50MHz there is no longer any significant difference. I_r rises linearly with the frequency, and the path of Z_q is inverse to this. The product $I_r \cdot Z_q$, which represents the part of the inherent noise voltage generated by the noise current, remains practically the same between 0.1 and 50MHz. With an antenna capacity of 10pF and a 10kHz band width, the value calculated is only 80nV, whereas U_r is already 220nV. Combined together in accordance with equation (4), these values give 235nV. We could thus dimension the passive antenna section to be quite considerably shorter and so lower in capacity. At 3.5pF, the noise fractions attributable to I_r would be about the same as those for U_r . The internal noise would then be about equal to the external noise - i.e. about 0.35 μ V. These calculations indi-



cate that the capacitive broad-band active antenna would not have been possible without depletion layer FET's.

MOSFET's have a markedly lower noise current than depletion layer FET's, while that of MESFET's is slightly higher. For both, the cut-off frequency of U_r - i.e. the curve bend, at which the $1/f$ rise begins, is at markedly higher frequencies. Should it prove possible to improve this behaviour in the foreseeable future, they would have the advantage of a higher rate of rise, and thus greater amplification in the first stage. It would make virtually everyone happier if the semi-conductor industry would include more U_r and I_r curves in their data sheets. It would then no longer be necessary to select an appropriate component through trial and error.

5. WAVE PROPAGATION

The frequency range of a capacitive broad-band active antenna can extend, for example, from 10kHz up to 30MHz. Because of this wide range, it is not possible to give a generally valid de-

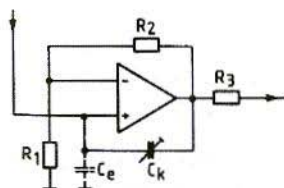


Fig.9: Principle of Compensation for Input Capacitance of an Amplifier

scription of wave propagation. However, it can be stated that the ground can be regarded as a conducting surface at low frequencies and as a dielectric interface at high frequencies. In between, mixed behaviour is present - a dielectric with conductivity, which is characterised by the occurrence of losses. Losses are greater when the polarisation is horizontal than when it is vertical. For particularly wide ranges, the ground wave must therefore be vertically polarised.

The polarisation initially corresponds to the alignment of the transmission antenna. But after one or more reflections to the ground or the ionosphere, a sizeable orthogonal polarisation fraction is present, so that for long-distance reception it is relatively immaterial what the transmission polarisation was.

Waves with a high wavelength (VLF, very long frequency, or long wave) are carried between the ground and the ionosphere as if in a waveguide and so retain their polarisation. As they are transmitted with vertical polarisation, a vertically active antenna is correct for reception. As losses occur, the field strength vector declines in the propagation direction. But its main fraction is vertical, as before. This applies only to the ground wave, the range of which is in any case reduced when ground losses arise. At shorter wavelengths - i.e. medium and short wave - reflection on the ground involves polarisation. Whilst horizontally polarised waves are always subjected to a 180° phase shift, with vertically polarised waves it depends on the angle of incidence. Shallow waves with a low angle of elevation are reflected through a 180° phase shift.

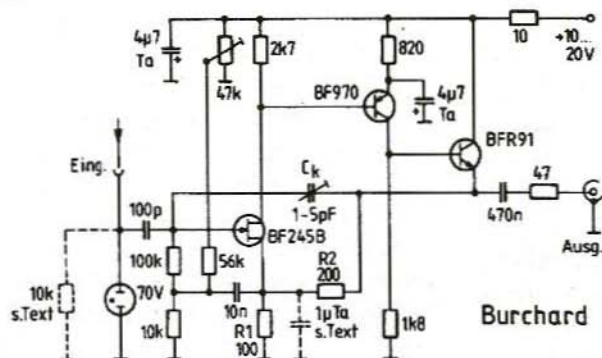


Fig.10:
Active Antenna
Amplifier Circuit

Eing. = Input
Ausg. = Output

Burchard

Those with a high angle of elevation come back in the same phase - i.e. they amplify the wave with direct incidence. At one specific angle, the Brewster angle, which depends on the frequency and the ground characteristics, nothing at all is reflected. The reception takes place as if the reception antenna were mounted in free space.

Direct and reflected waves are vectorially combined at the reception antenna. If the reflection phase position is 180° , this results in a reduction in the reception field strength. If not, there is an increase. Thus horizontally polarised waves are always weaker when received than vertically polarised waves. Only at very small angles of elevation (5° to 20°) are they approximately equal.

The longest ranges are obtained above sea water with good conductivity, $Q = 2(\Omega \cdot m)^{-1}$, and a high relative permittivity, $\epsilon_r = 81$. As 70.8% of the earth's surface consists of sea water, this is very convenient for signaling.

The reflection and attenuation capabilities of the ionosphere are connected to solar radiation, sunspot activity and the earth's magnetic field. There are daily cycles, seasonal cycles and sunspot

cycles. Nothing suggests that there are times when vertical polarisation of reception antennas could be unfavourable. Active antennas can also be constructed for horizontal polarisation and can also be directional. But that is outside the scope of this article.

The free space field strength of a 1 MW long-distance station is certainly considerable. It amounts to about 1 V/m at a distance of 10km.. The active antenna must still operate satisfactorily with such field strengths coming from one or more transmitters. This requires a level range free of inter-modulation of more than

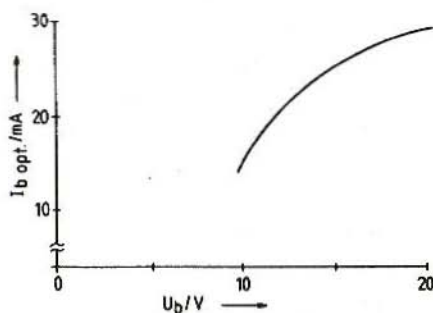


Fig.11: To ensure as little Distortion as possible, the Power Consumption is set to the Optimum Value for Every Operational Voltage



120dB. Even measuring such a value involves the greatest of difficulties.

6. PLANNING

Optimal results can be expected if the capacitive broad-band active antenna is measured for a given receiver. Here I would like to refer to the receiver based on [2], which I actually use for short-wave reception on an everyday basis. It has a noise band width of 7.6kHz, which means there is a difference of only 1.2dB in the noise, as against Fig.5. In relation to the input, the noise level is -107dBm, and the input power which is required for an inter-modulation product of the third order to be generated of the same value is -28dBm. The layout is designed for a broad-band input, 3 - 30MHz.

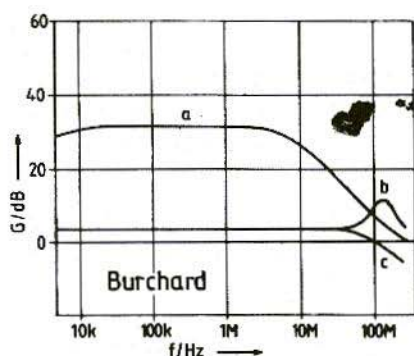


Fig.12: Amplification from Amplifier in Fig.10 without Negative Feedback (a) and with various Coupling Capacitors, with Generator 10pF (c) or 10nF (b)

If the noise field strength is approximately $1\mu\text{V/m}$ (Fig.5), then a conversion factor of 1 V/V/m should be about right. It can be obtained, for example, through a rod antenna with an effective length of 0.33m. and an amplification of 3. For an effective length of 0.33m., a rod about 0.5m. long is required. The external and internal noise levels at the amplifier output are then approximately equal to the noise background of the downstream receiver.

For short-wave reception of frequencies above 3MHz (high-frequency range), the amplifier must have an input resistance exceeding $10\text{k}\Omega$ if the source impedance corresponds to 5pF . Including the medium-wave, long-wave and extremely long wave ranges requires $100\text{k}\Omega$, $1\text{M}\Omega$ or $3.3\text{M}\Omega$. For the high-frequency (short-wave) and medium-frequency (medium-wave) ranges at least, no bootstrapping of the shunting resistor is required. The output impedance of the amplifier should be as low as possible, so that load variations have no influence. An ohmic series resistance of 50Ω then ensures that the subsequent receiver sees 50Ω . This "reciprocal matching rule" makes sure that the receiver behaves as at the signal generator, and that selective filters in the input are correctly blocked off.

To compensate for the input capacity, a circuit as in Fig.9 is required, in which R_1 and R_2 determine the amplification and R_3 determines the output impedance. The voltage distributor, R_1 , R_2 , is low-ohmic, so only the non-inverting input of the amplifier needs to have an FET input. Its unavoidable capacity, C_e , is compensated if:



Fig.13: Sandwich Format of Amplifier between two Octagonal Boards made of PCB Material

$$C_k = C_c(R_1/R_2) \quad 4$$

which is equivalent to neutralisation. Fig.10 shows a suitable amplifier. It has a three-stage layout and can be operated using operating voltages of between 10 and 20 Volts. As the operating voltage increases, the level control capability increases. For every voltage, there is an optimal operating current for minimum distortion, which can be read off from Fig.11. The measurements below were carried out at 15V and 25mA. The

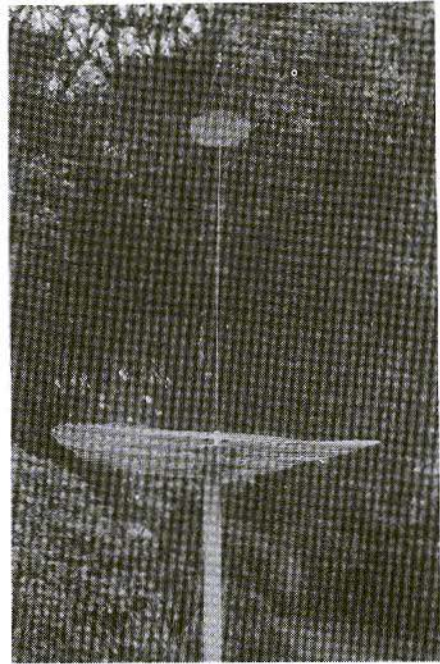


Fig.14: A Roof Capacitor and an Earth Plate can be used for Tests

current is adjusted using the trimming potentiometer, which compensates the considerable operating point tolerances of field effect transistors at the same time.

The circuit in Fig.10 scarcely differs from the one I used in 1974. Then the first stage was equipped with an NF3819 (i.e. 2N3819 in a metallic TO-18 housing), the second with an AF139, and the third with a BFX59. As these types of equipment can scarcely be obtained today, and the metal housing and its earthing do not seem to be absolutely necessary, I changed over to



the types of component given. A BFW11 could not have been operated here, so a BF245 was substituted for it, which is probably the reason why the noise in relation to the input was greater than Fig.8 predicts.

This amplifier has an open-loop voltage gain of about 30, which decreases linearly above 10MHz and is still about 7 at 30MHz. To measure it, the negative feedback resistance, R_1 , should be bridged over at 1 μ F. Due to the low series resistance, a wrapped capacitor or tantalum capacitor is necessary.

With negative feedback, the amplification is 3, and with a 50 Ω load it is 1.5 over the entire frequency range between 10kHz and 30MHz, as shown by Fig.12. The correct setting for C_k can be recognised by the fact that the amplification is the same for low and high coupling capacities at the signal generator.

The resonance type accentuation at 150MHz, with a large coupling capacitor, is connected to its inherent inductance, and can therefore be ignored.

This amplifier may display a tendency to oscillate with an open input and over-compensation of C_e . But oscillation will stop immediately if an antenna with a few picofarads is connected.

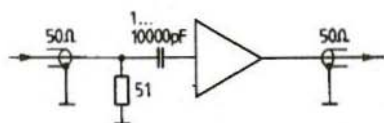


Fig.15: Test Circuit for Amplification, Noise and Inter-Modulation

7.

MECHANICAL ASSEMBLY

Two octagonal discs made of single-lined printed circuit board material are arranged one above the other, with interval columns. Their diameter is selected in such a way that a tin can could be put over them to protect them from the rain. The upper one has a banana socket in the middle, into which the antenna conductor is inserted. The lower one has a BNC socket as the high-frequency output and a banana socket as the operating voltage feed.

Using a coax cable for the power supply is initially dispensed with. Fig.13 shows this formation on a mast, from which the discone antenna mounted there was removed for the tests.

The amplifier circuit is mounted between the two detachable discs, which is possible because of the small number of components and the uncritical layout, with the help of a few Minimount supporting points.

Fig.1 suggests that a capacitive broadband active antenna should have a large earth plate. To investigate this, two 1/4" flat plugs are mounted on the upper octagonal disc, onto each of which a roughly cut earth plate made of wire netting, with a diameter of 50cm., can be placed with a flat plug at the end, onto which a roof capacitor with a diameter of 10cm. can be placed.

With the two wire netting arrangements, the antenna then looks like Fig.14.

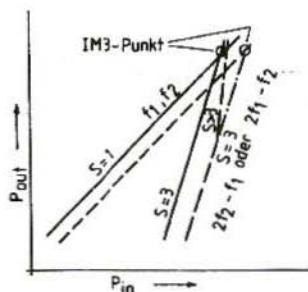


Fig.16: Fundamental Pattern of Signals generating Inter-Modulation with Frequencies f_1 and f_2 , and of Third-Order Inter-Modulation Products with Frequencies $2f_2 - f_1$ and $2f_1 - f_2$

Punkt = Point, Oder = Or

8. MEASUREMENTS AT THE LABORATORY TEST BENCH

If power from a high-frequency generator is transmitted through a resistance with an order of magnitude of $10\text{k}\Omega$ into the as yet uncompensated input of the amplifier ($C_k = 0\text{pF}$), then the input capacity of the amplifier can be determined on the basis of the cut-out frequency of the low-pass behaviour. It was measured here as 4.5pF . The banana socket in the printed circuit board material adds 3pF to this, and the glow lamp contributes 0.3pF . The input capacity to be compensated is thus 7.8pF . C_k must theoretically be set at 3.9pF . But in practice it is much simpler to set C_k in such a way that the same amplification is present with a small coupling

capacitor - i.e. about 10pF - at the signal generator as with a very big coupling capacitor - i.e. about 10nF . That was already a pre-condition for Fig.12.

The capacity of the antenna conductor can be measured with a universal measuring bridge. The result was 4pF , which correlated very well with equation (2). Adding the roof capacitor on increases it by 3pF , which also correlates very well with equation (1). The influence of the earth plate on the antenna conductor capacity is extremely small. 0.5 and 0.2pF increases were measured for the rod and the roof capacitor respectively.

The open-loop voltage gain and composite gain are measured in a circuit as per Fig.15. The results have already been indicated above (Fig.12). In this circuit, which is also suitable for measuring noise and inter-modulation, the input power supplied from the signal generator is fed to the 51Ω resistance. The capacitive broad-band active antenna picks up virtually no power, for which reason the signal generator must be artificially closed off. Although the typical power amplification measured with the circuit in Fig.15 is 3.5dB , it is actually very high!

The output of the circuit in Fig.10 can supply a maximum current of 15mA_{eff} . This is sufficient for a yield of 10mW ($+10\text{dBm}$) at 50Ω , or a peak voltage of $\pm 1\text{V}$. Although an oscillogramme indicates hardly any distortion for this output power, this does not supply proof that the 120dB inter-modulation intervals desired have now been obtained. Unfortunately, the true picture looks very much darker.



A tried and trusted measurement for the absence of distortion is the interval between the third-order inter-modulation products and those generating two signals of the same size. It can be demonstrated that the levels of these inter-modulation products rise three times as steeply as those for the generating signals. This is true as long as the characteristics can be described by a Taylor expansion as far as the third grade and all higher grades can be dispensed with. Fig.16 shows this relationship. The steepness gradient ($S = 1$ or $S = 3$) is thus initially valid for low-signal operation. But the curves are extended until they cross at the intersection point, and we then read off the input or output power at the interception point as the IM3 point at the input or the IM3 point at the output. The higher this point is, the better the amplifier! If

the low-level signal operation is ignored during measurement, then the curve for the inter-modulation product is steeper, because higher powers of the Taylor expansion are taken into account. This can be seen from the path of the dotted line in Fig.16, which initially runs parallel to the $S = 3$ curve, but then bends steeply upwards if, for example, limitation occurs. The IM3 point to be given here is located in the extension of the bottom part of the curve along the dot-dash line. The dotted inter-modulation product curve is derived from the solid line through negative feedback. The linearity is increased, so the IM products are decreased. The amplification is reduced, but the inter-modulation interval increases.

The measurement guidelines can be derived directly from this. The generating signals are to be reduced until the

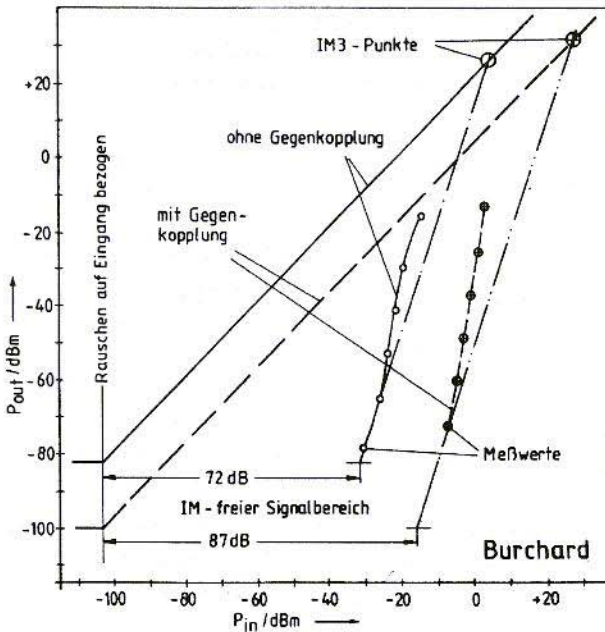


Fig.17:
Amplifier Readings
 according to Fig.10 at
 15V Supply and
 Measurement Freq's
 $f_1 = 19.8\text{MHz}$ and
 $f_2 = 20.2\text{MHz}$

Rauschen auf Eingang bezogen = Noise in relation to input,
Punkte = Points,
Ohne Gegenkopplung = Without negative feedback,
Mit = With,
Meßwerte = Readings,
IM-freier Signalbereich = "IM-free signal range"

inter-modulation products with a gradient of 3 decrease. If you have discovered one or more points on the $S = 3$ curve in this way, then carry on the extension with this gradient up to the interception point.

The signal generators for IM3 measurement must emit relatively high signal levels, and must therefore operate with little attenuation in the calibrated attenuator. If they are inter-connected through a simple T-piece, then power also goes from the one generator into the output of the other. A rectifier, for example, is connected up there to measure the input voltage of the calibrated attenuator. Due to its non-linearity, it also creates inter-modulation products, which can call the entire measurement process into question. A power combiner can provide a remedy (example: PSC2-2 from Mini-Circuits) by inter-connecting the generators, together with additional buffer amplifiers between the generators and the combiner, as well as an external calibrated attenuator after the hook-up. This last makes it possible to operate the generators with constant coupling, and thus with the inter-modulations conditional upon it. If they are too high, then Fig.16 gives an inter-modulation curve which runs parallel to the signal curve.

The IM products are measured through a selective receiver. That can be, for example, the short-wave receiver described in (2), or even a spectrum analyser. In the second case, we have the advantage that we can see all the products simultaneously, whereas the short-wave receiver must be tuned to them one after another. One would initially think that the selective receiver

itself must have better IM characteristics than the object of measurement. However, under favourable circumstances it is possible to overcome these obstacles. The question, as can be seen from Fig.17, is whether a measurement point on the $S = 3$ curve can still be reached using the level range of the receiver, which can still be measured with certainty and is free from inter-modulation. If we make the conservative assumption that the lowest reading with negative feedback (open circuit) is already on the $S = 3$ curve, in this case there is a signal range of at least 87dB free from IM. Without negative feedback, it is only 72dB. Thus the only requirement for the inherent inter-modulation of the receiver is that it should still play no part if the difference in levels is 70dB. This is the case for the receiver described in (2), and also for many spectrum analysers. We can now read off the IM3 points without negative feedback from Fig.17 at +4dBm (input) and 26dBm (output). With negative feedback, they improve to at least +27dBm (input) and +31dBm (output).

To measure still more extensive IM-free signal ranges with an inadequate receiver, a notch filter is required between the object of measurement and the receiver. It considerably lowers the generating signal, but allows the IM products to pass. As it can be implemented only passively, the distance between the two measurement frequencies must then be considerably increased.

The receiver noise amounts to -107dBm in relation to its input, so its fraction is -110dBm (0.7 μ V) at the input of the



active antenna. The active antenna noise and the receiver noise are geometrically combined in accordance with an equation similar to equation (4). Here we are measuring the voltage noise with a large coupling capacitor, with the receiver noise at -109dBm ($0.8\mu\text{V}$), plus the current noise fraction included at 5pF coupler C -104dBm ($1.3\mu\text{V}$) and at 10pF coupler C -106dBm ($1.1\mu\text{V}$). From this we can calculate $U_r = 0.4\mu\text{V}$ (-115dBm) in addition to $Z_q \cdot I_r(5\text{pF}) = 1\mu\text{V}$ (-107dBm) or else $Z_q \cdot I_r(10\text{pF}) = 0.5\mu\text{V}$ (-113dBm). The IM-free range for

the entire system of active antenna and receiver is naturally smaller than that for the individual components, namely 73dB with an antenna capacity of 5pF or 75dB with 10pF . The loss is small, because the two components have been tuned to one another.

The amplifier current noise turned out to be much greater than one would have expected from Fig.8. This may come from using a different type of FET. An experiment in which the component was replaced by a BF244 produced no significant alteration.

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Improved Feed for the Off-Centre Fed Dipole

Theoretical data suggest that the commonly used feedpoint for the off-centre-fed dipole (OCFD) may not be the best. The OCFD is an attractive multi-band antenna because it is simple, inexpensive, and requires no antenna tuner. Improving its performance simply by moving the feedpoint makes the antenna even more attractive. This note illustrates how the feedpoint influences antenna performance by analysing computer-modelled SWR data for three different feedpoint locations.

The OCFD, which is shown schematically in Figure 1, consists of a single wire radiator of length L fed off-centre a distance D from one end. The usual implementation uses a "1/3-feed", that is, the RF source is located one-third of

the way from the end, so that $D = L/3$. Why the feedpoint should be located there is not exactly clear. The 9th edition of the *Antenna Book* [1], for example, observes that there is not much theoretical justification for this choice. Nevertheless, the 1/3-feed is accepted practice for building an OCFD.

Design details for a 1/3-feed 3-band OCFD (80-40-20 meters) appear in the 17th edition of the *Antenna Book* [2]. A 4:1 current balun at the feedpoint matches this antenna to any length of 50 coax. More recently, Bill Wright, G0FAH, [3] described a four-band 1/3-feed OCFD (40-20-15-10 meters) fed with 300 ladder line. Matching 50 coax requires a 4:1 balun on 40-20-10 meters and a 1:1 balun on 15 meters. Four band operation therefore requires switching

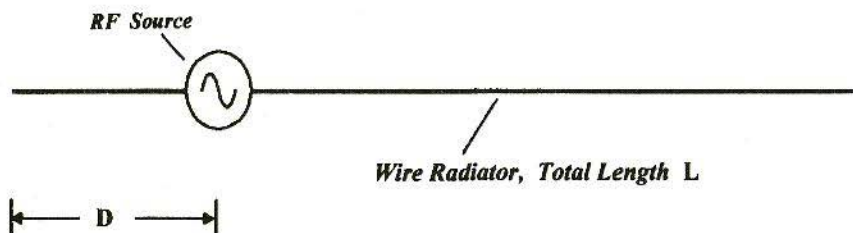


Fig.1: The Off-Centre Fed Dipole Antenna



baluns. Another minor limitation is that the ladder line length can be only an odd multiple of the wavelength at 21 MHz because the line is used as an impedance transformer. A simpler approach to achieving four-band operation is to feed the OCFD at a different point along its length.

A 21.03 meter (69 foot) long, 0.2053 cm diameter (12 AWG) OCFD was computer-modelled in free space. The dimensions are the same as those in the G0FAH design. Free-space results are a good approximation for antennas high enough above the ground (typically a significant fraction of a wavelength). The band-centre SWR was computed on 40-20-15-10 meters at the antenna input terminals for a feed system impedance $Z_0 = 200$. The theoretical values of

input resistance and reactance were used to calculate SWR (the antenna was not assumed to be tuned). Because $Z_0 = 200$ a 4:1 balun is required to feed the antenna with 50 coax. The results for three different feedpoints appear in Figures 2, 3 and 4.

Figure 2 plots SWR at the antenna terminals for the conventional 1/3-feed ($D = 6.98$ m). The 40 and 10 meter values are slightly over 2:1, while the 20 meter SWR is about 1.75. In marked contrast, the 15 meter SWR is off the scale (actual value > 20). It is this behaviour that makes a special feed system necessary on 15 meters, a complication which can be avoided by moving the feedpoint.

Figure 3 plots SWR when the OCFD feed is located 8.65 meters from one

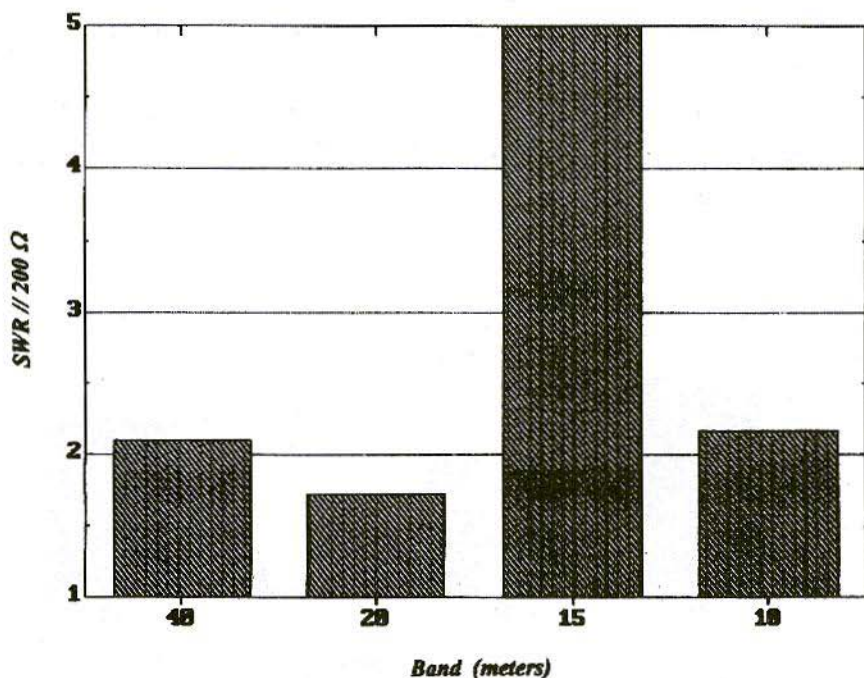


Fig.2: VSWR for Feedpoint at 6.98m

end. The 40-20-10 meter SWRs are somewhat higher than they are with the 1/3 feed, but the 15 meter SWR is very low (app. 1.2). Moving the feedpoint 1.67 m closer to the centre of the antenna results in a much better average SWR. And, more importantly, special matching is not required to achieve SWR < 2.5 at the antenna terminals on all bands. Balun and coaxial cable losses, which are inevitable, reduce the SWR at the coax input to even lower levels. For most installations, it is probably reasonable to expect SWR at the transmitter to be less than 2:1 on all bands.

One more example of the effect of feedpoint location appears in Figure 4, which plots SWR for a feedpoint 3.65 meters from one end. The values on

40-20-15 meters are excellent. The 40 meter SWR is only slightly above 2, and the 20 and 15 meter SWRs are below 2:1. The highest value occurs on 10 meters, where it is approximately 2.4. Since the SWR is reduced by feed system losses, it will be less than 2.4 at the coax input. And, because balun and cable losses increase with frequency, the SWR reduction will be greatest on 10 meters where it is needed most. Feeding the antenna 3.65m from one end may well provide the best overall four-band performance.

In a specific implementation, the OCFD, like any antenna, must be "tweaked" for optimum SWR. This is accomplished by adjusting the feedpoint location. Other antennas, nearby metallic objects, and the earth are typical factors that influ-

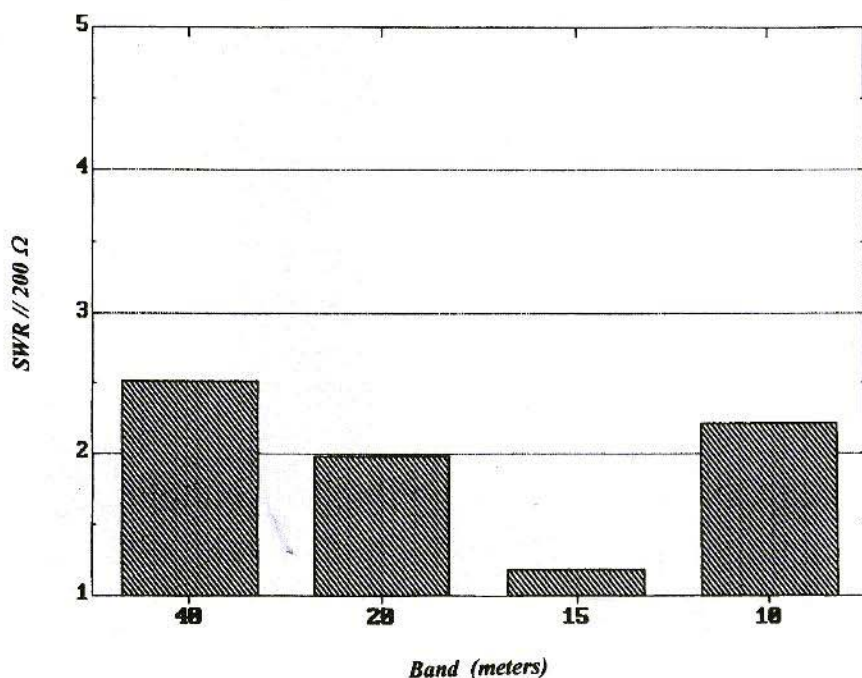


Fig.3: VSWR for Feedpoint at 8.65m



ence antenna performance. Since these factors are not included in the computer model, they must be dealt with empirically by adjusting the antenna on-site. The data presented here provide a starting point for experimenting with different feedpoints. Depending on the total antenna length L , height above ground, earth electrical parameters, and feed system Z_0 , it should be possible to operate a single OCFD on four or more bands without an antenna tuner or special feed arrangement. It is apparent is that the OCFD's SWR varies dramatically as the feedpoint is moved, and that the commonly used 1/3-feed is not necessarily the best. Other feedpoints may therefore produce a better antenna.

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- [2] *The ARRL Antenna Book*, 17th Edition, R. Dean Straw, N6BV, Editor, American Radio Relay League, Inc., Newington, CT, USA, 1994, page 7-20
- [3] Wright, Bill, *Four Bands, Off Centre*, QST, February 1996, page 65

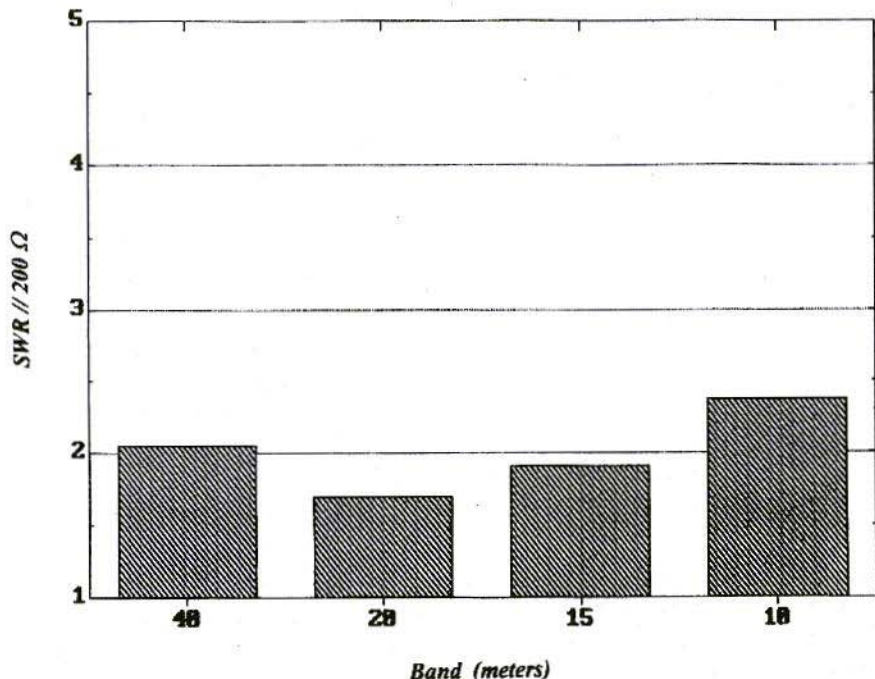


Fig.4: VSWR for Feedpoint at 3.65m

Denys Roussel, F6IWF

An Ultra Low-Cost HF SSB/CW Transceiver with 20W Output, an AGC Meter, S-Meter and Audio Filters

Part-1 - The Design

Note: Although VHF Communications Magazine, by its very name, normally specialises in VHF and up, this project was found to be of such quality that we decided to include it. We hope that you find it as interesting as we have and forgive the 'intrusion'. The Editor

1. INTRODUCTION

The idea to build an inexpensive transceiver came from a joke on the air with a friend of mine, Francis "F6AWN", about the high prices of modern radio equipment made in Japan. As I do not have any recent

commercial equipment, I was not very well informed about cost and I was somewhat surprised by prices. Some of these Japanese rigs cost more than £2000. Consequently, it is sometimes difficult for youngsters, beginners or radio-clubs to invest in the very latest equipment.

To help bring amateur radio activities to every body, it should be interesting to make a transceiver for the value of a banknote. I preferred to build a complete SSB/CW unit with medium power output and satisfactory performance, rather than a QRP station I watt crystal-controlled unit.

The challenge was set at building such a unit for less than DM 100, or approx. £45.

2. SPECIFICATIONS

2.1 Why SSB ?

A lot of QRP transceivers at low cost have been described in the amateur radio literature, few for phone operation. It is obvious that this mode is more attractive than a CW contact (of course for me!).

Nowadays, it is almost impossible to use a Double Side Band receiver. HF bands are very busy, and the other side band carries signals received simultaneously on a DSB receiver. In DSB transmit mode, the interference often spoils communications on adjacent frequencies.

2.2 On Which Band ?

During the period of sunspot activity, propagation conditions vary, and for this reason the lower frequencies 3.5 and 7 MHz are more reliable than the higher 10 to 28. While the 80 metre band performs well at night, but is not ordinarily useful in the day time, at the same time the 7 MHz band is the only band useful for short or long distance work with a good antenna. In addition, 40 metres is the better test for a receiver. The transceiver should operate on 40 metres, and if possible on 80 metres if it is feasible with the price target.

2.3 The Power Level

The greater part of QRP transceivers are running between 1.5 and 8 Watts. This power level

is too low to have easy contacts, a power of about 20W will give best performance.

It is 4 times lower than normal radio sets, i.e. 6dB down or 1 S-point. The power supply for 20W output is not so big and is still compatible with a battery operation.

2.4 Other points

The minimum requirements of a transceiver for comfortable operation, a transceiver are that it should have a loud speaker, an AGC stage, an S-meter and filters.

The Automatic Gain Control function is subject for discussion, some operators do not like using AGC, because it is impossible to correctly hear weak signals when stronger are present. This is quite true, but it is very difficult to scan the entire 7 MHz band entirely without AGC, the first broadcast station makes the operator jump and wake up neighbours! I have acquiesced to the wishes of both camps by fitting an AGC ON/OFF switch.

2.5 Complete Specifications (objective):

The purpose is not to have the best HF transceiver in the world, but to offer minimal characteristics to use it in actual conditions. The performance of the receiver should also be similarly constructed.

Some operators will consider that the basic specifications as are being considered here by no means indicate a



state-of-the-art transceiver, but like all such projects, it is necessary to relate it to the price you want to pay. Below are listed the basic specifications for the project:

General:

Range: 7 to 7.1 MHz
 Modes: LSB and CW
 Stability: better than 500 Hz/hour after 10 minutes warm-up
 Supply: 13.8 V DC nominal
 Current: <4A

Receiver:

Input intercept point: 15 to 20dBm
 Unwanted side band rejection: 40dB
 Noise figure: 20dB
 IF bandwidth wide SSB: 300 - 300 Hz
 IF bandwidth narrow SSB: 300 - 1800 Hz
 IF bandwidth CW: 300 Hz centred on 650 Hz tone from S9 to S9+40dB for 6dB change-
 AGC Range: from S9 to S9+40dB for 6dB change-
 Maximum AF output: approx. 1W

Transmitter:

Unwanted sideband rejection: 35dB typical
 Carrier rejection: better than 35dB
 Power output: approx. 20W
 IM3: better than -30dB at full power output
 Harmonics level: better than -40dB at full power output

3. DISCUSSION OF THE CONCEPT

There are two basic systems for generating SSB signals:

- a) The Filter method
- b) The Phasing method

a. The use of a band-pass filter is the conventional method. A good side band rejection can be obtained with filter using four or more crystals. Probably it is the only real solution to building a modern SSB transceiver today. In our case, the problem remains its price. A non-expensive crystal filter with crystals supplied for LSB and USB costs of the

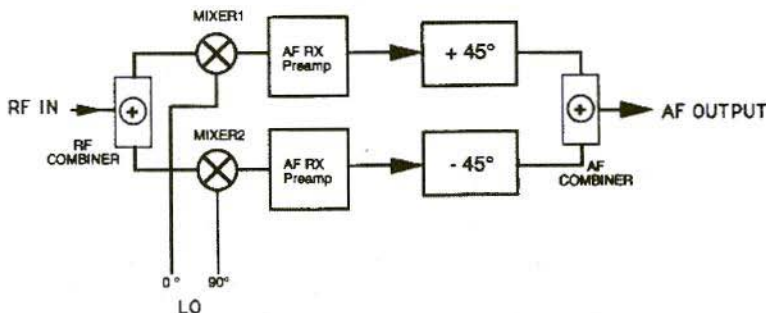


Fig.1: Receiver using the Phasing Method

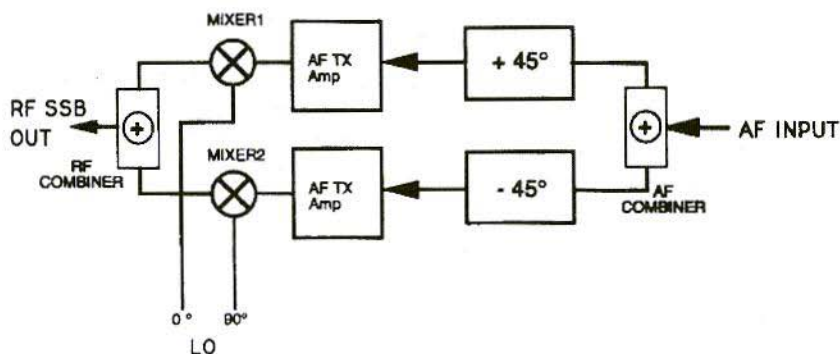


Fig.2: Transmitter using the Phasing Method

order of £25 at least, over 50% of the target price of our unit!

After several tests on scale filters using cheap crystals, I was disappointed with the results (too much disparity between elements). At this time, I read some articles by Mr. Oppelt (*VHF Communications*) and Mr R. CAMPBELL (*QST*) which persuaded me to find another solution.

b. The Phasing method was very popular in the sixties, especially for transmission. This solution is a good one on paper, but in practice the very accurate phase shifting required and the high precision components associated almost completely eliminated this process.

Today, a 1% metal resistor is inexpensive, it is possible to find 5% stable capacitors very easily, the development of logic circuits permits clocks to be built generating in perfect quadrature. This allows us to realise a different concept of an SSB transceiver working with phasing technology in direct conversion.

Direct conversion is the most simple (and the cheapest) process to employ to

listen to the HF bands. Mr. Campbell and some English amateurs have shown that it is possible to design such a Direct Conversion Receiver which eliminates the opposite side band with the phasing method.

In a receiver, all the functions are available to realise a transmitter, it is simply a matter of input/output connections and levels.

From this, I concluded that the design of a phasing transceiver remained theoretically possible. It is a solution to attempt the DM 100 target price and this unusual technology is for me a very attractive project.

4. PROBLEMS WITH DIRECT CONVERSION RECEIVERS

Direct Conversion receivers are the most simple, but are not without problems.

There are five major problems with these designs which are:

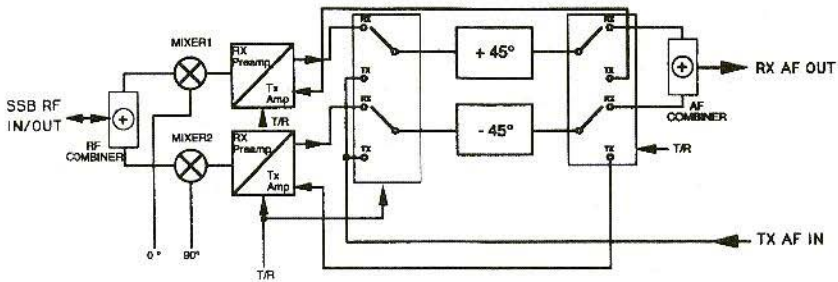


Fig.3: A Transceiver using the Phasing method

The RF hum
 The AM detection
 The AF hum
 The Microphone
 The AF loop oscillations

All these problems were explained very well in an article by Nic Hamilton (G4TXG) in Radio Communication, April 1991.

4.1 The RF hum

The RF hum comes from 50 Hz leaks which modulate the Local Oscillator, providing 50 Hz side bands. These side bands are detected by the input of the receiver causing a 50 Hz buzz in the loud speaker.

With an external mains power supply, the leakage of the LO is phase modulated in the rectifying bridge of the supply by the 50 Hz, re-radiated by external electric lines and received by the antenna! This phenomena is explained in the ARRL Handbook 93

4.2 AM broadband detection (also called in Europe the 'Radio Moscow' effect)

When very strong AM signals are

presented to the mixer, it rectifies (demodulates) the Amplitude Modulation. This signal is always present in output of mixers but rejected by the IF filter.

In a Direct Conversion receiver, it is the AF signal which is retained and amplified, so it is important to filter only the wanted bandwidth, especially on 7 MHz. The strongest station in western Europe is Radio Moscow on 7155 kHz, which is very difficult to eliminate.

4.3 The AF hum

The AF hum comes from 50 Hz leaks directly into the AF stages. In a Direct Conversion receiver, the gain distribution is completely different from that in superhet one. The AF gain is more than 100dB and the AF circuits are very sensitive to the 50 Hz leakage.

4.4 The Microphony

There are two types of microphony, AF and RF. The AF microphony can be easily cancelled by employing good mechanical construction. The RF microphony effect results from the LO leakage at the receiving mixer input, which



is reflected by the input filter. Mechanical actions on the input circuits changes the phase of the reflected LO and gives AF sound in the mixer output. Good capacitors and sealed coils are required to reduce this effect.

4.5 The AF loop oscillations

These oscillations appears when there is a power AF amplifier driving a loud speaker. High currents in the supply wires give voltage differences which are detected by low power stages. Special PCB designs are necessary to eliminate these self-oscillations. This problem is the most difficult to resolve when designing a Direct Conversion receiver with a high level AF output stage. I had to redesign the AF section printed circuit board twice because of this problem.

5. TRANSCEIVER BLOCK DIAGRAMS

5.1 A few words about the Phasing Method

This process was explained very well by Mr.Oppelt in VHF Communications 2/87. In short, two oscillators are generated in quadrature, giving 0 and 90 phase difference to the converted signals after mixing. After this, two AF phase shift networks (+45° and -45°) achieve the phase difference of 0° and 180° to remove the unwanted side band and keep the other.

If the phase difference is maintained to within 1 difference and levels to within 0.1% of magnitude, 40dB of side band rejection is possible theoretically. Table 1 gives the rejection levels for various phase accuracies.

Phase Accuracy (°)	Rejection (dB)
0.25	53
0.5	47
1	41
2	35
5	27
10	20
20	14

Table 1: Rejection Level versus Phase Accuracy

The block diagram in Fig.1 shows the AF Phase shift method for receive and the block diagram in Fig.2 the same method used for transmit. Both diagrams are very similar and it is a simple matter to combine them to achieve receive/transmit operation. The block diagram in Fig.3 shows this method employed in a transceiver.

5.2.1 Preamble

To achieve the target of DM 100 construction price, the maximum number of circuits should work in receive and transmit.

The solution researched during the phase development allows the use of most of the stages in both modes, excepting the transmit power amplifier and two operational amplifiers. Even the AF amplifier is used in the transmit

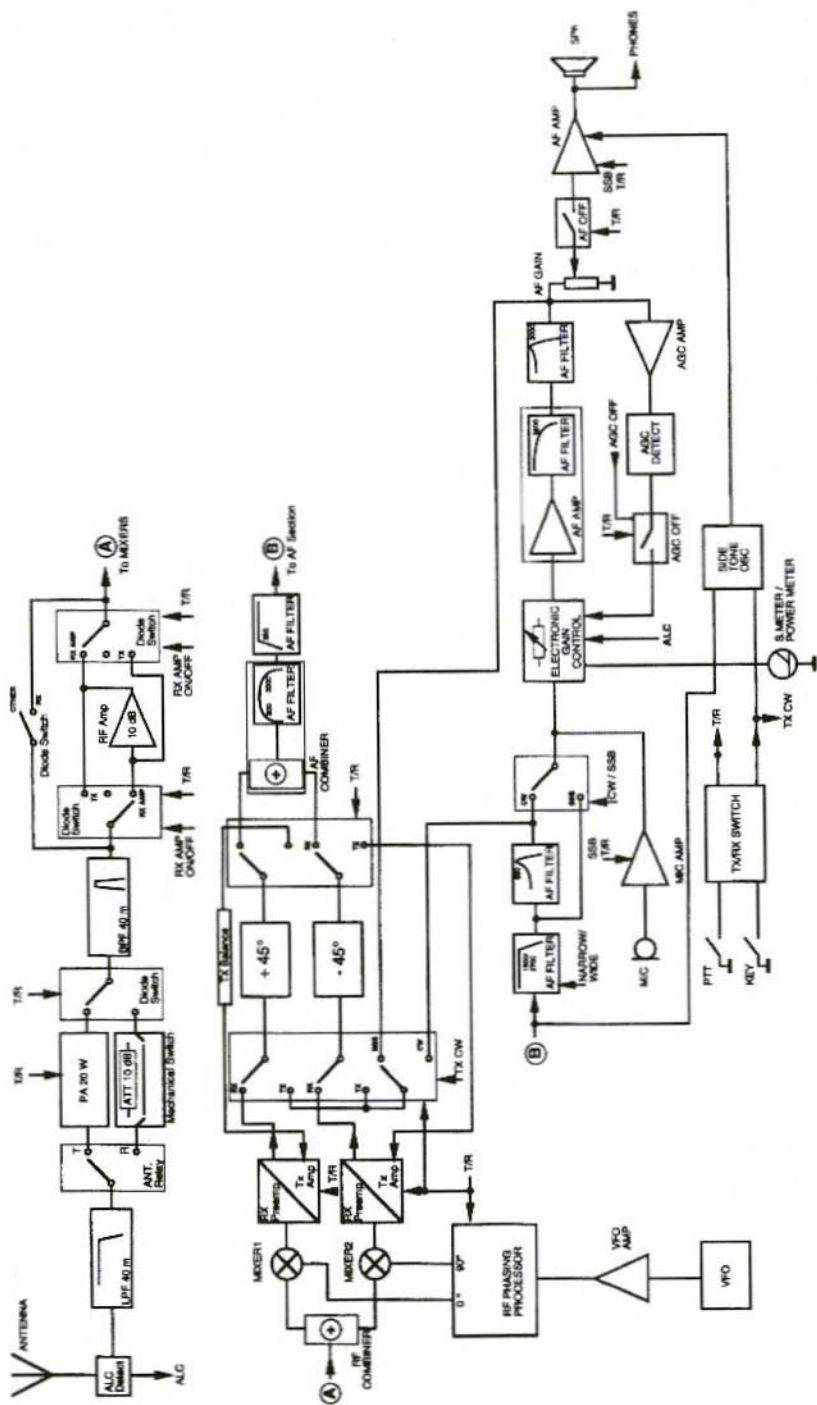


Fig.4: Block Diagram of the Complete Transceiver



mode for CW sidetone operation. A block diagram of the complete transceiver is shown in Fig.4.

5.2.2 Receive mode operation

The received signal from the antenna is fed by the transmit/receive antenna relay to the transmit low pass filter. Putting the LPF at this point provides a cleaner signal for the pre-mixer bandpass filter for only 0.5dB additional losses.

After a 10dB switchable attenuator (mechanical switch), the received signal passes through a diode switch and through the bandpass filter. More diodes are used to switch on or off the RF Pre-amplifier. This 10dB amplifier is not absolutely necessary, but it enables the transceiver to be operated using lower gain antennas (e.g. portable or mobile operation).

To protect the pre-amp from out of band strong signals, I preferred to place it after the bandpass filter. This solution is not the best for noise figure, but it is for intermodulation products. Because of the large losses produced in mixers, the global noise performance changes by only 1 dB, the preamp may be placed before or after the LPF.

The noise figure characteristic is not the most important consideration on the low HF bands, so it is preferable to place the preamp after the BPF. In normal mode the preamp is by-passed using a diode.

The signal is then divided into two equal components by a splitter. Both mixers receive a local oscillation from the RF processor (in quadrature for phasing operation), and the split RF

signal. The mixer outputs are amplified by two receive preamps before the $+45^\circ$ and -45° phase shifting via an AF switch. The Rx preamps are audio ones because with the direct conversion process, the AF signal is already present in the mixer output. A perfect amplitude and phase symmetry is required along this line.

After combining, there is only one side band. A first stage amplifies and coarsely filters the audio signal. A true HPF eliminates the low AF frequencies which are not necessary for voice recognition.

The next stage is an AF low pass filter. To reduce the number of DIL packages (cost), the same filter serves for SSB WIDE" and "NARROW" and helps the CW filter in narrow mode.

The AF is now filtered and must be amplified. An AGC function is required for the comfort factor discussed earlier. An AF electronic gain control is used to develop the AGC. Before the AGC detection, it is necessary to amplify and filter the audio again. Please note that the positioning of the AGC detection is after the complete filtering, because, so as not to cause desensitisation of the receiver by out-of-band signals. The filter function is split into two parts for better working in transmit mode.

The AGC can be switched "ON" or "OFF" for certain QRM, QRN operations. The S-meter is connected directly to the AGC line.

The AF signal is amplified to loudspeaker level using a standard IC, an AF switch is necessary to cut off the AF amplification during transmit.

5.2.3 SSB Transmit operation

In transmit mode, the signal passes through these stages in reverse.

The microphone voltage is amplified by a switched audio amp activated only in Tx. This AF signal is amplified and filtered by the second line of AF filters (the same ones used in receive) and the AGC is now used for the Automatic Level Control (ALC).

The AF switch cuts off the AF to the loudspeaker amplifier, and the AF is switched to the $+45^\circ$ and -45° phase shifters via the logic switches.

The two receive AF amplifiers are changed in impedance and the low impedance signals modulate the mixers. Only one side band RF signal remains after combining.

The receive preamplifier is used as a post mixer amplifier with diode switching. After filtering in the band pass filter, the RF signal is sent to the PA via another diode switch, amplified up to a level of 20W, switched again via the antenna relay and filtered by the output low pass filter.

The ALC detection is fed to the electronic gain control and gives 4 to 5dB of compression.

5.2.4 CW Transmit mode operation:

A Tx/Rx circuit detects the keydown, switches on a squarewave sidetone oscillator and switches the transceiver into transmit with an adjustable delay. This squarewave is sine converted by the AF receive CW filter and fed to phase shift networks.

In transmit, there is only one carrier which is the converted AF note in single side band. The ALC function is deactivated during CW Tx operation and the microphone amplifier is switched OFF. The AF tone is also sent to the loud speaker amplifier for sidetone generation.

The advantage is that, because the AF passes through the receive CW filter, the transmit frequency is the same as the receive one, without oscillator shift and spot button.

6. CIRCUIT DESCRIPTION

6.1 The Mixer (Fig.5)

The mixer is one of the most important part of a receiver. In a Direct Conversion receiver this stage should be able to offer resistance against intermodulation and also to the AM broadband detector effect. After some tests, I first eliminated active mixers (MosFETs and special IC circuits) which required sharp IF filter in the input circuitry.

Schottky diode ring mixers like SBL 1 or so gave the best results. The problem was that AM stations are on the air very often, even during the day.

The low threshold of Schottky is probably not best suited to this problem. (And not for price - remember the DM 100 !) To increase the threshold of the diodes, one possibility is to use silicon diodes and another is to put diodes in series. The performance against AM



Broadband detection is then better and the intermodulation too.

The input signal is split in TR2. The impedance here is 25Ω ($2 \times 50\Omega$ in parallel), and TR1 is necessary to match to the 50Ω RF band pass filter. Each transformer is wound on one ferrite bead and the diodes are 1N4148.

The mixer output is terminated by a 51 Ohm resistor for the RF (R001 and C001). The filter is constructed with 3 coils (3 ferrite beads) for cost reasons. Also, the core effect of a bead is important for reducing the possibility of 50 Hz leaks. This filter is very simple

(too simple) and must be changed for a more sophisticated one in the near future.

Performance of the mixer

This mixer was tested for losses and intermodulation (with an L0 power of +16dBm):

- 1) LO: 10 MHz; Input: 0.1 to 10 MHz
Loss: -5dB
- 2) LO: 100 MHz; Input: 0.1 to 100 MHz; Loss: -7dB
- 3) LO: 10 MHz; Input: 2 carriers near 5 MHz at -8dBm
3rd Intercept point: +22dBm (in)

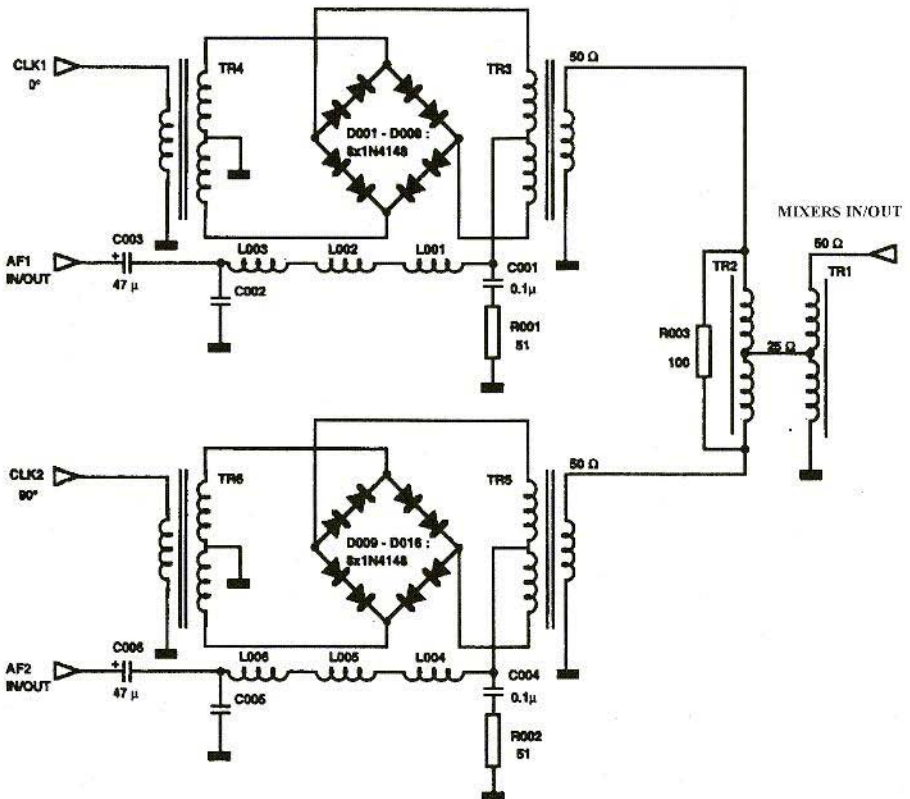


Fig.5: The Mixers

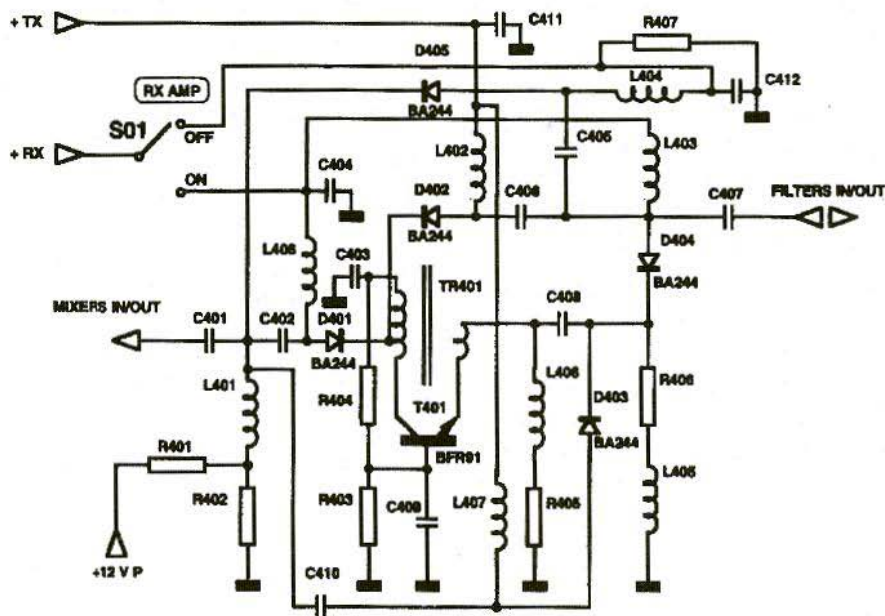


Fig.6: The RF Preamp

6.2 The HF Preamplifier

Requirements:

We need to amplify the signals in receive and also in transmit. In receive we must be able to switch the preamplifier on or off. In transmit a post mixer amplifier is used to increase the S/N ratio before the band pass filter and to present the required level to the Power Amplifier. A solution is to use two separate amplifiers with a diode switch, another is to use the same amplifier with a bidirectional switch (Fig.6).

A Norton stage is used in the amplifier. This kind of amplifier is very resistant against intermodulation and has very low noise characteristics (ca. 1dB). To save components, biasing of D401 and D402 is provided by the transistor current consumption (15mA).

During transmit +Rx is OFF, +Tx at 12 V, D402 and D403 conduct and T401 is supplied with DC. The mixed signal is output on the emitter of the BFR91 and amplified by about 10dB. D405 is biased in reverse by R401 and R402 to have low capacity and cancel the loop effect.

In receive +Tx is OFF, +Rx at 12 V. If Rx AMP is OFF, D405 conducts, and the other diodes are blocked. The Rx signal is feed to the mixer without amplification.

If RX AMP is ON, D404 and D401 conduct and T401 is supplied. The input signal is sent to the mixer with 10dB of gain.

All chokes are wound on a ferrite bead. TR401 is wound on two ferrite beads. Only one additional diode is necessary



to change this stage to a dual transistor system, and the price of a second amplifier is saved.

order preamplifiers, etc.) I reverted to the diagram of Mr.Lewallen and Mr.Hayward which was the most linear under strong signal conditions. This design was modified to ensure an Rx/Tx operation (see Fig.7):

6.3 The AF Amplifier and Mixer

After some tests on various circuits (low noise operational amplifiers, voice re-

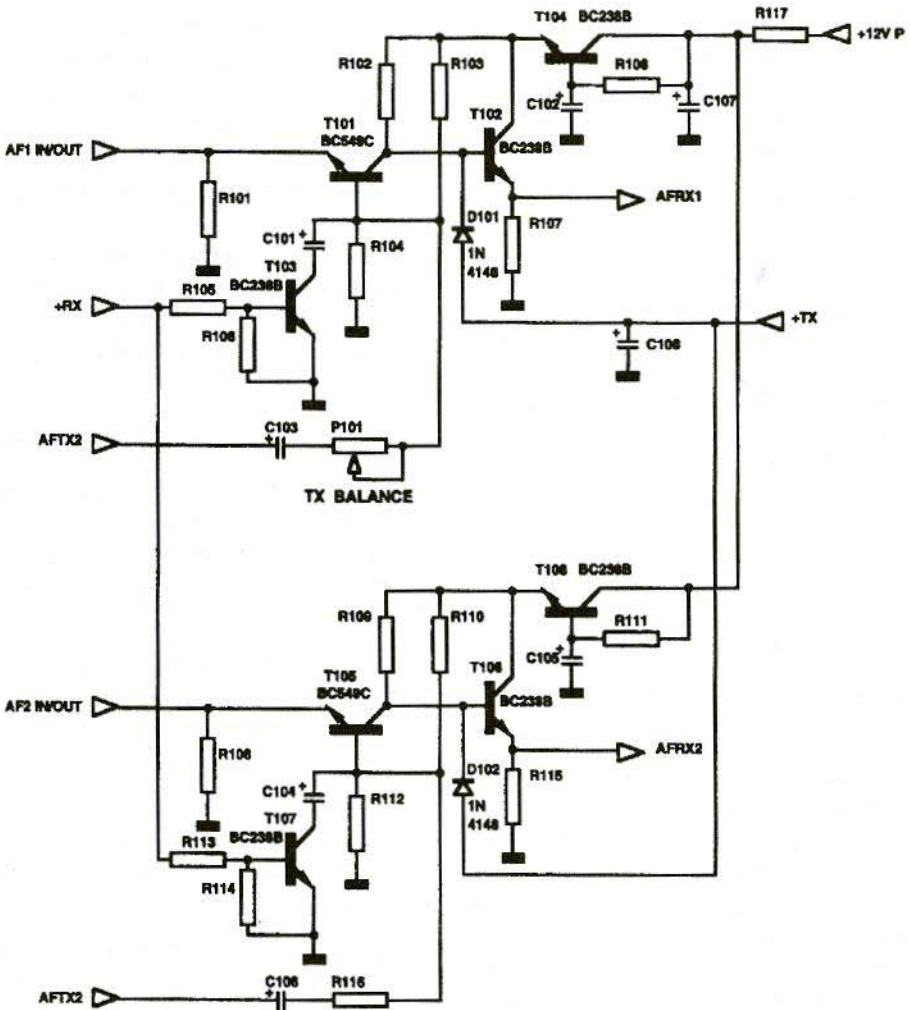


Fig.7: The AF Preamp and the Mixer Driver Stages

a. Receive mode:

+Tx OFF, +Rx = 12 V:

T103 conducts, the base of T101 is AC grounded by C101. D101 is blocked. This ground base amplifier is biased with 500mA to provide a 50 Ohm termination to the mixer

$$Z = 26/IE \text{ (IE in mA)}$$

A low noise transistor (BC549C) is used to ensure low noise performance. The emitter follower T102 provides a low impedance output. The Gain is fixed by R102 at about 40 dB. T104 is an active supply decoupler.

b. Transmit mode:

+Rx OFF, +Tx = 12 V:

T103 is blocked, D101 is conducts. The collector of T101 is AC grounded by C108. The transistor T101 becomes a common collector with input on the base through C103 and P101. Because the emitter is already connected to the mixer port, there is no need for any other switching. The amplitude TX balance is set by P101.

6.4 The VFO and Phase Shifter (Fig.8)

a. VFO

The VFO is a classic Colpits with a J31 0 and a separator stage T802 (BC238B). This diagram benefits from the noise improvements described by Mr.J.Jir-mann , DB1NV in VHF Communica-tions 3/93. The working frequency of the oscillator is 28 MHz to give the 7 MHz band after division by 4 in the phase shifter.

b) Phase Shifter:

The RF from the VFO is amplified by 7400 gates and drives a high speed 74F74 dual D flip-flop. The two sections of the 7474 are connected as a digital phase shifter with two outputs 0° and 90°. Because of propagation delay differences in the circuits, outputs are not absolutely in quadrature. Also, in phasing receivers described in the past, the RF Phase Shift Network was realised with passive components and tune-able. I believe that this tuning capability is important to compensate the existing phase differences in and after the mixer and to achieve good unwanted sideband rejection. C902 modifies the propaga-tion delay and adjust the phase differ-ence. Its value is determined experimen-tally during set up. R904 and R905 provide good termination to the mixer

6.5 AF Phase Shift Network (Fig 9)

The two + and -45° shift networks are made from Hilbert transformers (or all pass filters) according to the work of Mr.R.Oppelt (VHF Communications, 2/87).

This process works better than poly-phase or simple shifters with R and C which are sufficient to transmit but not for receive.

R201 to R219 should be 1% metallic film resistors. C203 to C209 are 5% MKH capacitors, 1 or 2% would be better but are not easily available.

The phase difference is theatrically better than 1. IC's are low noise (and low cost) type TL074. Two CMOS 4053 are used for AF switching. After

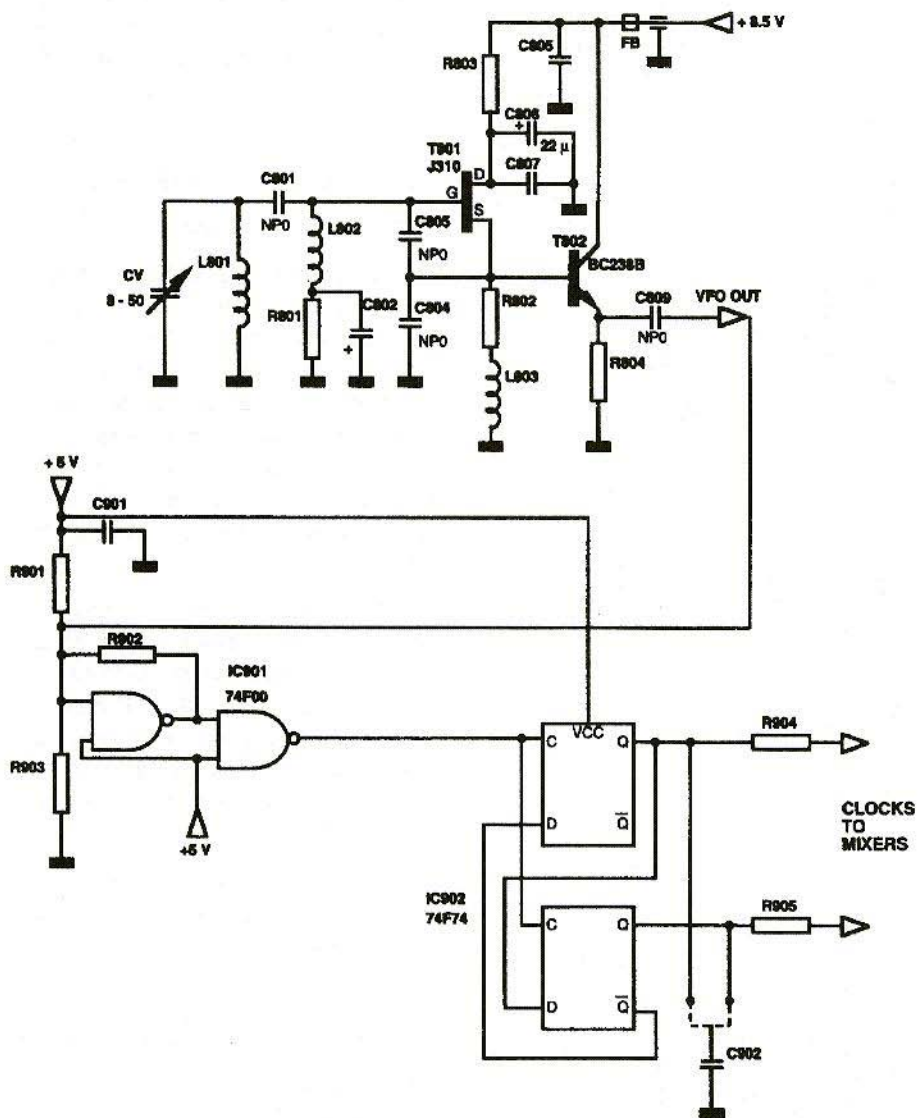


Fig.8: The VFO and RF Phaseshifter

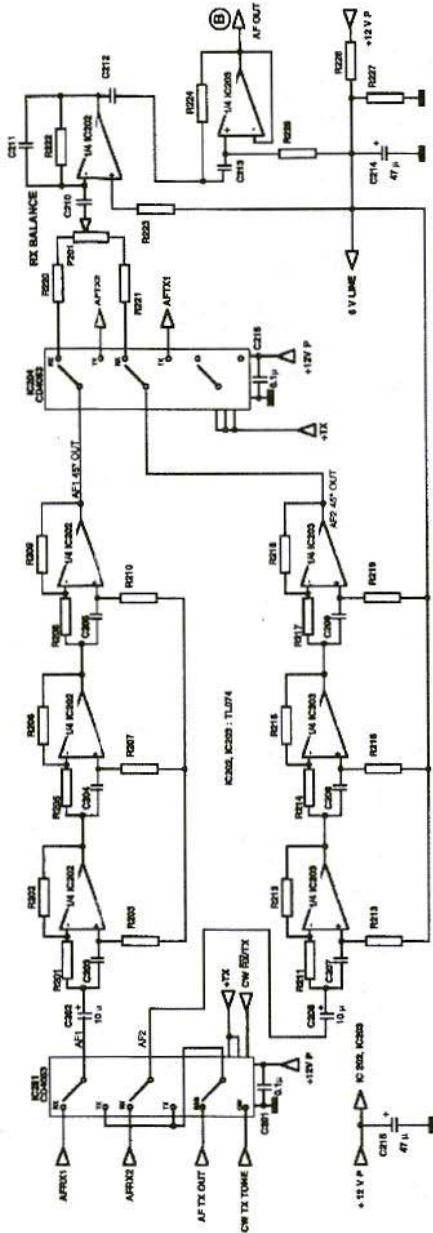


Fig.9: The AF Phaseshifter and Switch

combination in P201, the AF receive signal is amplified and filtered by two opamps. C210 and C211 form a coarse band pass filter and the last stage is a conventional high pass filter.

6.6 The AF Stages (Fig.10)

6.6.1 SSB ICW filter

The signal from the high pass filter is fed to a low pass filter (IC301A). This filter has switchable bandwidth. Most of the time, separate filters are used and switched to provide several bandwidths. In this design, resistors are switched on the same filter by a CMOS 4066 to give wide and narrow positions. This allows saving of an additional quad OP amp package.

In CW mode, IC301A is always in narrow position and assists the CW band pass filter. All CMOS switches are controlled by the +Rx. During Tx operation IC 301A is on "narrow" and IC302A and IC302B are open, providing isolation from the AF microphone signal.

This stage also serves to sine convert the square CW wave before modulation (CW Tx mode).

6.6.2 AGC/ALC/ AF Transmit filter:

RX operation:

In an SSB transceiver, the ALC detection can be carried out on the audio. In a Direct Conversion receiver, the AGC action has to be performed on the AF because there is no IF.

After several tests on specialised IC's, FET's, etc., I found that the most

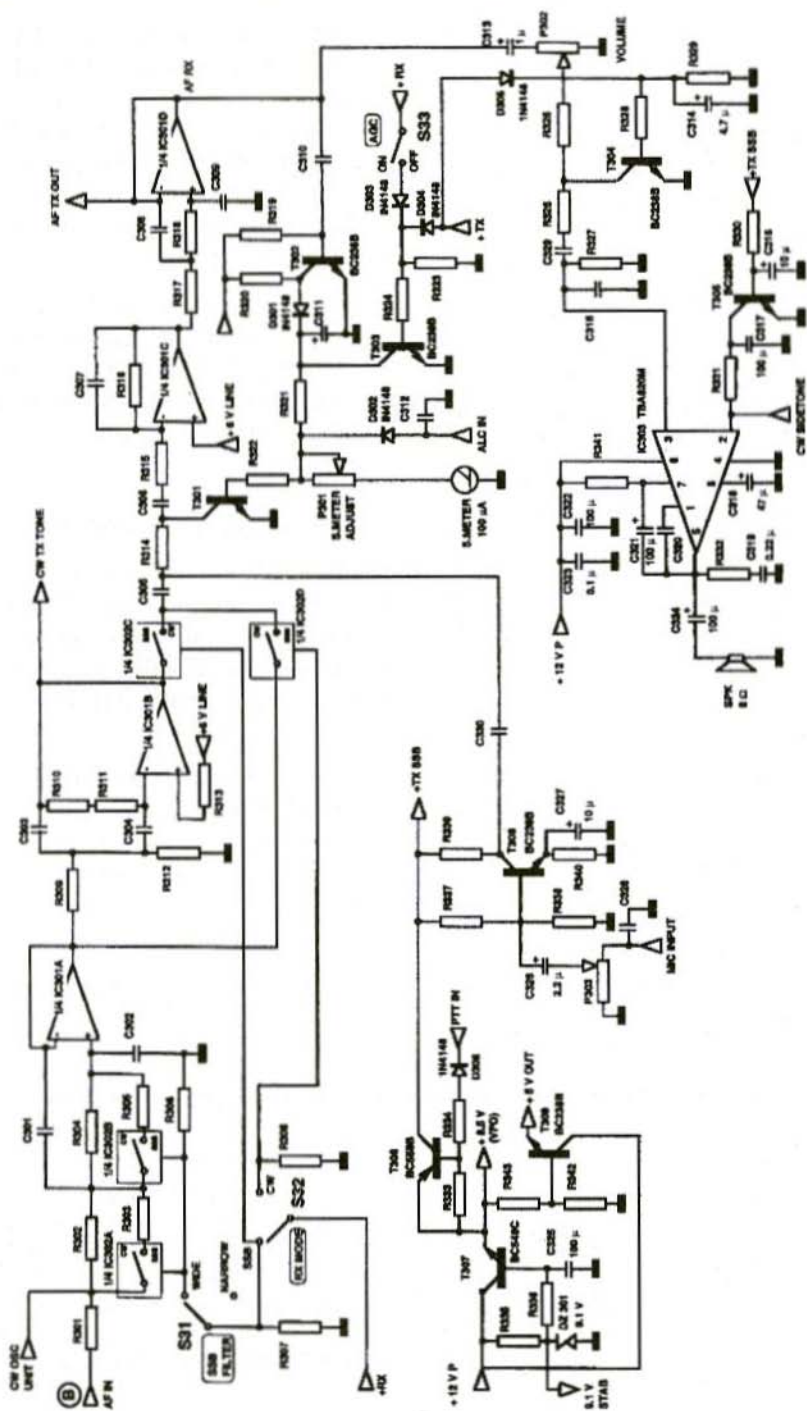


Fig 10: The AF Stage

efficient circuit was a standard transistor (T301) mounted in a variable resistor circuit.

After amplifying and filtering (1C301C-1C301D), the AF signal is sent simultaneously to the loud speaker amplifier and the ALC amplifier (T302). The cut-off frequency of both filters is 3 kHz. This function also takes place in Tx mode.

D301 rectifies the AF and C311 produces the ALC delay, the AGC can be set on or off by means of T303. AGC detection is after the complete filtering to prevent desensitisation by unwanted signals.

The S-meter is connected directly on the AGC line. A 100A model is required and there is no need for "zero" adjustment.

TX operation:

The microphone signal is amplified by T308, which is supplied only in Tx (T306 switch). The signal is amplified and filtered by IC301C and D (as with the Rx signals). The AF filtering is important for phasing transmitters because there is no crystal filter to limit the HF spectrum.

The AGC line is grounded in Tx, and T301 is now controlled by the ALC line through D302. With only a few components, this function provides an appreciable comfort during speaking as well as 4 to 5dB compression. The ALC voltage is also used to display a power indication on the S-meter.

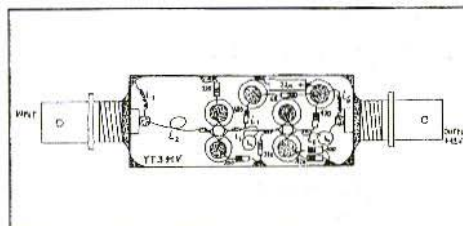
In CW, T308 is not supplied and the CW signal is sent directly to the AF phase shifters (CW Tx TONE). There is no AGC action but the power level is still displayed by the S-meter.

6.6.3 Loudspeaker Amplifier

A TBA820M was chosen for its low price and good performance. The most difficult task is to block AF to the loudspeaker during voice Tx operation, T304 and T305 ensure this mute function.

For CW side tone operation, the CW tone is injected in the chip via the gain setting pin-2. It is an unconventional method but the audio reproduction is satisfactory.

To be continued.



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Dr. Hans Sapotta, DF8FR

Push-Pull Receiver Front End Stages in Common Base Circuits

The Solution to the Inter-Modulation Problem

This article describes the wiring up of bipolar transistors in a common base circuit as receiver front end stages. A new common base circuit, the push-pull common base circuit, is presented. Simulation and test results show the unique inter-modulation behaviour of this circuit.

1. THE SINGLE-PHASE COMMON BASE CIRCUIT

This circuit is used for VHF and UHF receivers. It has good high-level signal behaviour which, contrary to widely-held opinions, is considerably better than that of field effect transistors in source circuits. Moreover, it has a low noise factor and adequate stability. Fig.1 shows the basic wiring diagram. The equation below makes the behaviour of the circuit clearer:

This equation describes the transmission behaviour. R_{gen} is the generator resistance, R_{load} the load resistance and I_E the

$$V_{out} = \frac{R_{load}}{R_{gen} + \frac{kT}{eI_E}} \cdot V_{in} \quad (1)$$

emitter current. So the amplification can be greater than one, only if the load resistance is greater than the generator resistance. A high output impedance is always required for this reason.

Thus this circuit comes into the picture only to a limited extent for broad-band applications. In narrow-band applications, the high output impedance required can always be attained using resonance elements.

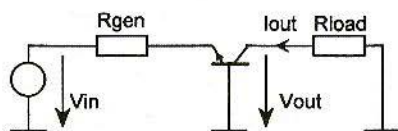


Fig.1: Outline Diagram for a Common Base Circuit

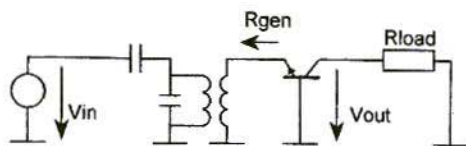


Fig.2a:
Common Base Circuit with
Coupling in through Transformer

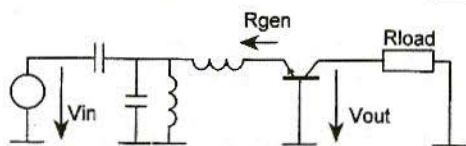


Fig.2b:
Common Base Circuit with
Inductive Coupling

Equation 1 also tells us something about the stability. The common base circuit frequently has a tendency to self-excitation, because a front-end coupling is selected which, as high frequencies are approached, displays a generator resistance to the transistor which tends toward zero. Moreover, these high frequencies are far from the desired operating frequencies. In these circumstances, the transistor is bound to start "singing".

So, a high-Ohmic drive should be selected, right up to frequencies on the other side of the transistor's transit frequencies. The input signal can be coupled in, for example, with an input transformer in the form of a resonant circuit. Another method uses a choke in series with the emitter connection.

The impedance of the input resonant circuit required in each case will be relatively high (500Ω to $2k\Omega$, depending on the frequency of resonance and the grade of service). The minimum noise values for a transistor lie at generator resistances of approximately 80Ω .

Due to the noise in the subsequent mixer stage, it is, of course, usually advantageous to go over to smaller source resistances of 50Ω and to put more emphasis on the amplification.

Fig.2 shows the two possible types of coupling in. The transistor itself has a very low input resistance at normal operating currents, e.g. 5Ω for an emitter current of 5mA . (This should not be confused with the generator resistance required!). In practice, it short-circuits the secondary winding of the input transformer. It can therefore be strongly advisable to give the secondary coil more turns than the primary coil. The resonance inductance is then formed from the primary coil, the leakage inductance and the secondary inductance.

The inductive coupling shown in Fig.2 has the disadvantage in practice that a very high series impedance must be formed at the emitter. The low generator resistance required can then be created once more through resonance transformation.

One very interesting characteristic of this circuit is its inherent linearity. If we look at the denominator in equation 1, what we have there is the sum of the generator resistance and the input resistance, kT/eI_E of the transistor. The input resistance, as previously stated, is relatively low, for which reason the second summand can be virtually ignored. The amplification can thus be given as the



ratio of two resistances - a non-linearity will then no longer occur.

But, unfortunately, the input signal controls the emitter current, and if the current from the input voltage source is of a similar order of magnitude to the emitter current, inter-modulation products can suddenly arise, in any quantity and of any magnitude. The third-order intercept point, which should describe the non-linear behaviour of the circuit, lies relatively high. However, inter-modulation products of the second order, and of higher orders, soon move into the foreground.

In Fig.3, the output power is plotted against the input power with two-tone modulation on the basic wave and the inter-modulation frequencies of the sec-

ond, third and fifth order. The emitter current should be 1mA and the generator resistance 50Ω. The data has been compiled without taking output-side compression into account. It is noticeable that the second-order inter-modulation at the $f_1 - f_2$ differential frequency dominates the inter-modulation behaviour. At the same time compression does occur and the curve for the carrier wave does not take a linear path.

The curves shown represent simulation results, for which reason it can be assumed that for initial levels below -90dBm the digital noise from the simulation program falsifies the results.

If it is desired to improve the high-level signal behaviour of the circuit, this can be attained in several ways. One possi-

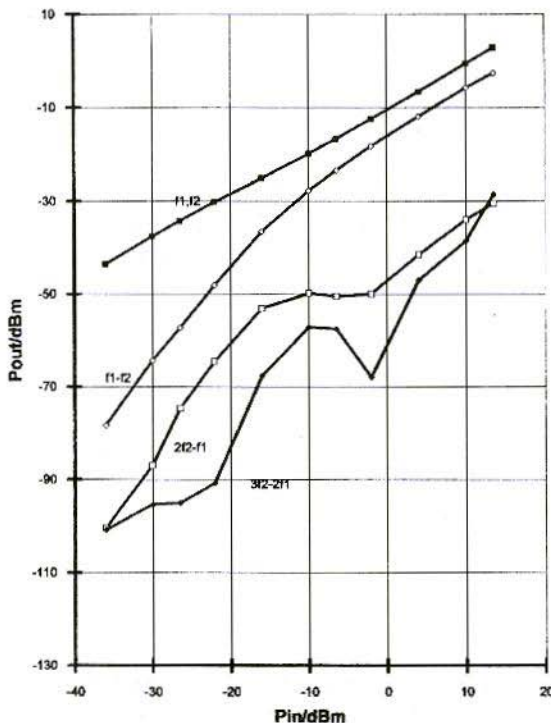


Fig.3:
Output plotted against Input
with Two-Tone Modulation of
Single-Phase Circuit;
 $I_E = 10\text{mA}$

bility is to increase the generator resistance, thereby increasing the first summand in the denominator of equation 1 and bringing about linearisation. The amplification is reduced at the same time. It is possible that you may move away from the noise matching point.

Another possibility is reciprocal coupling. However, since the output signal and input signal are in phase, the reciprocal coupling must be through a transformer. This usually means a lot of work and expense. Moreover, it can lead to self-excitation if the amplifier no

longer displays the behaviour intended for it at high frequencies.

Thus the simplest method for increasing linearity lies in increasing the emitter current. However, this puts additional stress on the battery of the Handy! In the car radio range, where it comes down to high-level signal stability and sufficient current is available, it is usual to have 10mA in the input stage. Of course, the noise factor of the amplifier stage increases as the emitter current increases.

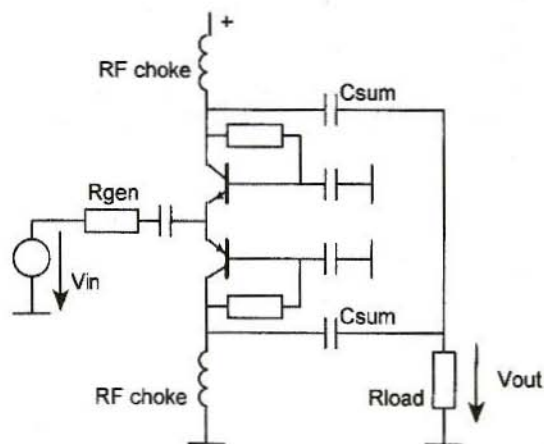


Fig.4a:
Push-Pull Common Base
Circuit with Linked
Emitters

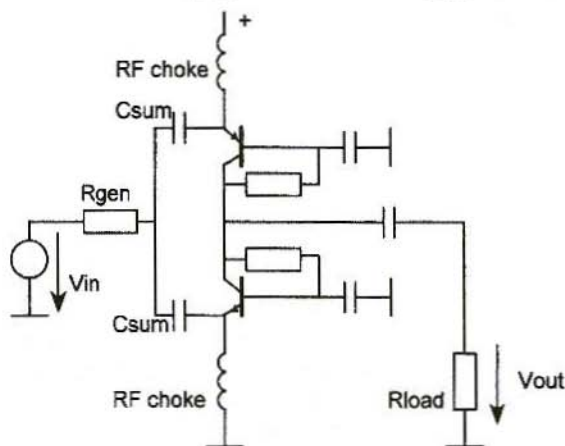


Fig.4b:
Push-Pull Common Base
Circuit with Linked
Collectors

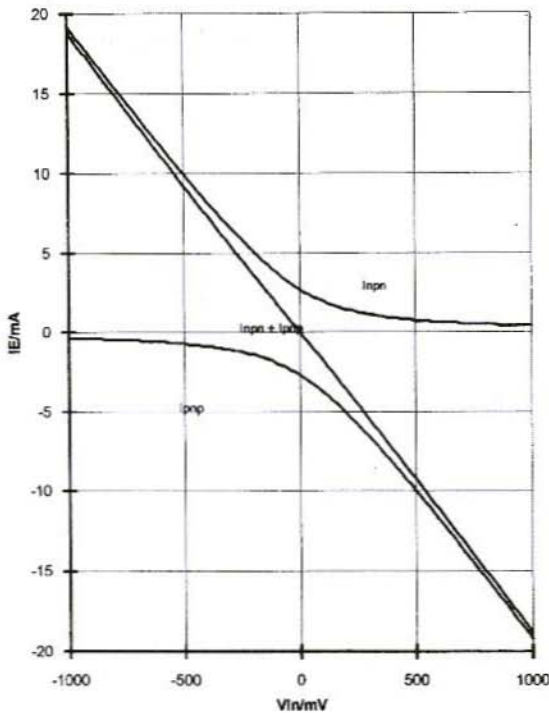


Fig.5:
NPN and PNP Emitter
Currents and their Total,
plotted against Input Voltage

2. A NEW CIRCUIT: THE PUSH-PULL COMMON BASE CIRCUIT

The problem with the push-pull common base circuit described above is the fact that the positive half-wave of the input current reduces the emitter current. This leads to non-linear effects.

It is thus an obvious move to wire up a second transistor, which has contrary conductivity, parallel to the input transistor. This then takes over the positive input current half-wave. The best results will be obtained if the emitter current is set to be identical through both transistors. This is a simple thing to do if the

two elements are wired up in series. For AC voltages, the transistors should be wired in parallel. For DC power supplies, in contrast, they should be wired in series.

Fig.4 shows two possible circuit options. The same conditions apply to the input network as with single-cycle circuits -i.e. low generator resistances at high frequencies should be avoided. In Fig.4, the input network is represented only symbolically, through the generator resistance, R_{gen} . There is no difference in principle between the functions of the two circuits. Practical aspects can thus return to the foreground in the selection of a circuit. Fig.5 shows the result of a simulation. We see the input voltage plotted against the currents through the

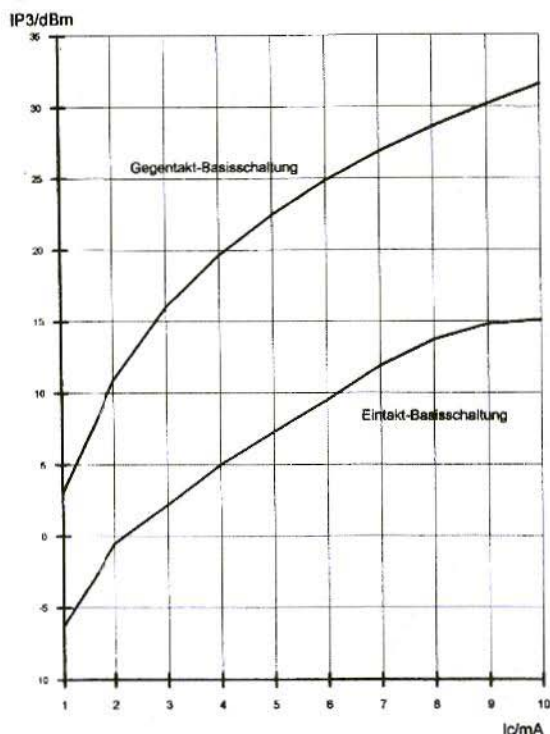


Fig.6:
 IP3 plotted against Emitter
 Current with $R_{Gen} = 50\Omega$
Gegentakt-Basisschaltung =
 Push-Pull Common Base
 Circuit
Eintakt-Basisschaltung =
 Single-Phase Common Base
 Circuit

NPN transistor and the PNP transistor, together with the sum of the two. For the summing function, we recognise the almost ideal linearity. Unfortunately, deriving the actual conditions mathematically does not yield as much as equation 1.

What seems so easy in theory actually is so, even in practise! Adding up the output currents in Fig.4a poses no problem, as bipolar transistors in the common base circuit at the collector junction represent almost ideal current sources. The only thing that needs to concern us is keeping the parallel circuit right down at the lowest differential frequency available. Decoupling with only a 1nF capacitor is not sufficient. A big capacitor should be wired parallel to

it. From the point of view of feed line inductances, the best results will be obtained using surface-mounted devices.

3. SIMULATION RESULTS

The circuit is now complete, and we are eager to see the simulation results, which have been obtained using Spice G26 and Hspice. In Fig.6, we see the third-order intercept point of the classic single-phase circuit contrasted with the intercept point of the push-pull circuit, plotted against the emitter current. The generator resistance is once again as-

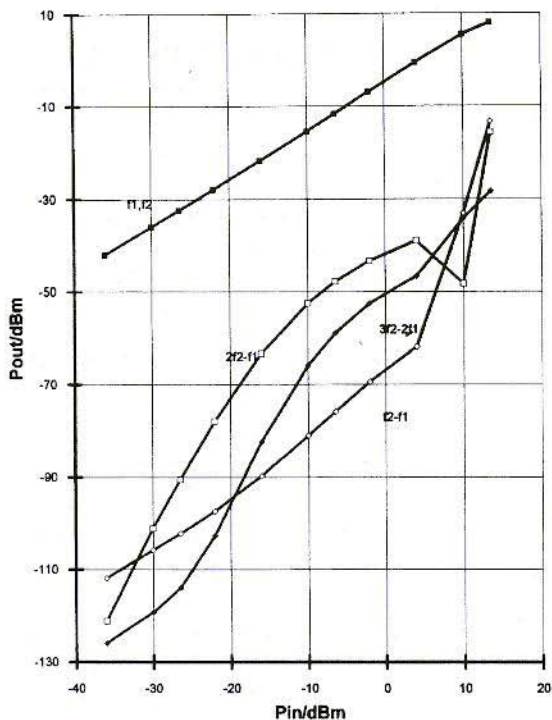


Fig.7:
Output plotted against Input
with Two-Tone Modulation of
Push-Pull Circuit: $I_E = 1 \text{ mA}$

sumed to be at 50Ω . With a constant intercept point, the emitter current is reduced by a factor of at least 3. At 10 mA , a push-pull circuit has an input intercept point of $+32 \text{ dBm}$, i.e. about 1.5 Watts . If we operate this circuit at, for example, 12 Volts , then the DC power consumption is only 120 mW . It is thus considerably lower than the intercept point IP_3 !

Analogously to Fig.3, Fig.7 shows the relationship between the output and the input with two-tone control for the carrier wave and the inter-modulation frequencies of the second, third and fifth order. We see that the inter-modulation behaviour is considerably smoother. The difference becomes particularly apparent for second-order inter-modulation be-

haviour. Of course, it makes no sense to extrapolate a second-order intercept point from these curves. It would be too strongly dependent on the modulation.

4. THE NOISE BEHAVIOUR

With regard to low-level signal control, both stages are wired up in parallel, so there is no problem determining the noise factor. In principle, the same noise factor will be obtained which a transistor would have which was being operated with double the current.

Since the hole mobility in silicon is less than the electron mobility, PNP transistors basically have a slightly higher noise factor. Thus in practise a noise factor will be attained which corresponds to the mean value between the noise factors for the PNP and NPN transistors. Since in the meantime transistors with emitter strip widths of less than 1m have become available, the basic track resistance no longer plays the dominant role in the noise figure, the difference in the noise factor is also scarcely of any importance any longer.

Other noise sources, such as base current Schottky noise and collector current Schottky noise, which technologically can not be prevented, determine the noise factor. It should be possible to attain values of about 1dB.

Of course, in this context we should take it into account that the total noise factor of a receiver today is determined more by the damping of the input circuit than by the active amplifier element. Especially by using miniaturised coils, damping values of 3dB and more can be brought about here, which are then added up to give the total noise factor. If we want to avoid this, the antenna must be permanently fixed to the input circuit. Then the source resistance of the antenna damps the input circuit, which makes it broad-band -and thus opens the way to inter-modulation.

If we leave the inter-modulation behaviour integrally constant against the frequency, when the push-pull circuit is used we can select a more permanent antenna coupling. Thus, with this precondition, even a reduction in the receiver noise factor is conceivable.

5. MEASUREMENT RESULTS

The measurement results now presented were obtained using transistors BFR93A (NPN) and BFT93 (PNP), which are not precisely babes in arms any more. But the results can be reproduced with newer components without difficulty.

The measurements were carried out, not in the amateur radio range, but in the VHF radio range. However, they are valid up to the 70cm band and above.

In spite of using professional equipment, it was not quite so simple to generate a two-tone signal free from inter-modulation. If we wire test transmitters together through power combiners, they are already generating inter-modulation on their own, particularly in the higher performance range which is of interest here. The test transmitters must be wired up to the power combiner through damping elements to dampen the power flowing from the second test transmitter into the first as well. But then the two-tone signal will only remain available at lower power. It was raised to acceptable level ranges again using a weakly modulated 40 Watt broad-band amplifier.

In Fig.8, we see the measurement result for a single-phase stage at $I_E = 10\text{mA}$. Numerous measurements are displayed, with the triggering level rising by 2dB each time. Simultaneously, the frequencies of the two transmitters were slightly shifted. Thus, at the right-hand side of the graph, we recognise the two signals from the carrier wave modulation, and to the left of that the inter-modulation



products. The third-order intercept point can be read off at +15.5dBm, which correlates very well with the value obtained from Fig.6.

If we look at the build-up of the second-order inter-modulation products at the left-hand side of the graph, it is apparent that they do not rise by 4dB, as they actually should according to theory. This result also correlates well with the simulation results from Fig.3. As described in the simulation, the carrier wave signals are compressed as the levels become higher.

By comparison, Fig.9 shows all the measurements for a push-pull common base circuit. The emitter current here was only 8mA, i.e. somewhat lower.

The third-order intercept point is +27.5dBm. This value also corresponds to that obtained from the simulation in Fig.6. The higher-order inter-modulation products are strongly damped. The difference is most marked in the second-order inter-modulation at the left-hand side of the graph. The ratio of the vertical scale here is 10dB/div.

6. PRACTICAL ASPECTS

In principle, the two circuits from Figs.4a and 4b behave identically, but of

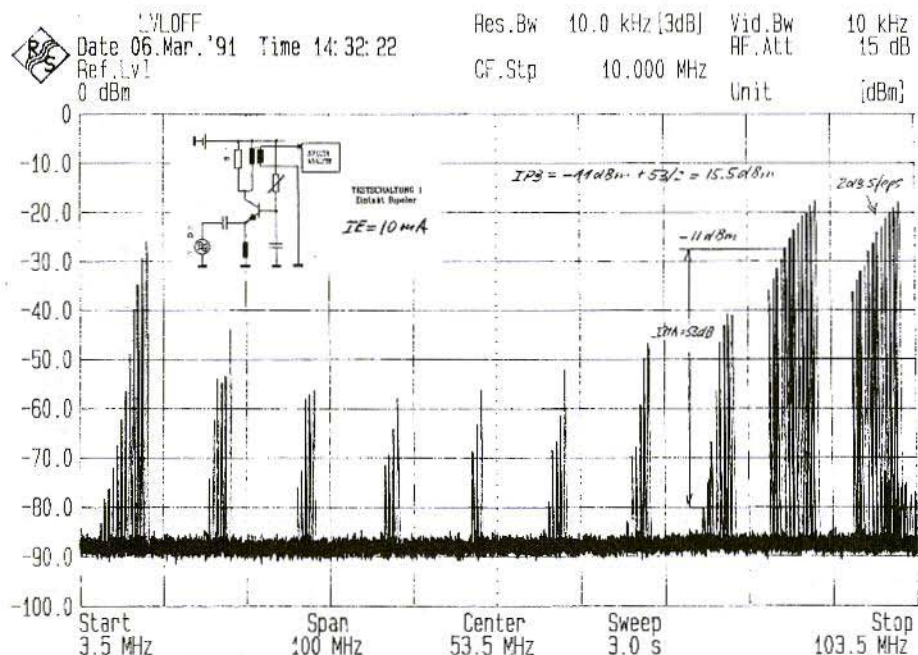


Fig.8: Measurement Result for a Single-Phase Stage with $I_E = 10\text{mA}$ (Input Level increased by 2dB each time and Frequency increased slightly) *Testschaltung* = Test Circuit; *Eintakt* = Single-Phase

course there is a difference in their practical significance. The common base circuit requires a high output impedance for its amplification to be maintained. Thus the chokes in the collector lines of the circuit according to Fig.4a are problematic. Their loss resistance should dampen the output resonant circuit connected only insignificantly. Even if one choke is used as a self-inductive coil, there is still a further high-quality choke left.

The circuit shown in Fig.4b is simpler here. In this case, the choke is in the emitter line, the emitter side of the circuit being low-Ohmic in itself. The loss resistances of the

chokes are wired up parallel to the generator resistance in the equivalent circuit diagram. Thus the chokes can have a negative effect in so far as their loss resistance is in the order of 50Ω . But for that to happen they must already be really poor.

The only problem with the circuit shown in Fig.4b is that it takes rather more effort to generate the base bias. If we use the resistances proposed from the collector to the base, we do obtain a certain amount of compensation for the dispersive current amplification. However, individual smoothing will nevertheless be required.

A more expensive way to generate the

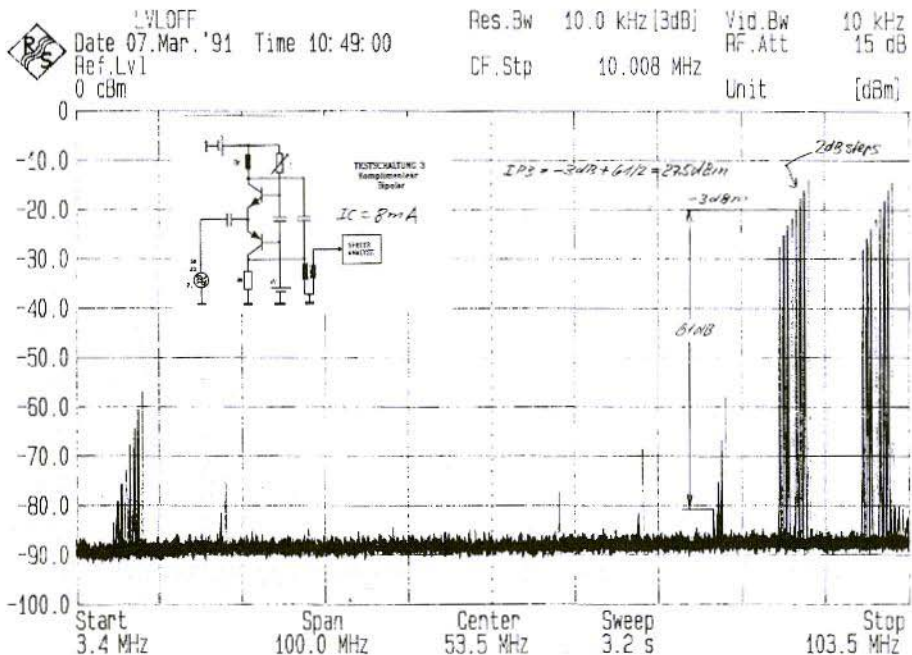


Fig.9: Measurement Results for a Push-Pull Common Base Circuit with $I_E = 8\text{ mA}$ (same format as Fig.8)
Testschaltung = Test Circuit; *Komplimentar* = Complimentary;
Spectranalyse = Spectrum Analysis



base bias is to use diodes individually, from the base against the emitter potential, to balance out temperature effects. Individual pairing of transistors is definitively not required. You should select types from the data book with approximately identical transition frequencies and noise factors. The inherent counter-coupling for the common base circuit means there is no need for a pairing such as used to be familiar earlier for low-frequency high-level stages in collector circuits.

Anyone still not obtaining adequate

intercept points in the watt range is recommended to try reciprocal coupling through a transformer, which is always possible. Should anyone wish to experiment with it, I am looking forward to the results.

One comment -any potential commercial users of this circuit should remember the patent rights aspect.

This article is a revised German version of a contribution to the Wireless Symposium in Santa Clara, U.S.A. in February, 1994.

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IF Amplifier with Wide Range of Adjustment

Stimulated by the suggestions for expansion of the spectrum analyser from DK6ZK proposed in [1], I tried to find an integrated circuit which has the function of a regulated intermediate-frequency amplifier. During the search, I came upon a well-known component from Motorola, the MC1350 gain block, an integrated monolithic intermediate-frequency amplifier.

1.

DESCRIPTION OF CIRCUIT

The biggest advantage of this circuit lies in the fact that both the amplification and the damping can be adjusted within wide limits using the AGC input. According to the data sheet [2], the power gain is 50dB and the AGC range -60dB.

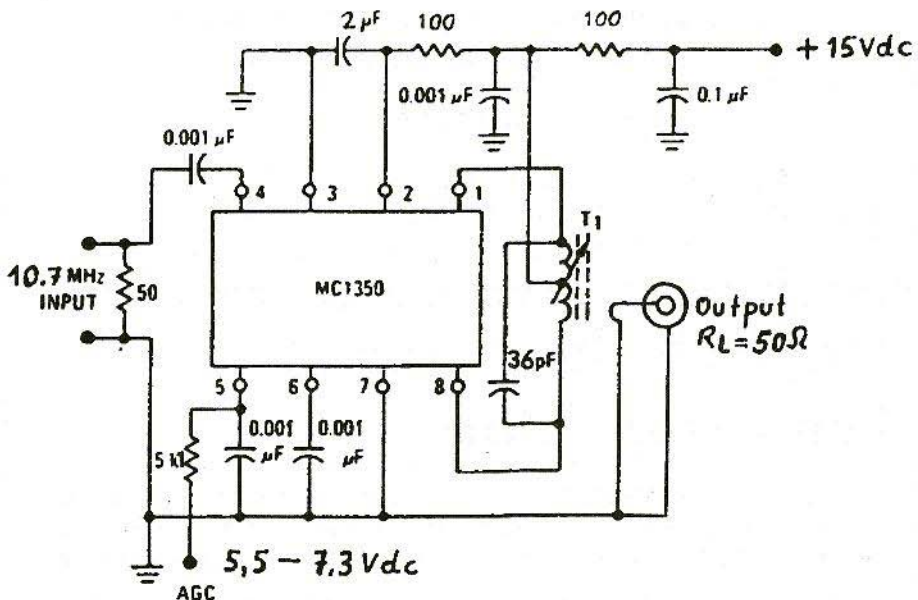


Fig.1: A 10.7 MHz Amplifier with Input and Output Matching

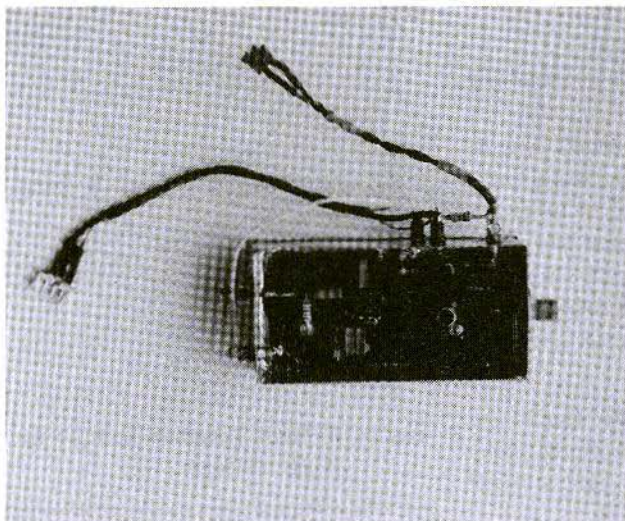


Fig.2: Prototype of Intermediate Frequency Amplifier

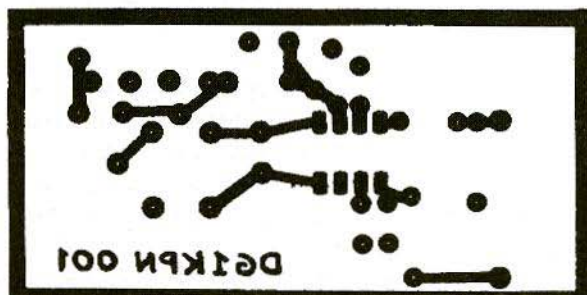


Fig.3: Layout of Printed Circuit Board DG1KPN 001

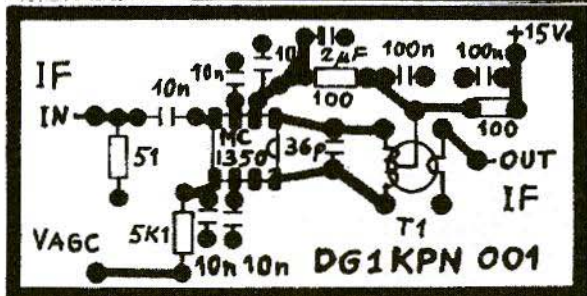


Fig.4: Component Overlay

Fig.1 shows the wiring diagram for the 10.7 MHz amplifier, following the data sheet. The amplifier input is automatically matched through a 50Ω resistance. The collector voltage for the output

stage of the amplifier must be fed through the centre tapping of T1 to pins-1 and 8 of MC1350. The coupling coil transforms the impedance to 50Ω.

2.

CIRCUIT ASSEMBLY

The transformer, T1, is wound onto a 6-mm. diameter coil former using 0.13 mm enameled copper wire. 24 turns are wound around the coil former and fixed on using mixed adhesive. 2.5 turns are wound over the primary winding in the vicinity of the centre tap as coupling, and these are also fixed on.

The inductance of the coil should be app. $6\mu\text{H}$, so that it can be brought into resonance at 10.7 MHz with a 36pF capacitor.

Fig.2 shows the amplifier prototype, and Figs.3 and 4 the layout and the component overlay for the printed circuit board, which should be soldered into a standard tinplate housing. The component side of the PCB is a continuous ground plane.

Signal inputs and outputs are fitted with SMA sockets. The 15V DC supply and the AGC voltage are fed through InF feedthrough capacitors.

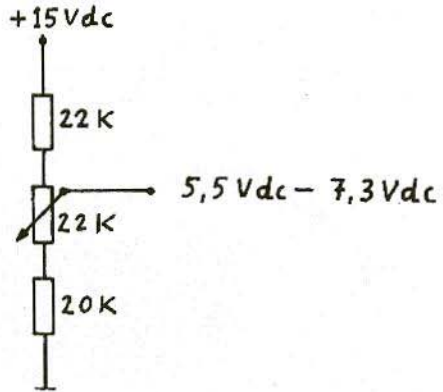


Fig.5: Voltage Potential Divider for AGC Voltage

3.

PUTTING INTO OPERATION

Once all the components have been mounted on the printed circuit board, the supply voltage is applied and the AGC input is connected to earth (maximum amplification). Precisely 10.7 MHz is then fed in at app. -50dBm. The output should then be approximately 0dBm. T1 is tuned for maximum power output.



Fig.6:
Screen Photo: 10 MHz
Signal rich in Harmonics
(Quartz Filter off; Top
Line -10dBm)



Fig.5 shows the resistance values for the voltage potential divider, which generates the AGC voltage required from an input voltage of 15V DC. This voltage must be variable between app. 5.5V DC and 7.3V DC.

The 22k Ω potentiometer of the spectrum analyser's frame wiring can be used for this purpose, by setting the supply for the 465 MHz intermediate-frequency amplifier to 15V DC permanently, as it no longer needs to be adjustable. So no additional potentiometer is required, which makes fitting into the spectrum analyser easier.

For a direct voltage of 5.5V DC at the AGC input, we obtain an amplification of app. 45dB. 0dB can be obtained at approximately 6.7V DC, and app. 60dB damping at 7.3V DC.

4.

FITTING INTO THE SPECTRUM ANALYSER

This intermediate frequency amplifier circuit is very general in its range of application. Here, for example, it is incorporated into the spectrum analyser created by DBINV. The intermediate frequency amplifier is simply looped in by means of a coaxial cable between the assemblies DBINV 006, IF out, the quartz filter, DBINV 008, and IF in.

TI can now be fine tuned in the spectrum analyser at the maximum amplitude of the zero mark using the display.

When the spectrum analyser is operating normally, the gain potentiometer is set in such a way that, by and large, the noise level peaks precisely in the vertical position on the display. At this point, the amplification is precisely 0dB.

Starting from this setting, we can now increase signals by app. 40dB or reduce them by app. 50 - 60dB, depending on their power - practical testing will tell you far more about this than a long description from me!

With my spectrum analyser, I can identify signals satisfactorily, even at -120dBm, with maximum amplification from the 10.7 MHz amplifier and in a selected band width of 1 kHz.

5.

COMPONENT LIST

1 x	MC1350
1 x	51 Ω
2 x	100 Ω
1 x	5.1k Ω
1 x	2.2 μ F
2 x	100nF
3 x	10nF
1 x	36nF
1 x	22k Ω potentiometer (see text)
1 x	22k Ω
1 x	20k Ω
1 x	coil former, 6mm diameter, with core, colour coded red, 50cm enameled copper wire, 0.13mm diameter
1 x	tinplate housing, 74 x 37 x 30mm
2 x	coax flanged bushes, SMA

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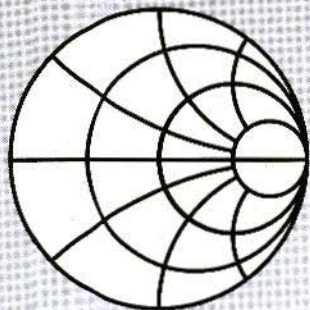
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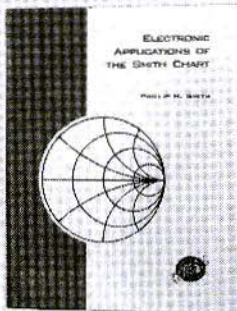
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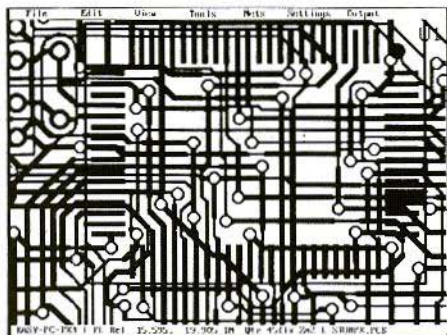
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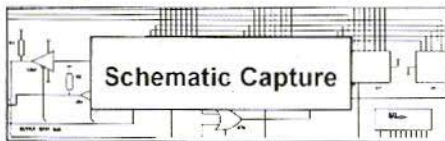
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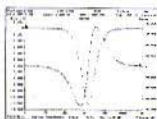
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