A Very High Dynamic Range LNA for 144 MHz

Peter Hoefsloot, PA3BIY

Abstract: The pre-amplifier described here uses the latest technology of pHEMT devices. This technology enables the development a pre-amplifier with a noise figure less than 0.3 dB and an Input 3rd order Intercept Point (IIP3) of +2.5dBm. The combination of such a low noise figure and high IP3 is unique, and to my knowledge not met by any commercial pre-amplifier for the 2 metre band. The amplifier is unconditionally stable.

Introduction

Low noise amplifiers have always fascinated me. Though critical by design, they were my first steps in building an all mode 144 MHz transceiver in the early 80's. In fact, the first pre-amplifiers that I built (1978) did just one thing: oscillate! My fascination originated form the (wrong) concept, that a low noise amplifier improves the signal to noise ratio. As I will explain in this article, it does not! Nevertheless, a low noise amplifier is very important at VHF and higher, in order to be able to detect the often very weak signals. However, in order to obtain a low noise figure, amplifiers often had to be designed in such a manner, that strong signal handling suffered. The advent of new devices enables us to design amplifiers which are able to combine a low noise figure with strong signal handling capability.

The following topics will be addressed in this article:

- Theory on signal-to-noise ratio's (describing the relation between ambient noise, pre-amp noise figure and signal to noise ratio);
- The importance of (high) dynamic range of a pre-amplifier and the available design options (short overview of available techniques, and the introduction of some interesting recent developed devices);
- Designing the ideal Pre-Amplifier (issues are described which must be met, in order to obtain low noise figure and linearity;
- Circuit description and construction of the 144 MHz LNA; Putting it to operation).

- Stability
- Important qualities of a pre-amplifier and how to measure them (Noise figure, IP3 and return loss measurements are described); Measurement results (2-tone test results, return loss and gain sweeps are shown in plots. Also the result of the noise figure measurement is presented);
- Conclusions

Einführung

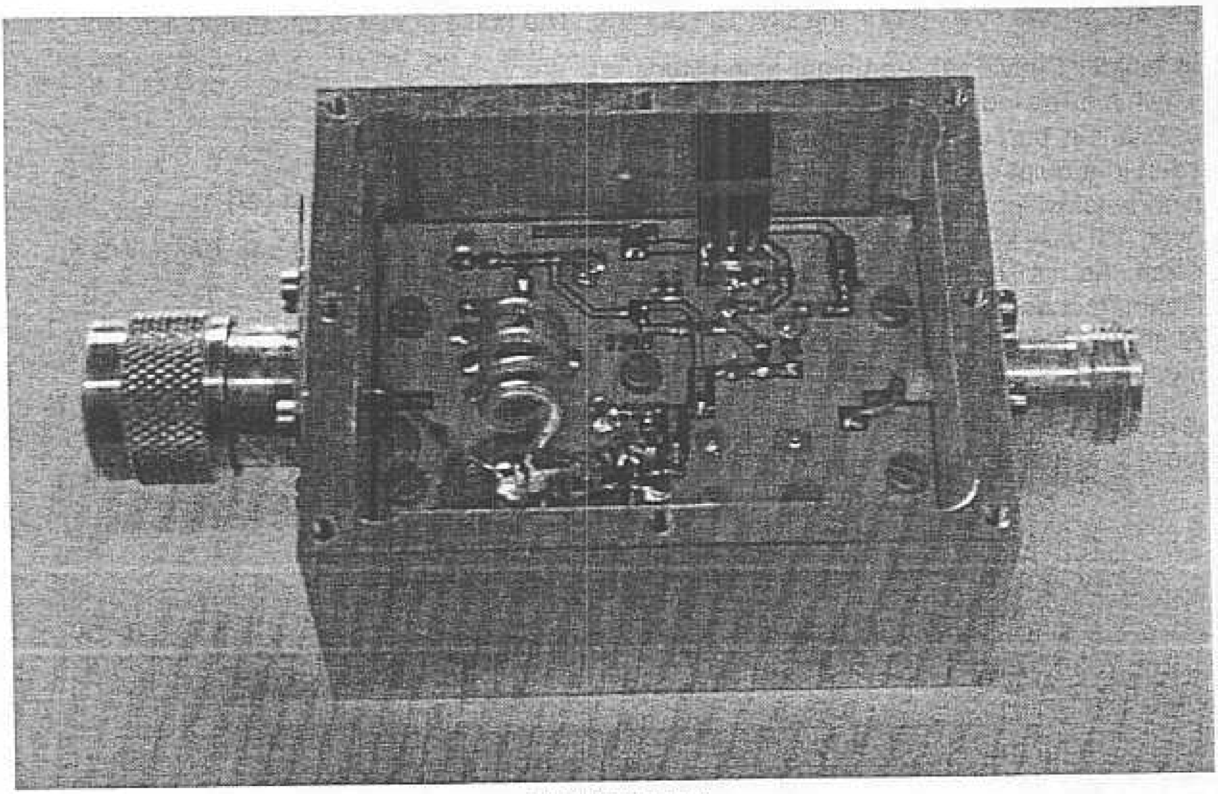
Rauscharme Verstärker haben mich immer fasziniert. Die Anziehungskraft kam von dem - mittlerweile als falsch erkanntem - Konzept daß diese Verstärker das Signal/Rausch-Verhältnis verbessern sollten. Ich werde im nachfolgenden Artikel erklären, daß dieses keineswegs der Fall ist. Trotzdem sind sie wichtig, um auf VHF und höheren Frequenzen schwache Signale zu verstärken. Oft wurde zugunsten der Rauschzahl das Großsignalverhalten geopfert. Nun ist es neue Transistoren, mit denen sowohl eine niedrige Rauschzahl als auch ein hoher IP3 erreichbar ist.

Folgende Themen werden behandelt:

- Details zum Signal/Rauschverhältnis
- Intermodulationsverhalten
- Design des idealen Preamp
- Stabilität
- Wichtige Kenngrößen und deren Messung (Rauschzahl, IP3, Rückflußdämpfung)
- Folgerungen

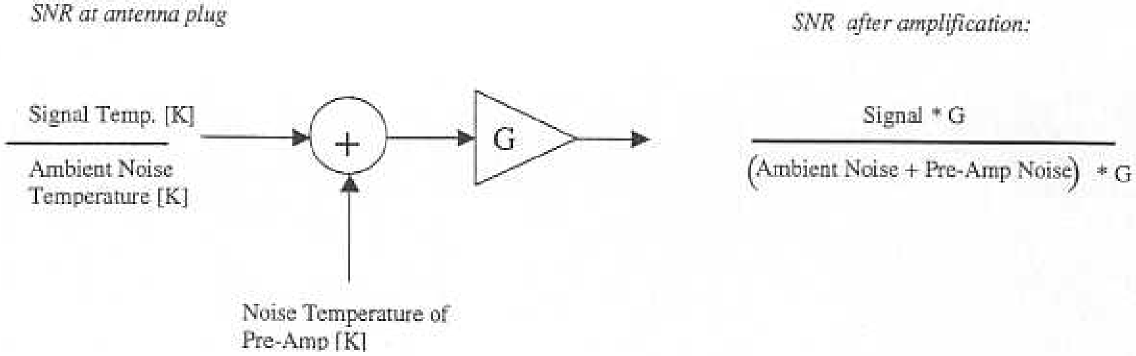
Signal-to-noise ratio

Low noise amplifiers are the logical choice of every first stage in VHF and up receivers. Their purpose: to minimise the deterioration of the signal-to-noise ratio as seen on the antenna connector. A pre-amplifier will never improve the signal-to-noise (SNR) ratio! The SNR at the antenna plug is the best we are ever able to obtain. Any noise added by an amplifier will deteriorate the SNR, so we want



144MHZ LNA

Bild/Figure 1: SNR degradation, due to pre-amplifier nois



to minimise that amount of noise. The ambient noise (man made, terrestrial or galactic noise) sets the limit to the SNR and can be expressed as dB above the thermal noise at T0 (T0=290 K) or as an equivalent temperature [K] at which the thermal noise would be of the same value. In order to calculate SNR and to perform calculations, it is often easier to describe noise and wanted signal as temperature [K] rather than in dB. The power produced by a noise source is related to the temperature of that source. An obvious example we can see every day in the sky: The Sun!

In table 1 I have set 3 ambient noise levels which may be found on 144 and 432 MHz. For calculation purposes I'll use a low level as seen in a quiet spot in the sky (EME), and a high level which is common in a city. Moreover, I'll use a very low value for ambient noise, as can be observed on 70 cm and higher frequencies when pointing into the sky. At these frequencies the Galactic noise is much lower than on 144 MHz.

In order to assess the SNR we need to transform it to an equivalent noise power and divide it by the ambient noise level. In figure 1 I have described the process between antenna and pre-amplifier in terms of noise power. The noise contribution of the pre-amplifier is transferred to the summing device (+). The amplification (G) is considered noise less, because the noise contribution is transformed to the input.

The bandwidth in the calculations is set to 2.4 kHz, being the most common used by amateurs.

The noise power is calculated as:

[1] $P_n = k \cdot T \cdot B$

Where:

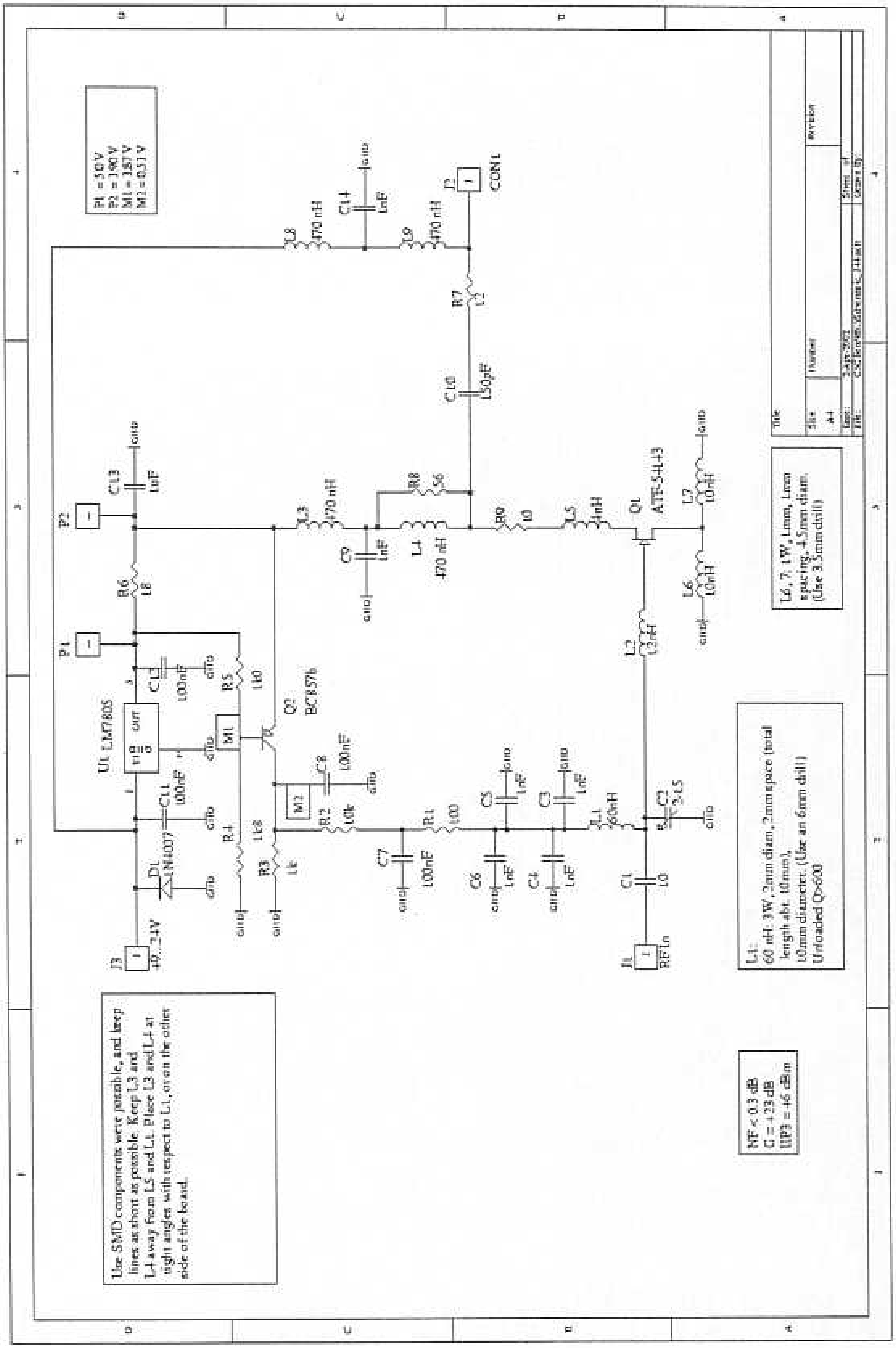
- P_n = Noise Power [W]
- k = Boltzmann ($1.38 \cdot 10^{-23}$ [Ws/K])
- B = Bandwidth [Hz]

A signal of -130 dBm equals $1 \cdot 10^{-16}$ W. Using the above formula, T equals 3019 K (assuming the noise is evenly distributed over a 2.4 kHz bandwidth).

Table 1: Various noise powers

Ambient Noise	Noise (dB)	Noisepower (K)
Low (432MHz)	0.15	10
Low (144MHz)	2.28	200
High (144MHz Urban)	8.97	2000

Bild/Figure 2: Circuit Diagram



Furthermore, for calculation purposes, let's assume a decent pre-amplifier, with a noise figure of 0.8 dB. The noise temperature of such a pre-amp is 59 K.

Using the ambient noise levels from table 1, the following SNR's at the antenna plug and after amplification can be calculated. For the SNR the amplification figure does not matter, as both noise and signal are equally amplified. We only need gain in order to amplify the nanovolt signals to levels were headphones can be exited, but this figure is not important when assessing the influence of pre-amplifier noise to the final SNR.

Table 2: Signal-to-noise-ratio when a wanted signal of 3019 K signal power is received. SNR is shown before and after amplification, applying various ambient noise levels. The Noise figure of the pre-amp was set to 0.8 dB.

As we can see from table 2, the importance of low noise temperatures increases as ambient noise levels drop. At 432 MHz EME it becomes obvious that a small decrease in noise temperature of the pre-amplifier has a great effect on the SNR after amplification. Even on 144 MHz EME it is worth while to strive for a low noise figure.

Lowering the noise figure further, will improve the SNR even more (I ought to say: "A lower noise figure gives less deterioration of the SNR"). Table 3 gives the SNR when a pre-amplifier with a noise figure of just 0.3 dB (21 K) is used.

Table 3: Signal-to-noise-ratio when a wanted signal of 3019 K signal power is received. SNR is shown before and after amplification, applying various ambient noise levels. The Noise figure of the pre-amp was set to 0.3 dB.

Comparing the SNR's after amplification, it is evident that 70 cm EME gains the most of low noise figures. A drop of noise figure from 0.8 to 0.3 dB gives 3.5 dB improvement in SNR (mind you, a drop in noise figure from, for example, 3.5 to 3.0 dB will not give such an improvement!).

Signal/Rauschverhältnis

Jeder Verstärker verschlechtert das von der Antenne angebotene Signal/Rauschverhältnis. Die Verschlechterung wird genau durch seine Rauschzahl definiert.

Um nun das bestmögliche Signal/Rauschverhältnis in einem Empfänger zu erhalten, müssen die ersten Stufen eine Empfänger sorgfältig auf niedriges Eigenrauschen ausgelegt werden.

Um einen Überblick zu bekommen, was überhaupt erreichbar ist, habe ich in Tabelle 1 die Rauschleistung, welche die Antenne aus der Umgebung empfängt, in Abhängigkeit von der Frequenz und der Umgebung aufgetragen. Diese Rauschleistung definiert das minimale Signal, was empfangen werden kann.

Abb1. zeigt den Zusammenhang zwischen der Rauschzahl eines Verstärkers und dem Signal/Rauschverhältnis. Ich nehme ein Signal von -130dBm an, was nach Formel [1] einer Temperatur von 3019K entspricht. Tabelle 2 zeigt die Verhältnisse bei einer Rauschzahl von 0,8dB (= 59K) auf.

Erniedrigt man die Rauschzahl auf 0,3dB (=21K) ergeben sich drastische Verbesserungen auf 70cm EME (+ 3,5dB SNR), immerhin noch 0,7dB auf 2m EME (ruhiges Land), aber nicht in der Stadtumgebung.

Dynamic range and available design options

An important feature of an amplifier, is its ability to cope with strong signals. Even though a pre-amplifier will have the lowest signal levels in the chain of amplification, any distortion it creates can never be deleted by subsequent components. Moreover, strict bandwidth limitation is not possible yet, due to the involved loss which influence the noise figure. Hence, the pre-amplifier needs to cope with strong out of band signals as well.

Let's have a look at some devices. Probably the most widely used device in low noise pre-amplifiers, is the MGF1302. If properly designed, a noise figure of 0.35 dB can be achieved at 144 MHz. However, its large signal properties are poor: the IIP3 is -7...-13

Table 2: S/N of a 3019K Signal w/ NF=0.8dB

Ambient Noise	Noise (dB)	Noisepower (K)	S/N @ Antenna (dB)	S/N @ RX-Output (dB)
Low (432MHz)	0.15	10	24.8	16.4
Low (144MHz)	2.28	200	11.8	10.7
High (144MHz Urban)	8.97	2000	1.8	1.7

dBm. Some "old fashioned" transistors do pretty good too, for example the BFT66, which can deliver IIP3 better than +10 dBm using loss less feedback technology [1, 2]. However, the noise figure that is obtained is relatively poor: about 1.5 dB. Some have experimented using high power GaAs FET's like the MGF1801. Though low noise figures were acquired (0.2...0.4 dB), the IIP3 figures remained relatively poor, given the type of device used: +1.5 dBm [3].

The advent of today's modern communication technology has seen the development of some interesting FET devices. Although specifically designed for PCS. (Personal Communication Systems like DECT at 1900 MHz) and GSM (900 and 1800 MHz), some of these devices work pretty good into VHF frequencies as well. One specific company that has created such kind of devices is Agilent (formerly Hewlett-Packard semiconductor division). A series of wide gatewidth PHEMT devices have been introduced that show remarkable features. Noise figures below 0.5 dB are obtained at 1 GHz, combined with exceptional large signal properties. IIP3's of +7 dBm are easily achieved at UHF and SHF frequencies. So what's the trade off? A rather high drain current of 40 to 80 mA! However, for desktop transceivers, power consumption is rarely a limitation....

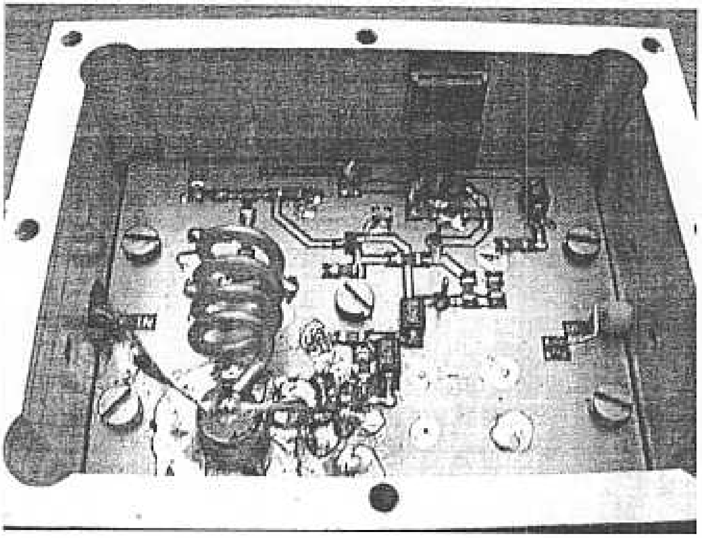
PHEMT stands for "Pseudomorphic High Electron Mobility Transistor". The HEMT structure can be visualised (simplified) being a 2-dimensional sheet, in which the electrons can move from source to drain. The properties of the layer are such that very high electron mobility and saturation velocity is achieved, giving rise to high fT, and few collisions with other electrons and atoms, hence low noise.

The structure of these pHEMT's differs from MESFET's, like the MGF1302, such that the gate structures are very wide. The ATF33143, for example, has a gate width of 800 m, as opposed to the MGF1302 which has a different structure, utilising relatively narrow gate widths. As a result of the wide (pHEMT) gate, the current consumption has increased, but without sacrificing noise figure. The ATF33143 (introduced 2 years ago) needed negative biasing on the gate. More recent devices have a slightly altered structure and

Table 3: S/N of a 3019K Signal w/ NF=0.3dB

Ambient Noise	Noise (dB)	Noisepower (K)	S/N @ Antenna (dB)	S/N @ RX-Output (dB)
Low (432MHz)	0.15	10	24.8	19.9
Low (144MHz)	2.28	200	11.8	11.4
High (144MHz Urban)	8.97	2000	1.8	1.74

Bild/Figure 3: Top view of the Prototype



are called E-pHEMTs. The "E" stands for "Enhanced" and designates the need for positive gate biasing. The ATF54143 needs +0.6V gate biasing, and draws a low (leakage) gate current (a few µA). For VHF applications, the positive biasing has complicated the circuit design somewhat, as we'll see further on.

Devices designed for PCS and GSM (0.9 to 2 GHz), have very high gains at low frequencies like 144 MHz. Gains in excess of 30 dB can be achieved, but are too much to be useful. Gain is fine, as long as it serves the purpose of overcoming the noise figures of subsequent stages. Anything more will only result in overloading mixer and IF stages, decreasing the dynamic range of the receiver. 30 dB of extra gain will decrease the usable dynamic range by nearly 30 dB as well!

So how can we decrease the gain of SHF devices without sacrificing the noise parameters?

1. Using an attenuator behind the amplifier;
2. Inserting loss less feedback across the device;
3. Use non-optimal loss less matching of the device.

An attenuator will decrease the gain, and diminish the load on the subsequent receiver stages. However, by putting an attenuator behind the pre-amplifier, the OIP3 is diminished by the same amount as well. Although the subsequent stages have some protection from overloading, the maximum attainable dynamic range is limited too. Limiting the gain of the pre-amplifier without sacrificing the OIP3 (using some form of feedback or non-optimal matching) will conserve the dynamic range.

Distortion is nearly always the effect of voltage or current limitations that appear in the output circuit of an active device. Limiting the voltage and/or current level in the output circuitry, using loss less feed-back for example, will decrease the gain, hence a higher IIP3 will be obtained. Loss less feedback can be created by inserting a small amount of inductance in the source leads of the pHEMT. Using simulations [4] a value of about 6 nH turned out to give enough feedback to decrease the gain to about 23 dB in stead of 30 dB. However, this form of feedback may also affect stability.

The linearity of a device depends on its ability to handle currents and voltage across the output terminals. Devices that either run at high voltages, or are biased at high currents, will usually perform well. The E-pHEMT devices used here, are biased at currents of about 70 mA (remember the MGF1302 running at 10 mA!). The addition of loss-less feedback in the source, gives some reduced gain, hence a higher IIP3.

Dynamischer Bereich

Sehr wichtig für die Auslegung des Vorverstärkers ist die Optimierung des Intermodulationsverhaltens. Intermodulation maskiert schwache Signale und wird durch starke Signale im Band und außerhalb des Bandes erzeugt. Naturgemäß ist die Bandbreite von Vorverstärkern recht groß.

In der Vergangenheit wurde oft der MGF1302 eingesetzt. Bei einer minimalen Rauschzahl von 0,35dB ist sein Eingangs-Interzeptpunkt ca. -13...-7 dBm.

Eine andere Option ist die Verwendung von bipolaren Transistoren wie z.B. dem BFT66 in einer Gegenkopplungsschaltung [1,2]. Dort wird ein IIP3 von +10dBm bei einer Rauschzahl von ca. 1,5dB erreicht.

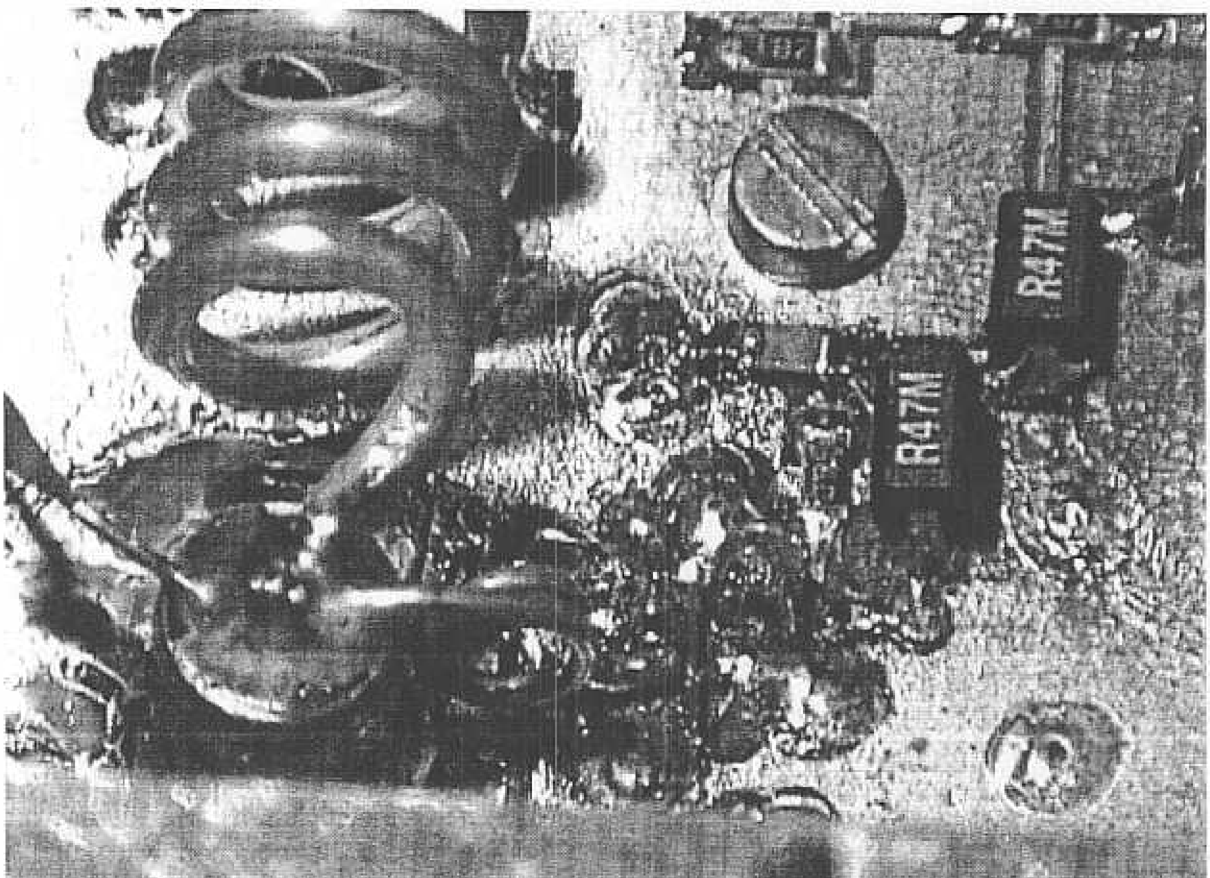
Auch Leistungs-GaAs-Fet wie der MGF1801 werden benutzt. Bei Rauschzahlen von 0,2...0,4dB ist der IIP3 ca. +1,5dBm [3]. Trotz hoher DC-Leistung ist das weniger als erwartet.

Inzwischen wurden moderne Leistung-HEMT's für die Anwendung in GSM-Basisstationen entwickelt. Speziell Agilent (früher als HP bekannt) vermarktet PHEMT's, die eine Rauschzahl von 0,5dB bei 1GHz und einen IIP3 von +7dBm bei dieser Frequenz erreichen.

Diese PHEMT's haben sehr große Breiten in der Gate-Struktur. Ein MGF1302 hat 250µm, ein ATF33143 hat 1600µm (=1,6mm!) und ein ATF51443 hat 800µm. Dadurch werden kleine Rauschzahlen bei großem Strömen möglich. Große Ströme implizieren niedrige Intermodulation. Typische Ströme für minimales rauschen liegen bei 60...80mA. Der ATF54143 ist ein E-PHEMT. Er braucht wie ein normaler Leistungs-MOSFET eine positive Spannung auf dem Gate.

Die Transistoren sind für den Bereich von 0,9...2,5GHz vorgesehen. Auf 144MHz zeigen sie einen exorbitante Verstärkung von mehr als 30dB. Das ist viel zu viel und ernied-

Bild/Figure 4: Close-up view of the E-pHEMT



rigt nur den Dynamikbereich des Empfängers.

Es gibt Methoden, um die Verstärkung zu erniedrigen:

1. Abschwächer nach dem Verstärker
2. Künstliche Verstimmung
3. Anwendung von "Lossless Feedback"

Ein Abschwächer erniedrigt die Verstärkung. Gleichzeitig wird aber auch der Ausgangs-Interzeptpunkt erniedrigt. Damit wird der Dynamikbereich erniedrigt. Also keine gute Idee.

Mit bewußter Fehlanpassung kann die Verstärkung auch erniedrigen. Z.B. kann man als Drainlast einfach einen 50Ω Widerstand nehmen.

Am besten ist Anpassung auf die optimale Last, die Spannungs- und Strombegrenzung gleichzeitig erfolgen läßt und die Anwendung von Induktivitäten in der Source. Mittels PUF [4] läßt sich zeigen, daß eine Induktivität von 6nH die Verstärkung auf 23dB sinken läßt. Allerdings können durch Sourceinduktivität leicht Instabilitäten oberhalb von 3GHz entstehen.

Stability issues

When I started this design, I thought that stability would be the least to worry about. It turned out to be close to a nightmare!

The device I used has a cut-off frequency well over 15 GHz. Hence, the maximum device gain at VHF is very high (30 dB). One would expect that the circuit might oscillate at VHF frequencies, but it did not. It oscillated like hell at 2...5 GHz.

The initial circuit I build was stable, but was put on a piece of board, without any "cutting edge" tweaking of the PCB. As soon as I changed to "nice circuits" (nicely designed board and a shiny aluminium case), things turned sour. Murphy had spotted me!

I made 3 new PCB designs with various results, except one: They all oscillated as soon as I put the lid on the box. However, in the last design I had incorporated some features to the circuit that did improve stability somewhat. But it still oscillated...

The part that had blotted my sight turned out to be the tuning capacitor! I had used a tuning capacitor which had an effective capacitor length of some 9mm (total length head to tail was 13mm). The series and parallel series resonances across the tuning capacitor was (one of) the causes of input impedance values which turned the circuit into an oscillator. Having substituted the tuning capacitor for a low inductance type, the oscillations ceased. Moreover, measuring the S-parameters of the pre-amp (130 to 10000 MHz), calcullations showed that the pre-amplifier is now unconditionally stable. Even with the lid on, which is the most difficult task to achieve! This is familiar to every microwave experimenter.

Stabilität

Als ich mit diesem Projekt anfing, dachte ich nicht, daß die Stabilität ein Problem werden könnte. Tatsächlich wurde es das Hauptproblem. Der FET, die ich benutze, er hat eine Grenzfrequenz von über 15GHz. Auf VHF ist die Verstärkung mehr als 30dB.

Die Oszillationen waren zwischen 2 und 5GHz, sobald ich den Deckel aufsetzte. Auf 144MHz selbst war der Verstärker stabil. Die induktive Gegenkopplung in der Source zusammen mit einem ungeeigneten Abstimmkondensator im Eingangsschwingkreis erzeugten einen Oszillator.

Durch Wahl eines geeigneten Trimmers und Serieninduktivitäten im Gate und im Drain wurde Stabilität erreicht. Das wurde durch Messungen der S-Parameter von 130MHz bis 10000MHz bei geschlossenen Deckel bestätigt.

Designing the ideal Pre-Amplifier

Let's get this straight: the ideal pre-amplifier can not be constructed using amateur equipment, but we can get very close to it these days! Some modern high frequency FET's have NFmin which are equal or less than 0.1 dB! (NFmin is the minimum noise figure which theoretically can be obtained with such a device using loss less matching to the optimum impedance). NFmin is acquired at a specific impedance. The transformation from the antenna impedance (usually 50Ω) to the impedance of NFmin introduces loss, which will add up to NFmin. If the impedance match to NFmin fails (i.e.: the transformation ends up with an impedance that differs from the optimum noise impedance), the noise figure of the device will also be higher than NFmin. The sensitivity for impedance mismatches is often expressed with noise circles. Any point on this circle will give a specific NF of the device at the specified impedance. The larger the distance between the successive circles, the less critical the requirements of the matching network will be.

To match to the optimum noise impedance, we need to design a circuit that will do the trick with the least amount of loss. The matching loss is a function of loaded- and unloaded-Q values of the matching network (Q is the ratio between the centre frequency and the 3 dB bandwidth of a resonant LC-network).

The unloaded Q (Q_u) is the bandwidth of a circuit without any loading by active or passive circuitry: just an LCR network. The loaded Q (Q_l) is the same network, but coupled to an active or passive system, which lowers the total bandwidth. The ratio between these 2 figures, determines the loss, according to the following formula:

[2] $$IL = 20 \cdot \log\left(\frac{Q_u}{Q_u - Q_l}\right)$$

As can easily be derived from the formula, a high value for Q_u and a low value for Q_l, minimises the losses. However, the selectivity of the LC network, is determined by the loaded Q (bandwidth!). A high value for the loaded Q, means small bandwidth (= high selectivity), but increases the network loss. If the active device has good strong signal parameters, we can tolerate poor selectivity, because the device is able to cope with strong out of band signals, hence design a matching network with low loaded Q.

Well designed lumped coils can obtain fairly high Q values of 300 to 600. Using a 3 dB bandwidth of 20 MHz, the insertion loss will amount 0.1...0.2 dB, giving an acceptable selectivity of -10 dB at 100 MHz, compared to 144 MHz.

Summarising the characteristics of the input matching network:

- matching antenna impedance to NFmin (NFmin is a device specific, frequency dependent, impedance);
- keeping matching losses as low as possible, given the practical constraints of available Q of the network.

Design

Es gibt heute Transistoren, die selbst nur eine Rauschzahl von 0,1dB auf VHF haben. Der Transistor will eine bestimmte Impedanz für die minimale Rauschzahl "sehen". Das wird durch Kreise konstanter Rauschzahl im Smith-Diagramm beschrieben.

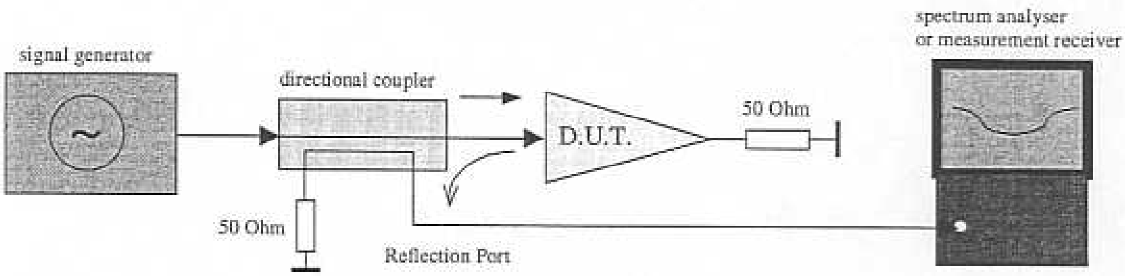
Die optimale Impedanz ist meistuns von 50Ω verschieden. Man braucht also eine Transformationsschaltung, welche diese Impedanz aus der Nennimpedanz von 50Ω herstellt. Die Verluste einer solchen Transformation werden durch Gleichung [2] beschrieben.

Es kommt also darauf an, ein möglich größes Verhältnis zwischen der Leerlaufgüte Q_u und der Lastgüte Q_l, zu erreichen. Spulen haben auf 2m Güten zwischen 300 und 600. Setzt eine 3dB Bandbreite von 20MHz an, ist die Lastgüte 144/20=7. Damit ist der Eingangsverlust 0,1...0,2dB.

Circuit Description

The LNA is designed straight forward (figure 2), however L2 and L5 need some special attention. The input matching network consists of C1, C2 and L1. L1 is a lumped air coil, made of 2 mm diameter silver plated copper wire. Using wire spacing of 2mm, the theoretical Q is estimated to be above 500. The inductance is 65 nH. The bandwidth (-3 dB) is set to about 20 MHz, a compromise between selectivity and loss. At 100 MHz the gain is reduced by 12 dB compared to 144 MHz. The calculated insertion loss amounts to 0.1....0.15 dB.

Bild/Figure 7: Input return loss measurement set-up



Konstruktion

Die Leiterplatte und der Bestückungsplan sind aus den Abb.14...17 ersichtlich. Alle Bauelement werden montiert und verlötet. Zuletzt wird der sehr FET ATF54143 (4p) eingelötet. L6 und L7 sitzen auf der Masseseite der Platine. Die Abb. 3 und 4 zeigen den Aufbau.

Adjustment

Although the pre-amplifier is stable, make sure to connect an antenna, or 50Ω dummy loads to in- and output. Connect a clean power supply to the circuit (9 to 24V DC) and measure the current at the power supply or the voltage across R6. The current should read between 65 and 70 mA, the voltage measured across R6 should read about 1.14V.

Next, insert the pre-amp in the RX chain and tune to a beacon. Adjust the input coil tune capacitor to maximum signal strength. I found that the minimum noise figure is a very broad minimum. Tuning for maximum signal strength equals the minimum noise figure within a few hundreds of a dB. The only way to squeeze the last hundreds of noise figure out of the pre-amp, is by using a calibrated set-up.

Abgleich

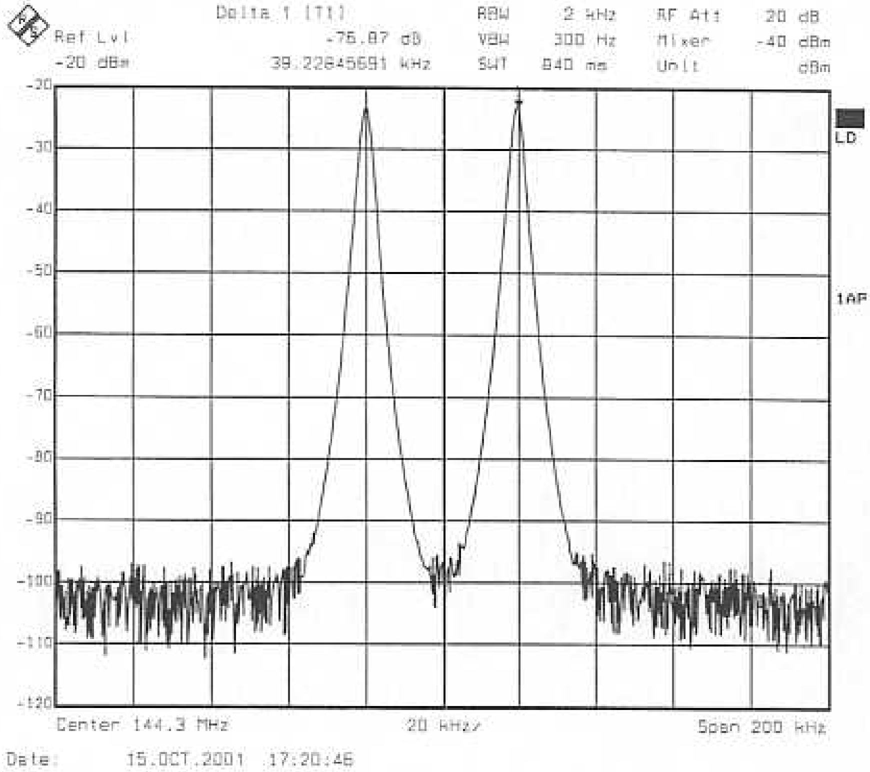
Die Betriebsspannung von 9...24V wird angelegt und verifiziert, daß der Strom 60...70mA beträgt. Die Spannung über R6 beträgt 1,14V.

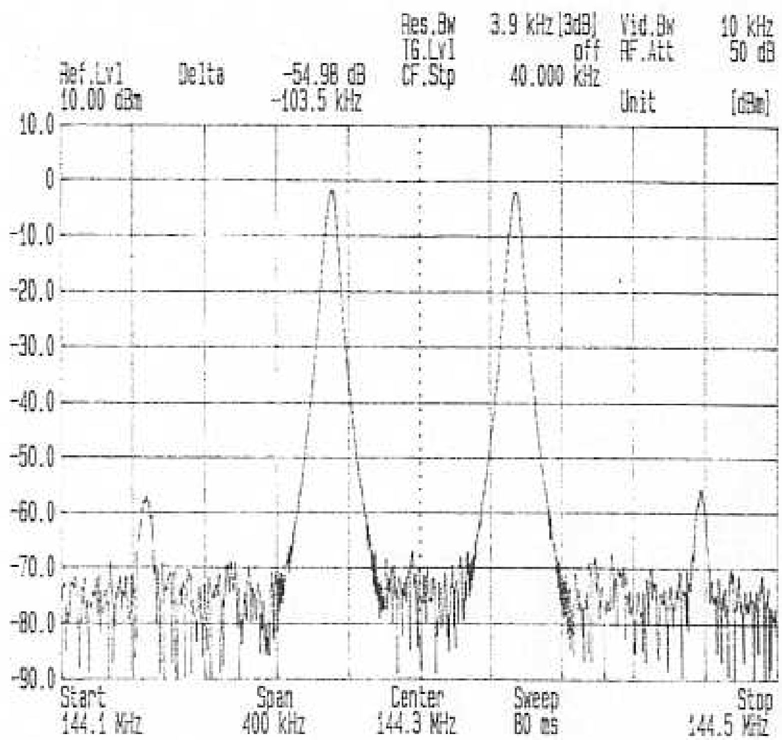
Der Eingangskreis wird an einer schwachen Bake einfach auf Maximum gedreht. Die Rauschzahl ist dann bis ein paar Hundertels dB im Minimum.

Measurements

In this chapter some considerations and pitfalls of measurement set-ups will be described. Obviously, the noise figure of a pre-amplifier is one of the most important quantities, however, some errors easily made, may lead to erroneous conclusions. The linearity of a pre-amplifier may seem unimportant, but, like the strength of a chain, the weakest part of a receiver chain determines the total strength. The in- and output impedance is an indication of how stable an amplifier is. Moreover, high output return loss values will keep subsequent amplifier stages quiet.

Bild/Figure 8: Input 2-tone signals





Bild/Figure 9: Output Intermodulation Products

Noise figure measurements and inherent errors

Fortunately, I am in the position to use professional noise measurement equipment from HP (HP8970A and HP346B noise source). Nevertheless, one should be careful not to make measurement errors. Measurement errors may be not obvious when starting a measurement. An example is the effective value of a periodic signal. If this is not a perfect sinewave, most voltage meters will give the wrong effective voltage. They are designed using the assumption that an alternating voltage will be a sinewave!

The input SWR of a low noise amplifier is often everything but 50Ω. Most designs will use a tuned circuit in order to transform the input impedance to optimal noise impedance of the active device. As a consequence, the optimal source impedance of an active device rarely matches the optimal power impedance. Hence, it is rare to find low noise pre-amps that exhibit a 50Ω input impedance. A poor input SWR of a pre-amplifier may influence the measurements dramatically (not the noise figure itself!!).

The nature of noise measurement equipment is to switch a known noise source ON and OFF. The ratio between the ON and OFF situation is the basis on which the noise figure calculations are performed. The noise source is followed by a fixed amount of attenuation. However, the impedance of the noise source itself may not be exactly 50Ω, due to the fact that a switch is involved (ON and OFF). As long as the impedance of the DUT is 50Ω the noise source mismatch is of no concern, because the equipment is calibrated using 50Ω impedance. However, a DUT with an input impedance other than 50Ω, will result in an error in attenuation factor between the 2 alternating noise states. This difference is usually ignored, but becomes crucial when the impedance deviates much from 50Ω or when the noise figures we want to measure are very small.

One example I recall, is a design from YU1AW [5]. This pre-amplifier had an excellent noise figure (it utilised a MGF1302), but had an input SWR of 1:100 or so. Such a SWR will not deteriorate the amplifier's performance, but measuring its noise figure, was quite a challenge! In fact, I was able to tune the input circuit to negative noise figures!! The only way to resolve this was by inserting a fixed (calibrated) attenuator between the noise source and amplifier. After subtracting the attenuator value of the measured value, I was able to obtain the real noise figure (0.35 dB for the YU1AW pre-amp).

If you are interested to get to the bottom of noise figure measurements and pit falls, it is worth reading an article by Rainer Bertelsmeier, DJ9BV [6].

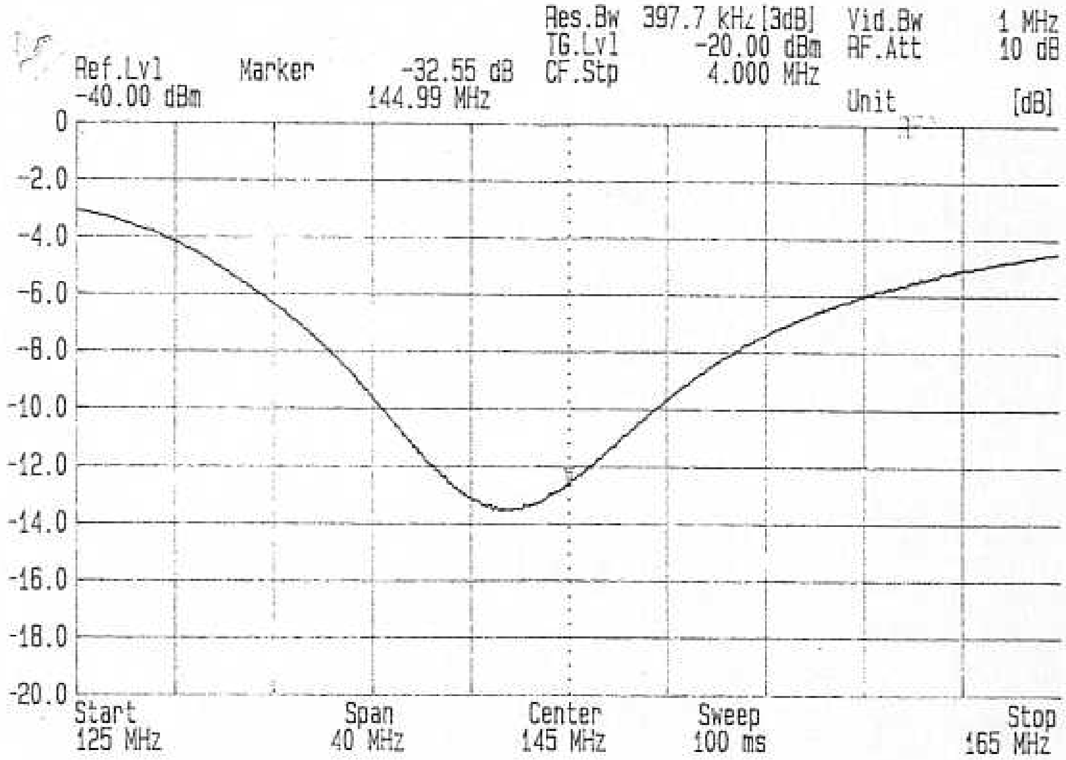
Inserting extra attenuation reduces the measurement error considerably and should be standard practice in low noise pre-amp measurements. Using a standard noise source (even a professional one), may lead to misleading (low) noise figures! Some noise figure measurement set-ups automatically measure the S-parameters of both noise source and DUT, and compensate. This may be the ultimate method, but requires a very expensive set-up.

Measuring IP3

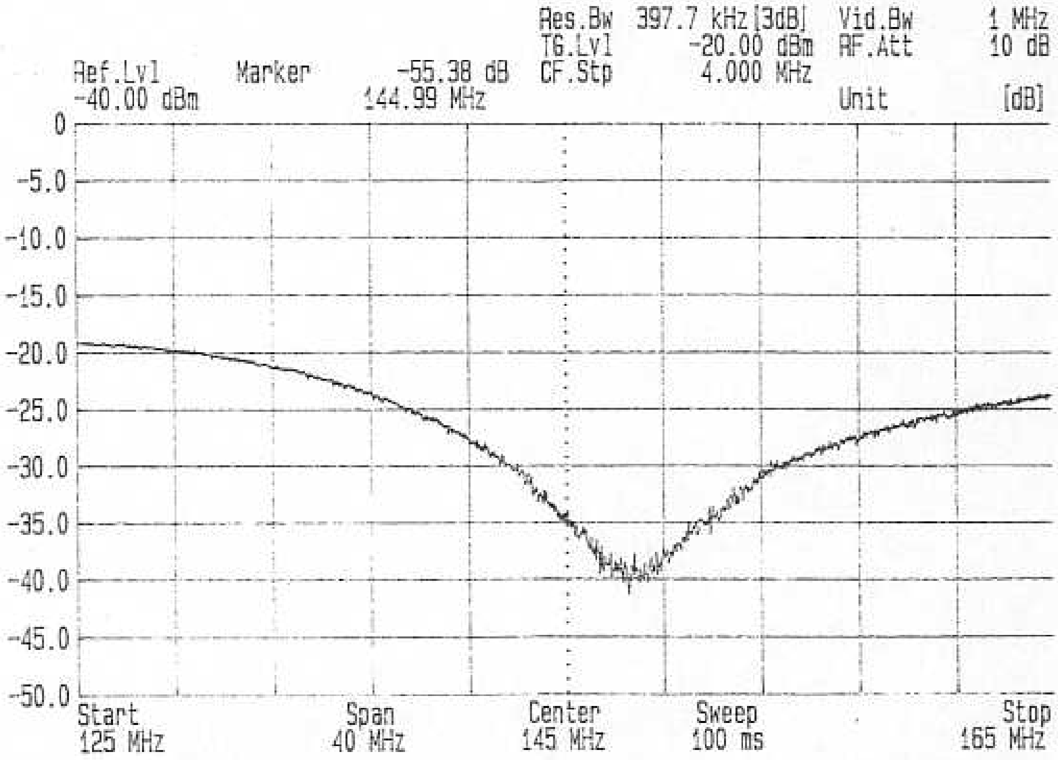
Linearity can be measured using 2 independent signal sources, isolated from each other by circulators and a combiner. The circulators are added in order to increase the isolation between the 2 signal sources and prevent intermodulation to emanate from the signal sources themselves (figure 5). In this set-up I used the FSEA spectrum analyser from Rohde & Schwarz (but any spectrum analyser having adequate dynamic range, may be used). Care must be taken not to overload the analyser, which will lead to the generation of IM products in the analyser itself.

The isolators are from Philips (surplus 154 MHz pagers systems, tuned to 145 MHz), which provide more than 40 dB isolation. Added to the isolation provided by the hybrid combiner (about 20 dB), both signal generators are isolated by more than 60 dB, which was sufficient to obtain more than 80 dB suppression of any 3rd order products.

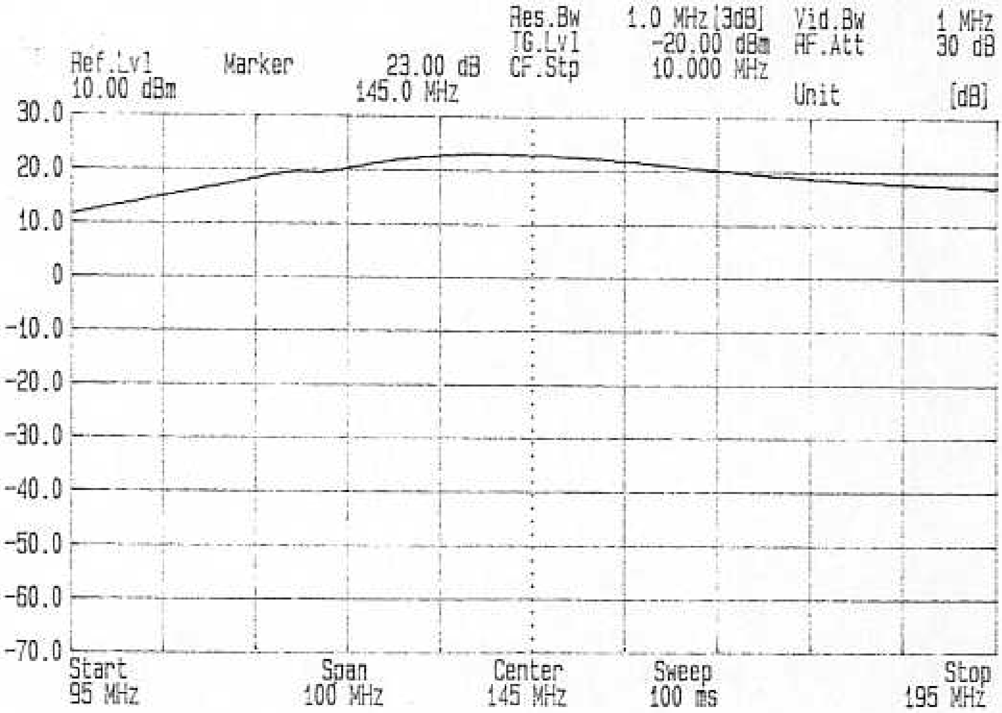
Bild/Figure 10: Input Return Loss



Bild/Figure 11: Output Return Loss



Bild/Figure 12: Gain plot from 95 to 195 MHz



In figure 6, the relation between IM, the fundamental signal and IP3 is shown. As can be seen the actual 3rd order IP point will never be reached by an amplifier, because saturation will have occurred long before. Instead, we need to measure the intermodulation products some 10 to 20 dB below the saturation point of the amplifier, and calculate the IP3. The Spurious Free Dynamic Range (SFDR) can also be calculated if the IIP3 and noise floor are known.

The equation for IP_3 is:

$$[3] \quad IP_3 = \frac{3 \cdot P_O - IMD_3}{2}$$

where

P_O Single tone output level in dBm

IMD_3 Two Tone 3rd order distortion level in dBm

From the intermodulation products as they appear at the output of the Device Under Test, the IIP3 can be calculated (as a tool I use a freeware program from Agilent: AppCad [7]).

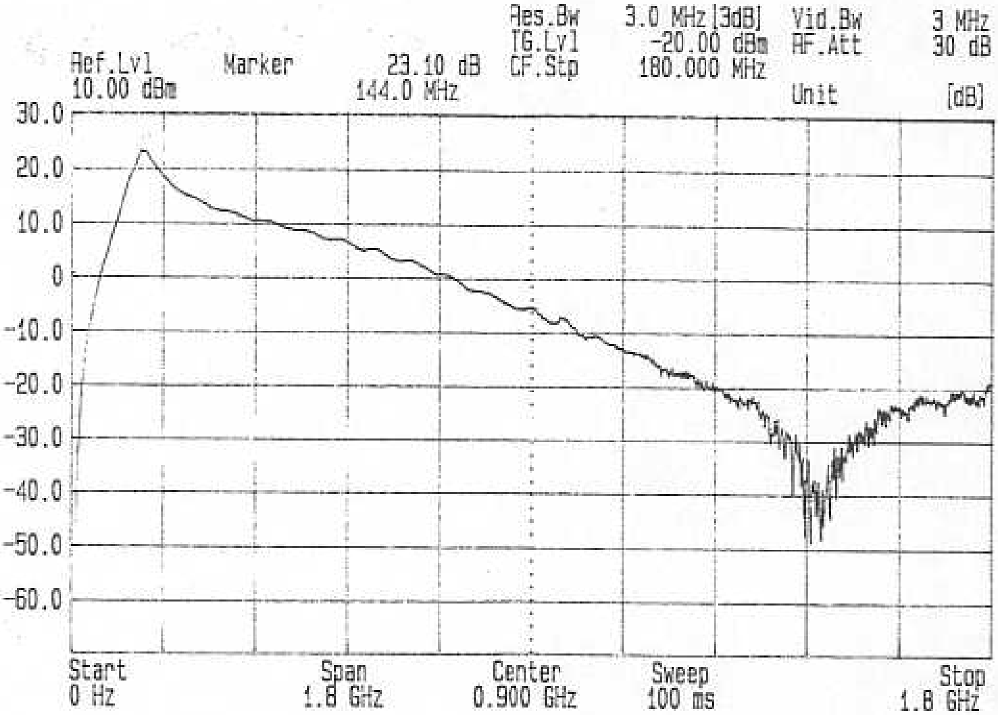
Measuring the return loss of amplifier ports

The input and output return loss are a measure of how well an amplifier matches to 50Ω impedance. The smaller the reflection (i.e. a high return loss level in dB), the better it matches to 50Ω. Is it an important value? Not for the actual noise figure. However, upon employing the LNA good matching guarantees that 50Ω interconnection lines can be used without having to take their length into account. The return loss can be measured using a directional coupler which is fed by a signal generator and terminated by the Device Under Test (DUT). The reflection port will show the amount of energy that is not absorbed by the DUT and reflected back into the directional coupler (Figure 7).

In stead of a spectrum analyser, a sensitive power meter (< -10 dBm) or a calibrated receiver can be used as well.

The set-up can be easily calibrated: Step 1: Let the output port open (i.e.: no DUT attached to the directional coupler) and read the strength of the signal at the reflection port of

Bild/Figure 13: Gain plot from 0 to 1800 MHz



the directional coupler (a value should be found equivalent to the specified coupling loss of the directional coupler. For example, if the coupling loss is 20 dB, and the signal generator is set to -20 dBm, a value of about -40 dBm should be found).

Step 2: Attach a 50Ω dummy load or 50Ω SMD resistor to the output of the directional coupler, and the reading should be at least 20 dB less than before because all energy should be absorbed in the 50Ω load and no reflections should appear at the reflection port (apart of the directivity of the coupler in use, usually 30 dB).

Messungen

Die Messungen beziehen sich auf folgende Eigenschaften:

- Rauschzahl
- Linearität (IP3)
- Eingangs- und Ausgangsanpassung

Rauschzahl

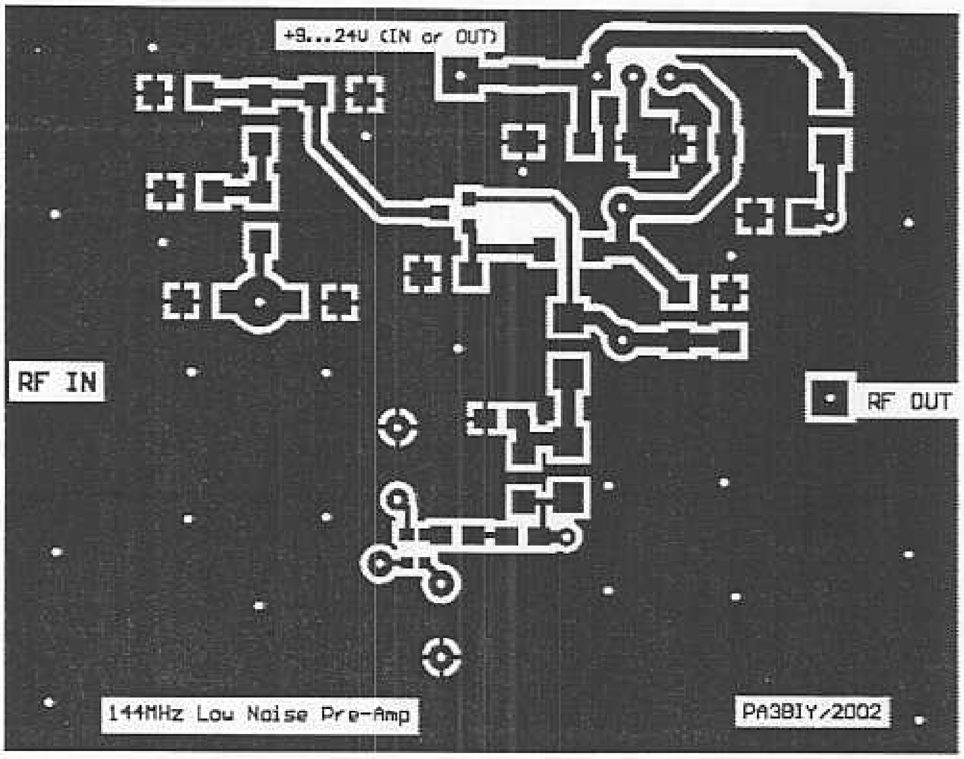
Normalerweise hat ein Vorverstärker ein schlechtes Eingangs-VSWR. Das geht den Meßwert eine Rauschzahlmessung sehr stark beeinflussen.

Die Rauschquelle schaltet periodisch zwischen Ein- und Auszustand hin- und her. Dabei ändert ihre Impedanz ganz wenig. Selbst Sie ihre Anpassung zeigt (> 30db Rückflußdämpfung) reicht eine kleine Änderung der Impedanz, um bei Meßobjekten mit hohem VSWR eine Änderung der Verstärkung herbeizuführen. Damit kann je nach Phasenlage ein größerer oder kleiner Y-faktor vorgetäuscht werden, was dann zu einem falschen Meßwert für die Rauschzahl führt. Speziell 15dB ENR Rauschquellen wie die HP346B oder die AILTECH 7615 sind dafür anfällig. Da hilft dann nur ein zusätzlicher 10dB Abschwächer.

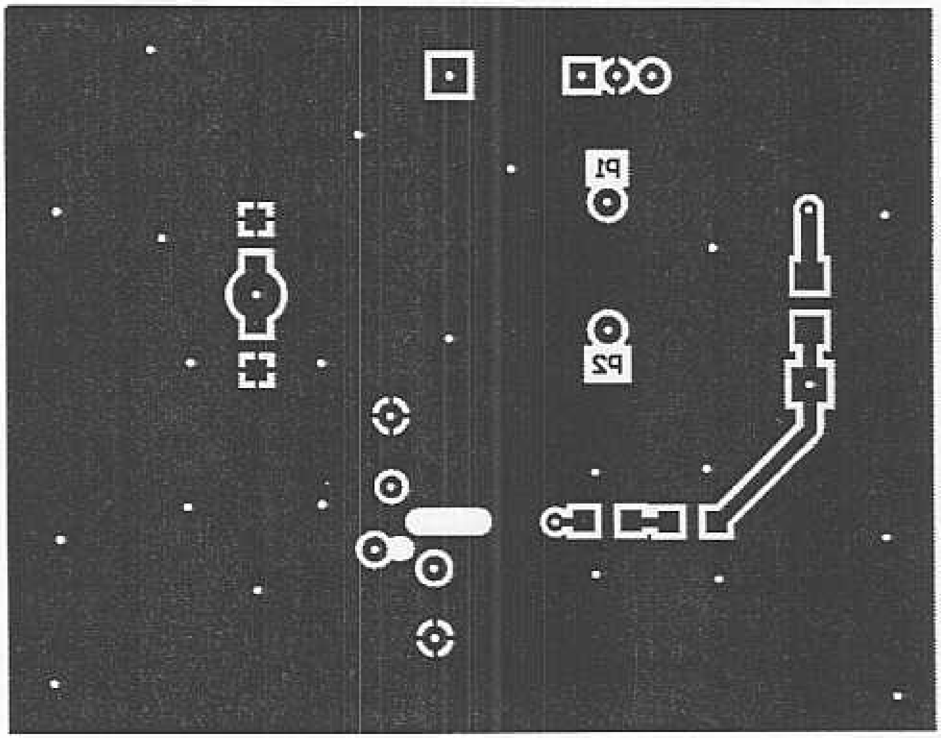
Das war vor Jahren die einzige Methode, um einen Vorverstärker nach YU1AW [5], der ein Eingangs-VSWR von 1:100 hatte, zu messen.

Wer sich da eingehender informieren möchte, sei auf den Artikel von Rainer Bertelsmeier, DJ9BV [6], verwiesen.

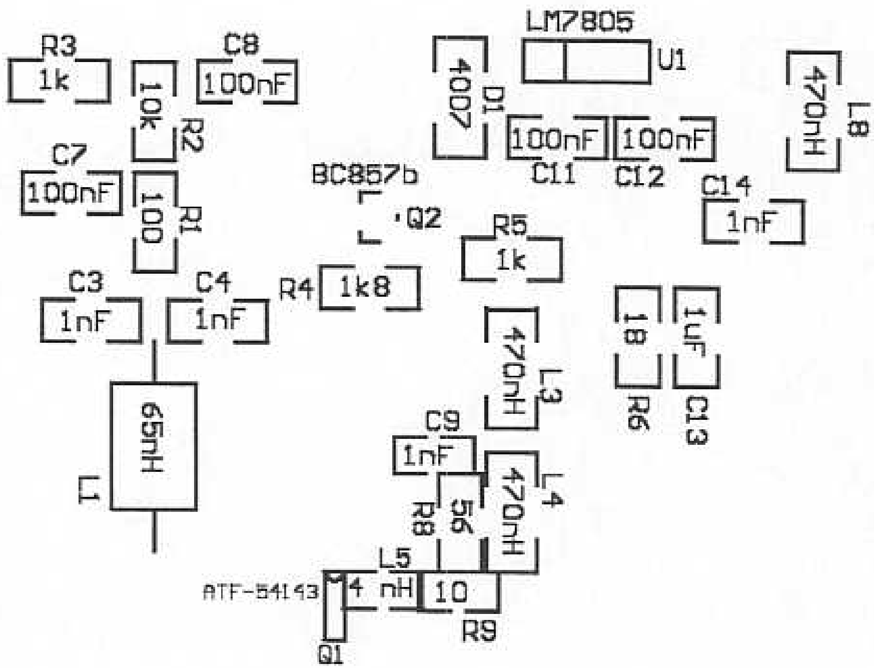
Bild/Figure 15: PCB Top View



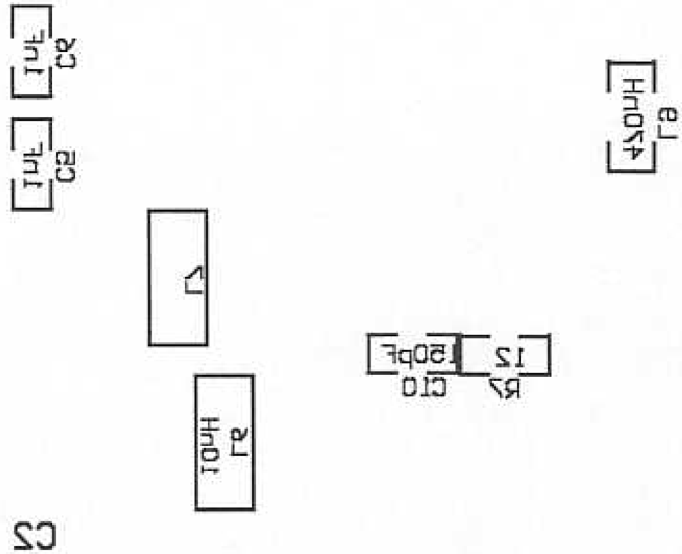
Bild/Figure 14: PCB Bottom View



Bild/Figure 16: Silk Screen Top



Bild/Figure 17: Silk Screen Bottom



Linearität (IPI)

Wie in BA,. 5 gezeigt wird die Intermodulation mit zwei entkoppelten Signalquellen und einem Spektrumanalysator gemessen.

Die Entkopplung wird mit zwei Isolatoren von Philips und einem Hybridkoppler auf mindestens 60dB gehalten. Das reicht aus, um keine Intermodulation im Dynamikbereich (> 77dB) des Spektrumanalysators zu sehen (Abb. 8).

Abb. 6 zeigt den Zusammen zwischen den Intermodulationsprodukten und dem Eintonsignalen. Die Produkte wachsen im linearen Bereich mit der 3-fachen Geschwindikeit wie die Einzeltöne, bis schließlich Sättigung eintritt. Aus einer solchen Kurve kann den IP3 und den intermodulationsfreien Bereich bestimmen.

Anpassung

Die Anpassung im Ein- und Ausgang macht den Einsatz des Verstärkers problemlos z.B. vor Filtern und ist außerdem ein Kriterium für Stabilität.

Die Rückflußdämpfung kann nach dem Aufbau im Abb. 7 bestimmt werden. Die Kalibration wird durch Anlegen einer Last von 50Ω und einer offenen Last bewerkstelligt.

Measurement Results

Intermodulation

From the intermodulation products as they appear at the output of the Device Under Test, the OIP3 can be calculated (I used a freeware program from Agilent: AppCad [7]). The IIP3 then follows by subtracting the gain (AppCad will perform this equation as well). The 2-tone test signal as applied to the pre-amplifier is shown in figure 8a. All IM products are 80 dB down. In figure 8b, the measured IM spectrum is shown as produced by the amplifier.

The measurements performed on the 144 MHz pre-amplifier shows 3rd order Intermodulation Products of -55 dBm, compared to the fundamental signals of -2 dBm. These values result in an Output IP3 of +25.5 dBm. At an amplifier gain of 23 dB, the IIP3 is +2.5 dBm (IIP3 = OIP3 - Gain).

Input and output return loss

The input and output return losses were measured using a professional S-parameter measurement set-up by HP. The input return loss shows a fairly wide dip, and peaks at -13.5 dB around 142 MHz. However, the return loss is still -12 dB at 144 MHz.

The output return loss has a more or less flat characteristic, which is the result of the heavy loading by R8 in the drain circuit. At 144 MHz the output return loss is better than 25 dB.

Noise figure

In order to prevent the gain error, a fixed (carefully measured!) attenuator of 10 dB was inserted between the noise source and the LNA. The noise LNA figure was measured to be 0.25 dB.

Gain

The gain is measured in a traditional manner (if you are lucky to have a spectrum analyser at your disposal!). Figure 11 and 12 show plots of different frequency spans. A slight rise in gain can be noticed for frequencies higher than about 1500 MHz. This is due to the series inductance of the tuning capacitor and the series inductance in the source. As can be seen from the plot, the gain is 23 dB around 145 MHz.

Summarising

Gain:	23 dB
NF:	0.25 dB
Input return loss:	12 dB
Output return loss:	>25 dB
3rd order input intercept point:	2.5 dBm
Supply voltage:	+9...+24V
D.C. supply current:	68 mA

Ergebnisse

Intermodulation

Abb. 9 zeigt die Intermodulationsprodukte eines intermodulationfreien Eingangssignal nach Abb. 8. Aus den Pegeln errechnet sich ein OIP3 von IP3=-2dBm+55/2=25,5dBm. Bei einer Verstärkung von 2,5dBm entspricht das einem IIP3 von +2,5dBm.

Anpassung

Abb. 10 zeigt die Eingangsanpassung. Sie ist auf 144MHZ 12dB, ein erstaunlich hoher Wert für die Rückflußdämpfung.

Abb. 12 zeigt, daß auf 145MHZ die Rückflußdämpfung besser als 30dB ist.

Rauschzahl

Die Rauschzahl des Prototypen wurde mittels HP8970B und einer HP346B+10dB Abschwächer mit 0,25dB gemessen.

Verstärkung

Die Verstärkung betragt auf 145MHz 23dB (Abb. 12). Der Breitbandplot ist in Abb. 13.

Conclusions

The merit of this pre-amplifier is in the combination of a high dynamic range with a very low noise figure. Optimal utilisation of both aspects can only be realised when the subsequent stages have a high dynamic range as well, otherwise only the noise figure will drop, sacrificing dynamic range. Modern E-pHEMT FET's enable us to construct pre-amplifiers with low noise figures and high IIP3, without the need for compromises on either aspect. Moreover, the Input- and Output impedance's can be matched very closely to 50Ω as well, without sacrificing performance. Some care must be taken in order to prevent oscillation in the SHF frequency range, given the high f_T of these devices, but carefully designed these amplifiers are unconditionally stable.

Schlußbemerkungen

Der Vorteil dieses Vorverstärkers besteht in der niedrigen Rauschzahl, des hohem IP3, der guten Ein- und Ausgangsanpassung und nicht zuletzt der absoluten Stabilität von 100 bis 10000MHz. Das wurde ermöglicht durch den Einsatz eines modernen E-PHEMT.

Acknowledgement

I'd like to thank Frank van Vliet (PA3FAQ) for reviewing this article and for his technical support considering E-pHEMTdevice issues. I'd also like to thank Jos van de List (PA0JOZ) and Rainer Bertelsmeier (DJ9BV) for their constructive comments.

[As an option, I can supply complete "do it yourself" kits, including the PCB, 2 x N connector, a milled aluminium box and all electronic parts (ATF-54143 pHEMT, Hi Q tuning capacitor, SMD parts, etc). I need at least 100 interested Hams before I'll order PCB's. If you are seriously interested, please send an email to: pa3biy@wanadoo.nl]

Danksagung

Ich möchte Frank van Vliet (PA3FAQ) dafür danken, daß diesen Artikel Korrektur gelesen hat. Jos van de List (PA0JOZ) und rainer Bertelsmeier (DJ9BV) danke für Ihre konstruktiven Beiträge.

Als Option biete ich die Bereitstellung von Teilesätzen an, die alle Teile einschließlich PCB und Gehäuse enthalten. Ich brauche mindestens 100 Interessenten, um eine PCB kommerziell fertigen zu lassen. Nur ernsthaft Interessierte melden sich bitte unter pa3biy@wanadoo.nl

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