Low Noise GaAs-FET Preamps for EME:

Construction and Measurement Problems

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1. Introduction

This article is based on an unpublished paper 'Limitations of Automatic Noise Figure Measurement Equipment' [1], which has been given to some EME-amateurs (i.e. DL6WU, DL9KR, K2UYH, W1JR and I5TDJ) and a paper, covering a lecture held during the annual VHF/UHF conference in Weinheim, West Germany, in September 1987[2]. It describes the problems and - hopefully - their solutions, which arise with the use of single GaAs-FET's for Low Noise Amplifiers (LNA's) on 432 MHz.

2. Noise temperature of receiving system

The signal to noise ratio in a moonbounce circuit depends on

1. Antenna gain
2. Transmitting power
3. Path loss
4. Noise power of receiving system

Path loss is given by the geometry, i.e. of fixed size - variation is only 2 dB because of the changing distance of the moon and other effects like polarisation fadind etc are not considered -. Transmitting power and antenna gain have financial, legal and size limitations, which allow only minor improvements: There is not much left after having installed 16 yagis or a 10 m dish and a 2 kW amplifier. Because of the low sky temperature on 432 MHz and up careful optimisation of the receiving system can lead to significant improvements in S/N-ratio (> 3 dB). This is of utmost importance for QRP-stations - 4 yagis only or low power - and for those QRO stations, who really want to catch the edge.

System noise temperature, which states the noise power of the receiving system, is given by:

$$T_{sys}=T_A+(L-1)T_L+LT_R \qquad [1]$$

- T_{sys}: System noise temperature
- T_A: Noise temperature of antenna system
- T_L: Physical temeprature of cable between antenna and first LNA
- T_R: Noise temperature of receiver

The optimisation of the first two terms in equation [1], which describe the influence of antenna and cable, is subject of a future lecture and has been discussed for the case of yagi antennas on 432 MHz in DUBUS 4/87[3].

The following paper describes how to achieve the lowest possible receiver noise temperature, which is a nontrivial task even with the availability of the most modern measurement equipment and of low noise single Gate GaAs-FET's.

3. LNA's with single gate GaAs-FET on 432 MHz

3.1. Properties of single gate GaAs-FET's

Single gate GaAs-FET's used for LNA's are devices, which have been designed and used preferebly on very high frequencies (> 2 GHz). EME-Amateurs have used them on low frequencies (144 or 432 MHz) also and sometimes have succeeded in achieving very low noise figures. Because of their extremely high gain on low frequencies the major problem of single gate GaAs-FET preamps on 144 and 432 is instability.

The properties of microwave devices are described by scattering parameters (S-parameter). Gain and stability can be calculated, if the S-parameters of a device or LNA are known:

S_{11}: Input Reflection Coefficient

S_{12}: Reverse-Gain

S_{21}: Forward-Gain

S_{22}: Output Reflection Coefficient

Conditions for stability:

1. $|S_{11}|<1$
2. $|S_{22}|<1$
3. $K=\dfrac{1+|D|^2-|S_{11}|^2-|S_{22}|^2}{2|S_{12}S_{21}|}>1$ with $D=S_{11}\cdot S_{22}-S_{12}\cdot S_{21}$

These conditions guarantee the unconditional stability of a device or LNA. If one condition is violated the device can oscillate.

For easy judgement of the stability of a device stabilty circles can be calculated from the S-parameters and drawn on a smith chart:

$$R=\left|\frac{S_{12}S_{21}}{|S_{ii}|^2-|D|^2}\right| \qquad \text{Radius}$$

$$C=\frac{(S_{ii}-DS_{kk}^{*})^{*}}{|S_{ii}|^2-|D|^2} \qquad \text{Center}$$

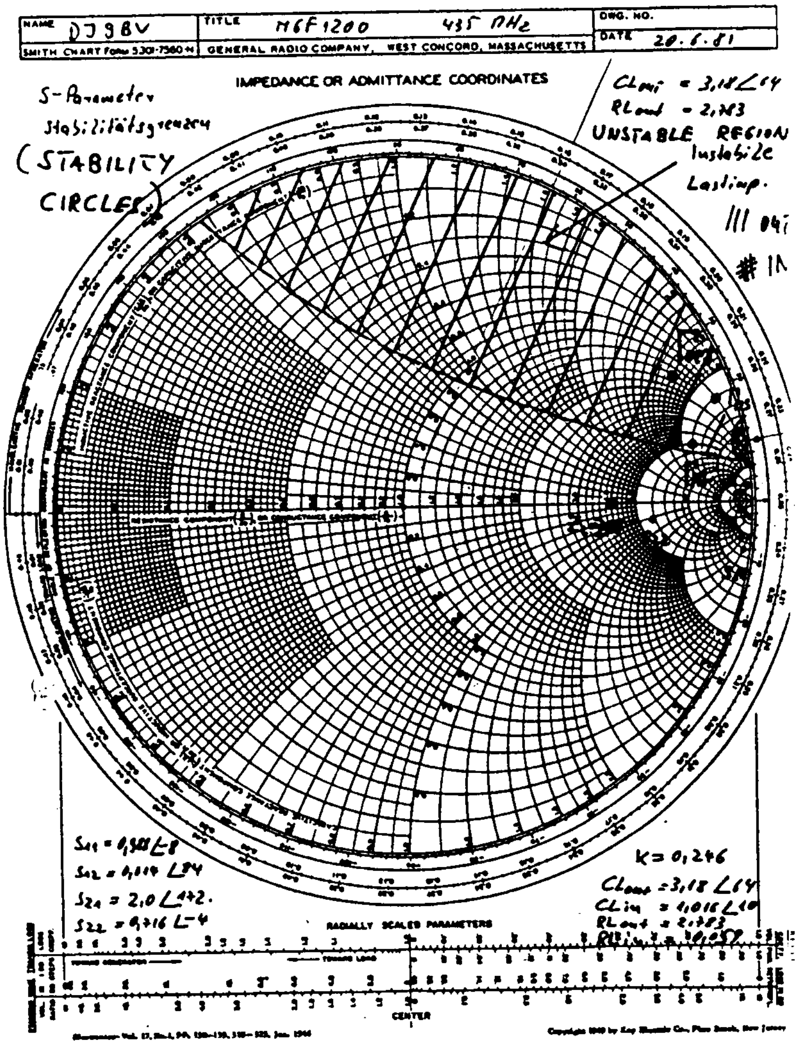
with:

Port	ii	kk
Output	22	11
Input	11	22

If input and output stabilty circles are outside of the unit radius of the smith chart unconditional stability is provided.

Figure 1 displays the stability circles for a GaAs-FET MGF 1200:

Figure 1:

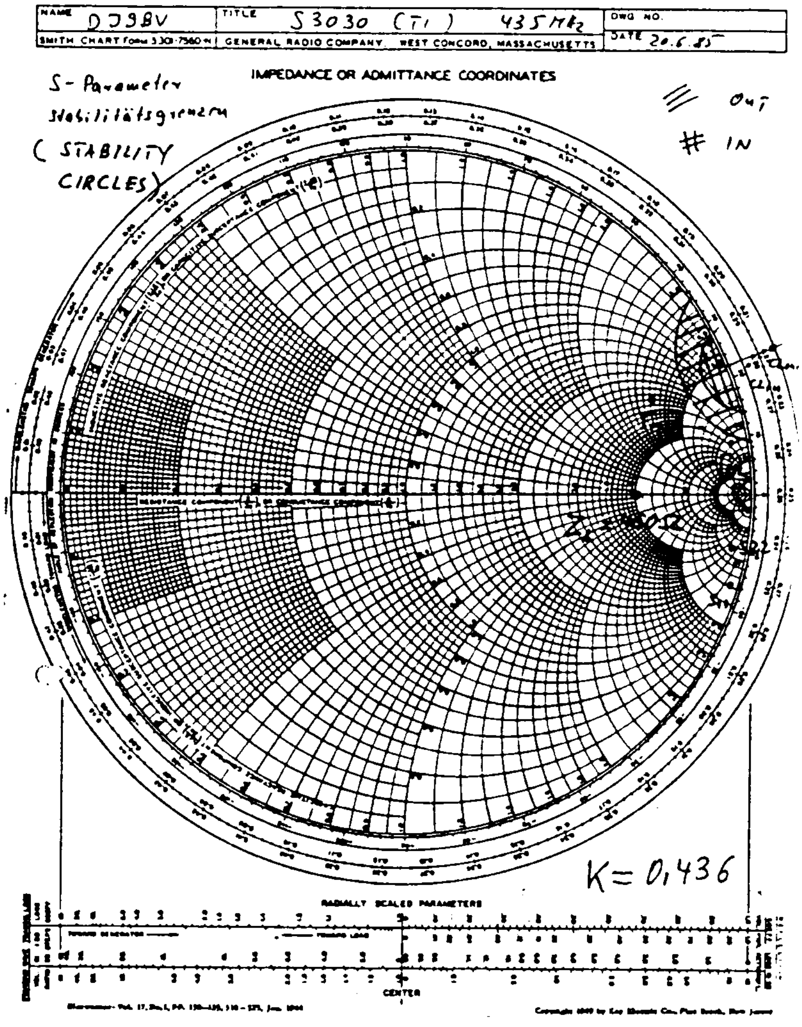


It's easy to see, that output matching with inductive reactance produces unstable amplifiers. The optimum reflection coefficient for maximum gain (Γ_{ML}) lies in the unstable region already. On the other hand output matching to 200 Ohms real impedance at the drain produces a stable amplifier, which cannot oscillate for any source impedance. This does **not** mean unconditional stability, which allows **any combination** of source and load impedances. That will be achieved if stability factor K > 1. That's not the case for most single gate GaAs-FET's for an operating frequency less than 4 GHz.

But for amateur use conditional stability for 50 Ohms load is sufficient, because that can be assured by output pads or cable loss. But instability with negative real input impedance ($|S_{11}|>1$) is not tolerable even for amateur use, because the slightest change in operating temperature or source impedance (i.e. antenna impedance) can lead to oscillations. So chances for successful EME-QSO's will be like roulette game.

For comparison Figure 2 displays the less critical situation for dual gate GaAs-FET's:

Figure 2:



The regions of instability are much smaller and therefore it's easier to construct stable LNA's with these devices. But the higher noise figure of these devices inhibits the application to very low noise preamps.

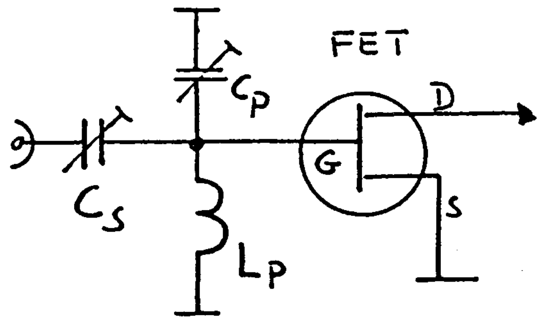
3.2. Input matching

Optimum source reflection coefficient Γ_{opt} for minimum noise figure is Γ_{opt} = 0,9/+21, for example. Translated to series impedance: 73 + j*249 Ohm. Or as parallel impedance: 922 ll +j*270 Ohm. These values are typical for single gate GaAs-FET's on 432 MHz.

The main function of the input circuit is to transform 50 Ohms real source impedance to the optimum source impedance above. This has to be done with minimal losses, because the matching losses degrade the noise figure of the device.

A well proven input circuit is given in figure 3:

Figure 3:



C_s transforms to the real part of optimum source impedance. Typical value for C_s is 1,4 pF. C_p provides in conjunction with L_p the adjustable imaginary part of optimum source impedance. Input losses calculate as:

$$L_K = \frac{1}{1 - \frac{Q_L}{Q_0}} \quad [2]$$

Table 1 gives the input losses for various combinations of loaded and unloaded Q:

Table 1:

Input circuit loss [dB]				
Q_L : Q Loaded	Q_0 : Q Unloaded			
	300	500	750	1000
10	0.15	0.09	0.06	0.04
15	0.22	0.13	0.09	0.07
18	**0.27**	**0.16**	**0.11**	**0.08**
20	0.3	0.18	0.12	0.09

Typical loaded Q value is 18 for a C_p value of 7 pF. To achieve input losses less than 0.1 dB unloaded Q values of more than 750 are required.

3.2.1. Output matching

Stability is the **main** concern in the design of output matching. A classical solution[4] for the best compromise between gain and stability is a 4:1 transmission line transformer at the drain, which provides a broadband 200 Ohms load at the drain with moderate gain (15 - 18 dB). This solution, originated by W6PO, is very well verified by the stability circles in figure 1. Output matching has no influence on noise figure NF!

But gain shouldn't drop too low, because the gain referenced noise measure M will be degraded:

$$M=\frac{F-1}{1-\frac{1}{G}} \qquad [3]$$

- M: Noise measure
- F: Noise factor
- G: Available gain of preamp

M + 1 is the resulting noise factor, if an infinite number of preamps of the same kind is cascaded. Or in other words, if you don't have a better preamp with lower noise figure in your system, you will never get a better noise factor than M + 1 for the whole system. Therefore it's wise to optimize M instead of F, because it's a lower bound to the system noise measure. This fact rules out any type of resistive broadband matching by a simple resistor in the drain circuit, because gain drops too low.

Most of the foregoing remarks are not new and have been described elsewhere in amateur literature. It's just for reminding some of the basic facts about preamps.

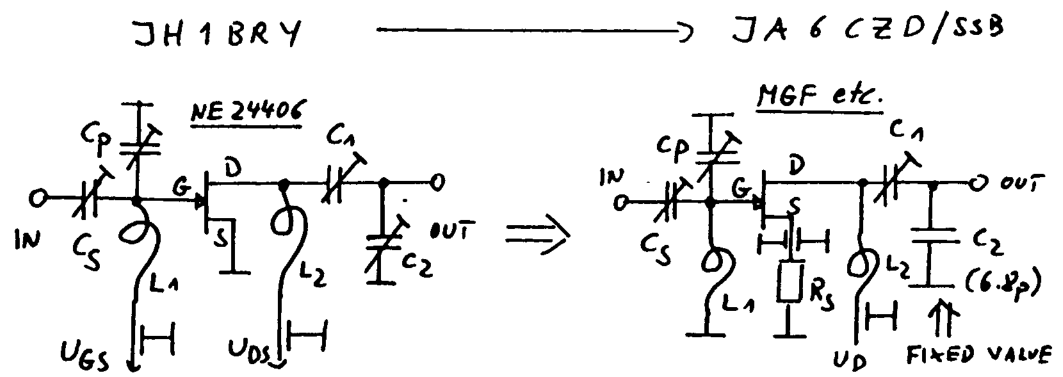
4. Construction of LNA's

4.1. An obsolete design

S. Sando, JH1BRY, has made one of the first designs of a single gate GaAs-FET LNA on 432 MHz[5].

The circuit is shown in figure 4:

Figure 4:



The GaAs-FET used is a NEC V244. L_1 und L_2 are simple wire loops with low unloaded Q (ca. 250). C_s and C_p are the capacitors for adjusting the source impedance. Output matching is adjusted by C_1 and C_2.

When replicating the design, JA6CZD changed air trimmer C_2 to a fixed capacitor of 6.8 pF. The same holds for the copy of this design by SSB-Electronics in West Germany, which are sold as the DX432 series.

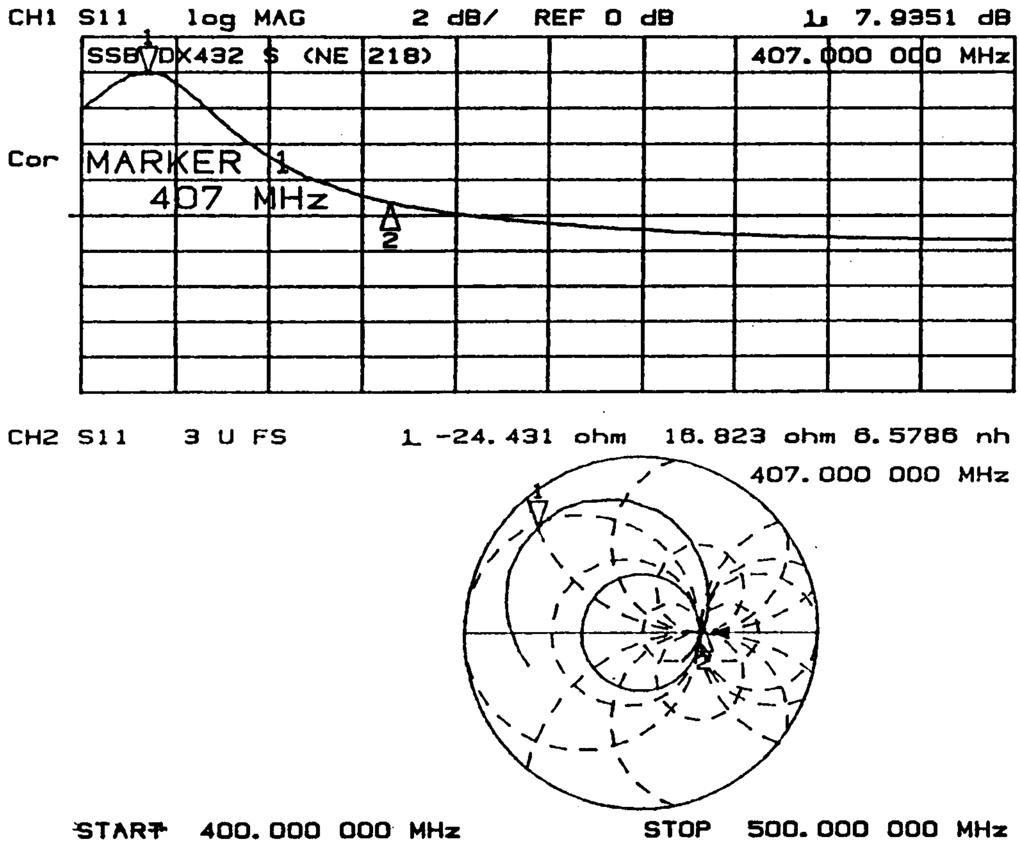
This little detail change shifts the simple and somewhat crude workbench prototype design of JH1BRY, which will work satisfactorely with careful tuning on a network analyzer, to a dangerous and not reproducible design, which should be avoided for the following reasons:

1. For modem GaAs-FET's the low unloaded Q of the input circuit produces too much losses. With the $2\mu m$-GaAs-FET V244, which has a mediocre noise figure of say 0.7 dB, an additional loss of 0.2-0.3 dB may be tolerable. But that's definitely not true for modern $0{,}3\mu m$-GaAs-FET-Triodes, which are capable of noise figures of 0.2 - 0.3 dB on 432 MHz.

2. If you tune C_1 the resulting transformation or drain load impedance depends on the variations of L_2. If L_2 is large, C_1 has to be tuned to a low value and hence to the unstable region (see Figure 1) because of the large resulting transformation ratio of C_1 to the fixed 6.8 pf capacitor. The result is a negative real input impedance and instability of the preamp. 8 different exemplars of the DX432 variety showed up with negative real input impedance ($|S_{11}|>1$). So Murphy holds again: What can go wrong will go wrong!

S_{11} for a SSB-Electronics DX432S preamp with a NEC NE218 device is shown in figure 5:

Figure 5:



$|S_{11}|>1$ holds for all frequencies lower than 441 MHz. **Return-Gain** (not **Return-Loss**) is 0.8 dB at 432 MHz and climbs to a maximum value of 7.93 dB at 407 MHz! That's a nice circuit for an oscillator, if the input power is reflected with a gain of 6.92.

4.2. A working design

The following design (Figure 6) is in use at the author's station and has been successfully duplicated by other EME-amateurs. It's main features are:

- Low loss input circuit

  A high-Q input circuit is made from a silvered stripline with optimal impedance (75 Ohms) and really high-Q Trim-Capacitors like Tekelec 5700/5800 series, which have twice the Q-factor than the Johanson 5200 variety.

- Transformer output match

  Ouput match is accomplished by a 4:1 broadband transformer for moderate gain (15-18 dB) and conditional stability ($|S_{11}|<1$ for 50 Ohms load, see figure 7a)

Figure 6 (Circuit Diagram):

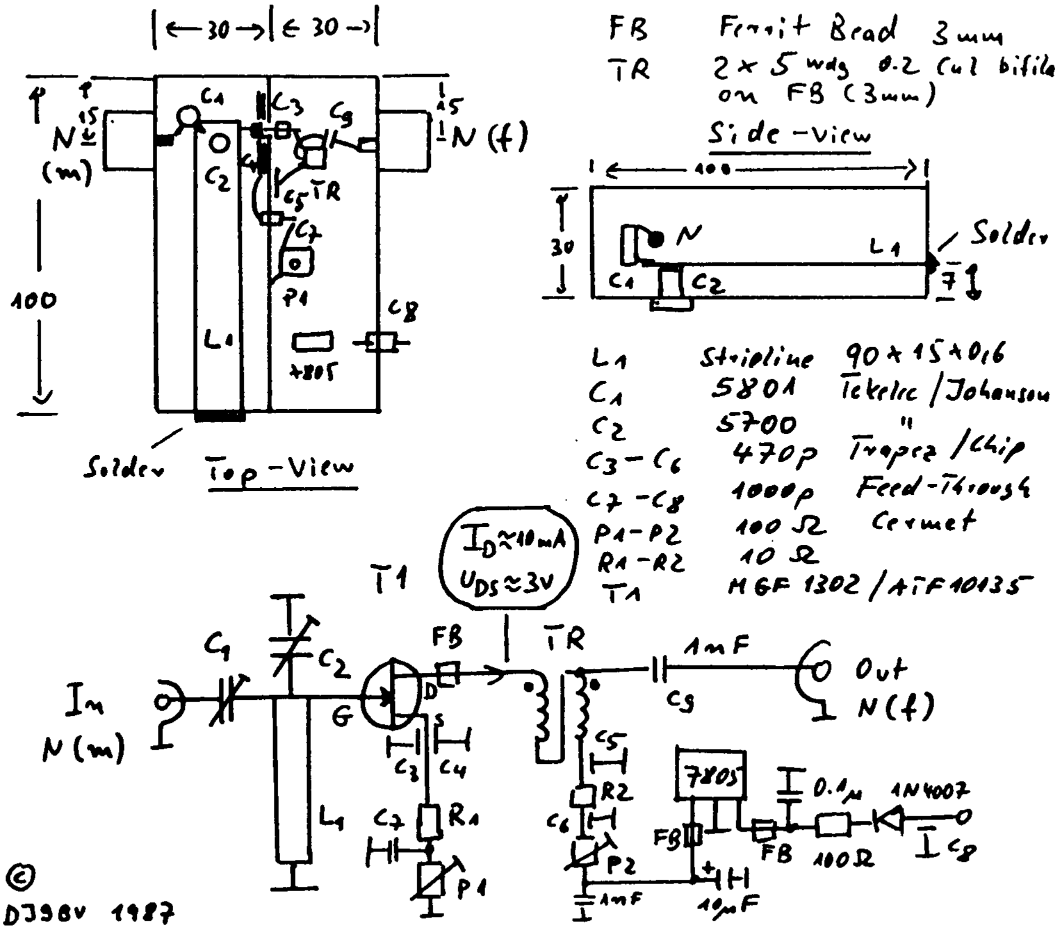


Figure 7 (Gain Curve):

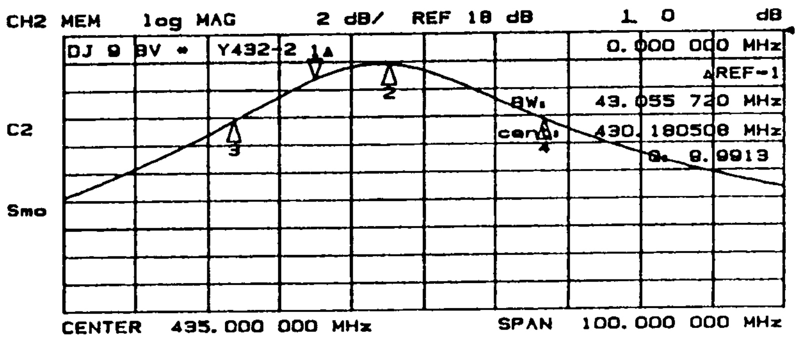
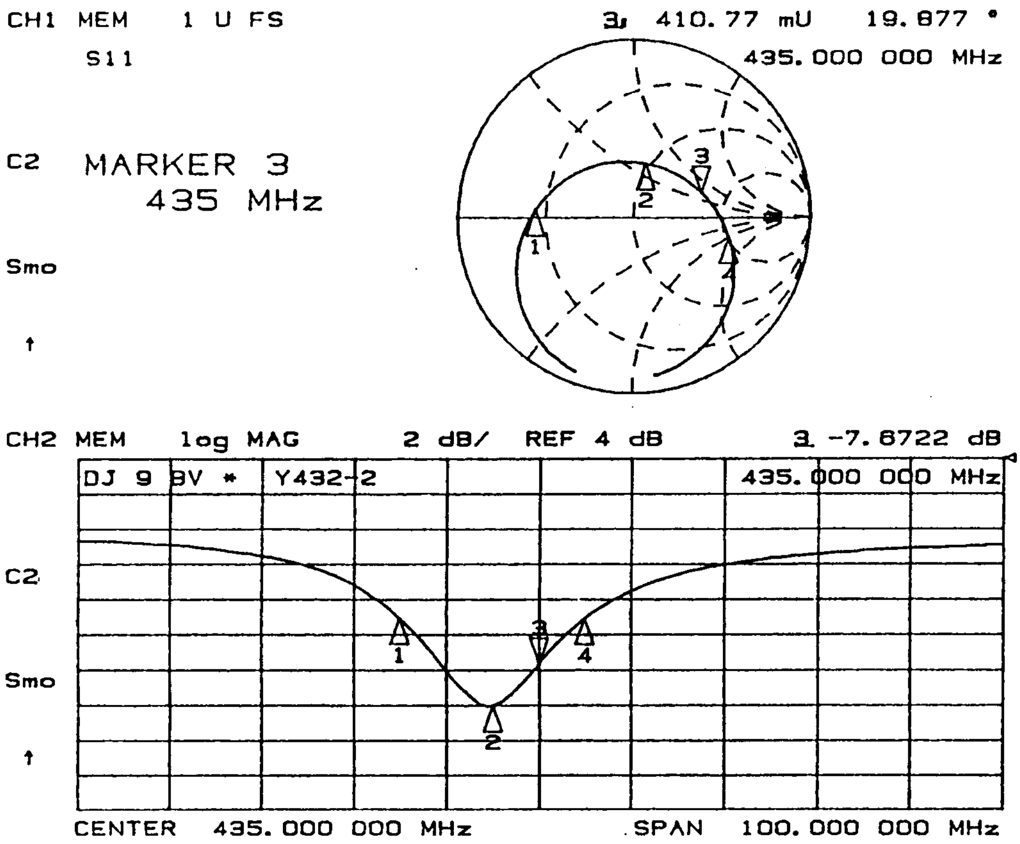


Figure 7 shows, that Q loaded has a value of 10. Therefore an input-loss of less than 0.1 dB is assured with a measured Q unloaded of about 850 for this type of input circuit.

Figure 7a displays the well behaved input return loss for 50 Ohms real load impedance:

Figure 7a (Input Reflection Characteristic):



Improvements to this design can be made by improving the unloaded Q of the input circuit even further by constructing a real input cavity with solderless junctions. DL9KR has made proposals about this topic and has built some prototypes.

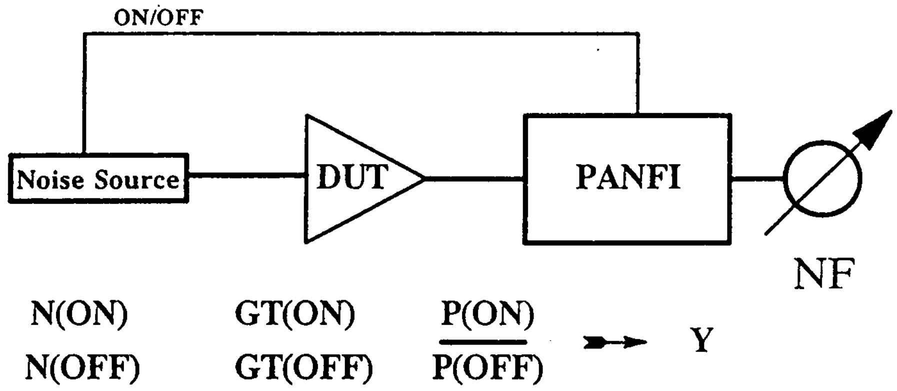
5. Measurement of noise figure of LNA's

5.1. Basics

Modern automatic noise figure measurement equipment from AIL or HP utilizes a switchable noise source and a PANFI (Precision Automatic Noise Figure Indicator), which measures the ratio of noise powers delivered by the DUT (Device under test) to the indicator.

This ratio is the Y-factor (Figure 8):

Figure 8:



Y-factor is defined as:

$$Y=\frac{P_{on}}{P_{off}} \qquad [4]$$

By referencing the output noise power to the input of the DUT by virtue of the transducer gain of the DUT, the following equation holds:

$$Y=\frac{GT(ON)\cdot(N_{on}+NDUT(ON))}{GT(OFF)\cdot(N_{off}+NDUT(OFF))} \qquad [5]$$

- GT(ON): Transducer gain of DUT with noise source ON
- GT(OFF): Transducer gain of DUT with noise source OFF
- N_{on}: Available noise power of noise source during ON-State
- N_{off}: Available noise power of noise source during OFF-State
- NDUT(ON): Noise power of DUT with noise source ON
- NDUT(OFF): Noise power of DUT with noise source OFF

Equation [5] is the exact formulation of the Y-factor, because not only the transducer gain but the also the unknown noise power NDUT are dependent on the reflection coefficient of the noise source. In normal applications an ideal noise source is assumed (i.e. $\Gamma_{on}=\Gamma_{off}=0$) and [6] is reduced to:

$$Y=\frac{N_{on}+N_{dut}}{N_{off}+N_{dut}} \qquad [6]$$

because now GT(ON) = GT(OFF) und NDUT(ON) = NDUT(OFF) = N_{dut} holds.

Replacing N=KTB and $F=1+T_{dut}/290$ and $ENR=T_{on}/290-1$ leads to:

$$F=\frac{ENR-Y\left(\frac{T_{off}}{290}-1\right)}{Y-1} \qquad [7]$$

This simplified equation holds under certain restrictions only, but is used by all commercial PANFI's from AIL and HP. Therefore this equipment will display false values for the noise figure NF if applied to Non Impedance Matched DUT's for example GaAs-FET-LNA's in the Sub-GHz-range (144-1296 MHz). It will be shown, that even with good but inevitably not 'ideal' noise sources the measurement error can have the same magnitude as the expected value of NF.

5.2. Sources of error for automatic noise figure measurements

5.2.1. Classical error sources

Before switching to the analysis of a subtle source of measurement error, which is typical for an amateur environment, we should remember the "classical" sources of error, which had been subject of application[6] or amateur literature[7] :

1. Spurious signals
   Bad setups are sensitive to external and internal signals, which fall into the bandwith of the DUT or the PANFI. Spurious signals produce erroneous values for the observed Y-factor. Cures are:
   - Shielding
   - Careful selection of mixing scheme
2. Noise contribution
   Subsequent stages add noise. This influence can be corrected by calculation.
3. Temperature error
   The noise source is calibrated to 290 K. Deviations from the standard temperature can be calculated and corrected.
4. Calibration error
   The noise source has a calibration error. This error is significant for absolute measurements only and can not be corrected. Typical calibration precision is around +- 0.3 dB.
5. Nonlinear detector
   Modern PANFI's exhibit typical nonlinearities of 0.05 dB.

5.2.2. Gain error

If the noise source is not ideal (i.e. $\Gamma_{on} <> \Gamma_{off} <> 0$), two additional measurement errors arise.

First mismatch error includes such effects as change of the ENR by reflection of the DUT's internal noise power and change of the DUT's internal noise power by the changing reflection coefficient of the noise source. We will neglect these effects.

Second gain error DG, caused by the changing source impedance of the switching noise source and the subsequent change in transducer gain of the DUT, lets the PANFI measure a wrong value of Y-factor. Then a wrong value of NF is calculated by the PANFI according equation [7]. Especially in the case of measuring non impedance matched LNA's in the sub-GHz-range (< 1 GHz) the gain error has a considerable magnitude of up to 1 dB for conventional noise sources!

For the gain error the following equation holds (see Appendix A):

$$DG = \frac{1-2|S_{11}||\Gamma_{off}|\cos(\phi_{11}+\phi_{off})}{1-2|S_{11}||\Gamma_{on}|\cos(\phi_{11}+\phi_{on})} \qquad [8]$$

The magnitude depends both on the input reflection coefficient of the DUT and the quality of match of the noise source. The magnitude of the gain error exhibits a cosine-type function, if the angle of the input reflection coefficient of the DUT is changed.

Let's look to the numbers!

We have calculated the error value limits for three popular noise sources, i.e. the AIL7615, the HP346B and the HP346A. The source reflection coefficients of the first two sources have been measured on a HP-Network-Analyzer 8753A. The values for the HP346A have been estimated from the data sheet and display the worst case for this source. The analysis has been carried out for a frequency of 432 MHz.

Table 2:

Reflection coefficient of some noise sources @ 432MHz				
Noise source	Ser.-Number	Γ_{on}	Γ_{off}	$\lvert\Gamma_{on}-\Gamma_{off}\rvert$
AIL7615	4044	0.015/+120	0.014/-112	0.027
HP346B	2037A00657	0.039/109	0.009/-61	0.048
HP346A	===	0.005/+90	0.005/-90	0.01

Figures 9,10 and 11 show the magnitude of gain error calculated by equation A3 (Appendix A) for the AIL7615, HP346B and HP346A for $|S_{11}|=1$. A second table shows the values of maximum gain error for various values of $|S_{11}|$.

Table 3:

Worst case gain error versus DUT's $\lvert S_{11}\rvert$ (432 MHz)				
DUT		Gain error ($\pm dB$)		
$\lvert S_{11}\rvert$	RL(dB)	AIL7615	HP346B	HP346A
1.2	1.6	0.29	0.50	0.11
1.1	0.8	0.26	0.46	0.10
1.0	0.0	0.23	0.42	0.09
0.9	-0.9	0.20	0.38	0.08
0.8	-1.9	0.17	0.34	0.07
0.7	-3.0	0.15	0.30	0.06
0.6	-4.4	0.12	0.25	0.05
0.5	-6.0	0.10	0.20	0.04
0.0	-∞	0.00	0.00	0.00

The worst case gain error obviously is greater than the display accuracy of the PANFI ($\pm 0.1db$) but even greater than the expected value for NF. When using a HP346B noise source you cannot measure noise figures under 1 dB if the DUT is a non impedance matched LNA - i.e. GaAs-FET preamps in amateur style on 432 MHz-. Even tuning to minimal noise figure is not possible, because the true minimum will be missed by just the same error. So one will arrive with a badly tuned preamp in most cases, because the PANFI displays a fictive value for NF with such a noise source. Many EME-amateurs have made this experience after 'optimizing' their preamp on a PANFI and installing this preamp into their system.

5.3. Analysis of unknown measurement setups for possible gain error

The main source of measurement error, which cannot be corrected easily, is gain error, if you want to measure the NF of DUT's with an input reflection coefficient $|S_{11}|>0.5$. That is true even for very good noise sources. It has been shown in a previous chapter, that for various types of amateur style GaAs-FET-Amplifiers on 432 MHz the magnitude of S_{11} shows values between -6 dB (0.5) and +3 dB (1.4)! Therefore the foregoing discussion is relevant for noise figure measurements of such devices. Table 3 shows the gain error to have nearly the same magnitude as the expected noise figure NF.

What should be done in this situation?

1. Make your own table of gain error for your favorite noise source after measuring its reflection coefficient with a network analyzer.
2. Use a network-analyzer to determine the input reflection coefficient S_{11} of each device you want to measure.
3. There are experimental means to prove the existence of gain error in a certain measurement setup. You have to insert a lossless variable phase line such as SAGE 6502 between noise source and DUT. By changing the phase you can verify equation [8].

   The poor man's approach is a 42 degree line consisting of a UG57B/U and a UG29B/U N-Connector. This line has a loss less than 0.05 dB on 432 MHz. In most cases it will produce a very nice change of the indicated noise figure.

   Also a set of two RG213 cable with 90 degrees and 135 degrees electrical length can produce very nice results of gain error. If possible, one should try at least two different phase shifts to avoid possible cancellations, which are caused by the multivalued cosine term in equation [8].

5.4. Minimisation of gain error

For reducing gain error the equation [8] suggests two possibilties:

1. Improvement of match of noise source
   The quality of the noise source ($\Delta\Gamma = |\Gamma_{on} - \Gamma_{off}|$) has to be as good as possible.
2. Improvement of match of preamp
   Preamps like the design of JA6CZD or the DX432 series from SSB-Electronics have negative real part of input impedance. This kind of preamp is instable even with 50 Ohms load impedance and the resulting gain error during measurement of noise figure is much higher than for well behaved constructions a la W6PO - i.e. with a HP346B noise source gain error is 0.46 dB with $|S_{11}| = 1.1$ compared with 0.3 dB gain error for $|S_{11}| = 0.7$). These types of amplifiers are badly tuned in most cases and will not provide the expected system temperature for the receiver.

Improvement of the match of the noise source can be done by two different means:

1 Insertion of an attenuator between noise source and DUT

2. Insertion of an isolator between noise source and DUT

Isolation of an attenuator is twice its attenuation. ENR is reduced by the attenuation. This method has been applied by HP to its noise source HP346A, which in fact is a HP346B with an additional internal 10 dB pad. A very good value for $\Delta\Gamma = |\Gamma_{off} - \Gamma_{on}| < 0.01$ is guaranteed. This noise source is recommended for non impedance matched DUT's. This method is especially suited for broadband sources.

For narrow band applications an isolator between noise source and DUT will improve the available isolation even further (26 dB typ.). The attenuation is only 0.3 dB. This high isolation effectively masks any change in reflection coefficient of the noise source (Appendix B). Figure 12 shows the gain error of an AIL7615 source plus an isolator. So even with very bad behaving DUT's ($|S_{11}| > 1$) gain error will be less than 0.1 dB. Therefore an isolator provides the best means for measuring non impedance matched DUT's in narrow band applications even with mediocre noise sources.

The following table gives some comparisons:

Table 4:

Worst case gain error for various noise sources @ 432 MHz						
DUT		Gain Error [$\pm dB$]				
$\lvert S_{11}\rvert$	RL[dB]	AIL7615	HP346B	HP346A	HP346B+Isolator	AIL7615+Isolator
1.4	3.0	0.34	0.6	0.12	0.07	0.035
1.2	1.6	0.29	0.5	0.11	0.06	0.032
1.0	0.0	0.23	0.42	0.09	0.05	0.030
0.8	-1.9	0.17	0.34	0.07	0.04	0.025
0.6	-4.4	0.12	0.25	0.05	0.035	0.02
0.0	$-\infty$	0.0	0.0	0.0	0.035	0.02

The following table shows the total error for relative measurements:

Table 5:

Total error for relative measurements ($\lvert S_{11}^{dut}\rvert \leq 1.0$)						
Error Type	Typ. (RSS) [$\pm dB$]			Maximum [$\pm dB$]		
	HP346B	AIL7615+Iso.	HP346A	HP346B	AIL7615+Iso.	HP346A
Y-Factor		0.05			0.1	
Gain	0.21	0.015	0.05	0.42	0.03	0.09
Total error	0.22	0.05	0.07	0.52	0.13	0.19

The table above is applicable to the case of **comparison** of different DUT's on the same PANFI. If you want to compare different DUT's on different PANFI's , i.e. . for **absolute** measurements, you have to use the following table:

Table 6:

Total error for absolute measurements ($\lvert S_{11}^{dut}\rvert \leq 1.0$)						
Error type	Typ. (RSS) [$\pm dB$]			Maximum [$\pm dB$]		
	HP346B	AIL7615+Iso.	HP346A	HP346B	AIL7615+Iso.	HP346A
ENR-Error		0.15			0.3	
Y-Factor		0.05			0.1	
Isolator	0.00	0.02	0.00	0.00	0.05	0.00
Gain	0.21	0.015	0.05	0.42	0.03	0.09
Total error	0.26	0.16	0.17	0.82	0.48	0.49

Appendix A: Gain error

Gain error is defined as:

$$DG=\frac{GT(ON)}{GT(OFF)} \tag{A1}$$

and we assume:

$$T_{dut}=TDUT(ON)=TDUT(OFF)$$

Then from equations [5] and [6] follows:

$$YTRUE=DG\frac{T_{on}+T_{dut}}{T_{off}+T_{dut}}$$

or with [A1]

$$YTRUE=DG\times YIND \tag{A2}$$

From the definition of transducer gain[8] :

$$GT=|S_{21}|^2\frac{1-|\Gamma_G|^2}{|1-\Gamma_1\Gamma_G|^2}\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2} \tag{A3}$$

with

$$\Gamma_1=S_{11}+\frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}$$

follows:

$$DG=\frac{1-|\Gamma_{on}|^2}{|1-\Gamma_1\Gamma_{on}|^2}\times\frac{|1-\Gamma_1\Gamma_{off}|^2}{1-|\Gamma_{off}|^2} \tag{A4}$$

which can be reduced to :

$$DG=\frac{1-2|S_{11}||\Gamma_{off}|cos(\phi_{11}+\phi_{off})}{1-2|S_{11}||\Gamma_{on}|cos(\phi_{11}+\phi_{on})} \tag{A5}$$

S_{11}	Input reflection coefficient of DUT
Γ_{on}	Reflection coefficient of noise source during state ON
Γ_{off}	Reflection coefficient of noise source during state OFF
ϕ_{11}	Angle of S_{11}
ϕ_{on}	Angle of Γ_{on}
ϕ_{off}	Angle of Γ_{off}

Equation [A5] is valid under the following assumptions:

1. $\left|\frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}\right| \ll |S_{11}|<2.0$
2 $|\Gamma_{off}|<0.05$
3. $|\Gamma_{on}|<0.05$

Appendix B: Gain error reduction by an isolator

If we reformulate equation [A1] we get:

$$DG_{iso}=\frac{G_{av}^{iso}(ON)}{G_{av}^{iso}(OFF)}\times\frac{GT_{dut}(ON)}{GT_{dut}(OFF)} \qquad [B1]$$

This equation can be simplified to:

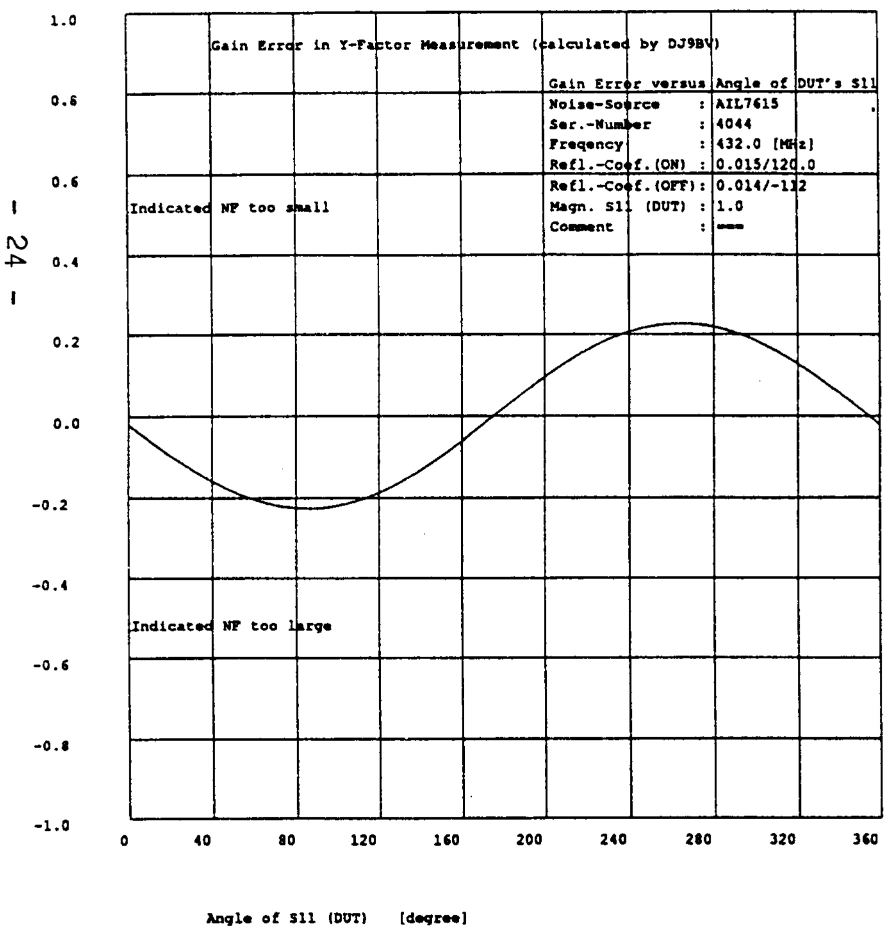
$$DG_{iso}=\frac{|1-\Gamma_{off}S_{11}^{iso}|^2}{|1-\Gamma_{on}S_{11}^{iso}|^2}\times\frac{|1-S_{11}^{dut}(S_{22}^{iso}+S_{12}^{iso}\Gamma_{off})|^2}{|1-S_{11}^{dut}(S_{22}^{iso}+S_{12}^{iso}\Gamma_{on})|^2} \qquad [B2]$$

The first term in equation [B2] denotes the mismatch error between noise source and isolator. The second term denotes the mismatch error between the system (noise source + isolator) and the DUT.

References

1. R. Bertelsmeier, "Limitations of automatic noise figure measurement equipment," *Unpublished Paper*, March 1987.
2. R. Bertelsmeier, "Ultrarauscharme GaAs-FET Empfangsvorverstaerker fuer EME," *32. Weinheimer UKW-Tagung, Scriptum der Vortraege*, Weinheim, West Germany, September 1987.
3. R. Bertelsmeier, "Effective Noise Temperature of 4-Yagi-Arrays for 432 MHz EME," *DUBUS*, vol. 4, 1987.
4. Eimac, "Some GaAs FET Preamplifiers," *EIMAC Application Note*, vol. AS-49-31, EIMAC.
5. S. Sando, "Very Low Noise GaAs-FET Preamp for 432 MHz," *HAM RADIO*, vol. 4, pp. 22-27, 1978.
6. Hewlett-Packard, "Fundamentals of RF and Microwave Noise Figure Measurements," *HP Application Note 57-1*, Palo Alto, 1983.
7. J. Gannaway, D. Holms, "Fussangeln bei Rauschzahlmessungen," *UKW-Berichte*, pp. 77-81, 1982.
8. Hewlett-Packard, "S-Parameter Design," *HP Application Note 154*, Palo Alto, 1972.

FIGURE 9



Y-Factor
Error
[dB]

FIGURE 10

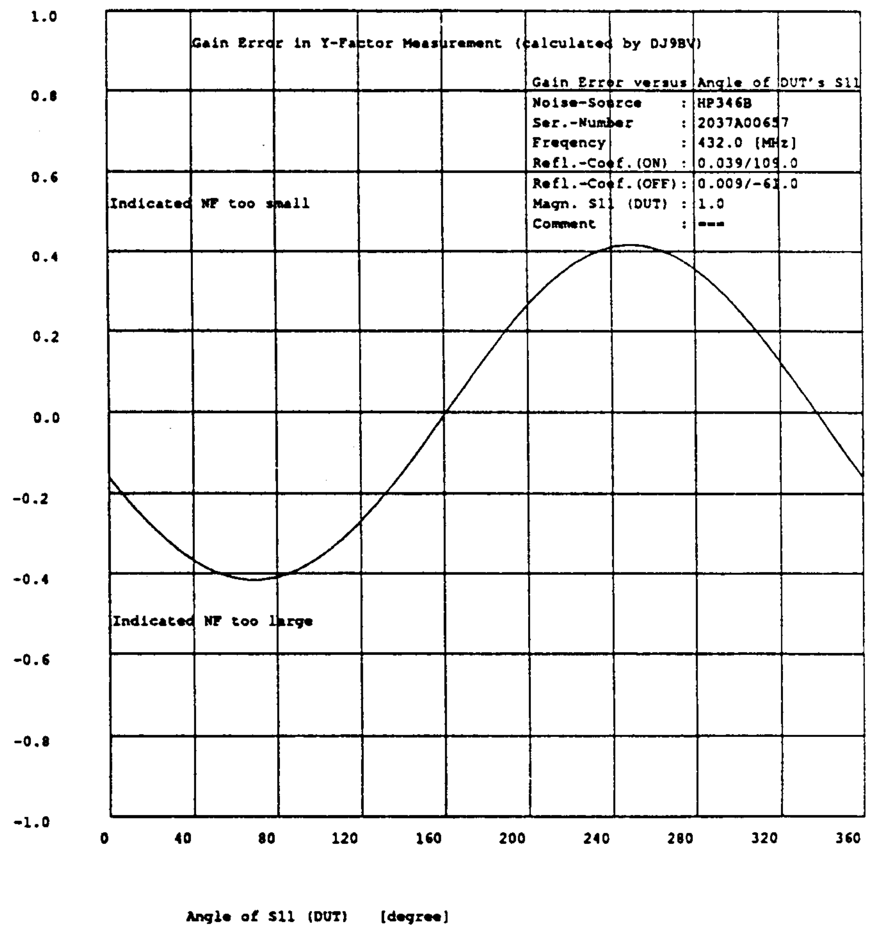
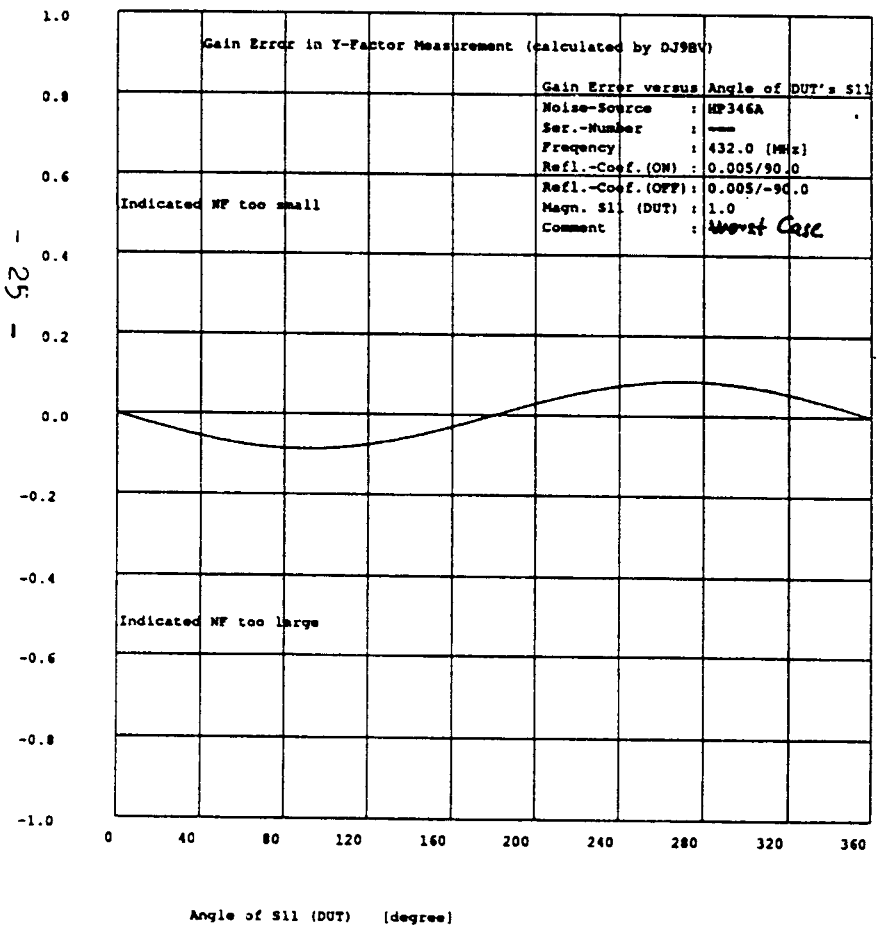


FIGURE 11



Y-Factor

Error

[dB]

FIGURE 12