A BROADBAND HOMODYNE NETWORK ANALYSER WITH BINARY PHASE MODULATION

Uwe Gärtner and Burkhard Schiek

Institut für Hoch- und Höchstfrequenztechnik, Ruhr-Universität Bochum
D-4630 Bochum, Federal Republic of Germany

ABSTRACT

An automatic homodyne network analyser system using cascaded binary phase shifters is described which operates over the full X-band with a high dynamic range due to an auxiliary modulation and linear detection technique. A general analytic solution of the complex, nonlinear system equations is given which allows the use of a coupled modulator/phase shifter structure. Measurement results are reported.

I. INTRODUCTION

Until now the research in homodyne systems for the determination of the complex scattering parameters of a device under test (DUT) has mainly focussed on five- or six-port techniques. A traditional drawback of this type of network analyser (NWA) is the restricted dynamic range due to a limited resolution of the power meters [1].

This paper presents a X-band NWA system based on a novel principle [2,3,4], which combines the simplicity of the RF-hardware used with the potential sensivity of heterodyne detection principles.

II. SYSTEM CONCEPT

A block diagramm of the NWA circuit is given in Fig.1. The object connected to the test ports of the system may either be the device under test (DUT) itself for transmission measurements or a test-set for combined transmission / reflection measurements.

A homodyne linear downconversion of the test signal (employing e.g. a balanced mixer) only yields the real part of its complex amplitude. However, amplitude and phase information are gained if the test signal is subjected to a digital triple phase modulation prior to frequency conversion using three decoupled binary phase shifters PS1, PS2, PS3 [2,3,4]. An additional amplitude modulation with a modulation frequency of e.g. f_m = 10 kHz together with a synchronous detection in the intermediate frequency (IF) range increases the sensivity of the system since it allows a low noise narrowband amplification of the IF-signal for a better noise reduction.

The position of the modulator is determined by technical considerations. The high backward isolation of the chain of phase shifters attenuates the L.O. signal leaking through the mixer which -if reflected at the modulator- yields a disturbing crosstalk.

Because of the use of three decoupled digital phase shifters the transfer function of the series connection of PS 1,2,3 can adopt 8 different values and thus the modulated test signal is converted to an IF - signal of 10 kHz with 8 different and in general complex amplitudes $\underline{V}_i$, i = 1...8.

A fast but approximate evaluation of the test signal from the measurement data is obtained -save for a complex system constant- with increasing accuracy from a weighted sum of 2, 4 or 8 of the complex amplitudes $\underline{V}_i$ = 1...8.

An exact solution of the system equations is time consuming and requires a minimum of three decoupled phase shifters. It yields the complex amplitude of the test signal as well as the characteristics of PS 1,2,3 -which are needed for the weighted evaluation- irrespectively of the amplitude and phase of the test signal. Thus it is very usefull as a calibration algorithm.

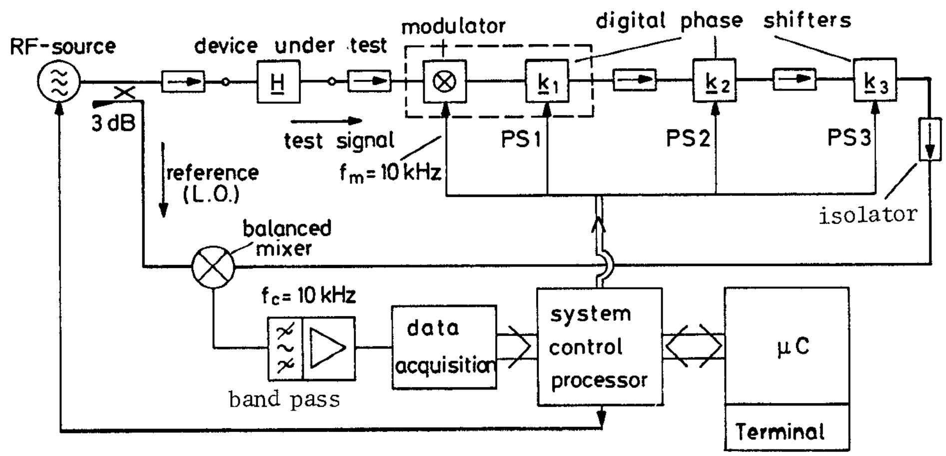

Fig.1: Computer controlled homodyne X-band network analyser system

0149-645X/85/0000-0725 $01.00 © 1985 IEEE 1985 IEEE MTT-S Digest

III. SYSTEM EQUATIONS

Consider a test signal with a complex amplitude $\underline{K}_0 \cdot \underline{H}$, where $\underline{K}_0$ is a system constant and $\underline{H}$ is the transfer function of the device under test (including error terms for crosstalk, mismatch etc. which have to be removed by calibration measurements), a modulator M operating at f_m = 10 kHz with the conversion factors $\underline{B}_u$ and $\underline{B}_\ell^*$ to the first upper and lower sideband and three cascaded and decoupled phase shifters with their respective characteristics $\underline{k}_{iu,\ell}$, i = 1..3, where the $\underline{k}_{iu,\ell}$ describe the phase shift and insertion loss changes of the binary elements for the upper and lower sideband frequency.

The complex amplitude of the IF-signal after a bandpass filtering with a center frequency $f_c = f_m$ is obtained as the sum of the complex RF-amplitude of the upper sideband and the complex conjugate of the RF-amplitude of the lower sideband. Therefore, the following 8 system equations are valid:

$$\begin{aligned}
\underline{V}_1 &= \underline{K}\,\underline{H}\,\underline{B}_u &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^* \\
\underline{V}_2 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{1u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{1\ell}^* \\
\underline{V}_3 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{2u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{2\ell}^* \\
\underline{V}_4 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{3u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{3\ell}^* \\
\underline{V}_5 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{1u}\,\underline{k}_{2u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{1\ell}^*\,\underline{k}_{2\ell}^* \\
\underline{V}_6 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{1u}\,\underline{k}_{3u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{1\ell}^*\,\underline{k}_{3\ell}^* \\
\underline{V}_7 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{2u}\,\underline{k}_{3u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{2\ell}^*\,\underline{k}_{3\ell}^* \\
\underline{V}_8 &= \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{1u}\,\underline{k}_{2u}\,\underline{k}_{3u} &&+ \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*\,\underline{k}_{1\ell}^*\,\underline{k}_{2\ell}^*\,\underline{k}_{3\ell}^*
\end{aligned} \qquad \text{(1a-1h)}$$

In general the equations (1a-h) form a system of 8 complex equations for the 8 complex unknowns $\underline{K}\cdot\underline{H}\cdot\underline{B}_u$, $\underline{K}^*\underline{H}^*\underline{B}_\ell^*$, $\underline{k}_{1u}$, $\underline{k}_{1\ell}^*$, $\underline{k}_{2u}$, $\underline{k}_{2\ell}^*$, $\underline{k}_{3u}$, $\underline{k}_{3\ell}^*$ since the conversion factor $\underline{B}$ as well as the phase shifter characteristics $\underline{k}_i$ may differ for the upper and lower sideband frequency. Nevertheless, if the modulation frequency is chosen small compared to the bandwidth of the phase shifters and if a decoupled amplitude modulator with symmetrical sidebands ($\underline{B}_u = \underline{B}_\ell$) is employed it is a good approximation to assume $\underline{B}_u = \underline{B}_\ell$, $\underline{k}_{iu} = \underline{k}_{i\ell}$, whereby (1a-h) degenerates to a system of 8 real equations in 4 complex unknowns.

A solution for this case as well as for the case $\underline{B}_u \neq \underline{B}_\ell$, $\underline{k}_{iu} = \underline{k}_{i\ell}$, has already been given [4].

Assuming a mutual coupling of M and PS1 the conversion factors $\underline{B}_{u,\ell}$ will no longer be independent of the setting of PS1 i.e.

$$\underline{V}_1 = \underline{K}\,\underline{H}\,\underline{B}_u + \underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^* \qquad (2a)$$

$$\underline{V}_2 = \underline{K}\,\underline{H}\,\tilde{\underline{B}}_u\,\tilde{\underline{k}}_{1u} + \underline{K}^*\,\underline{H}^*\,\tilde{\underline{B}}_\ell^*\,\tilde{\underline{k}}_{1\ell}^* \qquad (2b)$$

where the ~ denotes the modification due to coupling. This has to be taken into account by assuming a modification of the characteristic $\underline{k}_{1u,\ell}$

$$\underline{k}_{1\,u,\ell} = \tilde{\underline{B}}_{u,\ell} / \underline{B}_{u,\ell} \cdot \tilde{\underline{k}}_{1\,u,\ell} \qquad (3)$$

Then a nonlinear system with at least five complex unknowns $\underline{K}\,\underline{H}\,\underline{B}_u$, $\underline{K}^*\,\underline{H}^*\,\underline{B}_\ell^*$, $\underline{k}_{1u}$, $\underline{k}_{1\ell}^*$, $\underline{k}_2$, $\underline{k}_3$ will result.

IV. AN ALGEBRAIC SOLUTION

An exact and general solution to the complex system equations (1a-h) may be given as follows.

We solve for $\underline{k}_{1\ell}^*$ ($\underline{k}_{1u}$) and obtain

$$\underline{k}_{1\ell}^* \,(\underline{k}_{1u}) = \underline{u} \;+/-\; (-/+)\; j\cdot\underline{w} \qquad (4a)$$

with

$$\underline{u} = \underline{b} / 2\cdot\underline{a} \;; \qquad \underline{w} = \sqrt{4\cdot\underline{a}\cdot\underline{c} - \underline{b}^2} \;/\; 2\cdot\underline{a}$$

and

$$\underline{a} = \underline{V}_{34} - \underline{V}_{17} \;; \quad \underline{b} = \underline{V}_{45} + \underline{V}_{36} - \underline{V}_{18} - \underline{V}_{27}$$

$$\underline{c} = \underline{V}_{56} - \underline{V}_{28} \;; \quad \underline{V}_{ij} = \underline{V}_i \cdot \underline{V}_j, \; i = 1..8.$$

and further

$$\underline{k}_{2u} = (\underline{V}_5 - \underline{V}_3\,\underline{k}_{1\ell}^*)/(\underline{V}_2 - \underline{V}_1\,\underline{k}_{1\ell}^*) \qquad (4b)$$

$$\underline{k}_{3u} = (\underline{V}_6 - \underline{V}_4\,\underline{k}_{1\ell}^*)/(\underline{V}_2 - \underline{V}_1\,\underline{k}_{1\ell}^*) \qquad (4c)$$

while $\underline{k}_{2\ell}^*$ and $\underline{k}_{3\ell}^*$ are determined from (4b,c) by replacing $\underline{k}_{1\ell}^*$ by $\underline{k}_{1u}$.

Finally with the help of (4a) the complex amplitude of the test signal is obtained from (1a,b):

$$\underline{K}\,\underline{H}\,\underline{B}_u = 1/2\,(\underline{V}_1 \;+/-\; j\,(\underline{V}_2 - \underline{u}\,\underline{V}_1)\,/\,\underline{w}) \qquad (5a)$$

or

$$\underline{K}\,\underline{H}\,\underline{B}_\ell = 1/2\,(\underline{V}_1 \;-/+\; j\,(\underline{V}_2 - \underline{u}\,\underline{V}_1)\,/\,\underline{w}) \qquad (5b)$$

The unknown system constants $\underline{K}\cdot\underline{B}_u$, $\underline{K}\cdot\underline{B}_\ell$ are determined by replacing the DUT by a known through connection or -for high precision measurements- using calibration standards for vector error correction procedures [5].

The ambiguity in the evaluation of $\underline{k}_1$ and thus in all other unknowns is resolved if the characteristic of at least one phase shifter is roughly known.

V. AN APPROXIMATE SOLUTION

Since the evaluation of equ. (5a,b) is time consuming it is advantageous to determine the test signal with the help of a different algorithm, a linear combination of 2, 4 or 8 of the complex quantities $\underline{V}_i$, i = 1..8 [2].

Using e.g. only PS1 and PS2 one obtains $\underline{V}_{1,2,3,5}$ and may form a linear combination

$$\begin{aligned}
\underline{H}_R &= \underline{V}_1 + \underline{p}_1\,\underline{V}_2 + \underline{p}_2\,\underline{V}_3 + \underline{p}_1\,\underline{p}_2\,\underline{V}_5 \\
&= \underline{K}\,\underline{H}\,\underline{B}_u\,(1 + \underline{p}_1\,\underline{k}_{1u})\,(1 + \underline{p}_2\,\underline{k}_{2u}) \\
&\quad + \underline{K}^*\underline{H}^*\underline{B}_\ell^*\,(1 + \underline{p}_1\,\underline{k}_{1\ell}^*)\,(1 + \underline{p}_2\,\underline{k}_{2\ell}^*)
\end{aligned} \qquad (6)$$

It is obvious that the right choice of the $\underline{p}_i$ is given by

$$\underline{p}_i = -\,1 \,/\, \underline{k}_{i\ell}^* \qquad (7)$$

which leaves

$$\underline{H}_R = \underset{\sim}{\underline{K}}\,\underline{H} \qquad (8)$$

where $\underset{\sim}{K}$ is a complex system constant.
Since the error term in (6) resulting from a poor choice of the $\underline{p}_i$ is of second order only good estimates for the $\underline{k}^*_{i\ell}$ have to be provided.

VI. DATA ACQUISITION AND SYSTEM CONTROL

In order to acquire the complex IF - amplitude $\underline{V}_i$, i = 1..8, a syncronous modulation and sampling process is used.
Following the setting of the phase shifters a quartz stable square wave signal of f_m = 10 kHz is applied to the modulator. After the decay of the transient response of the bandpass filter the sinusodial filter output is sampled and digitally converted each quarter of a cycle to yield the real- and imaginary part of $\underline{V}_i$ (the chosen settling time is 15 cycles and 3 cycles for sampling). The data acquisition unit includes a software controlled prescaler amplifier to enhance the dynamic range of the system.

The network analyser set-up is linked to a microcomputer system (µC) by a special subprocessor unit (Fig.2) which is devoted to control frequency, RF - phase shift, modulation and data acquisition with automatic gain control.

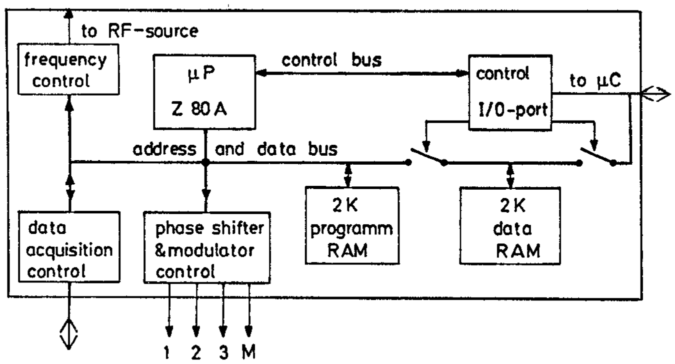

Fig.2: Subprocessor for automated data acquisition and system control

The measurement values are stored in a 2 k-Byte data buffer. After completion of one data acquisition cycle the microcomputer gains access to the data storage area using bus multiplexers.

VII. DATA PROCESSING AND MEASUREMENTS RESULTS

For an increased system speed only PS1 and PS2 are used together with a weighted evaluation of the measurement data $\underline{V}_{1,2,3,5}$ (see V.)
As a prerequisite estimates for the characteristics $\underline{k}_{1\ell}$, $\underline{k}_{2\ell}$ have to be available.
These are provided by a single measurement cycle over the frequency range employing 3 phase shifters and the algebraic solution for $\underline{k}^*_{1\ell}$, $\underline{k}^*_{2\ell}$ of equ. (4a). They are stored together with possible additional calibration terms derived from error correction measurements.

The characteristic k_1 of PS1 (decoupled PIN-diode "switched line" phase shifter, $\underline{k}_{iu} = \underline{k}_{i\ell}$) is given in Fig.3 for 32 discrete frequencies in the range of 8.5 - 11.5 GHz as given by the exact solution of equ. (4a).

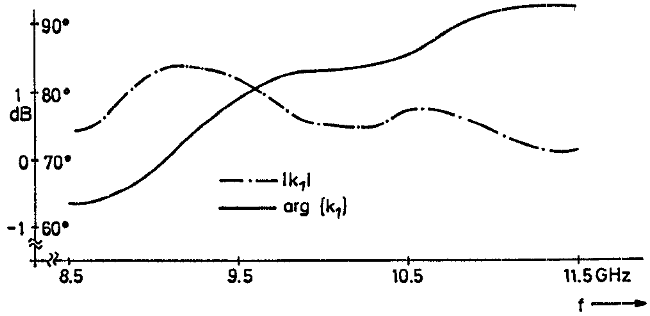

Fig.3: Phase shift and insertion loss changes of a pin-diode "switched line" phase shifter

To check the broadband measuring capabilities of this system the transfer phase of a precision waveguide phase shifter (HP X 855A) has been measured in the range of 8.5 - 11.5 GHz for the settings of the DUT of 0°, 45°, 90°... .
The maximum deviation of the calculated values from the reading of the DUT were limited to +/- 3° over the whole frequency range but this includes the non-ideality of the reference phase-shifter.

The dynamic range of the system is demonstrated by Fig.4. The DUT has been realized by a series connection of two waveguide attenuators and a precision waveguide phase shifter. Phase linearity errors are limited to approximately +/- 2° for an insertion loss range α_{DUT} of 0-65dB.

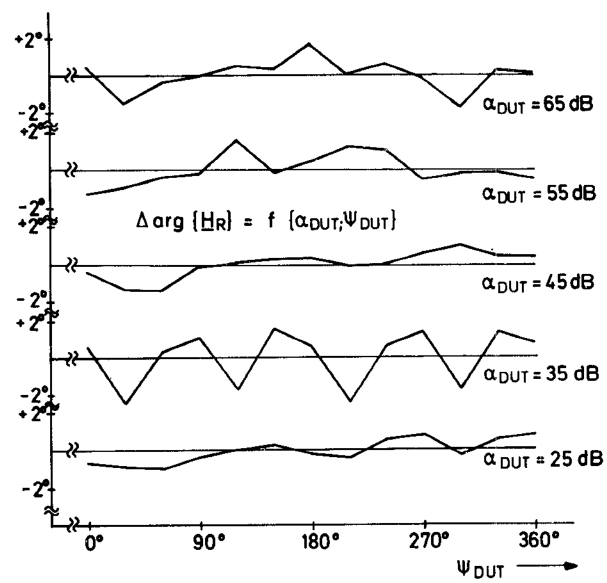

Fig.4: Deviations of the calculated phase $\{\underline{H}_R\}$ from linearity versus the transfer phase Ψ_{DUT} of the DUT for different insertion losses α_{DUT}; weighted evaluation

No serious attemps have yet been made to improve the signal to noise ratio which at present is limited by the interference of the computer clock pulse. An improvement of several decades is expected using an optimized IF-circuitry. Finally refinements in the RF - set-up like the use of a bi-phase (180°) modulator and an increased backward isolation should yield results comparable to heterodyne systems employing a downconversion at the fundamental frequency.

VIII. ACCURACY LIMITS DUE TO SYSTEMATIC ERRORS

The non-ideal behavior of the network analyser circuit due to testport mismatch, directivity and crosstalk can be minimized by standard means if the vector voltmeter operation of the circuit itself is exact. The latter is mainly influenced by the coupling of two neighbouring phase shifters used.

The only important interaction of the phase shifters stems from a switching state dependent reflection coefficient $\underline{\Gamma}_{o1}$+/-$\Delta\underline{\Gamma}$, $\underline{\Gamma}_{i2}$+/-$\Delta\underline{\Gamma}$ of the output port of e.g. PS1 and the input port of PS2. A numerical example shows that a $|\Delta\underline{\Gamma}|$ of -20 dB together with an isolation of 20 dB will yield worst case phase and amplitude errors of 0.06° and 0.009 dB respectively.

Measurement errors due to poor weighting factors are within the same limits if arg $\{\underline{k}_1\}$, $|\underline{k}_1|$ are known within an error limit of +/- 5° and 0.5 dB respectively. This is absolutely assured if they are provided by the exact solution of equ.(4).

IX. CONCLUSIONS

The homodyne network analyser system reported exhibits broadband measuring capabilities as well as a high sensivity. Phase errors werelimited to +/-2° for an insertion loss of up to 65 dB and should be improvable.

On account if its simple RF-structure the system may serve as a model for a millimeter wave network analyser to be developed.

APPENDIX

A short outline of the algebraic solution equation (4a) shall be given here.

Multiplying (1a) with $\underline{k}_{1\ell}^*$ and substracting it from (1b) yields

$$\underline{V}_2 - \underline{V}_1\,\underline{k}_{1\ell}^* = \underline{K}\,\underline{H}\,\underline{B}_u\,(\,\underline{k}_{1u} - \underline{k}_{1\ell}^*\,) \tag{A1}$$

Applying the same procedure to equs. (1c,e), (1d,f), (1g,h) one obtains

$$\underline{V}_5 - \underline{V}_3\,\underline{k}_{1\ell}^* = \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{2u}\,(\,\underline{k}_{1u} - \underline{k}_{1\ell}^*\,) \tag{A2}$$

$$\underline{V}_6 - \underline{V}_4\,\underline{k}_{1\ell}^* = \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{3u}\,(\,\underline{k}_{1u} - \underline{k}_{1\ell}^*\,) \tag{A3}$$

$$\underline{V}_8 - \underline{V}_7\,\underline{k}_{1\ell}^* = \underline{K}\,\underline{H}\,\underline{B}_u\,\underline{k}_{2u}\,\underline{k}_{3u}\,(\,\underline{k}_{1u} - \underline{k}_{1\ell}^*\,) \tag{A4}$$

Dividing (A2) as well as (A3) and (A4) by (A1) we gain three functional relations:

$$\underline{k}_{2u} = f_2\{\underline{k}_{1\ell}^*\};\ \underline{k}_{3u} = f_3\{\underline{k}_{1\ell}^*\};\ \underline{k}_{2u}\cdot\underline{k}_{3u} = f_{23}\{\underline{k}_{1\ell}^*\}$$

(f_2 and f_3 are given in (4b,c)).

Therefore one may solve a quadratic equation in $\underline{k}_{1\ell}^*$

$$f_2\{\underline{k}_{1\ell}^*\}\cdot f_3\{\underline{k}_{1\ell}^*\} = f_{23}\{\underline{k}_{1\ell}^*\} \tag{A5}$$

which results in the solution $\underline{k}_{1\ell}^*$ = u +/- jw (4a). Exactly the same solution is obtained for $\underline{k}_{1u}$ if equs.(1a,c,d,g) are multiplied by $\underline{k}_{1u}$ instead of $\underline{k}_{1\ell}^*$ and if the same procedure as outlined above is followed. It is seen that both unknowns $\underline{k}_{1u}$, $\underline{k}_{1\ell}^*$ only differ by the sign of the root in (4a).

For the special case of $\underline{k}_{1u}$ = $\underline{k}_{1\ell}$ it may be shown that u as well as w are real quantities and therefore $\underline{k}_1^*$ = u +/- jw and $\underline{k}_1$ = u -/+ jw as to be expected. The ambiguity of the sign still has to be resolved by some information about e.g. $\underline{k}_1$, for instance whether the phase change of PS1 is positive or negative.

With the knowledge of $\underline{k}_{1u\ell}$ the rest of the unknowns $\underline{K}\,\underline{H}\,\underline{B}_u$, $\underline{K}\,\underline{H}\,\underline{B}_\ell$ are obtained from (1a,b). and $\underline{k}_2$, $\underline{k}_3$ from (4b,c).

The same derivation holds if $V_1 \ldots V_8$ are real ([4], equs.2,7-10).

ACKNOWLEDGEMENT

The authors wish to acknowledge the Deutsche Forschungsgemeinschaft for financial support and H. Severin, R. Frese as well as A. Baumeister, J. Ney, H. Osterwinter and R. Pawellek for scientific and measurement support.

REFERENCES

[1] C. A. Hoer, "Performance of a Dual Six-Port Automatic Network Analyser", IEEE Trans. Microwave Theory Techn., vol. MTT-27, pp. 993-998, December 1979.

[2] U. Gärtner, B. Schiek, "Homodyne Network Analysis by Digital Phase Modulation and Baseband Sampling", Electronic Letters, vol.20, No.2, pp. 73-75, Jan. 1984.

[3] U. Gärtner, B. Schiek, "Homodyne Network Analysis by a Digital Triple Phase Modulation and Baseband Sampling Technique", Conference Digest CPEM 1984 Delft, Netherlands, pp. 134-136.

[4] U. Gärtner, B. Schiek, "An Automatic Homodyne Network Analyser with Phase Shift Keying and Baseband Sampling", Proc. 14th EuMC 1984, pp. 813-818.

[5] "Automating the HP 8410B Microwave Network Analyser", application note 221A, Hewlett Packard, June 1980.

[6] R. J. King, "Microwave Homodyne Systems", London, England: Peregrinus, 1978.