JOURNAL ON

COMMUNICATIONS

VOLUME XLII MAY 1991

MICROWAVE MEASUREMENT

Products-Services

News-Events

JOURNAL ON COMMUNICATIONS

A PUBLICATION OF THE SCIENTIFIC SOCIETY FOR TELECOMMUNICATIONS, HUNGARY

Editorial office
Gábor Áron u. 65.
Budapest, P.O.Box 15.
Hungary, H—1525

phone (361) 135-1097
(361) 115-2247
fax (361) 135-5560

Subscription rates
Hungarian subscribers
1 year, 12 issues 2900 HUF, single copies 360 HUF
Individual members of Sci. Soc. for Telecomm.
1 year, 12 issues 480 HUF,single copies 60 HUF
Foreign subscribers
12 issues 90 USD, 6 English issues 60 USD, single copies 15 USD
Transfer should be made to the Hungarian Credit Bank,
Budapest, H—1853, A/C No. MHB-323-10229

JOURNAL ON COMMUNICATIONS is published monthly, alternately in English and Hungarian by R. E. Phys. Soc., Fő u. 68. Budapest, H—1027, phone and fax (361) 201-8682. Publisher Klára Láng. Type-setting by DIAMANT Ltd. Printed by HUNGAPRINT, Budapest, Hungary

HUISSN 0866-5583

EDITORIAL

After decades of progress, microwave engineering still keeps growing, and microwave applications enter many fields of everyday life. Terrestrial and satellite communications became integral parts of telecommunication technology, and directly reach costumers in the form of satellite broadcasts or microwave distribution of TV programs. In addition to traditional applications in communications and radar engineering, a wide range of novel microwave equipment is appearing in the industry or even in a well-equipped kitchen.

An important and inseparable part of microwave technology is *microwave measurement engineering* which directly affects the quality of microwave products and related services. The continuous development of microwave measurement engineering is fuelled by three major factors:

- Advances in *microwave circuit technology* which resulted in an ample supply of microwave solid state devices and microwave integrated circuits, and made feasible the economical realization of sophisticated microwave subsystems. As a consequence, technical features of equipment (reliability, size etc.) have been considerably improved.
- New and severe *demands* are posed by *advanced technological processes* and *novel applications.* This challenge is stimulating unremitting efforts in test engineering developments. A recent example is the on-wafer measurement of monolithic microwave integrated circuits (MMICs), representing a key problem in advanced technology.
- Likewise important is the penetration of up-to-date *computation methods and devices* into measurement engineering, an especially significant step in microwave measurements. In addition to general control and data aquisition techniques, efficient error-correction methods and digital data processing procedures provide new possibilities for evaluation and interpretation of measurements. Good examples for the latter techniques are the virtual time domain capabilities via Digital Fourier Transform and the near-field antenna measurements.

The primary aim of microwave engineering is the evaluation of equipments and systems operating in the microwave frequency band. Originally, this comprised microwave measurements which were carried out in radar engineering and communications. Generally speaking, these measurements included two basic types: microwave signal measurements (power, frequency, etc.) and measurements of devices, circuits and subsystems (e.g. reflection coefficient, S-parameters). As developments in this field expanded and the procedures turned out to be fruitful and efficient, the results spreaded into various fields of science, industry and other applications. Several specific techniques evolved in the following fields:

Material measurements which give information on the structure and features of various materials. The range of measurements extends from a simple permittivity measurement to microwave spectroscopy, and these techniques are equally important in industry and in related branches of chemistry, physics, biology or medicine.

Measurements of basic physical parameters (mainly position and velocity), these measurements being related to the radar engineering, and find important applications in the industry.

Radiometric measurements which provide information on our environment or on the outer space, e.g. on microwave background radiation.

And finally *(microwave) remote sensing* which provides not only comprehensive information on our environment but became an important tool in the search for natural resources.

In the growing world of microwave measurement engineering there is a vivid activity motivated by different approaches and represented by different (somewhat overlapping) groups groups of professionals:

In the first group there are laboratories in which research are carried out on standards and high-precision measurements. This fundamental work constitutes a solid ground for the complex structure of microwave test engineering. This is shown by the fact that some important parameters of high-quality instruments are traceable to basic standards.

A second group of research and development laboratories (both in universities and in industry) addresses new measurements problems, and their development work results in advanced measurement methods and evaluation procedures. These solutions are often of direct practical importance, and on the other hand, they are of benefit for the measurement industry.

The third group is made up from manufacturers of microwave instrumentation, who are developing and producing test equipment and systems for microwave engineering. Their products and services represent the microwave measurements for most of the users.

It would have been impossible to present every aspect of microwave measurement engineering in a single issue of the Journal on Communications, so the guest editor had the responsibility to select topics which would be interesting and useful for the readers.

Based on an *URSI Review,* a survey is presented on selected topics of advanced microwave measurements, namely on network measurements and antenna measurements. Being in the forefront of microwave network measurements, network analyzer problems are discussed in two papers. One of them provides insight into network analyzer calibration and error-correction techniques while the other describes the design and results of a novel six-port network analyzer arrangement. A paper is devoted to remote sensing, providing not only an introduction but also reporting valuable research achievements.

The editor hopes that the ideas and results of these papers will be found worthwhile by the readers and will promote further progress and development.

I. KÁSA

István Kása

was born in Budapest. He graduated in electrical engineering from the Technical University of Budapest in 1960. He received the specialist degree in microwave engineering and the "dr. techn." degree also from this University. He received the Candidate of Technical Sciences degree from the Hungarian Academy of Sciences in 1974. Since 1960, he has been with the Research Institute for Telecommunications (TKI), Budapest. Dr. Kása was working on several major R & D projects in the fields of microwave measurements, passive and active microwave integrated circuits, microwave receivers and microwave communication.

Dr. Kása is the author of about 40 technical papers and two books. He is an associate professor at the Technical University of Budapest.

ADVANCED TECHNIQUES IN MICROWAVE MEASUREMENTS — A SURVEY

The paper provides a survey of recent results in advanced microwave measurements, specifically on (linear) network characterization and antenna measurements. The survey is based on an recent URSI review.

1. INTRODUCTION

As a consequence of the rapid and continuous progress in microwave measurement engineering, a number of technical papers have been published since 1987. This article is intended to provide a survey of recent results in linear network measurements and antenna measurements. The selected topics are in the forefront of microwave measurement technology, and are of importance and interest for a wide group of microwave engineers. The survey is based mainly on an URSI Review [*Ref A., 1990*].

2. NETWORK MEASUREMENTS

Reflection coefficients and S-parameters are basic parameters applied to describe and qualify microwave devices, components and subsystems. Presently used methods of reflection and S-parameter measurements rely basically on two types of measurement systems:

- up-to-date automatic vector network analyzers including down-converters or sampling heads and IF complex ratiometers,
- six-port type network analyzers utilizing microwave signal combiners and amplitude (power) detectors.

Both techniques widely utilize computerized control and evaluation procedures. In the following section both types will be discussed.

2.1. Vector reflectometers and network analyzers

A tuned reflectometer technique was established for the calibration of impedance measuring equipment and standard mismatches at K_u-band frequencies (12.4 to 18.0 GHz) using precision waveguide, matched termination, and quarter-wave short circuits [*Swarup et al., 1988*]. Three standard mismatches of nominal voltage standing-wave ratio (VSWR), 1.10, 1.20, and 1.30, were designed and developed for K_u-band frequencies. The standard mismatches have a variation of +0.01 in VSWR about the nominal value [*Yadava et al., 1988*]. *Powell and Miller* [*1987*] presented a technique to further reduce the reflection error and still enable wide frequency band automatic calibrations. Further, a technique was desribed which may be used to determine the reflection coefficient of power standards.

Scalar and vector homodyne detectors for use in homodyne network analyzers were investigated at the Ruhr University, Bochum. A millimeter-wave network analyzer system using this technique offers a dynamic range of 110 dB [*Gaertner et al., 1989*]. The total system error correction of a dual homodyne network analyzer controlled by a binary phase shifter, including determination of the phase shifter behavior, can be obtained by measuring only two arbitrary reciprocal networks [*Eul, 1989*].

A network analyzer based on the well-known unidirectional type was modified so that the transmission port, where the output of the device under test (DUT) is connected, can be set in either of two different unknown states. All four complex scattering coefficients can then be determined without reversing the DUT, thus providing a wider dynamic range in measuring S_{21}, but reduced sensitivity in measuring S_{12} and S_{22} if the transmission is low [*Eul and Schiek, 1988a*]. A general theory for network analyzer calibration was presented in which known calibration procedures such as the TSD and "through-reflectine" (TRL) methods are included as special cases. This theory also includes more general concepts like the "through-match-reflect" and "through-attenuation-network" methods which have no inherent limit in bandwidth and no lower frequency limit [*Eul and Schiek, 1988b*].

Integrated quadrature detectors in homodyne network analyzers covering frequencies from 90 to 100 GHz were developed. These detectors consist of two beam-lead diodes mounted one-eighth of a wavelength apart on two separate finline structures inside a waveguide [*Groll, 1989*]. A homodyne W-band (90 to 110 GHz) network analyzer was constructed employing an integrated quadrature detector. The error-corrected reflection coefficient is measured at 50 frequencies in 1 second [*Neumeyer, 1987, 1988*].

A computer-controlled scalar network analyzer employing quasi-optical measurements and harmonic generation for frequencies from 108 to 220 GHz was developed at lower cost than other systems [*Braun et al., 1988*].

Transistor parameters can be measured using a test fixture with a stripline substrate. Calibration standards for error correction were constructed from the same design [*Groll and Neumeyer, 1987a*]. The working principles, accuracy, frequency range, and measuring speed of different types of commercial millimeter-wave network analyzers were compared [*Groll and Neumeyer, 1987b*].

At Duisburg University, *Gronau and Wolff* [*1989*] presented a simple method for de-embedding the scattering parameters of microstrip broadband devices in the frequency range of 1 to 25 GHz. The method can be used with commercial ANAs that have a time domain option. Uncertainties of +0.05 percent in magnitude and +5° in phase can be achieved.

At the Technical University of Hanover, the TSD calibration method of Speciale, where no absorber is required, has been investigated. Its applicability for calibrations in stripline was rather poor; however, *Kindler* [*1987*] successfully developed a method using three different offset short-circuits.

When a setup is measured using a network analyzer, measurement errors arise from spurious transmissions and reflections. These systematic errors can be compensated for by using a correction circuit. *Ladvánszky* [*1987*] determined the conditions necessary for an error circuit to model the systematic errors of a measurement performed by a network analyzer.

A calibration method using an imperfectly matched stripline termination was developed for a commercial network analyzer at frequencies up to 18 GHz, and an error analysis was given [*Kindler, 1988*]. To achieve reliable measurements of the characteristic impedance of microstrips up to 18 GHz, the errors induced by reflections at the microstrip-to-coaxial transitions were minimized by a time domain method that employs a commercial network analyzer.

2.2. Six-port reflectometers and network analyzers

At the Technical University, Munich, a fast W-band six-port network analyzer operating at frequencies from 90 to 100 GHz, with beam-lead diodes integrated into finlines as power detectors, was constructed and experimentally proved [*Neumeyer, 1987*]. A new method of six-port calibration was developed in which five reflecting terminations with constant magnitude and unknown phase were used to determine the six-port-to-four-port reduction parameters [*Neumeyer, 1989*].

The submillimeter, quasi-optical, multiport reflectometer technique developed in the PTB has been simplified. Two four-port reflectometers, one using a quasi-optical technique and the other employing an oversized waveguide have recently been developed. They use only one field probe instead of three (Michelson interferometer-type) and only two thermistor power meters instead of four, as used with the submillimeter six-port reflectometer reported earlier. The new reflectometers permit determination of the complex reflection coefficient of phaseable reflective loads (e.g., samples of dielectric microwave materials) at about 390 GHz with an uncertainty of $+0.02$ percent for the modulus and $+2^{\circ}$ for the phase [*Stumper, 1988, 1989b*].

Six-port techniques were developed in several standards laboratories in China and used for measuring quantities such as reflection coefficient, permittivity, and permeability [*Hu and Chen, 1988; Xu and Chen, 1988; Lu, 1988*]. The frequency ranges for some of the reflectometers were extended to the millimeter-wave band.

Berman et al.[*1987*] compared a number of six-port junction reflectometer circuits proposed in the literature. The comparison was based on a geometrical criterion applied to data plotted in the complex plane. It is desirable to know how well a measured reflection coefficient is resolved from system noise, and how this resolution depends on the value of reflection coefficient and experimental design. *Hunter and Somlo* [*1988*] summarized the effects of departure of the waveguide propagation constant and characteristic impedance from their nominal values.

Five companion papers addressed various aspects of on-line accuracy assessment for the dual six-port ANA, ranging from background and theory to experimental results. *Engen* [*1987*] stated that a major challenge for the microwave metrologist is to provide an accuracy assessment for the ANA, and that a common problem is nonideal connector performance. It is impossible to characterize the six-port system accurately by a single number. Instead, the on-line accuracy assessment provides an estimate of the combined effects of these different error sources on each individual measurement result. *Hoer and Engen* [*1987*] described a technique for calibrating ANAs so that the scattering parameters of two-port devices having any combination of connectors can be measured, This technique is a generalization of the TRL calibration technique in which the "through" is replaced with a second length of precision transmission line that provides a "line-reflection-line" (LRL) calibration technique. *Judish and Engen* [*1987*] used statistical methods to evaluate the random errors in dual six-port measurements of reflection coefficient and scattering parameters. *Hoer* [*1987*] derived expressions for estimating the systematic errors in dual six-port or four-port measurements of reflection coefficient and scattering parameters which are caused by imperfections in the transmission line standard used to calibrate the system. *Juroshek* [*1987*] presented results showing uncertainty estimates obtained in quasi-real time.

The Microwave Research Laboratory (MRL) of Ecole Polytechnique in Canada has been active in computer-aided measurement of microwave devices and circuits. Several papers were published on the calibration of a six-port reflectometer [*Ghannouchi and Bosisio, 1987, 1988a, 1988b, 1988c*]. Based on reflection coefficient measurements, the six-port technique was also used to perform microwave permittivity measurements. A detailed analysis of the accuracy of complex reflection coefficient measurements using the six-port technique was published [*Ghannouchi and Bosisio, 1988d*]. Transformation of the Smith chart into a form convenient for presentation of an impedance measured by the six-port technique was developed by *Kaliouby and Bosisio* [*1987a*].

In telecommunications measurements, the six-port technique was used to measure the scattering parameters (small signal characterization) of microwave devices and circuits [*Kaliouby and Bosisio, 1987b*]. *Ghannouchi et al.* [*1989a*] published a new method for the large signal characterization of microwave transistors based on the ability of the six-port technique to perform simultaneous measurements of impedance and power flow. Large signal characterization is required for nonlinear transistor modeling, as well as for oscillator and high-power amplifier design. In order to perform pulse and swept-frequency microwave measurements, a computer-aided design (CAD) and computer-aided measurement procedure for a preferred frequency-balanced six-port junction were developed [*Ghannouchi et al., 1989b*]. A pulsed microwave measurement system was developed using a six-port reflectometer with time resolution of about 1 μs in both repetitive and single-shot modes of operation [*Demers et al., 1989*].

A new six-port reflectometer that meets theoretical criteria for optimum performance was developed at the Royal Signals and Radar Establishment (RSRE) in the U.K. It consists of a symmetrical five-port junction and a directional coupler. The implementation of this network for use at millimeter wavelengths was facilitated by the design of a broadband symmetrical, five-port, E-plane waveguide junction [*Cullen and Yeo, 1988*]. The same network was used in a dual six-port network analyzer in which phase-switching in the feed to the reflectometers was replaced by the use of a biphase bimodulation technique [*Judah and Wright, 1988, 1989*]. An analysis of the residual error arising from calibration of a reflectometer with a sliding termination showed that there is little need for a highly matched sliding element [*Griffin and Hodgetts, 1987*]. An application of quasi-optical techniques to enable both co-planar and crosspolar reflection to be measured with a multistate analog of a six-port reflectometer was reported [*Lederer et al, 1987*].

3. ANTENNA MEASUREMENTS

Gillespie [Ref. B, 1988] edited a special issue of *IEEE Transactions on Antennas and Propagation* on near-field scanning techniques which contains 21 papers on planar, cylindrical, and spherical near-field scanning. Gain, pattern, and polarization measurements were covered, and some error analyses were presented. The out-of-band response of antennas is important in EMI applications, and *Hill and Francis [1987]* presented above-band patterns of a phased-array antenna obtained from planar near-field scanning. They also derived a statistical theory to explain the observed gain reduction of antenna arrays at out-of-band frequencies. *Cown and Ryan [1989]* studied the potential application of near-field scanning techniques for making bistatic radar cross section (RCS) measurements. The modulated scatterer technique is one possible method for reducing the time required for probe movement in near-field scanning, and *Hajnal [1987]* developed a modulated, crossed-dipole scatterer for simultaneously obtaining two components of the electric field.

The generation of a field having a known strength is required for field probe calibration. *Kanda and Orr [1987]* measured the near-field gain of open-ended waveguides and pyramidal horns in an anechoic chamber and compared these results with theory. *Uno and Adachi [1989]* did experimental and theoretical studies of the range requirements for far-field pattern measurements, using both small probes and large antennas for reception. *Corona et al. [1989]* performed similar calculations of the range requirements for symmetrical and antisymmetrical aperture antennas.

Time-domain measurements are gaining popularity as a means for evaluating antennas. *Kunz et al. [1987]* described the Lawrence Livermore facility, which uses a monocone transmitting antenna with absorber. The facility can also be used in frequency-domain measurements for frequencies from 100 MHz to 18 GHz. *Morgan and McDaniel [1988]* described the transient scattering range at the Naval Postgraduate School. This facility operates in a free-space environment without a ground plane.

Mishra and Kashyap [1987] described a multipurpose facility at the NRC for EMC measurements and for calibrating antennas and probes. *FitzGerrell [1988]* measured the input impedance and gain of vertical monopoles and found that a small ground plane with resistively loaded radial wires gave approximately the same results as the theory for an infinite ground plane. *Johannsen [1988]* described a ground station for satellite antenna-pattern and crosspolarization measurements.

Spar Aerospace Ltd. and McGill University developed a test facility for spacecraft antennas *[Whitehouse et al., 1986]*. A field scanning system was designed to obtain three-dimensional patterns of the amplitude and phase near radiators and scatterers in the X- and K-bands *[Illott et al., 1986]*. Measurement of fields in the vicinity of large reflector antennas was shown to be useful in predicting the far-field pattern *[Wegrowicz et al., 1989]*.

A new spherical near-field test facility is under development at the David Florida Laboratory in Ottawa *[Wood, 1987, 1988a]*. Many user-friendly, user-interactive, and graphics features can be used for antenna measurements to determine far-, near-, and very near-fields. A probe-correction algorithm was developed in conjunction with this system *[Wood, 1988b]*.

Antenna far-field predictions usually require measurement of both amplitude and phase in the near-field of the antenna. Above 60 GHz, the phase measurement may be less accurate than required. A phase retrieval algorithm has been developed to recover aperture phase and amplitude from a single-intensity planar scan *[Anderson et al., 1989]*. This permits accurate far-field predictions and also provides information to enable alignment of the reflector antenna.

A 3-m-wide compact antenna test range for millimeter-wave antenna measurements up to 200 GHz has been built *[Parini et al., 1989]*. In a controlled environment, accurate measurements may be made on antennas that are large in terms of wavelengths. The measurement time in less than that of conventional near-field/far-field ranges, and the path loss of only 10 dB enables a wide dynamic range to be maintained with the low-power millimeter-wave sources. A compact test range for use over an octave bandwidth around 1 GHz employs an array of seven Yagi antennas *[Wong and Excell, 1989]*. A plane wave could be synthesized in the near-field region in front of the array. Over a volume of 0.6 m^3, the maximum deviation from a plane wave was 2.5 dB.

The use of a near-field antenna array to generate a uniform plane-wave field is mathematically equivalent to near-field scanning. *Hill [1988]* synthesized a plane wave with a near-field circular array of straight-line sources. *Bennett and Farhat [1987]* studied the relationship of errors in pattern measurement to planar wavefront quality.

Advances have been made in the simulation of EM fields in practical environments, where strong coupling exists between antennas and the wiring used for measurements. These simulation techniques were based on transmission-line modeling and were used to develop better calibration procedures *[Christopoulos et al., 1988]*. *Braun and Neumeyer [1989]* expanded the dynamic range of a receiver/transmitter system for measurement of antenna radiation patterns at 220 GHz to 40 dB. A personal computer was used to control the system.

Hollmann [1989] developed a measurement technique for use in anechoic chambers to determine the far-field gain of two identical pyramidal horn antennas with an uncertainty of +0.1 dB. This technique includes the localization of the amplitude center of a horn with an observation point in the far field. *Schone et al. [1988]* and *Heidrich and Wiesbeck [1989]* investigated polarimetric radar cross-section antenna measurements. The antenna parameters of input impedance, polarimetric gain, polarization decoupling, and scattering were deduced from the amplitudes of the scattered field.

Measurements of gain and polarization on a contoured beam antenna at five independent near-field test ranges (spherical, cylindrical, and compact) were compared by *Lemanczyk and Larsen [1988]*. Accurate polarization measurements using the spherical near-field technique were reported and used by *Larsen and Salomon [1988]* to determine cross-polar discrimination in a dual-polarized satellite communications system. The theoretical and practical aspects of spherical near-field antenna measurements, including accurate calibration of standard-gain horn antennas, were presented in a comprehensive monograph by *Hansen [ed. 1988].*

An indoor random-field method to measure the radiation efficiency of a small antenna was proposed *[Maeda and Ueyama, 1988].* Range distance requirements for aperture antenna gain measurement were discussed for the case where the transmitting and receiving antennas have comparable apertures *[Uno and Adachi, 1987].* The theory of cylindrical near-field techniques for measuring antenna radiation patterns was presented, and probe misalignment was discussed *[Iisugusa, 1989].*

The absolute gain of a standard horn in an anechoic chamber was measured at 4.8 GHz for a nominal gain of 20 dB, with a total uncertainty less than 0.24 dB *[Qian 1989]. Hunter and Campbell [1987]* described the calibration service of the Australian National Measurement Laboratory for standard-gain horns at frequencies from 2.4 to 18 GHz. The three-antenna method was used in an anechoic chamber with computer control of antenna separation, and automatic data acquisition.

REFERENCES

Ref. A [1990] "Review of Radio Science." (Editor E. Hyde) Ch. 1. "Electromagnetic Metrology" (Ed. M. Kanda) *URSI Publishing*, Brussels, 1990, A1-A39.

Ref. B [1988], "Special issue on near-field scanning techniques" (E. S. Gillespie, ed.), *IEEE Trans. Antennas Propag., AP-36.*

Anderson, A. P., G. Junkin, and J. E. McCormack [1989], "Near-field far-field predictions from single-intensity-planar-scan phase retrieval", *Electron, Lett.*, 25, 519.

Bennett, J. C., and K. S. Farhat [1987], "Wavefront quality in antenna pattern measurement: the use of residuals", *Proc. IEEE, 134 (H).*

Berman, M., P.: I. Somlo, and M. J. Buckley [1987], "A comparative statistical study of some proposed six-port junction designs", *IEEE Trans. Miscrowave Theory Techn., MTT-35,* 971.

Braun, G., J. Detlefsen, M. Lange, and B. Neumeyer [1988], "A computer controlled 108—220 GHz scalar network analyzer suited for quasioptical measurements", 18th European Microwave Conf., EuMC'88, Stockholm, *Proc.*, 930.

Braun, G., and B.Neumeyer [1989], "Measurements of antenna radiation patterns with a computer controlled 108—220 GHz transmitter/receiver system", IEEE AP-S Intl. Symp. & URSI Radio Science Meeting, San Jose, *Proc.*

Christopoulos, C., A. Mallik, P. Naylor, and P. B. Johns [1988], "Computer simulation in EMC design and testing of vehicles", 18th Intl. Symp. on Automotive Technology and Automation, Florence, Paper 88125.

Corona, P., G. Ferrara, and C. Gennarelli [1989], "Measurement distance requirements for both symmetrical and antisymmetrical aperture antennas", *IEEE Trans. Antennas Propag., AP-37,* 990.

Cown, B. J., and C. E. Ryan [1989], "Near-field scattering measurements for determining complex target RCS", *IEEE Trans. Antennas Propag., AP-37.*

Cullen, A. L., and S. P. Yeo [1988], "Designing a symmetrical five-port waveguide junction", *Proc. IEE, 135 (H),* 17.

Demers, Y., R. G. Bosisio, and F. M. Ghannouchi [1989], Wideband microwave pulsed reflectometer using a six-port junction, IEEE Instrum. & Meas. Tech. Conf., Washington, D. C., *Conf. Rec.*, 529.

Engen, G. F. [1987], "On-line accuracy assessment for the dual six-port ANA: background and theory", *IEEE Trans. Instrum. Meas., IM-36, 507.*

Eul, H. J. [1989], "Establishing the complex measurement ability of a homodyne network analyzer via self-calibration", IEEE MTT-S Intl. Microwave Symp., Long Beach, *Symp. Dig.*, 1187.

Eul, H. J. and B. Schiek [1988a], "Error-corrected two-state unidirectional network analyzer", *Electron. Lett., 24,* 1197.

Eul, H. J., and B. Schiek [1988b], "Thru-match-reflect: one result of a rigorous theory for de-embedding and network analyzer calibration", 18th European Microwave Conf., EuMC'88, Stockholm, *Proc.*, 909.

FitzGerrel, R. G. [1988], "Monopole impedance and gain measurements on finite ground planes", *IEEE Trans. Antennas Propag., AP-36,* 431.

Gaertner, U. F., J. Schneider, and B. Schiek [1989], "Homodyne network analysis — improving the accuracy and sensitivity by redundant phase coding and double modulation", *Archiv Elektr. Uebertragung, AEU, 43,* 16.

Ghannouchi, F. M., and R. G. Bosisio [1987], "A new six-port calibration method using four standards and avoiding singularities", *IEEE Trans. Instrum. Meas., IM-36,* 1022.

Ghannouchi, F. M., and R. G. Bosisio [1988a], "An alternative explicit six-port matrix calibration formalism using five standards", *IEEE Trans. Microwave Theory Tech., MTT-36,* 494.

Ghannouchi, F. M., and R. G. Bosisio [1988b], "An improved sliding load formulation for use in six-port calibration methods", *Alta Freq., 57,* 497.

Ghannouchi, F. M., and R. G. Bosisio [1988c], "The six-port reflectometer and its complete calibration by four standard terminations", *Proc. IEE, 135 (H),* 285.

Ghannouchi, F. M., and R. G. Bosisio [1988d], "A comparative worst-case error analysis of some proposed six-port designs", *IEEE Trans. Instrum. Meas., IM-37, 552.*

Ghannouchi, F. M., R. G. Bosisio, and Y. Demers [1989a], "Load-pull characterization method using six-port techniques", IEEE Instrum. & Meas. Tech. Conf., Washington, D. C., *Conf. Rec.*, 536.

Ghannouchi, F. M., S. Wang, and R. G. Bosisio [1989b], "A preferred balanced six-port junction in microstrip line", SMBO Int. Microwave Symp., San Paulo, *Proc., 1,* 2850.

Griffin, E. J., and T. E. Hodgetts [1987], "Estimating the residual error from calibrating a reflectometer with a sliding termination", *Proc. IEE, 134 (H),* 560.

Groll, H. P. [1989], "A new two-port network analyzer with a homodyne system and integrated quadrature detectors for the frequency range of 90 to 100 GHz", *ISRAMT'89 Digest,* Beijing.

Groll, H. P., and B. Neumeyer [1987a], "Measurement of microwave transistor-S-parameters in its final environment", 5th Int. School on Microwave Physics and Tech., Varna, *Proc.*, 224.

Groll, H. P., and B. Neumeyer [1987b], "Comparison and results of various S-parameter measurement systems in the mm-wave range", 5th Int. School on Microwave Physics and Tech., Varna, *Proc.*, 155.

Gronau, G., and I. Wolff [1989], "A simple broad-band device deembedding method using an automatic network analyzer with time-domain option", *IEEE Trans. Microwave Theory Tech., MTT-37,* 479.

Hajnal, J. V. [1987], "Compound modulated scatterer measuring system", *Proc., IEE, 134 (H).*

Hansen, J. E. (ed.) [1988], "Spherical near-field antenna measurements", London, Peter Peregrinus, Ltd., 387 pp.

Heidrich, E., and J. Wiesbeck [1989], "Features of advanced polarimetric RCS antenna measurements", IEEE AP-S Int. Symp. & URSI Radio Science Meeting, San Jose, USA, *Proc.*

Hill, D. A. [1988], "A circular array for plane-wave synthesis", *IEEE Trans. Electromagn. Compat., EMC-30,* 3.

Hill, D. A., and M. H. Francis [1987], "Out-of-band response of antenna arrays", *IEEE Trans. Electromagn. Compat., EMC-29,* 282.

Hoer, C. A. [1987], "On-line accuracy assessment for the dual six-port ANA: treatment of systematic errors", *IEEE Trans. Instrum. Meas., IM-36,* 154.

Hoer, C. A., and G. F. Engen [1987], "On-line accuracy assessment for the dual six-port ANA: extension to nonmating connectors", *IEEE Trans. Instrum. Meas., IM-36.*

Hollman, H. [1989], "Accurate gain measurements of horn antennas in the shortened far field", *IEEE Trans. Instrum. Meas., IM-38,* 617.

Hu, X. P., and C. R. Chen [1988] "An 8 mm dual six-port standard set-up for S-parameters", *The Annual Report for 1988,* National Inst. of Metrol., Beijing.

Hunter, J. D., and Somlo, P. I. [1988], "Rectangular waveguide characterization for reflectometry", *IEEE Trans. Instrum. Meas., IM-37,* 439.

Hunter, J. D., and P. W. Campbell [1987], "Microwave standard-horn calibration system at the National Measurement Laboratory", *IEEE J. Electr. Electron. Eng., 7,* 261.

Iisugusa, K. [1989], "Estimation of errors due to imperfect alignment of a probe line in cylindrical near field antenna measurement", 6th Int. Conf. on Antennas and Propagation, ICAP 89, *Proc.*, 1, 359.

Illott, P., P. Markland, and T. Parlasek [1986], "An intermediate and near field volume scanning X and K band facility", Symp. Antenna Meas. Tech. Assoc., *Proc.*, 69.

Johannsen, K. G. [1988], "Measurement of satellite communication parameters: satellite antenna measurements", *IEEE Trans. Instrum. Meas. IM-37.*

Judah, S. K., and A. S. Wright [1989], "A biphase bimodulation dual sixport network analyzer", *IEEE/SEMO Int. Microwave Symp. Dig.,* 305.

Judish, R. M., and G. F. Engen [1987], "On-line accuracy assessment for the dual six-port ANA: statistical methods for random errors", *IEEE Trans. Instrum. Meas., IM-36.*

Juroshek, J. R. [1987], "On-line accuracy assessment for the dual six-port ANA: experimental results", *IEE Trans. Instrum. Meas., IM-36.*

Kaliouby, L., and R. G. Bosisio [1987a], "A new transformed Smith chart for real-time impedance measurement using the six-port concept", *IEEE Trans. Instrum. Meas., IM-36,* 493.

Kaliouby, L., and R. G. Bosisio [1987b], "A new two-port scattering matrix measurement technique using a sliding load", *IEEE Trans. Microwave Theory Tech., MTT-36,* 1028.

Kanda, M., and R. D. Orr [1987], "Near-field gain of a horn and an openended waveguide: comparison between theory and experiment", *IEEE Trans. Antennas Propag., AP-35, 33.*

Kindler, K. [1987], "Errors in stripline calibration of automatic network analyzers", *Frequenz, 41,* 197.

Kindler, K. [1988], "Stripline calibration of automatic network analyzers with TSD method", *Arch. Electron. Ubertragtechnik., 42,* 262.

Kuncz, K. S., H. G. Hudson, J. D. Breakall, R. J. King, S. T. Pennock, and A. P. Ludwigsen [1987], "Lawrence Livermore National Laboratory Electromagnetic Measurement Facility", *IEEE Trans. Electromag. Compat., EMC-29,* 93.

Ladvanszky, J. [1987], "Exact measurement of the reflection matrix of a microwave circuit by means of a network analyzer", *Budavox Rev., 2,* 1.

Larsen, F. H., and R. Salomon [1988], "Determination of cross-polar discrimination (XPD) of a contoured beam reflector antenna by spherical near-field measurements", 11th ESTEC Antenna workshop on Antenna Meas., ESA WPP-001, Gothenburg, *Proc.,* 321.

Lederer, P. G., M. A. Wood, J. P. Ide, T. E. Hodgetts, and L. C. Oldfield [1987], "A quasi-optical multi-state reflectometer", *British Electromag. Conf. Dig.,* 19/1.

Lemanczyk, J., and F. Holm Larsen [1988], "Comparison of near-field range results", *IEEE Trans. Antennas Propag., AP-36,* 845.

Maeda, A., and T. Ueyama [1988], "Note on EMI measurement at open field test site. V. optimum antenna height for EMI sweep search measurement over wide frequency band", 1988 *IEEE Int. Symp. on EMC,* 284.

Mishra, S. R., and S. Kashyap [1987], "A multipurpose EM field measurement and antenna calibration facility", *IEEE Trans. Instrum. Meas., IM-36,* 1010.

Morgan, M. A., and B. W. McDaniel [1988], "Transient electromagnetic scattering: data acquisition and signal processing", *IEEE Trans. Instrum. Meas., IM-37,* 265.

Neumeyer, B. [1989], "Impedance measurement in the mm-wave range without frequency conversion", Thesis, Technical University Munich.

Neumeyer, B. [1987], "A fast W-band six-port network analyzer", 17th European Microwave Conf., EuMC'87, Rome, *Proc.,* 349.

Neumeyer, B. [1988], "A homodyne W-band network analyzer with integrated quadrature detector", 18th European Microwave Conf., EuMC'88, Stockholm, *Proc.,* 333.

Parini, C. G., A. D. Olver, P. McNair, and C. J. Prior [1989], "The design, construction and use of a millimeter-wave compact antenna test range", ICAP 1989, *IEE Conf. Pub. 301,* 345.

Powell, R. C., and A. C. Miller [1987], "Determination of the reflection correction when using a symmetrical two-resistor power splitter to calibrate a power sensor", *IEEE Trans. Instrum. Meas., IM-36.*

Qian, P. Z. [1989], "Accurate measurement of the absolute gain of standard-gain horn in an anechoic chamber", Int. Symp. on Electromagn. Metrol., ISEM'89, Beijing, *Proc.*

Stumper, U. [1988], "Measurement of the complex permittivity of microwave materials with a submillimeter waveguide four-port reflectometer", Conf. on Prec. Electromagn. Meas., Tsukuba, *CPEM'88 Dig.,* 141.

Stumper, U. (1989b], "Six-port and four-port reflectometers for complex permittivity measurements at submillimeter wavelengths", *IEEE Trans. Microwave Theory Tech., MTT-37,* 220.

Swarup, Ram., P. S. Negi, R. S. Yadava, and K. Chandra [1988], "Calibration facilities for microwave attenuators at National Physical Laboratory", New Delhi, India, Asia Pacific Microwave Conf., *Proc.,* 263.

Schone, G., S. Rieger, and E. Heidrich [1988], "Wideband polarimetric radar cross section measurement", IEEE Ap-S Int. Symp. & URSI Radio Science Meeting, Syracuse, *Proc.,* 2, 537.

Uno, T., and S. Adachi [1987], "Range distance requirements for antenna measurements", 5th Int. Conf. on Antennas and Propagation, ICAP'87, *Proc.,* 355.

Uno, T., and S. Adachi [1989], "Range distance requirements for large antenna measurements", *IEEE Trans. Antennas Propag., AP-37,* 707.

Wegrowicz, L., R. A. Pokuls, P. A. Ilott, P. J. Markland, and T. J. F. Pavlasek [1989], "Far field prediction of reflection antennas by partial measurements", URSI Radio Sci. Meeting, San Jose, USA, *Proc.,* 46.

Wong, K. T., and P. S. Excell [1989], "A compact range fed by an array of log-periodic dipole antennas", IEE Int. Conf. on Antennas and Propag., ICAP 1989, IEE Conf. Pub. 301, 351.

Wood, P. J. [1987], "A near-field test system for very large antennas", IEE Int. Conf. on Antennas and Propag., IEE Conf. Pub. 371, *Proc.*

Wood, P. J. [1988a], "Concepts of the new spherical near field measurement system at the David Florida Laboratory", 10th Annual Ant Meas. Tech. Assoc., *Proc.,* 14/11.

Wood, P. J. [1988b], "A probe correction algorithm for spherical antenna measurements", 10th Annual Antenna Meas. Tech. Assoc., *Proc.,* 3/9.

Xu, B. C., and Z. W. Chen [1988], "Development of a Ka-band six-port reflectometer", NSDMM'88, *Proc.,* 258.

Yadava, R. S., P. S. Negi, R. Swarup, and K. Chandra [1988], "Precision measurement of microwave impedance at National Physical Laboratory", New Delhi, India, Asia Pacific Microwave Conference, *Proc.,* 253.

MINIATURE BROADBAND LUMPED SIX-PORT REFLECTOMETER

V. BILIK, V. RAFFAJ AND J. BEZEK

SLOVAK UNIVERSITY OF TECHNOLOGY, FACULTY OF ELECTRICAL ENGINEERING
ILKOVICOVA 3, 812 19 BRATISLAVA, CZECHOSLOVAKIA

A six-port reflectometer structure is described which may consist solely of lumped elements. The structure is dissipative but very broadband, capable of covering bandwidths up to 1000 : 1. The paper outlines the device operation principle, gives the design guidelines, and presents experiments with a GaAs monolithically integrated version operating in the range 75 MHz to 8 GHz.

1. INTRODUCTION

The six-port reflectometer (SPR) [1] is a device to measure vector reflection coefficient (complex impedance) in terms of four scalar quantities — powers of four waves obtained by properly combining samples of the wave incident on, and reflected from a measured load. Unlike conventional heterodyne network analyzers (HENA), it is not necessary in SPR technique to measure phase of microwave signals. This considerably simplifies the microwave hardware and associated circuitry as well as reduces cost. Moreover, the technique can straightforwardly be applied to very high frequencies (mm to infrared wavelengths). A pair of six-port reflectometers with an additional power divider and a phase shifter compose the dual six-port network analyzer (SPNA), which can be used to the measurement of both reflection and transmission coefficients [2]. The measurement accuracy is comparable to that of HENA and depends primarily on the used power detectors. The highest accuracy is achieved with bolometric sensors (metrological applications); the price paid is the speed of measurement. Fast six-ports employ diode detectors. The main drawback of SPNA as compared with HENA originates from the broadband nature of the detection, resulting, like in scalar network analyzers, in lower dynamic range and increased sensitivity to spurious signals (e.g. source harmonics).

Today, six-ports have found their main application in metrology of microwave power, attenuation, and impedance. Their use in research laboratories and industry is limited. Some of the possible reasons are the narrowband operation of the existing six-ports, their excessive dimensions at lower frequencies (under 1 GHz), and lack of the supporting software.

The SPR described in this paper is both very broadband and capable of operation down to low frequencies. Because of being built from lumped elements, its size can be small regardless of frequency.

Section 2 of the present paper gives a brief account of six-port techniques. Section 3 presents the basic ideas behind the lumped SPR, explains its operation and outlines the design procedure. Section 4 is devoted to the developed monolithically integrated GaAs SPR implementation; gives its description and expected parameters obtained by computer modeling. In Section 5, the capability of the device is illustrated by presenting experimental results.

2. SIX-PORT REFLECTOMETER

The basic building block of the six-port reflectometer is a linear six-port (Fig. 1), one port of which is terminated in the device under test (DUT), characterized by reflection coefficient ρ, one port is connected to signal generator,

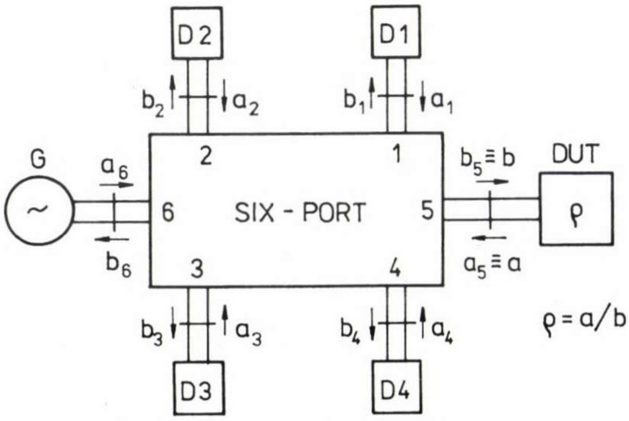


Fig. 1. Block diagram of six-port reflectometer

and the remaining four ports are connected to detectors D1 to D4. Waves b_1 to b_4 entering the detectors can be expressed as combinations of the waves $b=b_5$, incident on the DUT, and $a=a_5$, reflected from the DUT:

$$b_i=A_ia+B_ib=bA_i\,(\rho-q_i) \qquad i=1..4 \quad (1)$$

A_i, B_i are complex quantities characterizing SPR (they depend only on the S-parameters of the six-port and the reflection coefficients of the detectors),

$$\rho=a/b \quad (2)$$

is the reflection coefficient of the DUT, and

$$q_i=-B_i/A_i \quad (3)$$

are termed q-points of the reflectometer. Outputs of the detectors are supposed to be proportional to the powers carried by the waves incident on the detectors

$$P_i=|b_i|^2=|A_ia+B_ib|^2=P_o|A_i|^2|\rho-q_i|^2 \quad (4)$$

where $P_o=|b|^2$ is the power incident on DUT. One of the detectors, called *reference detector* (here denoted as No. 4), should ideally respond only to wave b, incident on the DUT. For the reference detector, it is more appropriate to write (4) in the alternative form

$$P_4=P_o|B_4|^2|\mathrm{d}\rho+1|^2 \quad (5)$$

where $\mathrm{d}=A_4/B_4=-1/q_4$. For an ideal reference port, $\mathrm{d}=0$.

Introducing *normalized powers* $p_i=P_i/P_4$, one can write

$$p_i = P_i/P_4 = C_i\left|\frac{\rho-q_i}{\mathrm{d}\rho+1}\right|^2 \qquad i=1..3 \quad (6)$$

where $C_i = |A_i/B_4|^2$. Three equations (6) are the working equations of the six-port reflectometer. They relate reflection coefficient ρ of the DUT to the measured quantities—powers. SPR itself is fully characterized by 11 frequency-dependent parameters: four complex quantities (q_i, d) and three scalars C_i. These or other related sets of 11 parameteters are called *calibration constants.* They are obtained in the process of six-port *calibration* in which a set of known loads (calibration standards) are connected in place of the DUT. Once the calibration constants are known, the measurement consists in simultaneous solution of three equations (6) for an unknown ρ. Graphically, each of the equations represents a circle in the ρ-plane; the solution is given by their common intersection point. The centers and radii of the circles are functions of six-port parameters and normalized powers p_i. The centers are approximately (if $d=0$ then exactly) coinciding with q-points q_i. This best visualizes both the meaning and the importance of the q-points.

The general guidelines for an optimum SPR design, i.e. the one which assures the most precise measurement of ρ at a given power measurement error, have been derived by Engen [1]. He has shown that the three q-points are the essential SPR parameters. Their magnitudes should be about 1.5 and they should be equally distributed around the unit circle in the complex ρ-plane (optimum phase differences of 120°). Thanks to calibration, however, the six-port is capable of accurate measurement even for large deviations from these criteria. For instance, it is sufficient that the phase differences of two q-points do not drop below approximately 45°.

The measurement accuracy of SPR is affected mainly by detector transfer curve nonlinearity, temperature dependence of detector sensitivities, generator frequency inaccuracy (unimportant with synthesized sources), inaccurately defined calibration standards, noise, interference, spurious r.f. signals. The SPR measurement error can be represented by an uncertainty circle with radius $\Delta\rho$ around the true value in the ρ-plane. Excluding the high-precision six-ports for metrological use, the typical value of $\Delta\rho$ is 0.01.

3. LUMPED-ELEMENT SIX-PORT REFLECTOMETER

In the design of most of the known six-port reflectometers, lossless distributed circuits such as directional couplers, hybrids, power dividers, symmetrical 5-ports, etc. have been used. The main drawback of such designs is the relatively narrow operation band—typically one octave to one decade.

Using lumped-element circuit design, new types of six-port structures became possible. Their main advantage is a very wide range of operation of up to 1000 : 1.

In the six-port reflectometers with diode detectors, relatively low input power levels are sufficient (less than −10 dBm). Therefore it is not always necessary to require the six-port to be loss-free, especially when an important property like bandwith can be traded for. A typical example of this type of philosophy is the reflectometric directional bridge which provides an extreme bandwidth at the price of absorbing about 75 percent of the input power. It turned out that this approach is fruitful also in the SPR technique.

The proposed six-port reflectometer can conveniently be split into a combiner and a power divider. The combiner provides three linear combinations of the waves incident on and reflected from the measured load; these determine the q-points of the SPR. The power divider, connected between the combiner and the DUT, provides a sample of the incident wave for the reference detector.

3.1. Basic combination circuit

The operation of the combination circuit (Fig. 2) resembles that of the reflectometer bridge. Two upper arms of the bridge are matched ($Z = Z_o$, Z_o being the reference impedance, e.g. 50 Ω). The third arm consists of a circuit with reflection coefficient Γ (henceforth called *reflector*).

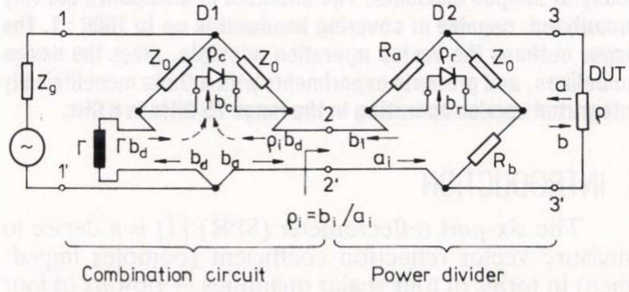


Fig. 2. Basic lumped-element six-port reflectometer

The fourth arm is the input of the power divider which, if properly designed, has the input reflection coefficient (ρ_i) proportional to that of DUT, i.e.

$$\rho_i = t^2 \rho \tag{7}$$

where $t = b/b_i$ is input-to-test-port transfer coefficient of the power divider. Due to symmetry of the bridge, wave $b_d = a_i$ entering the power divider is equal to the wave incident on reflector Γ. Wave b_c incident on the detector D1 (henceforth called *combination detector*) is the *difference* of the waves returned from the power divider and the reflector (because the two waves enter the detector "from opposite sides"). Therefore

$$b_c = A b_b \rho_i - A b_d \Gamma = A b_d t^2 (\rho - \Gamma / t^2) \tag{8}$$

where A is a proportionality constant. This intuitive reasoning has been confirmed by circuit analysis which yields the proportionality constant

$$A = [\Gamma(\rho_c - 1) - 2]^{-1} \tag{9}$$

with ρ_c being reflection coefficient of the combination detector. The comparison of Eqs. (8) and (1) gives immediately the q-point as

$$q = \Gamma / t^2 \tag{10}$$

Hence the value of the q-point is governed both by reflection coefficient Γ of the reflector and transfer coefficient t of the power divider. The preferable choice for the reflector is a reactance of $|\Gamma| = 1$.

One obvious method of making the three necessary wave combinations is a consecutive switching of three different reflectors Γ_1, Γ_2, Γ_3. Such a "switched" SPR has been described in [3]. SPR of this type, designed for the range 1 MHz to 1 GHz, using PIN diodes as the switches, has been constructed from common components and successfully tested, operating satisfactorily up to 700 MHz [4]. However, complications introduced by the three-throw switching circuit (complexity, limitation of usable frequency range, introduction of transients) justify a different combination circuit (Section 3.3.).

3.2. Power divider

As power divider (right-hand side of Fig. 2), again the bridge circuit can be employed, making use of the property of isolation between the diagonals of the balanced bridge $R_a R_b = Z_0^2$. Since the wave reflected from the DUT feeds a diagonal, it does not reach the reference detector D2 which then responds only to the incident wave b. Assuming $R_a R_b = Z_0^2$ the power divider is governed by the equations (using the notation of Fig. 2)

$$b = a_i t \quad (11a)$$
$$b_r = a_i(1-t) = b\,(1-t)/t \quad (11b)$$
$$b_i = a_i(1-t)^2\,\rho_r + at \quad (11c)$$

where ρ_r is the reflection coefficient of the refence detector D2. The transfer coefficient is

$$t = Z_o/(R_a + Z_o) = R_b/(R_b + Z_o) \quad (12)$$

Eq. (11b) proves that the reference detector, regardless of its mismatch, responds only to the incident wave b. The input reflection coefficient is

$$\rho_i = b_i/a_i = t^2\rho + (1-t)^2\rho_r \quad (13)$$

Substituting this in to (8), we have

$$b_c = Ab_d t^2 \left[\rho - \Gamma/t^2 + \frac{(1-t)^2}{t^2}\,\rho_r \right] \quad (14)$$

giving

$$q = \Gamma/t^2 - \frac{(1-t)^2}{t^2}\,\rho_r \quad (15)$$

Strictly, *Eq.* (15) becomes identical with *Eq.* (10) only if $\rho_r = 0$. Practically, however, the effect of reference detector mismatch is not significant (see below).

In our six-ports, reactance reflectors ($|\Gamma| = 1$) have been used and q-point magnitudes $|q_i| = 2$ have been considered the most appropriate. For this case, using (10), $t = 0.707$. Using then (12) with $Z_o = 50\ \Omega$, the values $R_a = 20\ \Omega$, $R_b = 125\ \Omega$ have been decided on, yielding actually $t = 0.714$ and $|q_i| = 1.96$. If $|\rho_r| < 0.5$, the magnitudes of q-points are affected by the second term in (15) by an amount not exceeding 5%.

3.3. Three-detector combination circuit

In the effort to avoid switching elements, a different type of combination circuit has been discovered. Essentially, it is a three-fold bridge with tripled left-hand pair of arms (each containing one reflector), and three associated combination detectors (Fig. 3). To avoid cross-coupling between reflectors and non-associated detectors, three compensation impedances Z_c are introduced, ideally equal to the impedance of the detectors (if not, the values of q-points would be affected). Such a combination circuit is actually a highly symmetrical structure as shown by its redrawing in Fig. 4. This is the heart of the proposed SPR.

The above qualitative considerations have been supported by analytically analysing the simplest circuit variant in which all impedanances, including those of detectors, are equal to Z_o [5]. While having a theoretical value, this case does not represent the best practical solution. The main reasons are the poor input match (SWR about 3) and the strong variatons with frequency (15 dB) of powers to the combination detectors. For the treatment of more general cases, a computer modeling program has been developed which can analyse the complete reflectometer including parasitic elements. Modeling has shown that the optimum choice for the combination detector impedance is $Z_c = R_c = 6 \times Z_o = 300\ \Omega$; the optimum bridge arms resistance (R in Fig. 3) is $R = 2.8 \times Z_o = 140\ \Omega$. This choice assures the input SWR less than 1.3 and the detector power variations less than 3 dB.

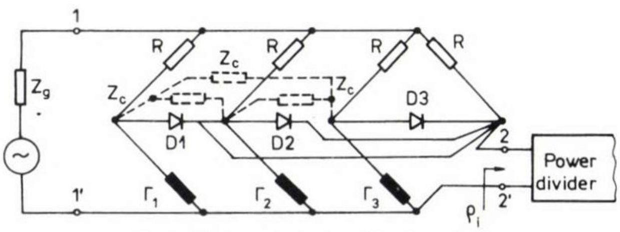


Fig. 3. Three detector combination circuit

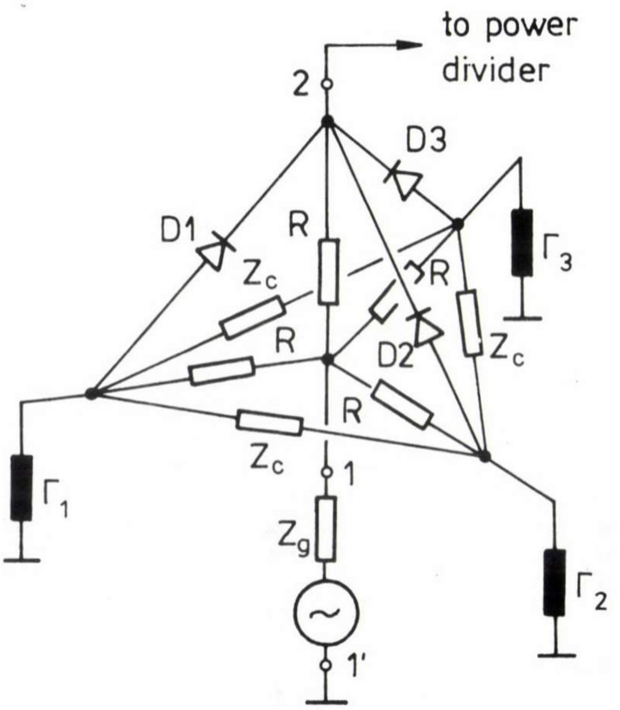


Fig. 4. Three detector combination circuit redrawn

3.4. Reflectors

Given that the six-port network be free of any parasitics, the ultimate bandwidth limitation on its performance is imposed by the ability of the reflectors to maintain three distinct reflection coefficients. The reflectors must be designed in such a way that phase differences of Γ_i do not drop below a specified minimum value (practically about 45°) over the whole band of SPR operation. The broadest possible band is obtained with lumped-element reflectors. The extra merit is the possibility of using such SPR down to very low (even acoustic) frequencies.

The case treated here will be that of $C-L-C$ reflectors (two capacitors, C_1 and C_2, and an inductor L). The idea is that for a single lumped element, Γ can rotate only by 180° over all frequencies. The situation is illustrated in Fig. 5, showing the phase angles of Γ_i. When logarithmic frequency scale is used, all three curves have the same shape. Changing the element values results only in shifting the curves along the frequency axis. Graphically, the curves should be positioned so as to maximize their mutual vertical distances over the required frequency range

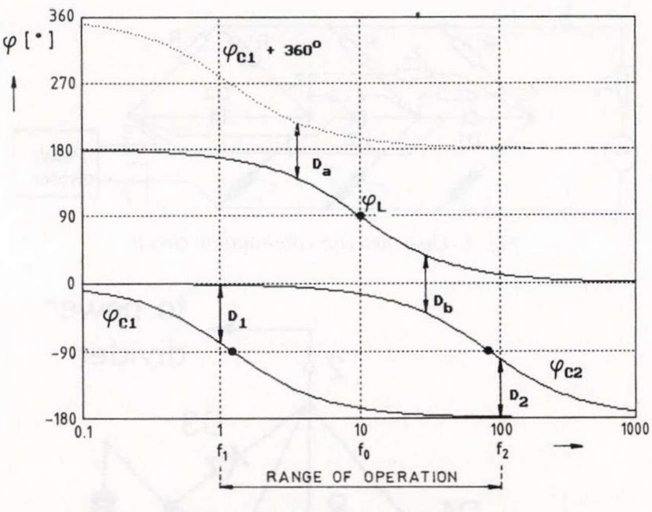


Fig. 5. Phase angles of C—L—C reflectors: an illustration to graphical reflector design

$f_1 \leq f \leq f_2$. The best choice is to center φ_L at the mid-frequency $f_o = (f_1 f_2)^{1/2}$ and shift φ_{C1}, φ_{C2} by the equal distances in opposite directions so that the phase differences at band edges (D_1, D_2) equal the local minima D_a, D_b. It turns out that these minimum phase differences are not less than 55° over a 1000 : 1 bandwidth, or 77° over a 100 : 1 bandwidth. Multi-element reflectors (e.g. series or parallel resonance circuits) can be treated in a similar manner. They are theoretically even better but the values of some elements may be too large or too small to realize.

Surprisingly, the optimum design of the $C—L—C$ and $L—C—L$ reflectors can be performed according to a closed-form analytical expression. The derivation is relatively lengthy and can be found in [5]. For the numerical analysis and design of more general reflectors (including short, open, inductor, capacitor, series and parallel resonance circuits), a computer program has been developed.

4. MONOLITHICALLY INTEGRATED GaAs LUMPED SIX-PORT REFLECTOMETER

A variety of six-ports based on the described principle have been constructed and tested, covering the bands of 100 kHz to 100 MHz, 200 kHz to 20 MHz, and 1 MHz to 700 MHz. Under development is a microwave implementation employing GaAs monolithic integration technology [6]. The circuit diagram of the reflectometer is shown in Fig. 6. The left-hand dashed box contains the combination circuit, the right-hand one the power divider. Reflectors of the type $L—S—C$ (inductor $L_1 = 9.15$ nH, series resonance circuit $L_2 = 0.34$ nH, $C_2 = 18.5$ pF, capacitor $C_3 = 0.69$ pF) have been optimized for the two-decade range 200 MHz to 20 GHz. The combination circuit and power divider are realized on separate GaAs chips with dimensions 1 mm × 1 mm and 1 mm × 0.5 mm, respectively. Four additional chips of Si zero-bias Schottky diodes are used as detectors. The diodes used in combination detectors should be matched to $R_c = 300\ \Omega$. To assure an acceptable match over the whole frequency band, the series resistance $R_s = 150\ \Omega$ had to be provided on the combination circuit chip. The price is the drop of detector sensitivities at higher frequencies. However, the SPR was actually not expected to work above 10 GHz. Since the match of the reference detector $D4$ (to 50 Ω) is not critical, the diode was connected directly. Prior to dc amplification, the detector output voltages ($V1$ to $V4$) are low-pass filtered by integrated resistors R_f and capacitors C_f exterior to chips. All components are placed on two alu-

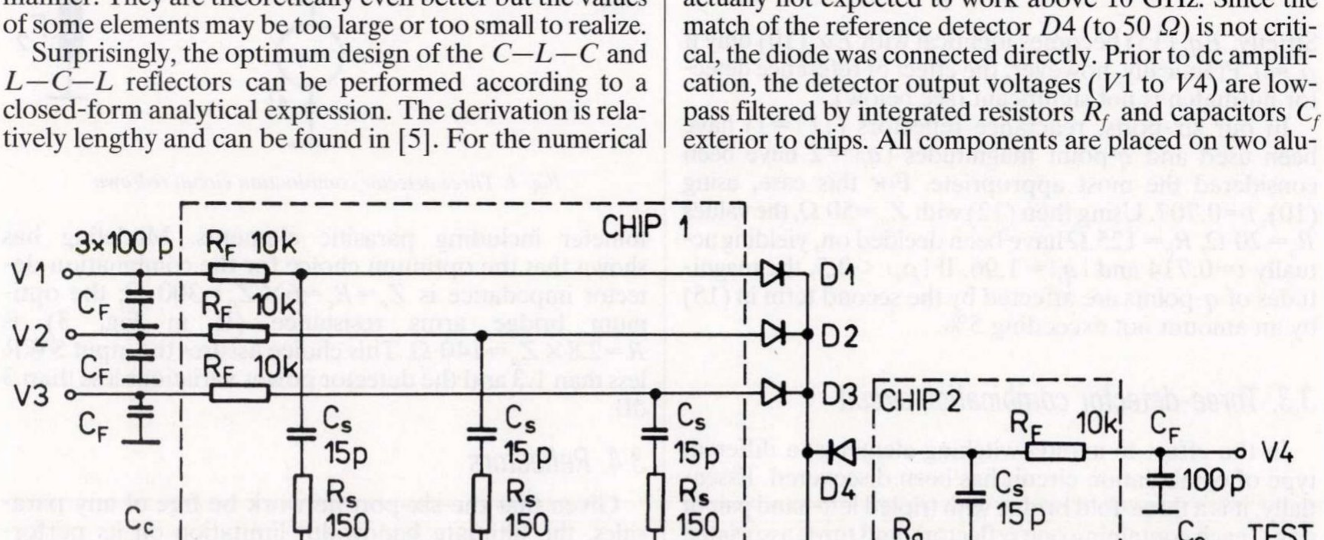


Fig. 6. Circuit diagram of the monolithic six-port reflectometer

mina substrates bonded to 10 mm × 10 mm kovar plate and separated by a grounding ridge with bypass capacitors C_f (Fig. 7). This sub-assembly is placed in a metallic box fitted to the block of pre-amplifiers, forming thus a compact six-port probe with dimensions of 80 mm × 50 mm × 22 mm and a weight of 100 g (Fig. 8). SMA type input and test port connectors are used. The figure also shows three of the calibration shorts and the microwave sub-assembly.

Results of computer modeling of the SPR including equivalent circuits of the detectors are illustrated in Figs. 9 to 11. Fig. 9 shows the magnitudes and Fig. 10 the phase angles of q-points. The magnitudes vary between 1.3. and 3.3.; the deviation from the nominal value 2 is caused by the combination detector impedances differing from 300 Ω. The minimum phase difference of q-points is 94°.

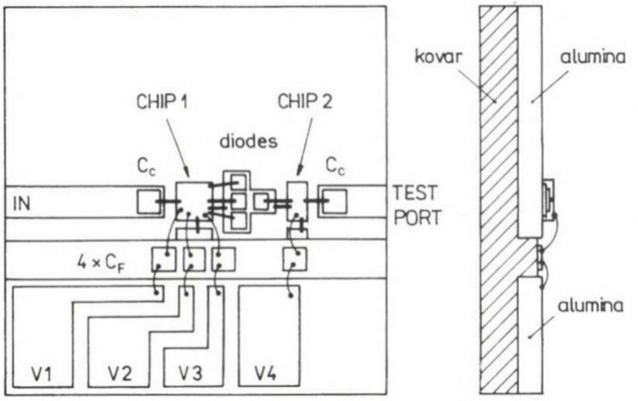


Fig. 7. SPR microwave sub-assemly layout

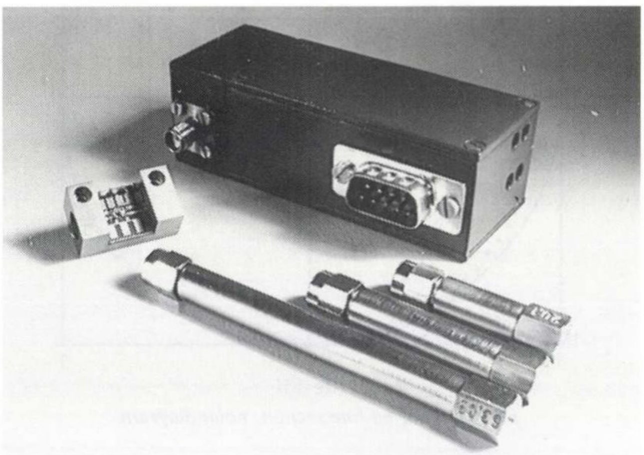


Fig. 8. Assembled six-port probe with microwave sub-assembly (left) and three of the calibration shorts (front)

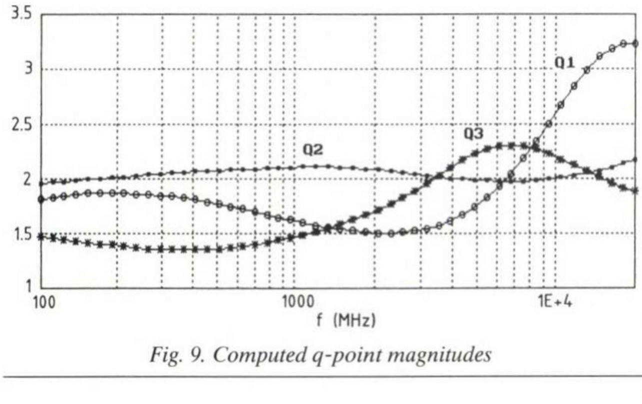


Fig. 9. Computed q-point magnitudes

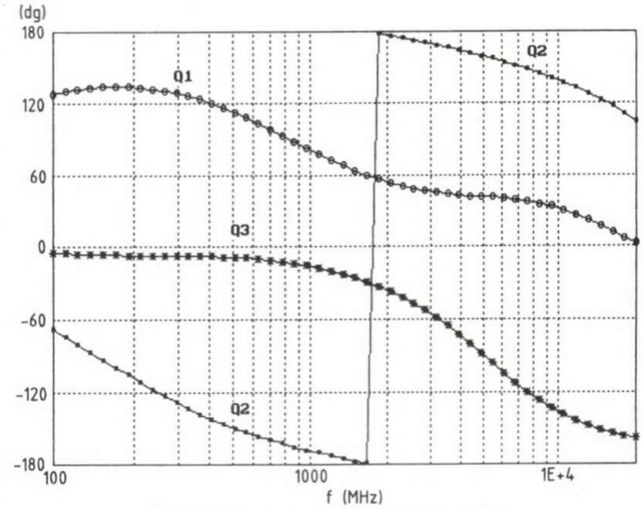


Fig. 10. Computed q-point phase angles

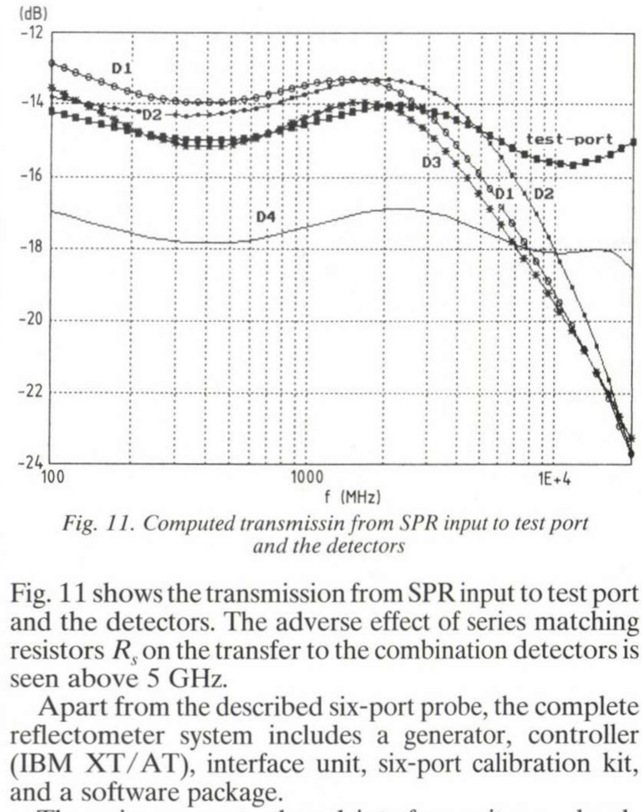


Fig. 11. Computed transmissin from SPR input to test port and the detectors

Fig. 11 shows the transmission from SPR input to test port and the detectors. The adverse effect of series matching resistors R_s on the transfer to the combination detectors is seen above 5 GHz.

Apart from the described six-port probe, the complete reflectometer system includes a generator, controller (IBM XT/AT), interface unit, six-port calibration kit, and a software package.

The microcomputer-based interface unit was developed as a dedicated XT/AT PC card. It contains 8 dc amplifiers, an 8-channel analog multiplexer, and a 12-bit A/D converter for the measurement of detector output voltages. The unit also generates tuning voltage for the sweeper (12-bit D/A converter) and several TTL control voltages (e.g. for switching off the signal power). Net detector voltages are calculated by subtracting the dc offset of the analog circuitry. The output data can be the result of a single measurement or the average of up to 256 samples. The sampling rate is approximately 1500 8-channel samples/s.

The generator may also be controlled directly from the computer via an HP—IB bus.

The calibration kit contains matched loads, a set of open- and short-circuited line sections with known electrical lengths, and a 3 dB pad. The total number of 13 line sections covers the frequency range of 100 MHz to 18 GHz.

The developed software facilitates SPR calibration, reflection coefficient measurement, and data processing. Maximum number of frequency points depends on the computer memory (typically 200 points). The measured results are displayed in polar form (Smith chart) or in rectangular coordinates (magnitude, log magnitude, SWR, phase vs. frequency). The measurement plane can be software-extended; recalibration to eliminate effects of transitions to other types of transmission lines is supported. The data can be converted to various formats (e.g. impedance), Fourier-transformed, stored and re-loaded, printed and hard-copied. The measurement speed is typically 100 points/s.

5. EXPERIMENTAL RESULTS

Figs. 12 and 13 show the magnitudes and phases of q-points in the frequency range of 75 MHz to 9 GHz as obtained by the calibration process. Fig. 12 shows that the magnitudes have the expected values up to approximately 6 GHz, then q_1 rises very steeply indicating that the wave reflected from the DUT does not reach detector D1. Modeling including parasitic elements, indicates that the interconnections between the chips are decisive.

The phases (Fig. 13) vary faster with frequency than one would expect from the reflection coefficient of a lumped inductor or capacitor. This is because of the distance between SPR chips and the measurement plane (ca. 25 mm). The phase differences are of importance: above 6 GHz, the phases of q_2 and q_3 come too close to each other; the measurement error is bound to increase. This is clearly demonstrated in the measurement of an offset short (Figs. 14 and 15). While below 5.5 GHz, the magnitude error is less than 0.01 (assuming the true value $|\rho|=1$), the error increases rapidly above this frequency.

The capabilities of the SPR are demonstrated in Figs. 16 to 20. Fig. 16 is the inverse Fourier transform of the reflection coefficient of the above offset short, simulating thus time-domain reflectometry (response to the unit impulse). The ripples are caused by the frequency-limited nature of the stimulus (0.1 to 8 GHz with steps of 0.1 GHz).

Fig. 17 represents the measurement on a rod antenna of 19 mm length and 2 mm diameter above a conducting plane. The results are referred to $Z_o=75\ \Omega$. A 50 Ω-to-75 Ω adapter was connected to the SPR, then recalibration was performed with a 75 Ω matched load, an open and a short. Finally, the measurement plane was software-extended right to the antenna input plane.

Fig. 18 shows an example for a resonator measurement. The 90 mm long coaxial resonator is filled with water. Since both Q-factor and resonance frequency are strongly affected by impurities, the method can be used to monitor water contamination. A similar method of testing crude oil quality has been reported using, however, an HP 8510 network analyzer.

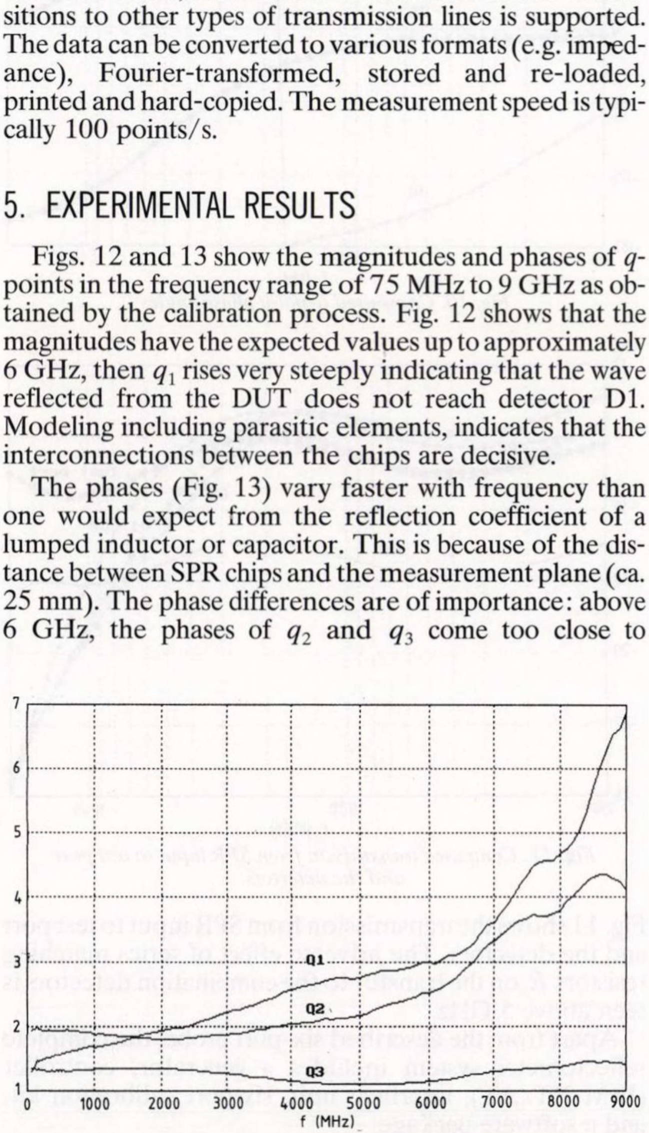


Fig. 12. Measured q-point magnitudes

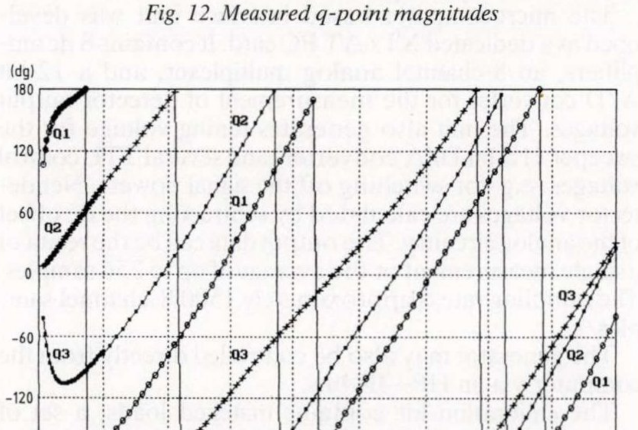


Fig. 13. Measured q-point phases

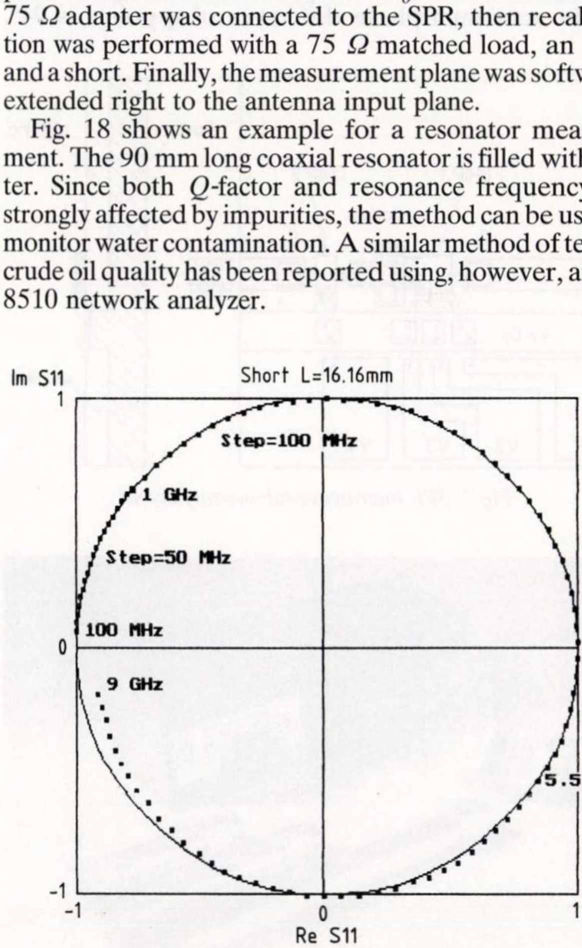


Fig. 14. Shorted line section: polar diagram

Fig. 15. Shorted line section: magnitude of reflection coefficient

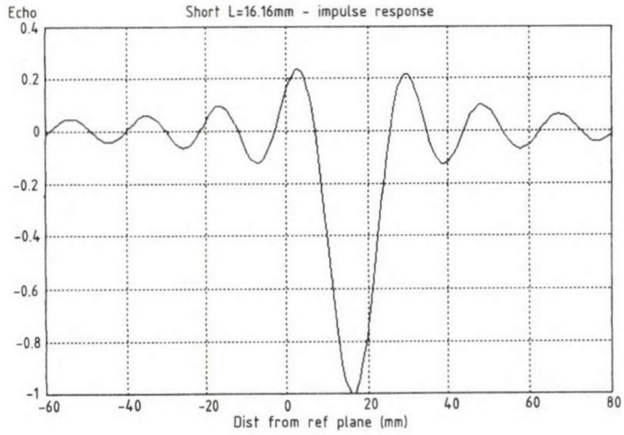


Fig. 16. Shorted line section: impulse response

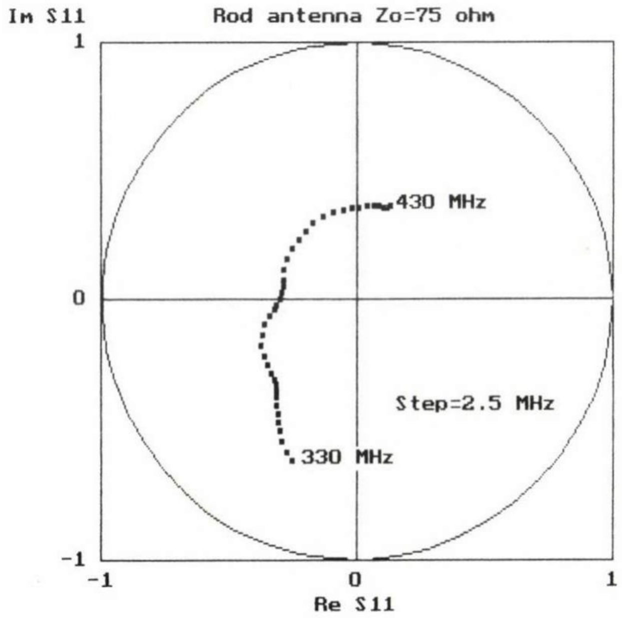


Fig. 17. Reflection coefficient of a rod antenna (referred to $Z_0 = 75\ \Omega$)

Figs. 19 and 20 demonstrate the broadband measurement of medium reflection coefficients and supply information on the developed SPR. Fig. 19 shows the loci of the input reflection coefficient S_{11} of the reflectometer as measured by a device of the same type. Fig. 20 shows the loci of the test port reflection coefficient S_{22}.

The experiments conducted so far have indicated that a typical magnitude error expressed as radius $\Delta\rho$ of uncertainty circle is $\Delta\rho=0.01$ to 0.02. The corresponding phase error ranges from $0.6^\circ/|\rho|$ to $1.2^\circ/|\rho|$.

6. CONCLUSIONS

A new promising type of a very wideband and miniature vector six-port reflectometer has been described. The basic properties characterizing the developed reflectometer and distinguishing it from other types are as follows:

1. *Extreme bandwidth.* The developed SPR can achieve bandwidths as large as 1000 : 1, typically 100 times more than other known six-port types. This property is a decisive advantage compared with all existing SPR types.

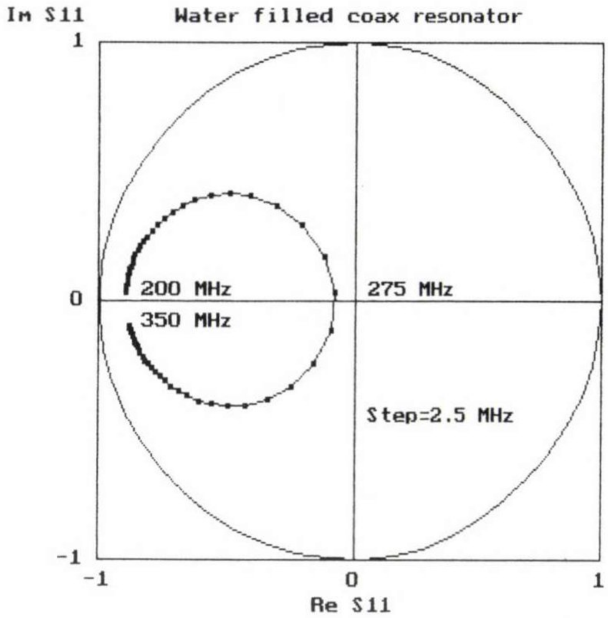


Fig. 18. Resonance loop of a water-filled coaxial resonator

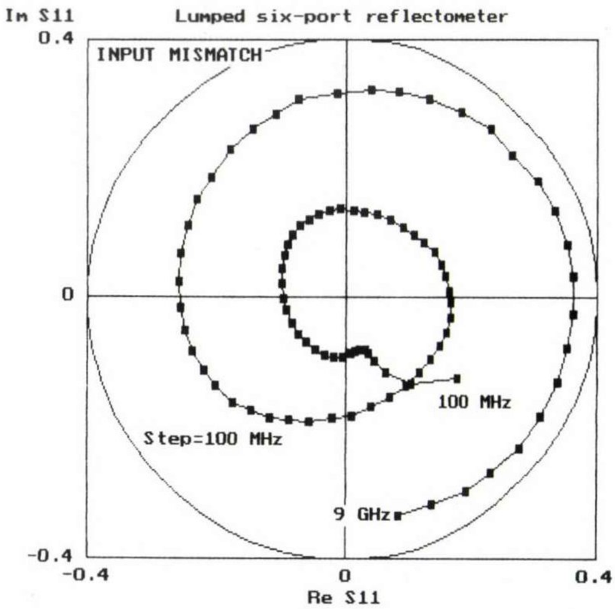


Fig. 19. Input reflection coefficient of the monolithic SPR

2. *Losses.* The SPR absorbs a considerable portion of the power supplied by the generator—typical attenuation between generator and test-port is 15 dB. In most applications, the losses are not a limiting factor.

3. *Small dimensions.* Since the SPR may consist solely of lumped elements, it lends itself to hybrid or monolithic integration.

4. *Versatility.* The principle of lumped SPR can be implemented in any frequency range where lumped elements with required values are realizable (several Hz to tens of GHz). The particular 3-decade band of operation is determined by 3 significant reactances (reflectors). This makes it possible, at least at lower frequencies, to construct a universal assembly (integrated circuit) to which external reflectors, optimized for a required frequency band, are connected.

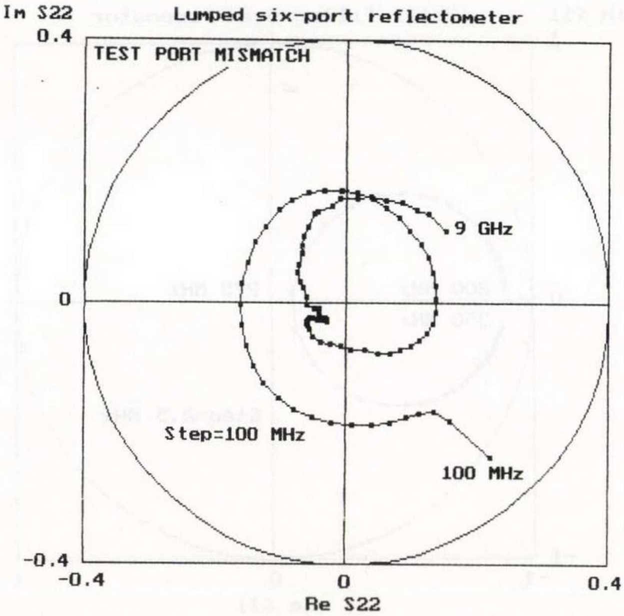


Fig. 20. Test port reflection coefficient of the monolithic SPR

5. *New applications.* Miniature dimensions, versatility, and low cost of the lumped SPR open new applications to vector reflectometers which were not practical before. An example is a successfully tested water level meter [7].

Practical capabilities of the system have been experimentally demonstrated with a GaAs MMIC device operating in the range of 75 MHz to approximately 8 GHz.

REFERENCES

[1] Engen, G. F., "The Six-Port Reflectometer: An Alternative Network Analyzer." *IEEE Trans. Microwave Theory Tech.*, MTT—25, 1977, No. 12, pp. 1075—1079.

[2] Hoer, C. A., "A Network Analyzer Incorporating Two Six-Port Reflectometers." *IEEE Trans. Microwave Theory Tech.*, MTT—25, 1977, No. 12, pp. 1070—1074.

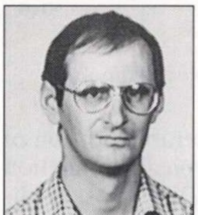
[3] Bezek, J., Bilik, V. and Raffaj, V., "Breitbandiges Sechsportreflektometer mit den Konzentrierten Parametern." *Proc. of 32nd Int. Wiss. Kolloquium,* Heft 2, Ilmenau 1987, pp. 113—116.

[4] Bezek, J., Raffaj, V. and Bilik, V., "Application of the six-port principle to the measurement of complex impedance in lower frequency bands." *Proc. of Elektrotechnika-88,* Sec. Electronics, Faculty of Elec. Eng. of Slovak Univ. of Technol., Bratislava 1988, pp. 98—100 (in Slovak).

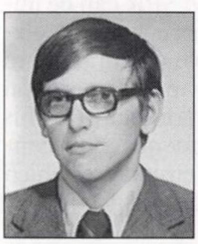
[5] Bilik, V. et al., "Wideband lumped-element six-port structures." *[Research Report F426/88].* Bratislava, Faculty of Elec. Eng. of Slovak Univ. of Technol., 1988 (in Slovak).

[6] Bilik, V., Raffaj, V. and Bezek, J., "A new extremely wideband lumped six-port reflectometer." *Proc. 20th European Microwave Conf.*, Vol. 2., Budapest 1990, pp. 1473—1478.

[7] Bezek, J., Raffaj, V. and Bilik, V., "Application of the lumped six-port reflectometer to water level measurement." *Proc. of Mikon—91,* Rydzyna (Poland) 1991 (to be published).



Vladimir Bilik
was born in 1948. He graduated from the Slovak University of Technology in Bratislava, Czechoslovakia, in 1972, where he also received his PhD in 1979. He has been with the Faculty of Electrical Engineering of the University since 1972. In 1986/87 he spent an academic season as a senior lecturer at the University of Jos, Makurdi Campus, Nigeria. Since 1984, his professional activities have included research in the field of automated microwave S-parameter measurements, especially six-port reflectometers and network analyzers.



Vladimir Raffaj
was born in 1950. He received his MSEE from the Slovak University of Technology in Bratislava in 1975 and his PhD from the same University in 1983. He has been with the Faculty of Electrical Engineering of the University since 1975. His professional activities are focused on application of microcomputers, experiment planning, and automation of microwave measurements.

Jan Bezek
was born in 1947. He graduated from Slovak University of Technology in Bratislava in 1970 and joined the University as research scientist. In 1987, he defended his PhD thesis on microwave measurements of dielectric properties of materials. His activities have also included the measurement of dielectric resonators, automatic S-parameter measurements, and applications of six-port reflectometers.

ERROR CORRECTION OF MICROWAVE NETWORK ANALYZERS

J. LADVÁNSZKY

RESEARCH INSTITUTE FOR TELECOMMUNICATIONS
H—1525, BUDAPEST, P.O.B. 15, HUNGARY

Exact calibration conditions for measuring the scattering matrix are given, and a simplified error model suitable for measuring the scattering matrix of active two-port devices is presented. The procedure is realized by a computer controlled network analyzer, and the computer program is illustrated by measurement examples.

1. INTRODUCTION

Measurement setups based on complex ratio detection — called network analyzers — are widely used for measuring the scattering matrix elements of microwave circuits. However, spurious transmissions and reflections arising in network analyzers are sources of measurement errors, necessitating error correction. Error correction has been treated by numerous authors [1] by assuming that systematic measurement errors can be modelled by an error network inserted between an error-free instrument and the device to be tested [2]... [4]. However validity conditions of this model have not been examined. In Section 2, sufficient conditions for modelling systematic errors by an error network are presented. In order to determine the parameters of the error network, calibrating circuits with known scattering matrices are used. We show in Section 3 that in general, at least one *non-reciprocal* circuit with a known scattering matrix is necessary for the calibration. Since non-reciprocal calibrating circuits are practically not available, conditions for calibrating by *reciprocal* circuits are presented.

As a summary of our results the algorithm of the calibration and the error correction are given in Section 4. A computer program is introduced in Section 5 for the calculations and for controlling the measurement setup. Finally three examples are presented in Section 6 to verify the program.

2. THE ERROR NETWORK

The block diagram of a two-port measurement setup is shown in Fig. 1. The measured device is connected to the S-parameter test set S through the fixture F. The test set S transmits the generator signal to the measured device X through the fixture F and receives the signal reflected by the measured device to the test input T of the ratio detector Q. The generator signal is transmitted also to the reference input R of the ratio detector through the test set S. The measured device changes the signal transmission from the generator to the input T, but the transmission to the input R remains unchanged. Thus the ratio of the two signals is proportional to the parameter to be measured.

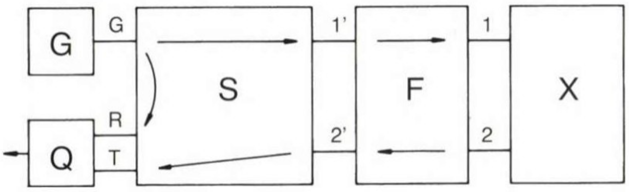


Fig. 1. Block diagram of a measurement setup using a network analyzer. G-generator, Q-complex ratio detector, X-device under test, S-test set

The spurious transmissions and reflections of the test set S, ratio detector Q and the fixture F cause measurement error. In numerous publications [2]... [4], the measurement errors are modelled by the four-port E shown in Fig. 2 which is inserted between the measured device X and the instrument which is assumed to be ideal.

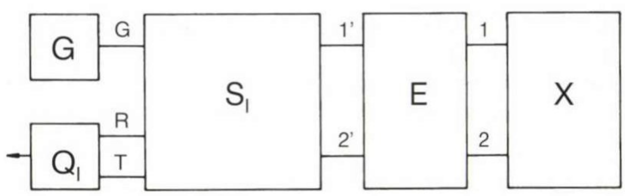


Fig. 2. Modelling of the measurement error by an error network. S_I-ideal test set free of spurious transmissions and reflections, Q_I-ideal ratio detector

The scattering matrix element of the two-port which is to be measured can be selected by switches of the test set S. In accordance with the four scattering matrix elements, four internal signal paths can be switch selected. In the ideal case, transmission exists only between three pairs of ports in each switch position, while all other transmissions and treflections are equal to zero. The signal paths corresponding to the measurement of S_{21} are shown in Fig. 1. In our previous examinations we showed [7] that the validity conditions of the model shown in Fig. 2 are as follows:

- when switching the test set S, one only of the main signal path transmissions is allowed to change at a time,
- the reflections at ports 1 and 2 of the test set S and the spurious transmission between them are not allowed to change while switching,
- all spurious reflections have to be much less than unity, and all spurious transmissions have to be much less than the transmissions along the main signal paths.

If the above conditions are satisfied then the measurement errors originating from four signal paths within the test set can be modelled by a single error circuit.

3. CALIBRATION BY RECIPROCAL CIRCUITS

The parameters of the four-port E have to be determined to enable the correction of measurement errors. Thus calibration is needed by measuring circuits having known scattering matrices. In all previous publications, it was assumed that the parameters of the error network E can be determined using reciprocal calibrating circuits. We shall prove that this is not valid in general, hence further restrictions concerning network E have to be fulfilled, as given by the following theoretical investigation.

The error network E has a scattering matrix $\mathbf{S}_E$ which can be partitioned into 2×2 submatrices:

$$\mathbf{S}_E=\begin{bmatrix}\mathbf{A} & \mathbf{B}\\ \mathbf{C} & \mathbf{D}\end{bmatrix} \tag{1}$$

Let $\mathbf{S}_X$ and $\mathbf{S}_M$ denote the actual and the measured scattering matrices. With the notation introduced in Eq. (1) the relation between $\mathbf{S}_X$ and $\mathbf{S}_M$ is as follows:

$$\mathbf{S}_M=\mathbf{A}+\mathbf{B}\mathbf{S}_X(\mathbf{I}-\mathbf{D}\mathbf{S}_X)^{-1}\mathbf{C} \tag{2}$$

where $\mathbf{I}$ denotes the 2×2 identity matrix and $(\mathbf{I}-\mathbf{D}\mathbf{S}_X)$ is assumed to be non-singular.

For the error correction, we have to determine the elements of the **A, B, C, D** matrices from the measured scattering matrices $\mathbf{S}_{Mi}$ (i = 1,2,...) of the calibrating circuits which have known scattering matrices $\mathbf{S}_{Ci}$ (i = 1,2,...). Then we should write $\mathbf{S}_{Mi}$ and $\mathbf{S}_{Ci}$ instead of $\mathbf{S}_M$ and $\mathbf{S}_X$.

From the 16 elements of matrices **A, B, C, D** only 15 are independent because multiplication of **B** and division of **C** by the same complex number leave the transformation given by Eq. (2) unchanged.

We introduce the transformed scattering matrix

$$\mathbf{S}_T=[\mathbf{C}^{-1}\mathbf{D}\mathbf{B}^{-1}+(\mathbf{S}_M-\mathbf{A})^{-1}]^{-1} \tag{3}$$

where the matrices $\mathbf{C}^{-1}\mathbf{D}\mathbf{B}^{-1}$ and $\mathbf{A}$ can be determined using reciprocal calibrating circuits [7]. A circuit is reciprocal when it has a symmetric scattering matrix: $\mathbf{S}_{Ci}=\mathbf{S}_{Ci}^T$ where T denotes transposition.

The relation between $\mathbf{S}_T$ and $\mathbf{S}_X$ can be expressed from Eq. (2):

$$\mathbf{S}_T=\mathbf{B}\mathbf{S}_X\mathbf{C} \tag{4}$$

This is a linear equation which can be rewritten in the following form:

$$\begin{bmatrix}S_{T11}\\ S_{T21}\\ S_{T12}\\ S_{T22}\end{bmatrix}=\mathbf{K}\begin{bmatrix}S_{X11}\\ S_{X21}\\ S_{X12}\\ S_{X22}\end{bmatrix}=$$

$$=\begin{bmatrix}b_{11}c_{11} & b_{12}c_{11} & b_{11}c_{21} & b_{12}c_{21}\\ b_{21}c_{11} & b_{22}c_{11} & b_{21}c_{21} & b_{22}c_{21}\\ b_{11}c_{12} & b_{12}c_{12} & b_{11}c_{22} & b_{12}c_{22}\\ b_{21}c_{12} & b_{22}c_{12} & b_{21}c_{22} & b_{22}c_{22}\end{bmatrix}\begin{bmatrix}S_{X11}\\ S_{X21}\\ S_{X12}\\ S_{X22}\end{bmatrix} \tag{5}$$

where b_{ij} and c_{ij} denote the elements of **B** and **C**.

Thus the problem of the error correction is reduced to the inversion of Eq. (5).

The elements of the first and the last column can also be determined by using reciprocal calibrating elements [7]. However, only the sum of the inner columns can be determined because $S_{C12}=S_{C21}$. This may be sufficient if the measured circuit is reciprocal because in this case, Eq.(5) can be simplified:

$$\begin{bmatrix}S_{T11}\\ S_{T21}\\ S_{T22}\end{bmatrix}=\begin{bmatrix}k_{11} & (k_{12}+k_{13}) & k_{14}\\ k_{21} & (k_{21}+k_{23}) & k_{24}\\ k_{41} & (k_{42}+k_{43}) & k_{44}\end{bmatrix}\begin{bmatrix}S_{X11}\\ S_{X21}\\ S_{X22}\end{bmatrix} \tag{6}$$

However, the measured circuit (for instance, a transistor) may not be reciprocal, thus we must find further relationships between the elements of the **K** matrix. These would be obtained if the error network E would be reciprocal. But generally, E is not reciprocal [7].

The elements of the **K** matrix are the products of the elements of matrices **B** and **C** according to Eq. 5. The **B** and **C** matrices comprise 8 elements in all, multiplication of **B** and division of **C** by the same number leave **K** unchanged. Hence, the matrix **K** has 7 independent elements at most. Based on Eq. (5), the number of the independent elements is exactly 7. In any column and row of the matrix **K,** there are 3 independent elements, and the elements of the 1st and the 4th column are already known. Thus the knowledge of one element of the 2nd and 3rd columns is necessary and sufficient to determine all other matrix elements. Therefore, in order to determine the elements of **K,** that is to calibrate, in general a non-reciprocal circuit is necessary. As non-reciprocal circuits having specified scattering matrices are not available, calibration by reciprocal circuits have to be investigated as well. We have seen that the calibration is only possible if one element of the critical 2nd or 3rd columns is known.

The latter assumption is fulfilled in the practice so that **B** and **C** will be diagonal matrices, i.e. the elements in the cross diagonal (b_{12}, b_{21}, c_{12}, c_{21}) represent the spurious transmissions of the test set S, and practically, these can be neglected [7], thus meeting the 3rd condition in Section 2.

4. ERROR CORRECTION

In this Section, a calibration and error correction algorithm based on the theoretical results is presented. Calibrating circuits realizable: double matched termination (two 50 Ohm resistors), double short-circuit, and a transmission line of known electric length. Their scattering matrices are the following:

$$\mathbf{S}_{C1}=\mathbf{O},\qquad \mathbf{S}_{C2}=-\mathbf{I},\qquad \mathbf{S}_{C3}=e^{-j\phi}\begin{bmatrix}0 & 1\\ 1 & 0\end{bmatrix} \tag{7}$$

In order to simplify the notations, two special matrix operations are used. Let $\mathbf{X}=[x_{ij}]$ and $\mathbf{Y}=[y_{ij}]$ be two matrices of equal dimensions. The direct multiplication and division of **X** by **Y** is defined as follows:

$$\mathbf{X}\cdot\mathbf{Y}=[x_{ij}y_{ij}]\qquad \mathbf{X}:\mathbf{Y}=\left[\frac{x_{ij}}{y_{ij}}\right] \tag{8}$$

In the error correction, the diagonal character of **B** and **C** is utilized. We form a new matrix **H** from the elements of **B** and **C**:

$$\mathbf{H}=\begin{bmatrix}b_{11}c_{11} & b_{22}c_{11}\\ b_{11}c_{22} & b_{22}c_{22}\end{bmatrix} \tag{9}$$

When B and C are diagonal, Eq.(2) can be rearranged into the following form:

$$\mathbf{S}_X=[\mathbf{D}+\mathbf{H}(\mathbf{S}_M-\mathbf{A})^{-1}]^{-1} \tag{10}$$

For the error correction, matrices **A, D** and **H** have to be identified based by utilizing Eqs. (7) and (10):

$$\mathbf{A}=\mathbf{S}_{M1} \tag{11}$$

$$\mathbf{H}=(\mathbf{S}_{C2}^{-1}-\mathbf{S}_{C3}^{-1}):[(\mathbf{S}_{M2}-\mathbf{S}_{M1})^{-1}-(\mathbf{S}_{M3}-\mathbf{S}_{M1})^{-1}] \tag{12}$$

$$\mathbf{D}=\mathbf{S}_{C2}^{-1}-\mathbf{H}(\mathbf{S}_{M2}-\mathbf{S}_{M1})^{-1} \tag{13}$$

We can see from Eq.(9) that the matrix **H** has only 3 independent elements, while the **A** and **D** matrices have generally four each. Hence, the system is characterized by 11 complex numbers at a given frequency.

5. COMPUTER PROGRAM

For performing the calculations in Eqs. (10) to (13), and for controlling the measurement setup, several computer programs have been written requiring a storage capacity of approximately 30 kbyte. The basic version of the program consists of three parts, for the setting of the initial values, for the calibration, and for the measurement of the unknown devices. Extended program versions are available for plotting, for determination of circuit parameters defined by the user, and for storage of the measured data. The stored parameters provide initial data for circuit analysis and optimization programs written for larger computers.

The second part of the program allows several kinds of calibrations. If the measured device is a one-port, the calibration is simple [5]. In this case matched termination, short circuit and open circuit are used as calibrating elements. In order to increase the measurement accuracy, the control of a sliding load, the circle fitting for measured data [6] and the capacitive model of the open circuit are also included in the program. In the case of two-port measurements, three calibrating two-ports according to Section 4 are used, providing 12 measured data at each frequency. The number of the error parameters is only 11, therefore one of the measured data is used for checking the calibration.

6. EXAMPLES

In this Section, three examples are presented which verify the necessity of the error correction.

Example 1

It is known that the reflection coefficient of a passive device has a magnitude less than one. However, due to measurement errors, the measured reflection coefficient may have a magnitude more than one. In our experiment, we measured the input reflection coefficient of a 10.25 cm long 50 ohm coaxial transmission line shorted at the end. S_M=1.2 was measured at 3.6 GHz, which resulted in a value of S_X=0.98 after error correction, in agreement with the expectation.

Example 2

The phase of the reflection coefficient of a trimming capacitor shunting a microstrip line is shown in Fig. 3. Due to the measurement errors, the phase response is irregular, while the points obtained after the error correction fit on a straight line with good accuracy, agreeing with the theoretical expectation. We found that the average distance of the measured points from the line of regression decreased to its tenth due to the error correction.

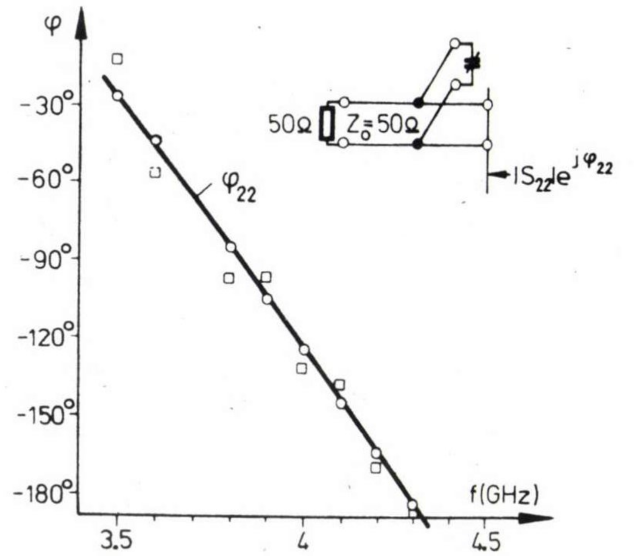


Fig. 3. The phase of the reflection coefficient of a trimming capacitor shunting a microstrip line.
Squares: measured data, circles: data after error correction.

Example 3

The S_{21} element of the field effect transistor CFY 11 is shown in Fig.4 in the 2 to 12 GHz frequency range. The operation point is V_{DS}=4 V, I_D=30 mA. The error correction presented in Section 4 has been applied. Similarly to Example 2, spread of the measured data is larger than that of the corrected data. We can also observe that the error correction is especially effective at high frequencies.

7. CONCLUSION

In this paper, conditions are given for modelling the measurement errors of a network analyzer by an error

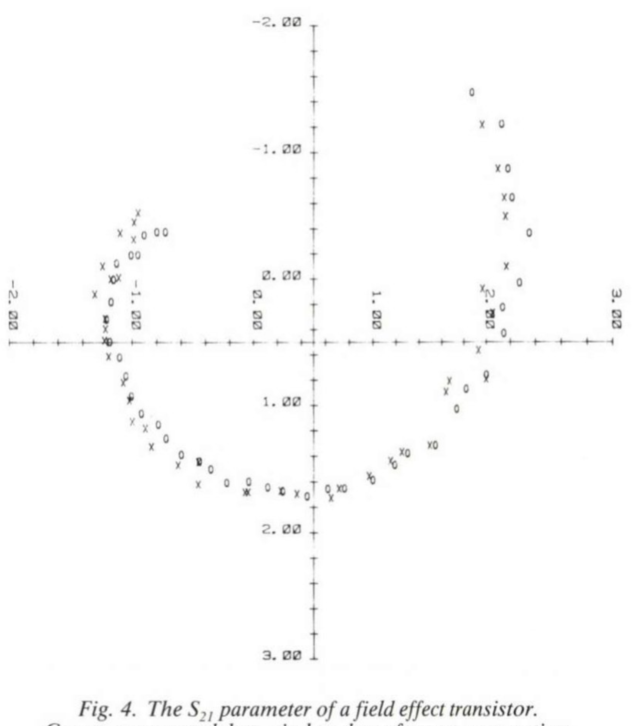


Fig. 4. The S_{21} parameter of a field effect transistor.
Crosses: measured data, circles: data after error correction.

network, and for the identification of model parameters by reciprocal circuits. An algorithmically effective error correction procedure has been developed and implemented in a computer program.

When applying this procedure, quality of the calibrating circuits is essential. Fabrication of good calibrating circuits is difficult and expensive. Therefore a question arises: What is the minimum number of data that we have to know a priori of the calibrating circuits? That may be a direction for further research.

8. ACKNOWLEDGEMENT

The author is indebted to his teacher, Dr. A. Baranyi, who initiated the research presented in this paper. The author also thanks his colleagues, especially Dr. I. Kása, for their kind help.

REFERENCES

[1] Adam, S.F., "Automatic Microwave Network Measurements", *Proc. of the IEEE,* April 1978, pp. 384—391.

[2] Gelnovatch, V.G., "Computer Program for the Direct Calibration of Two-Port Reflectometers for Automated Microwave Measurements", *IEEE Trans. on MTT,* Jan. 1976, pp. 45—47.

[3] Speciale, R.A., "A Generalization of the TSD Network-Analyzer Calibration Procedure Covering n-Port Scattering-Parameter Measurements Affected by Leakage Errors", *IEEE Trans. on MTT,* Dec. 1977, pp. 1100—1115.

[4] Fitzpatrick, J., "Error Models for Systems Measurement", *Microwave Journal,* May 1978, pp. 63—66.

[5] Bianco, B., Corana, A., Ridella, S. and Simicich, C., "Evaluation of Errors in Calibration Procedures for Measurements of Reflection Coefficient", *IEEE Trans. on IM,* Dec. 1978. pp. 356—358.

[6] Kása I., "A Circle Fitting Procedure and Its Error Analysis", *IEEE Trans. on IM,* March 1976, pp. 8—14.

[7] Baranyi A. and Ladvánszky J., "On the Exact S-parameter Measurement of Active Devices", *Proc. of the 10th European Microwave Conference,* Warszava, 8—11th Sept. 1980, pp. 278—282.

János Ladvánszky

was born in Budapest in 1955. He received the electrical engineer, candidate of science and dr. univ. degrees in 1978, 1988 and 1990, respectively, from the Budapest Technical University and the Hungarian Academy of Sciences. He has been with the Research Institute for Telecommunications since 1978. His main interests are microwave measurements, transistor modelling, circuit theory and neural networks. He is an associate editor of the Journal on Communications and a member of the Committee for Telecommunication Systems of the Hungarian Academy of Sciences.

ACTIVE MICROWAVE REMOTE SENSING

I. BOZSÓKI, B. FARKAS, É. GÖDÖR,
S. MIHÁLY AND R. SELLER

TECHNICAL UNIVERSITY OF BUDAPEST,
DEPARTMENT OF MICROWAVE TELECOMMUNICATIONS
H—1111 GOLDMANN TÉR 3., HUNGARY

The paper summarizes active microwave imaging. Basic types of ground-based, airborne and spaceborne active microwave sensors are discussed. An X-band scatterometer and a side looking airborne radar (SLAR), developed at the Department of Microwave Telecommunications, Technical University of Budapest, are described.

1. INTRODUCTION

Remote sensing is a method for getting information from an object without having physical contact with it [1], by using electromagnetic waves reflected or radiated by the object. In most cases, the Earth is the object. Remote sensing then provides us with an overall view of the Earth's surface and consecutive shots of any given point. These pieces of information are indispensable in many fields such as agriculture, meteorology, geodesy and cartography, civil engineering, water management, urban development and geology [2]. In a broader sense, remote sensing concerns everyone interested in the environment he or she lives in.

Air photography can be regarded as the first remote sensing tool in the visible band. Satellite imagery has long been an established technique. Landsat and Spot type orbiting satellites serve us on a commercial basis producing multispectral images in the optical and infrared bands.

Radar imagery is a relatively new imaging technique for surveying natural resources, a technique which is still not commercial but has considerable potential for application in resource studies. Moreover, not relying on illumination by the Sun, radar has the capability of producing an image unaffected by haze, clouds and independent of day or night [3].

However, radar observation of targets, cannot be considered in a traditional way according to the appearance of objects which the human eye is accustomed to, though many elements of a radar image are recognizable by size and shape. The interpreter must understand the nature of radar imaging in order to draw useful information from the image [4]. That is why we started at the Department of Microwave Telecommunications to build a scatterometer and a side looking airborne radar (SLAR), having gained support from the National Bureau of Technical Development and the Department of Agriculture [5], [6]. An airborne microwave radiometer has also been developed as it is reported in [7].

In this paper, a short review of radar remote sensing is given first. By this method, the differential radar cross section (σ^o) of the targets as well as dimensions and positions of independent picture elements are presented. The role of polarization and phase of the reflected signal is shown next. Finally, an X-band scatterometer and an X-band real aperture SLAR are described, together with results of measurements. An overview of planned future activities concludes the paper.

2. IMAGING RADARS

Basic imaging tools of microwave active remote sensing are the SLAR and the SAR (Synthetic Aperture Radar) [3, 8]. Both types are pulse radars mounted on airborne or spaceborne platforms (airplanes or satellites). These instruments regularly scan the surface in a direction which is perpendicular to the track formed by the carrier on the surface of the Earth. Consecutive scans are achieved by the motion of the carrier on its orbit, giving a strip-like image, i.e. the swath aside the track.

2.1. Methods of Imaging

Fig. 1 shows the imaging geometry of a SLAR and a strip-mode SAR. If one needs an azimuth resolution in the order of 10 m, a SLAR can be used only at a moderate flying altitude resulting in a few kilometer wide swath. On the other hand, spaceborne SAR can give the same resolution even from several hundred kilometers altitude and a swath-width of 100 km. A SAR on board of an airplane at a height between 5 and 10 km gives pixels with a few meter resolution. This resolution is adequate for most of the applications, including reconnaissance.

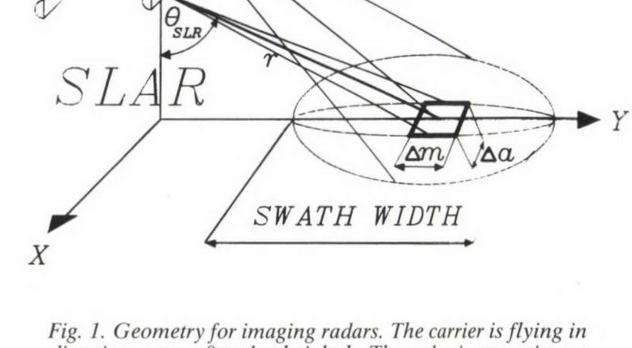


Fig. 1. Geometry for imaging radars. The carrier is flying in direction x at y=0 and at height h. The radar is operating at wavelength λ and bandwidth B_{tr}

SLAR — For real aperture imaging pixel dimensions are given by

$$\Delta a = \frac{\lambda}{L} r \; ; \; \Delta m = \frac{c}{2B_{tr} \sin\theta_{SLR}} .$$

SAR — For imaging with a synthetic aperture the radar positions are from x_1 to x_n and

$$\Delta a \geq \frac{L}{2} \; ; \; \Delta m = \frac{c}{2B_{tr} \sin\theta_{SA}} .$$

2.2. Information Content of the Reflected Signal

An image obtained by an active microwave sensor is composed of individual small parts, the pixels. Brightness of a pixel depends on the reflection characteristics of the surface illuminated by the radar, while the resolution is determined by the parameters of the radar. The cross track resolution depends on the width of the spectra of the transmitted signal and the along track resolution is affected by the along track beamwidth of the antenna and/or the processing of the received signal.

A target in electromagnetic scattering problems is often described by its 'polarization properties' [9]...[11]: either the scattered voltage (electrical field strength) or the power can be measured.

The received voltage in an orthogonal horizontal-vertical measurement system is given in complex analytic form by

$$z = \mathbf{S} \cdot \mathbf{a} \cdot \mathbf{b}$$

where

a, b– polarization vectors of the transmitted and received electromagnetic fields of general elliptic polarizations,

S– complex scattering matrix which describes the relation between incident and scattered electric field vectors with vertical and horizontal components (E), as given by

$$\begin{bmatrix} E_h^s \\ E_v^s \end{bmatrix} = S \begin{bmatrix} E_h^i \\ E^i \end{bmatrix} = \begin{bmatrix} S_{hh} & S_{hv} \\ S_{vh} & S_{vv} \end{bmatrix} \begin{bmatrix} E_h^i \\ E^i \end{bmatrix}$$

A more convenient way to describe a polarization state in terms of power is the use of Stokes vectors instead of using the polarization vectors. In this case, the received power is given in the following form:

$$P = \mathbf{M} \cdot \mathbf{g}(\mathbf{a}) \cdot \mathbf{h}(\mathbf{b})$$

where $\mathbf{g}(\mathbf{a})$ and $\mathbf{h}(\mathbf{b})$ are the Stokes-vectors of the polarization states **a** and **b** as directly derived from the polarization vector. **M** is the 4×4 real-value Stokes matrix which represents the target in terms of power measurement. The latter expression gives ways of treating target parameters and showing their reflection characteristics in an image form.

A well calibrated radar system is capable of measuring the complete **S**, which is then transformed into **M.** The Stokes-matrix contains the information about a measurement with any possible combination of the polarization states of the transmitting and receiving antennas.

Finally, the ability of a certain target to reflect electromagnetic waves incident at any polarization is expressed by the differential radar cross section,

$$\sigma_{pq}^o = \lim_{r \to \infty} \frac{4\pi r^2}{A_o} \frac{\overline{|E_{pq}^s|^2}}{|E_o|^2},$$

$\overline{(\ldots)}$ denotes the ensemble average which should be taken when evaluating the scattered E^s field at the position of the receiving antenna of p polarization. The scattered field is excited by the incident electric field E_0 of q polarization at the target position. The target area over which the actual measurement is carried out is denoted by A_0 and it is at a distance r from the radar.

Radiometric correction and interpretation techniques of the scene given by a microwave image can be improved significantly by using a microwave scatterometer. A microwave scatterometer is a small calibrated radar capable of measuring the differential radar cross-section of a selected area of the surface [12].

2.3. Statistical Description of Distributed Targets

The linear dimensions of a pixel are much larger than the wavelength of the transmitted signal. The reflection of the signal from a pixel can be regarded as the sum of signals, reflected by individual scattering centers [12].

An average-like quantity can be defined as the signal strength of the reflection of a pixel, and the standard deviation as "noise". This explains the noisy-like image which is typical of radar images, and called the speckle.

Speckle can be reduced by integrating N independent samples of the same pixel. The improvement of the signal-to-noise ratio or the reduction of speckle is equal to N. Another method is filtering; instead of the signal itself, the average of the signal of the given and of the neighbouring pixels is displayed at the position of the given pixel.

2.4. Geometry of Radar Observation

When measuring σ^0 of a pixel, the position of that pixel in the image has to be identified.

The distance of a target to the radar can be measured by the time delay between transmission and reception, and targets can be separated by the resolution of the radar in range. The radial velocity gives additional possibility, in case of coherent radars, to separate targets according to their Doppler frequency. This can be utilized to improve the resolution in the direction of the target motion (in azimuth), and/or to separate moving targets from clutter induced by stationary targets [13].

3. X-BAND SCATTEROMETER

In order to get reference data, a dual-polarized scatterometer has been developed at DMT.

According to Fig. 2, the scatterometer consists of an analog microwave and a digital data-processing unit [14]. The former contains the microstrip antenna [15] and the microwave VCO driven FM/CW radar, with an I/Q detector providing two output signals. The signals are digitized at twelve bits and stored for preprocessing [16].

The main aim of preprocessing is to obtain correct values characterizing the power of the received signal which is the transmitted FM signal backscattered by the observed target, namely by the area of the surface illuminated by the antenna [17]. As the frequency of the transmitted signal changes linearly with time, one can expect a sine wave at the output of the detector if a point-target is observed. The actual detected signal has some error due to the non-uniform frequency response of the microwave unit which has to be subtracted from the detected raw signal.

After getting the corrected and RMS detected values for the calibrating delay-line and the observed area, the ratio of the received powers can be obtained which is then used for retrieving the radar cross-section of the illuminated surface. It does not account for the effect of the non-uniform antenna pattern, so additional calculation should be performed to extract the differential radar cross-section (σ^0) whih characterizes the backscattering of a surface element.

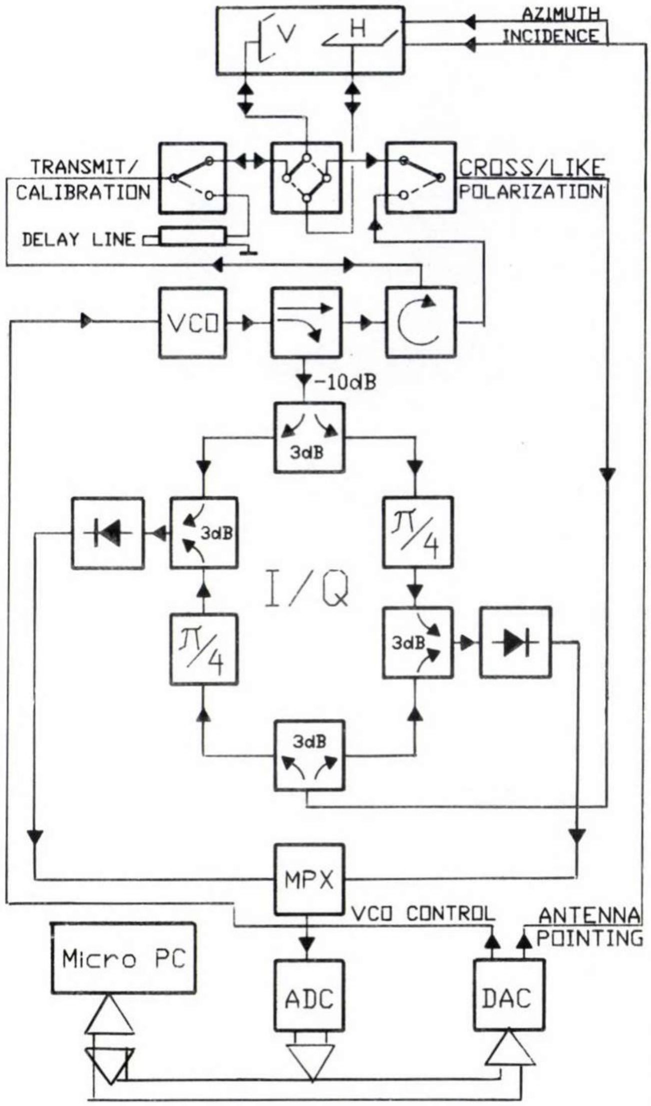


Fig. 2. Block-diagram of dual polarized X-band scatterometer

4. X-BAND SLAR SYSTEM

In 1987, an experimental X-band SLAR started to be developed for applications mostly in agriculture. In Summer 1990, the radar was complete to take the first SLAR images in Hungary [5], [18], [19].

4.1. Objectives

The operating characteristics of the SLAR system were matched to the sensor requirements on a regional basis to provide data for smaller agricultural enterprises.

An additional objective was to gain information on and learn how to use microwave remote sensing [20], distributing the results in Hungary and establish international relations in this field.

Basically, the SLAR system to be discussed, when mounted on board of a smaller airplane, is able to detect reflections of $\sigma^0 \geq -20$ dB with a resolution between 10 and 20 metres at an altitude from 1 to 2 km.

4.2. System Description and Operation

The SLAR system has three main blocks as shown in Fig. 3: X-band radar with antenna, fast data-processor, control PC with peripherals.

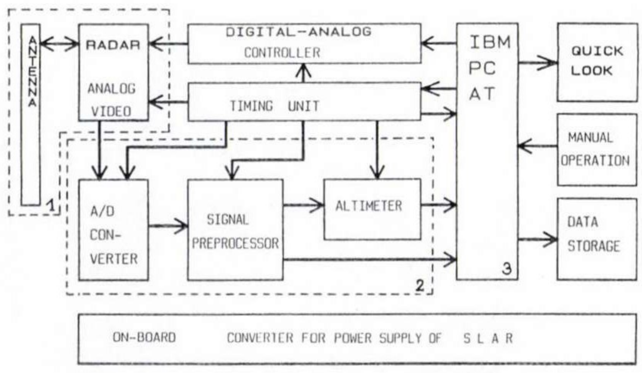


Fig. 3. Block-diagram of SLAR

The radar produces the pulse-modulated transmitter signal with output power P_t and pulse width τ_t, and receives it in the proper time-slot.

The fast data-processor provides real-time storage of the received signal after each pulse transmission, for later processing at a lower speed. A digital solution proved to be better than the analog one discussed in [21].

The PC controls the radar and allows preprocessing of the stored signal, performs the necessary corrections, provides a quick-look facility of the image and stores it for post-processing.

4.3. Imaging Geometry

The antenna radiation pattern is a narrow fan-beam pointing in a direction which is perpendicular to the direction of airplane motion. Return signal is received from the elements of the surface strip which lie within the footprint of the antenna in a consecutive series according to their radial distance. The continuous return signal is sampled at 20 MHz rate for additional processing. As the carrier platform moves on, the scanning of a new line is started and later added to the whole swath of the image.

4.4. Speckle Reduction

The width of the antenna footprint is between 10 and 20 m. An airplane with a velocity of 50 m/s changes its position by 1 metre during 20 ms. If the pulse repetition time is in the same range, several independent return signals are available from the same pixel of a given line. This makes possible the integration of a few samples from nearly the same pixel which results in reduction of speckle.

4.5. Radiometric Correction

The value to be measured is the $\sigma^0(\Theta)$ of the pixels, which has an inherent angular dependence. This and the Θ dependence of the radial distance and the antenna pattern causes a complex dependence of the received power on Θ as given by

$$P_r = \frac{P_t G_o^2}{(4\pi)^3} \frac{\lambda_t c\tau_t}{2} \frac{\beta_{eff}}{h^3} \frac{g^2(\Theta)\cos^3(\Theta)}{\sin(\Theta)} \sigma^0(\Theta),$$

where

Fig. 4. SLAR images taken along river Danube as merged on a map

P_r — received power
P_t — transmitted power
G_0 — maximum antenna gain
λ_t — wavelength of transmitted signal
τ_t — pulse width of transmitted signal
β_{eff} — effective beamwidth of antenna in azimuth plane
$g(\Theta)$ — normalized antenna gain in elevation plane

The retrieval of the true σ^0 is called the radiometric correction, which is carried out by the sensitivity time control (STC) of the receiver. It is a signal generated in digital form by the control PC and fed into the SLAR after digital to analog conversion. The procedure results in a received signal level which is determined only by reflectivity of the observed pixel but not affected by other variables.

4.6. Digital Signal Preprocessing

The received analog signal corrected by STC is sampled in real time with a rate equal to the transmitted pulsewidth of 50 ns and converted into 8 bit values. It is then transferred into a fast intermediate memory. As the scanning of one line means the reception of the return signal in the window between 5 to 50 μs, there is 20 ms left for transcribing the values from the intermediate memory into the long term memory. It can take place at a much slower rate so the speed conversion allows the application of a PC-AT for the rest of the data processing tasks.

During data transfer many operations, such as coherent integration, speckle reduction, storing of flight-parameters can be performed on the data stream. Manual control of the whole system is through the keyboard of the PC.

4.7. Quick-Look Facility

Fast evaluation and selection of the received images can be done by viewing the image appearing on the high resolution display of the control computer. After landing of the airplane the complete data set collected during a flight can be stored in a ground based host computer where additional processing can take place.

4.8. Geometric Distortions and Corrections

A major type of distortion of a SLAR image is due to the difference of the slant range, measured between the antenna and actual pixel, and the horizontal range, measured between the observed pixel and the ground track of the airplane route. Elimination of this factor can be achieved by post processing of the images.

Geometric distortions are also induced by the irregularities in the motion of the aircraft resulting in changes of flying altitude, velocity and variation of the airplane angular position. With detecting the instantaneous height using the leading edge of the return signal and some more information provided by the navigation system of the aircraft, if it has any, improvement can be achieved with complex correction algorithms.

Interpretation could start when the complete post processing with the necessary correction are finished.

4.9. Airborne Operation

For initial experiments, the SLAR system was mounted on an AN-2 type airplane. The system installed consists of two subsystems:

- externally mounted antenna
- rack-mounted indoor units

In the indoor subsystem the microwave unit, the digital signal processor and the control PC-AT with display are included. The microwave unit is connected to the external antenna via a flexible waveguide and a rotary joint. The main data of the subsystems are listed in Table 1.

Table 1. Dimensions and weights of main units of SLAR

Subsystem	Dimensions [cm]	Mass [kg]
Internal racks with display	50×50×110	50
Power converter	35×35×20	20
Antenna	10×20×300	20

The side-looking antenna is mounted on the body of the airplane parallel with its longitudinal axis so as to offset the radiating surface of the slot line antenna at 25 to 30 degrees from nadir. Such pointing of the antenna results in a position of the fan-beam providing higher gain for return signals from far-range at a low grazing angle, and suppressing signals from near-range below the airplane. With an optimal offset angle, the antenna pattern contributes remarkably to the radiometric correction.

Power supply of the SLAR is provided by a DC/AC converter, converting the on-board DC voltage of 22 to 30 V to the required AC voltage of 220 V, 50 Hz (300 W AC system consumption).

The microwave radiation at the very surface of the antenna, 67 μW/cm^2, is lower than the limit of 0.1 mW/ m^2, allowing continuous work given by the Hungarian Standard MSZ 1626-86.

Table 2. Operating Characteristics of X-band SLAR

Operating frequency	f_0=9.4 GHz
Transmitted Peak-Power	P_t=3.5 kW
Transmitted Pulse-width	τ_t=50 ns
Receiver Bandwidth	B=16 MHz
Receiver Sensitivity	$P_r$min=−90 dBm
Antenna Azimuth Beamwidth	B_{eff}=0.01 rad
Effective Pulse Repetition Frequency	F_i=50 Hz
Velocity of Aircraft	v_r=50 m/s
Flying Altitude	h=1000 m
Average Geometric Resolution	$\Delta x\ \Delta y$=10 m×10 m
Integrated Samples/Pixel	N_i=8
Sampling Rate	f_s=20 MHz—8 bits
Number of Pixels/Row	N_r=512 or 640

4.10. Navigation Support

As the SLAR is capable of imaging the surface at poor weather conditions it is necessary to instrument-navigate the aircraft to the area to be surveyed. Navigation data recorded synchronously with the images are very useful for the geometric correction.

In an additional phase of the development, complete record of aircraft attitude, position, flight velocity, yaw,

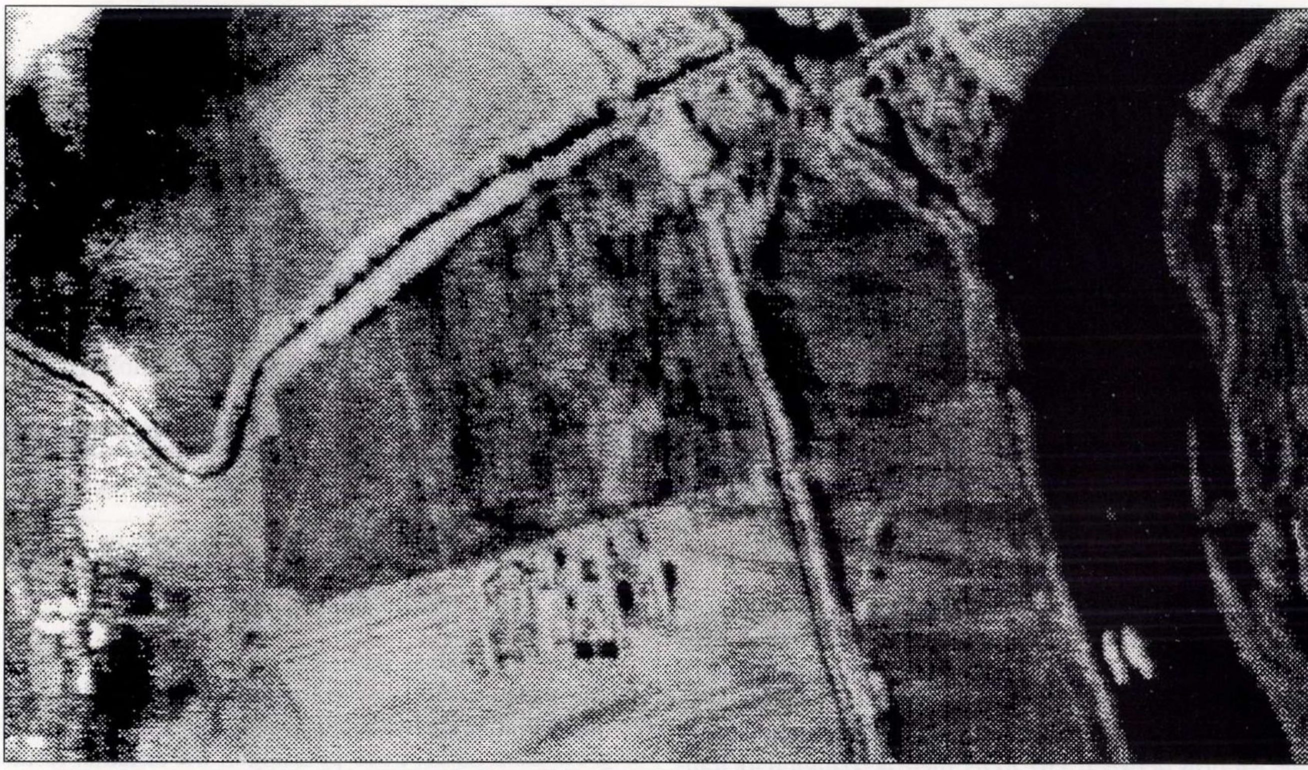


Fig. 5. SLAR image after average correction for near-range showing river Danube (right) and agricultural fields with roads (left)

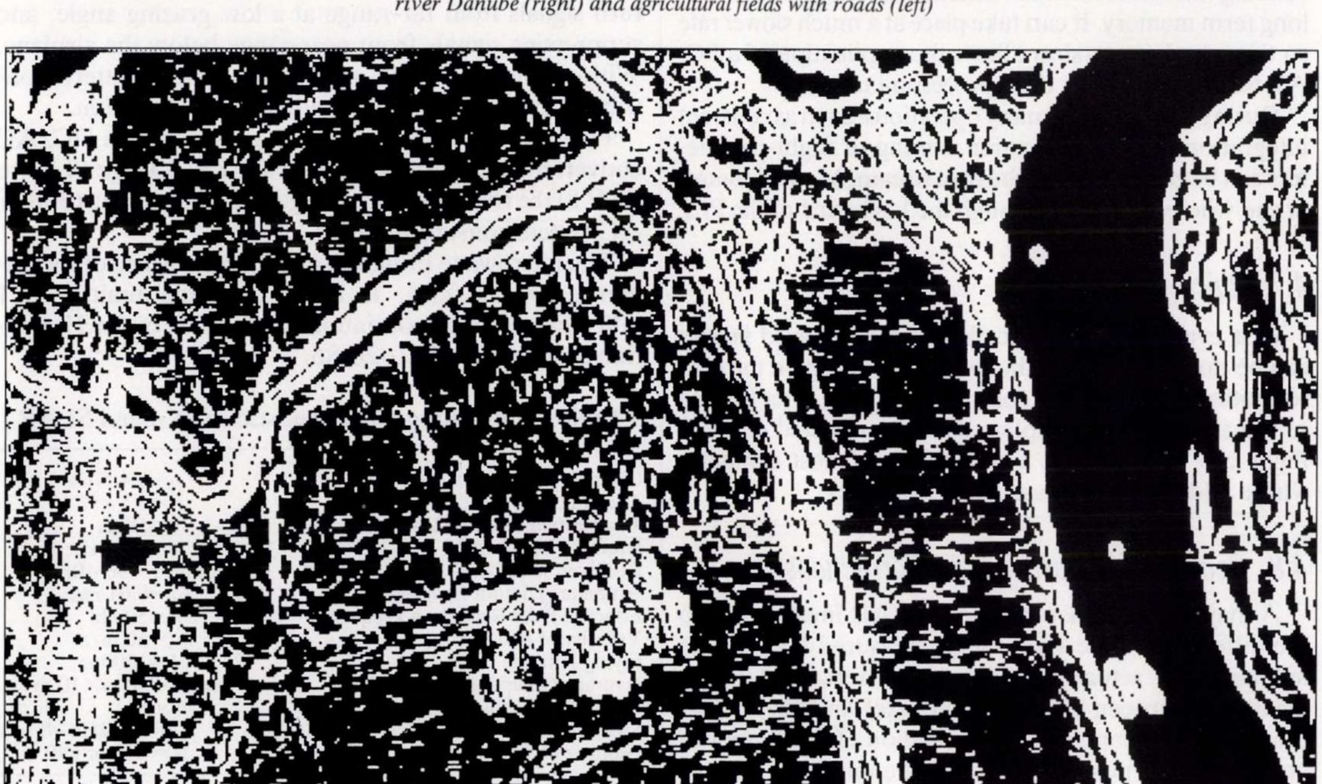


Fig. 6. The previous image after enhancing the edges

pitch and roll should be introduced. An LN-2000 GPS navigation system seems to be promising for this purpose.

During an interesting experiment, it could be tested how far a SLAR can be helpful for the navigator while he watches the weather independent radar image of the terrain below appearing on the screen of a display provided for this purpose.

4.11. Experimental Results

On 25 July 1990, the first experimental imaging with the SLAR was completed. There was no intention to check the effect of radiometric correction or adaptive control of receiver gain, so during the first flight, only uncorrected raw data were stored.

Fig. 4 shows the print of the first uncorrected SLAR images, taken in southward and northward flight-directions over the area lying between the two small villages of Dömsöd and Rácalmás, Hungary. The radar images have been merged on a simple tourist map, with no correction. Despite the uncorrected raw data, a very good coincidence is observed with the cultivated agricultural fields, water surface and channels, roads, etc. A rather heavy roll of the airplane caused an intensity variation of the image in the track direction, justified by the antenna pattern change as the airplane turned around its longitudinal axis. Naturally, this had no effect on the geometric structure of the image as the positions of the pixels are based upon the slant range between the target and the antenna.

4.12. SLAR Image Processing Experiments

Some initial processing experiments have been carried out on the images to test how far the information content of the images can be extracted.

The raw output image of the system is an eight-look averaged byte image in which each pixel density is represented by eight bits. The small details shown prove in practice the theoretically calculated resolution. Even two buoys can be resolved on the surface of river Danube beside other features as the river bank, which is cluttered with peninsulas, a channel and the rails, further agricultural areas with inhomogenities in the crop-fields.

As a first step of the processing, additional correction was performed in the near-range part of the image. This helped to eliminate the distorting effect of range and antenna pattern variations on the amplitude of the received signal as shown by the image of Fig. 5. It is seen that this was not completely successful because the antenna pattern results in a rather strong drop of the signal at low angles of incidence so the received signal was too weak to show significant components at some parts. In spite of this effect, the image shown in Fig. 6, which is the result of edge detection, shows details having been enhanced this way.

By using more adequate cross calibration with RCS data obtained by short-range scatterometry, the radiometric correction could be improved, yielding images with small radiometric errors [22].

5. CONCLUSIONS

Having been involved for the last decade in the field of active microwave remote sensing, an X-band truck mounted scatterometer and an X-band SLAR were developed, followed by field and in-flight measurements. The obtained results are encouraging, and could help to understand the information obtained by microwave remote sensing satellites (e.g. ERS-1, Almaz) in order to extend the research work towards more sophisticated sensors.

Not less important are the international cooperations based on application of the sensors.

Finally, the knowledge gained has allowed us to introduce microwave remote sensing into education at the Technical University of Budapest — however not in the sense of G. B. Shaw: Who knows makes, who does not, teaches.

6. ACKNOWLEDGEMENTS

The authors wish to thank the pilots performing the flights for their creative cooperation.

The financial support of the following organizations has to be thanked for:

National Commitee for Technical Development (OMFB),

Department of Agriculture (MÉM),

Institute of Geodesy, Cartography and Remote Sensing (FÖMI),

Ministry of Education (MM),

in projects 7—12—0127/OMFB, 978/1987/FÖMI and 3—531/1985/MM.

Image processing experiments are by courtesy of the Institute für Höchstfrequenztechnik und Elektronik, Universität Karlsruhe, Germany. Special thanks to Mr. Daniel Kähny and Professor Dr. Werner Wiesbeck.

REFERENCES

[1] F. F. Sabins, W. H. Freeman and Co., *Remote Sensing. Principles and Interpretation.* San Francisco. 1978.

[2] Cs. Ferencz, "Tasks and Problems of Modelling in Remote Sensing", *Communications,* in Hungarian, January 1990, pp. 9—15.

[3] F. T. Ulaby, R. K. Moore and A. K. Fung, *Microwave remote sensing.* Vol. I—II—III, Addison/Wesley, London, 1981—82—86.

[4] J. W. Trevett, Chapman and Hall, *Imaging radar for resources surveys.* London 1986.

[5] I. Bozsóki, É. Gödör and B. Farkas: "Microwave Remote Sensing", *Report for Remote Sensing Centre FÖMI,* in Hungarian, Budapest, Vol. I/1—2, 1983, 364 pages, Vol. II, 1984, 94 pages.

[6] I. Bozsóki, B. Farkas, É. Gödör, S. Mihály and R. Seller, "Information Content of Remote Sensing Data and their Evaluation", *Report to National Bureau of Technical Development, OMFB,* Vol. I—II—III, Budapest, 1986, 1987, 1989, 312 pages.

[7] G. Íjjas, "Aerial Survey of Soil Moisture by Airborne Microwave Radiometer", *Communications,* in Hungarian, December 1990, pp. 319—323.

[8] A. Farina, "Overview of SAR Systems", *Proc. of Space Meeting RIENA,* Rome, 1987, pp. 1—14.

[9] F. T. Ulaby and C. Elachi, *Radar Polarimetry for Geoscience Applications,* Artech House, 1990.

[10] S. Riegger and W. Wiesbeck, "Wide-Band Polarimetry and Complex Radar Cross Section Signatures", *IEEE Proc.,* May 1989, pp. 649—655.

[11] Z. Wirth, "Zusammenhang zwischen Huynen Parametern und geometrischen Eigenschaften von einfachen Objekten", Thesis for M. Sc., Institut für Höchstfrequenztechnik und Elektronik, Univ. Karlsruhe, Dept. of Microwave Telecom., Budapest Univ. Tech, 1990.

[12] L. Krul, "Scatterometer Systems", *Proceedings of an ESA-EARSeL Workshop,* Alpbach, Austria, 1981, pp. 19—27.

[13] A. Freeman and A. Currie, "Synthetic Aperture Radar (SAR) Images of Moving Targets", *GEC Journal of Research,* No 2, 1987, pp. 106—115.

[14] S. Mihály and I. Bozsóki, "X-band Scatterometry in Agriculture", *Proceedings of IGARSS,* Edinburgh, 1988, pp. 991—996.

[15] F. Völgyi, "High-Efficiency Microstrip Antenna Arrays", *Proc. of 17th EuMC,* Rome, 1987, pp. 747—752.

[16] S. Mihály, B. Farkas, R. Seller and I. Bozsóki, "X-band Scatterometry in SLAR Image Evaluation", to be published in the *Proceedings of IGARSS,* Helsinki, 1991.

[17] S. Mihály and N. T. Ngoc Minh, "Two-Dimensional Surface Generation with FFT", *Proceedings of 20th EuMC,* Budapest, 1990, supplement, p-4. 15.

[18] S. Zsámboki, I. Bozsóki, G. F. Remetey, K. S. Rao and G. Íjjas, "Airborne Microwave Remote Sensing Programme for Agriculture in Hungary. Aims and Present Results", *5th Symp. of WGRS,* Budapest, 1988, pp. 245—253.

[19] B. Farkas, R. Seller, S. Mihály and I. Bozsóki, "PC-Controlled X-Band SLAR", to be published in the *Proceedings* of IGARSS, Helsinki, 1991.

[20] L. Dávid and P. Szalai, "The State and Ways of Development of Education of Remote Sensing of the Earth", Evaluating study report in Hungarian, National Bureau of Technical Development, OMFB/23-8302-Et, 1984.

[21] I. Selmeczi, I. Bozsóki and T. Hetényi, "Speed-Converter for SLAR Video Signals", *Proceedings of IGARSS*, Amherst, 1985, pp. 1020—1026.

[22] S. Mihály and B. Farkas, "Active Microwave Reference Measurements", *Remote Sensing Applications* in Hungarian, MTA RKK Békéscsaba, 1987, pp. 126—135.

István Bozsóki

graduated in electronic engineering in 1953. Until 1956, he was teaching at the faculty of Military Engineering at the Technical University of Budapest, then at the Department of Microwave Telecommunications. He has become an Associate Professor and was head of the department between 1981 and 1990. He was involved in the design of microwave circuits between 1961 and 1967 at the Beloiannisz Co. and at the Dept. of Microwave Circuit Development at ORION. He received the Dr. Univ. degree in 1965 and has been the Candidate of Techn. Sciences since 1976. His main fields of activity were parametric amplifiers, microwave oscillators, theory and practice of radio communications and radar techniques. Currently, he is involved in research of microwave remote sensing. He is member of several technical committees. He got the Puskás Tivadar award in 1986, in addition to several further awards.

Botond Farkas

graduated in communications engineering in 1965. He became an assistant professor, then a senior assistant of the Department of Microwave Telecommunications at the Technical University of Budapest. His main fields of activity are development of microwave circuits and systems, microwave measurement systems, radars and microwave remote sensing.

Éva Gödör

graduated in telecommunications engineering in 1961 and received the specialized microwave engineer degree in 1966. She got her Dr. Univ degree in 1971. She is a senior assistant of the Department of Microwave Telecommunications at the Technical University of Budapest. Her current education and research activities are related to microwave communications systems and microwave remote sensing systems.

Sándor Mihály

graduated in 1984 in communications engineering. He was a full time postgraduate student until 1987 at the Department of Microwave Telecommunications at the Technical University of Budapest where he has been an assistant professor since 1987. He is involved in the development of active microwave sensors, and his education activity includes the theory and practice of microwave radio measuring systems.

Rudolf Seller

graduated in communications engineering in 1987. After graduation, he became a research fellow at the Department of Microwave Telecommunications where he has been an assistant professor since 1990. His current field of activity includes the general problems of active microwave remote sensing and the development of hardware and software for microwave radio measurement systems and radars.

NEW GENERATION OF HEWLETT—PACKARD NETWORK ANALYZERS

The HP 8510C is the next significant step in a seven-year evolution of microwave network analyzer technology from Hewlett—Packard. The HP 8510C *(see Fig. 1.)* gives breakthrough RF performance to 50 GHz in coax, and advancements in system measurement speed an usability.

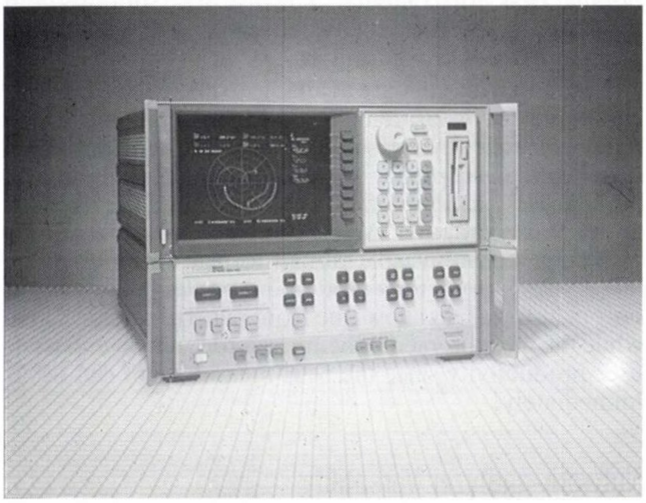


Fig. 1. The HP 8510C Microwave Network Analyzer

In one continuous sweep, from 45 MHz to 50 GHz, the HP 8510C and its companion HP 8517A S-parameter test set deliver RF performance and dynamic range beyond what has been available previously to 40 GHz.

The HP 8510's built-in, high speed computer provides the capability to characterize and effectively remove the impact of systematic errors through accuracy enhancement techniques. Effective directivity and source match can be improved to as much as 60 dB. The data processing speed of the system is such that a fully error-corrected, 401 point trace of data is updated in under one second. This virtual "real time" display of error-corrected data means that a test device can be easily adjusted while it is being measured.

Powerful measurement enhancement functions are also available. Data averaging can be employed to narrow the effective receiver IF bandwidth, extending dynamic range and increasing signal-to-noise ratio.

Built-in storage provides the capability to save and recall up to eigth different front panel states, eigth several calibrations and eight separate measurements in nonvolatile memory.

Time Domain Analysis

The time domain capability (available as Option 010 on HP 8510C) converts frequency domain data to the time domain by means of the inverse digital Fourier transform. Thus same time domain data can also represent the distance domain if the relationship between time and distance is known. Time domain is a powerful tool that allows further insight into the various elements of a circuit which can be studied as a function of location. With the gating function, the unwanted responses outside the gate can be removed. These responses can be "filtered-out" in the time domain and the gated responses can then be viewed in the frequency domain.

"No-Compromise" RF Performance to 50 GHz

Until now, with microwave network analyzers, a tradeoff existed between RF performance and frequency range *(see Table 1)*. Above 26.5 GHz, the benefit of gaining extra frequency range had been at the expense of overall system accuracy and stability. These two important areas are impacted by uncorrected (raw) system RF performance.

Table 1. Frequency/Performance Tradeoff with Existing 26.5 GHz and 40 GHz Test Sets

Test Set	*Frequency Range*	*Raw Permance Directivity*	*(Max Frequency) Port Match*
HP 8515A	*45 MHz to 26.5 GHz*	*26 dB*	*14 dB*
HP 8516A	*45 MHz to 40 GHz*	*11 dB*	*8 dB*

The exceptional raw RF performance delivered by the HP 8517A test set, with better than 20 dB directivity and 16 dB port match, means that measurement calibrations will last days instead of hours. It also means that the system will produce better measurements when hardware such as cables, adapters, fixtures and wafer probes are placed between the device and the test set (somethings virtually everyone must use for real-world measurements).

It is a commonly misunderstanding that only the characteristics of the additional lossy hardware determines uncorrected performance such as system directivity. In an error-corrected system, the sum of all the "errors" in the system are quantified during the calibration process. For example, the directivity of the system will be determined by the sum of test set directivity, cable loss and adapter return loss. Minimizing the test set raw directivity reduces the system's sensitivity to changes induced by environmental effects. Higher overall system accuracy and stability result in higher confidence in the measurement results.

Another key RF performance factor is system dynamic

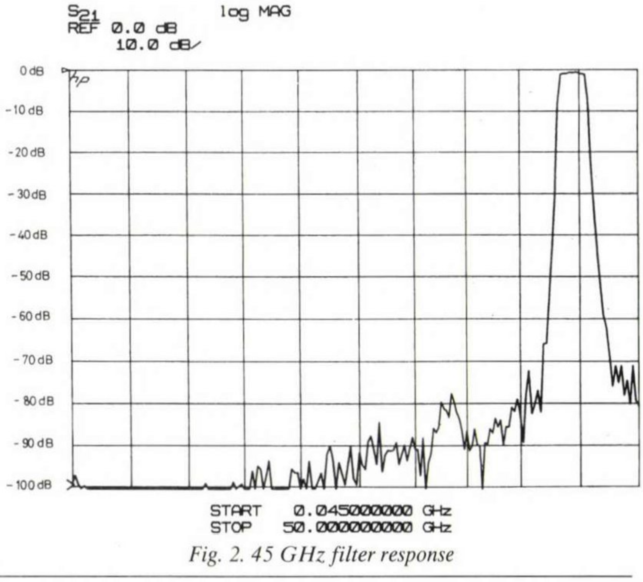


Fig. 2. 45 GHz filter response

range. The HP 8517A 50 GHz test set also sets a new level of achievement here. Depending on the test set used, 80 to 100 dB of dynamic range is available *(see Fig. 2)*.

Faster Measurement Updates with Higher Frequency Accuracy

The HP 8360-series synthesized sweeper is a powerful new engine for the HP 8510C. Its analog "ramp" sweep mode is unmatched when measurement speed is important, such as when tuning the response of a microwave component. A fully error-corrected, 401-point measurement can be updated in less than a second. The swept frequency accuracy of ramp sweep mode has been extended by an order of magnitude to 0.1% of span. This means the fastest sweep mode is now a viable mode for more applications, such as production final test, resulting in greater throughput and productivity.

For the ultimate in frequency accuracy, the "step" sweep mode provides synthesized frequency of each measurement point with 1 Hz resolution. Overall speed in the step sweep mode is improved by continuously ramping the frequency between measurement points. This eliminates the time needed to re-phaselock the system at each point. This reduces the time to move between frequency points to under 10 msec. Now a fully error-corrected 201-point measurement is updated in less than 4 seconds, which is four to six times better than with previous systems.

Corrected Power Flatness at the Test Port

The HP 8510C is the first microwave vector network analyzer capable of delivering flat power vs frequency at its measurement port. Using the HP 8360's unique user flatness correction feature makes this possible. This provides constant power at the measurement port for each frequency. With this capability, it's now possible to measure the compression level of active components. Since the measurement port is at a known, constant power level, it can be used to quantify or "calibrate" the transmission return path of the network analyzer test set. This means absolute power can be measured directly with the network analyzer for the first time.

Full Color Display and other Features Enhance Usability

With the large (19 cm diagonal) color CRT all four S-parameters of a test device can be viewed simultaneously on a single full-size graticule in a unique "overlay" format. The "split" format can be used to view all four S-parameters on separate cartesian, smith or polar graticules. The active marker value for each S-parameter can also be displayed simultaneously, allowing the user to instantly see the results at critical frequencies. Also, all five marker readouts for a single S-parameter may be viewed at the same time. With 16 customizable colors to choose from, it's easy to differentiate between different measurement traces and other display features such as the graticule, memory traces, reference lines, markers and annotation.

With the built-in 3 1/2-inch disc drive, measurement data can be stored for eventual transport to just about any popular desk-top computer. The disk drive supports both the LIF (for HP 9000 series computers) as well as the popular MS-DOS file formats.

Numerous features are available to make the job of documenting the measurement results easier and more productive. Hardcopy measurement results can be produced on either a direct plotter or graphics printer, including the popular HP LaserJet series printers, without the need for an external computer. Measurement results can also be printed in a customizable tabular format. Additional serial (RS-232) ports for a plotter and printer come with an output buffer. This allows the front panel to remain active while the measurement is being recorded which enhances user productivity since the next measurement can occur immediately.

Network-analyzer Versatility

The HP 8510C is HP's most versatile network analyzer. Test sets, software and custom-configured systems are available to characterize the magnitude and phase of transmission, reflection coefficients and group delay of many categories of components. Measurement applications include:

- passive and active devices such as filters, cables, connectors, amplifiers;
- on-wafer testing of transistors and GaAs MMICs;
- frequency translation devices such as mixers;
- pulsed-RF components (HP 85108A Pulsed-RF Network Analyzer System);
- millimeter-wave measurements (with the HP 85106A Millimeter-wave Network Analyzer Subsystem);
- light-wave components such as laser diodes, optical fiber and photodiodes;
- antenna pattern and radar cross section (RCS) measurements (HP 85301A Antenna Measurement System);
- dielectric measurements of materials.

Information from Hewlett—Packard

INTELSAT EARTH STATIONS FROM SCIENTIFIC ATLANTA

Introduction

Scientific Atlanta is headquartered in Atlanta, Georgia, USA and employs 3400 people. Scientific Atlanta designs, develops and manufactures products and systems to address the markets in four business sectors:
- Satellite Communications,
- Cable Television systems,
- Instrumentation,
- Government systems.

In 1990, the company reported record performance in sales, earnings and new orders. Sales for 1990 were $614.3 million, up 12 percent over 1989. In 1991, Scientific Atlanta expects to exceed all these results.

Scientific Atlanta is a full line original equipment manufacturer (OEM) of satellite communications equipment, building and integrating turnkey satellite earth stations from INTELSAT standard A through Z. Scientific Atlanta is manufacturer of antennas (1.8 to 18 meter), LNAs, up/downconverters, modems, video and message receivers and exciters, monitor and control systems, etc. Consequently, Scientific Atlanta builds earth stations from end to end, providing near complete quality control over the entire earth station. On the other hand, systems integrators without the same manufacturing capability must rely on the quality of products delivered by their suppliers. Scientific Atlanta relies on itself, thus offering the customer the assurance that all equipment and systems are fully tested and reliable.

The total systems management approach established Scientific Atlanta as the number one provider of satellite communications. This approach has been applied once again in the corporation's development of its large earth station systems. Due to its end-to-end systems management concept, Scientific Atlanta represents a single interface to customers.

In conclusion, Scientific Atlanta's telephony engineering experience and technical expertise, coupled with extensive satellite experience and financial strength, provide a unique combination to deliver an efficient, cost-effective and reliable satellite communications system, tailored specifically to meet the needs of the Hungarian INTELSAT earth station.

Earth stations in operation

Scientific Atlanta is the world's largest supplier of earth stations and equipment. To date, Scientific Atlanta has to its credit over 47,000 installed operational sites, located in over 120 countries. Notable among its accomplishments are the recent A station contracts in Oman, Papua New Guinea, Gabon, Guam, and other earth stations in the United Kingdom, Greece, Cyprus, the United States, Canada, Mexico, Chile, Indonesia, Africa, India, Europe, Thailand, and the Philippines.

Following the revised INTELSAT standards for Standard A Stations, Scientific Atlanta has entered this market as a prime contractor turnkey supplier. This is primarily due to the new standard more closely adhering to digital specifications, an area in which Scientific Atlanta has considerable experience as a leading supplier and manufacturer of INTELSAT specified satellite communication equipment. It has supplied and installed INTELSAT based Domestic Satellite Networks (DOMSATs) around the world, maintaining a close working relationship with INTELSAT, with a long history of design experience dating back to the design of the first transportable earth stations for the Early Bird satellite.

For many years in the late 1970s and early 1980s, the major elements in a large A station were the 32 meter antenna, and FDM/FM Ground Communication Electronics (GCE). Scientific Atlanta's FDM/FM equipment was the world standard delivered to a large number of international systems integrators. Now that the very large antennas are no longer required, Scientific Atlanta has become a full line manufacturer and turnkey system supplier of INTELSAT Standard A earth stations.

Following is a list of Scientific Atlanta A-stations and related equipment.

Gabon — 18-Meter A Station

In this French Equatorial African Nation, Scientific Atlanta provided, in formal competition with French and Japanese suppliers, an 18-meter standard A station on a turnkey basis, utilizing its GCE equipment for international voice links between Gabon and Europe. This station has recently been commissioned and is joining the worldwide INTELSAT network. Scientific Atlanta also provided a 15-meter master station in Libreville supplying voice and video for the DOMSAT network.

Oman — 21-Meter A Station

Scientific Atlanta is currently under contract to provide a 21-meter INTELSAT Standard A station in Oman. After selection in competition against 8 international competitors, Scientific Atlanta is supplying IDR equipment, DCME, civil works and installation for this turnkey digital A station.

Guam — 15-Meter A Station

Scientific Atlanta has started work on this station after winning in an international tender competition. This will be a digital A station including IDR equipment.

Papua New Guinea — A Station

Scientific Atlanta won the contract in competition with many international suppliers. This will be a 15.5 meter Standard A digital station with IDR equipment.

Algeria — A Station Retrofit

The operating A station in being retrofitted with digital products, including Scientific Atlanta's advance computer earth station controls and control room electronics.

United Nations — Voice Communication Network

For the U.N.'s voice communications network, Scientific Atlanta provided and installed the international gateway station at the U.N. headquarters in New York City. This gateway links the U.N. with its European Offices and U.N. forces in the Middle East, Africa, and Asia.

AFRTS — Global Network

The Armed Forces Radio and Television Service (AFRTS) of the U.S. Government has a global satellite network similar to that of the U.N. For this, Scientific Atlanta has supplied uplinks and downlinks for the worldwide distribution of video.

Latin America — Digital Gateway Stations

Scientific Atlanta is currently under contract to provide international gateway stations for IDR services in Guatemala and El Salvador to communicate with the U.S. via digital links.

Italy — 18-Meter Station

This 18-meter C-band international voice station includes Scientific Atlanta's computer controlled tracking, RF and telemetry electronics.

Design features

Scientific Atlanta has a long history of developing excellent system designs which are tailored to each customer's specific needs, and optimized relative to the functions and parameters that are most valued for the customer's particular application. In attempting to provide this service, the following design goals have been identified:

- Design an excellent quality earth station which meets customer's requirements for IDR trunks while providing exceptional reliability and availability.
- Assure that the final system design provides ample room for future traffic expansion, without substantially increasing initial acquisition costs, and provide for the smooth, eventual addition of other IDR carriers, modems, converters and DCME (Digital Circuit Multiplication Equipment).
- Use 1 : N protection switching to the maximum extent possible within the availability constraints in order to achieve full performance at acceptable cost.
- Apply a computer-based monitor and control system, based on Scientific Atlanta's Earth Station Controller tailored to the specific earth station design, assuring the maximum amount of control and information flow while remaining versatile and user friendly.
- Assure that the AOR earth station design takes into account the eventual addition of the IOR earth station such that maximum advantage can be obtained from their co-location. Stations for the AOR and IOR satellite were developed together so that AOR station equipment can easily be expanded, at extremely modest cost, to accomodate the needs of the IOR station. Further, the existence of the AOR station has the effect to greatly reduce the spares and test equipment needed for the IOR station, a sharing arrangement being provided by the co-location of the two systems. The fact that the two stations utilize identical equipments results in an advantage which will substantially reduce overall life cycle cost.

Design evolution and trade-offs

Many of the satellite earth station design concepts, pioneered in the early to mid 1960's, continued to dominate the industry well into the 1980's. With the advent of a mature solid-state technology, plus the advancement of microcircuits to aid the performance of satellites, earth station requirements have become much easier. The INTELSAT IESS-201 specification requirements for Revised Standard A Earth Stations can be met with antenna diameters between 15 and 18 meters. Low-noise amplifier performance has improved to the point where noise temperatures between 40 K and 60 K are achieved with thermoelectrically cooled or even uncooled GaAs FET LNAs with High Electronic Mobility Transistor (HEMT) front ends. Cryogenic cooling, or the use of parametric amplifiers, is virtually unheard of, and HPA sizes have shrunk to 400 to 700 watt range TWTs, for all but TV applications, bringing along the important advantage of 575 MHz bandwidths as well. With the single exception of the HPA output stage, all subsystems are now solid-state, with a concomitant extraordinary improvement in reliability that now makes the use of 1 : N protection switching the only sane approach to provide redundancy. For example, while a typical MTBF for a cryogenically cooled parametric amplifier might be in the 5000 to 10,000 hour range, a modern GaAs FET—HEMT LNA MTBF will typically be between 150,000 to 250,000 hours. The ability to achieve excellent availability performance with 1 : N protection has been a major contributor to the continuing reduction in the cost of INTELSAT A stations. This technology advancement is only slightly less important to A station design than the advent of microelectronic circuitry.

As a result of technology changes in recent years, the critical design trades have been reduced to:

Antenna Size vs. LNA Temperature vs. HPA Power

Given the selection of antenna size to be 18 meters, G/T can be met with 40 K to 50 K uncooled low-noise amplifiers. EIRP can be met, for carriers per HPA which is 2 more than used in the design, with TWT HPAs. The system redundancy design uses 1 : 2 upconverter switching, 1 : 3 downconverter switching, 1 : 5 demodulator switching, and 1 : 5 DCME switching.

The redundancy switches selected will inherently accomodate the following future expansion:

Upconverter	+2	Demodulator	+3
Downconverter	+1	DCME	+2
Modem	+3		

The TWT high-power amplifiers can each accomodate up to 5 IDR carriers for a total of 10 carriers distributed between two polarizations. The design uses only 5 carriers total, so that 110 percent expansion is possible without adding HPAs or increasing their size.

DCME circuits are lightly loaded: with a full capacity of 150 voice circuits per DCME, only 100 are used on each of the two USA and one Canadian DCMEs. Two additional DCMEs with 150 voice circuits each are also available for expansion. The DCME circuits have been designed as multi-destination links, each link having one transmit modulator plus two receive demodulators. One ESC circuit for each modulator and each demodulator is also provided, and each 2048 Mbit/s trunk at the DCME/terrestrial interface is provided with echo cancellation.

Scientific Atlanta believes that these design concepts offered are extremely efficient, and provide high traffic capacity plus exceptional station and carrier availability at a modest cost. Much of this efficient system design is based on the excellent quality and performance of Scientific Atlanta's equipment and that of a few vendors whose standards are required to be no less than those of Scientific Atlanta.

Standard A station characteristics

Typical standard A station characteristics provided by Scientific Atlanta are as follows. Provision of IDR service, including ESC interfaces, according to INTELSAT

IESS-308 (R-5) by a Revised Standard A Earth Station per IESS-201. As an option, television transmission and reception in NTSC, PAL and SECAM formats per IESS-306 are also provided.

The station is equipped with an 18-meter C-band antenna having a 4-port circular polarization, 1.06 : 1 axial ratio feed, limited motion kingpost 180 degree azimuth pedestal, and smoothtrack/steptrac/program track automatic tracking servo that uses a trajectory estimation algorithm (TEA).

The IDR modems and demodulators are protected by two 1 : 5 automatic protection switches which are capable of expansion to include additional modems and additional demodulators. Converters are 1 : 2 protected for two uplink carriers and 1 : 3 protected for three downlink carriers. The IDR HPAs are TWT amplifiers in a 1 : 2 protection switch.

The optional TV subsystem includes one equalizer per exciter to provide for satellite equalization. The TV uplink utilizes 1 : 1 redundant 3 kW klystrons which are hybrid combined with the IDR HPAs for transmission on either of two polarizations. TV reception is provided by a 1 : 1 protected TV receiver, and TV transmission is achieved via a 1 : 1 configuration utilizing Scientific Atlanta's world standard TV exciter.

Low-noise receivers use two 40 K uncooled GaAs FET LNAs and a 1 : 2 LNA protection switch. The Scientific Atlanta protection switch monitors LNA status and provides the switching and control logic for protection.

A test translator is provided to convert a 6 GHz transmit sample to 4 GHz for insertion to the off-line LNA which has an output returned to the console RF panel. In addition, LNA input and output couplers provide for test and evaluation of the off-line LNA from the equipment room. Station test equipment used with these LNA couplers make it possible to perform gain/bandwidth measurements on the off-line LNA.

The addition of an uninterrupted power supply (UPS) tends to insulate an earth station from a wide variety of power fluctuations as well as the more obvious power failures. Consequently, Scientific Atlanta normally advises the use of a UPS, especially one designed around modern, high speed microprocessor technology. Such a UPS offered as an option to assure that earth station availability performance is preserved, even in the presence of unanticipated power anomalies.

Scientific Atlanta in Hungary

Scientific Atlanta decided to submit a proposal for the Hungaria INTELSAT standard A earth station tender, issued by the Hungarian Telecommunication Company.

Scientific Atlanta has selected the Hungarian Research Institute for Telecommunication TKI as partner to support its proposal. This partnership would be capable to quickly address any issues or needs in Hungary during the implementation stage of the earth station, and world provide responsiveness to the customer's needs.

This Partnership is justified by the facts that Scientific Atlanta is committed to being the worldwide market leader in the supply of international earth stations and satellite networks, and TKI is committed in satellite communications engineering and support services in Hungary.

In the system design, ease of future expausion has been considered as a primary goal. For example, when the time comes to add modems or converters for the IDR links, it will be found that supporting equipment is already in place in the system, and that expansion may be accomplished with virtually no loss of traffic. Expansion is especially easy because of the use of 1 : N switching plus numerous spare HPA input combining ports.

The Scientific Atlanta modem and satellite converter subsystems, offered to support the IDR links have recently been selected also for the largest domestic, all-digital satellite communications network in the world by Compañia de Telefonos de Chile. This program is presently in the production phase, and will result in the delivery of 123 modems and 51 converters, along with associated protection switches for both. The existence of such a large program offers the unique advantage of a large production capability, operating near the peak of its efficiency.

Many of the equipments selected for use are manufactured by Scientific Atlanta to exacting corporate standards. Scientific Atlanta guarantees to its customers that they are receiving the highest quality equipment, and that most of that equipment has been tested by years of service in existing earth stations. This guarantee is only possible because of the many years during which Scientific Atlanta has been the foremost supplier of earth stations in the world, providing full performance at a relatively modest cost.

W. T. Carlin

■ EUREKA — TOGETHER FOR THE FUTURE

The Eureka initiative was launched in 1985 by 19 EC and EFTA countries and the Commission of the European Communities to raise the productivity and competitiveness of Europe's industries and national economies on the world market through closer cooperation among enterprises and research institutes in the field of advanced technologies. The EUREKA framework is built to harness the dynamism and innovative strength in Europe's industry and research. The ground rules provide a simple set of criteria for establishing EUREKA projects. The projects must:

- involve at least two partners from different EUREKA countries
- aim at securing a significant technological advance in the product, process or service concerned
- aim at applications in the civilian sector.

The success of the EUREKA program is shown by nearly 400 projects wih a total cost of more than 7,5 billion ECU involving some 2000 participating organizatione.

Considering rapid political and social changes the EUREKA Ministerial Conference in 1990 has strongly supportod the (increased) participation of companies and research institutes from Central and Eastern Europe in the EUREKA projects. To promote this idea in May 1991 a congress was held in Budapest under the auspices of the Netherlands EUREKA Chairmanship and the Hungarian government. More than 600 participants — representing about 100 companies, and research institutions as well as political and administrative corporations, — were exploring the conditions for promoting cooperation within the EUREKA framework. The speakers of the plenary session provided key information on the cooperation mechanisms. Specific industrial and technological opportunities were discussed in three parallel sessions:

- information and communication technology
- biotechnology
- enviromental and energy technologies.

Special attention was paid to the experience gained by companies and institutes from Central and Eastern European countries already participating in eight EUREKA projects.

Hungarian institutions are involved in three of these projects. The Research Institute of Computor Science and Technology (SZTAKI) is developing software modules in the FAMOS-ALCUT program for flexible design and qualitiy control for the leather industry (EU 335). The Central Research Institute for Atmospheric Physics is working in the EUTRAC program (EU 7). A new small private enterprise, SEMILAB Co. is a partner for Portuguese and Spanish companies in the qualitiy control of solar cell production (EU 366). As a perspective cooperation the participation of the Technical University of Budapest and the Central Research Institute of Physics in near to be acknowledged in the basic research subprogram of JESSI (Joint European Submicron Silicon, EU 127). Further cooperation possibilities were explored at partner search sessions. During the congress the EUREKA Database containing various informations on existing projects was directly acessible through tho ECHO computer in Luxemburg.

All this can help to include the existing technological skills of the Central and Eastern European countries into the European Economic Space and to further improve it by cooperation in the EUREKA projects. As a first step, the information on the EUREKA framework has been successfully disseminated at the congress. The second step of organizing the national EUREKA offices in these countries has to follow, and a third step of providing more flexible rules for the non-member countries is also necessary. This will be proposed at the next Ministerial Conference by Dr. Andriesse, Minister of Economic Affairs of the Netherlands. He expressed his opinion that Central and Eastern Europe can offer

- markets with growing potential
- skills and abilities in research and development
- low cost labour (at least for the time being).

He added that cooperation in environmental issues is of immediate mutual interest.

A key speaker of the congress Dr. Agnelli, Fiat SpA, Italy, emphasized the necessity of promoting cooperation projects with Eastern companies for protecting the most valued heritage of Central and Eastern European research centres. Professor Pungor. Hungarian Minister without portfolio, President of the National Committee for Technological Development, expressed strong Hungarian commitment for expanding scientific and technical cooperations. ■

■ SWEDISH PLANS, HUNGARIAN HOPES

The editor of the Journal on Communications asked Mr. Ulf Sandberg, Hungarian representative of the Ericsson concern, and Mr. István Fodor, managing director of the Swedish—Hungarian joint venture company Ericsson Technics Ltd. to give an interview about Hungarian involvments of Ericsson.

Mr. Sandberg stated that the establishment of a suitable telecommunications infrastructure is of utmost importance for the Hungarian economy. Consequently, it was a right decision to choose two leading European manufacturing companies, Ericsson and Siemens, to deliver the digital main exchanges to be installed in Hungary, thus qualifying the Hungarian telecommunication network to become an integral part of the European telecommunication system. Due to the central location of Hungary, it will then be possible to extend its telecommunications infrastructure well beyond the Hungarian borders. Hungary, as a transit country for telecommunications, would thus share the profit originating from international telephone traffic. It is a challenge for Ericsson to participate in this development work at all levels of the telecommunications network.

Speaking about the five AXE digital main exchanges scheduled for Hungary by Ericsson this year, Mr. Sandberg noted that these will be installed in downtown Budapest and in the towns of Szentendre, Komárom, Tatabánya and Biatorbágy. The 20.000-line Budapest downtown switch will have a remote subscriber stage with 2000 lines in the suburb of Ferencváros. Ericsson has also been awarded the task of extending the International Exchange and the Mobile Swithing center in Budapest. In addition the Rural Areas of Gyöngyös and Mohács are to be supplied with AXE switches. The exchanges to be delivered to Hungary will represent the latest generation of the AXE family, both from the hardware and the software aspect.

The operation of the new Ericsson exchanges will require considerable expertise, and Mr. Sandberg pointed out the strategic importance of expert training. Ericsson maintains six training centres worldwide, two of these being in Europe, Stockholm and Dublin. Training covers the complete product range of Ericsson, starting at the technician's level and going up to the system engineer's level. The direct contact with the end-user is of extreme importance for both parties concerned. In 1991, 50 people form the staff of the Hungarian Telecommunications Company will receive AXE training at various levels in Dublin. Ericsson is also willing to help Hungarian experts obtain the level of expertise needed for applying future high technology systems such as ISDN.

Mr. Fodor said that the staff of the company Ericsson Technics Ltd. will be presently trained in Budapest by Swedish specialists. As for investment planning, no final decision has yet been reached about the location of company headquarters. However, a computer aided planning procedure concerning the layout of assembly shops and laboratories, further the concept of financial and economical management and the structure of the staff is already being elaborated in Stockholm. Thick folders are pouring in from Stockholm every week, requiring an intensive effort of domestic processing and evaluation. Hungarian amendments are duely acknowledged by the Swedish party, resulting in a creative and democratic type of cooperation. Up to the end of 1993, the AXE program offers following job opportunities for the Hungarian technical staff:

marketing	15 people
engineering	100 people
hardware manufacturing	100 people
software development	100 people

The software house in Budapest will be responsible not only for developments for the Hungarian application but also for further development of software packages related to AXE subsystems. The first phase of Hungarian manufacturing is scheduled for 1991. Following 1992, the scope of Hungarian activities will be extended mainly by manufacturing of magazines and system final testing, but it would now be premature to go into details on this issue. ■

■ IEEE ORGANIZES INFORMATION THEORY SYMPOSIUM IN BUDAPEST

The Information Theory Society of the Institute of Electrical and Electronics Engineers (IEEE) sponsors the IEEE INTERNATIONAL SYMPOSIUM ON INFORMATION THEORY
at the Budapest Convention Centre in Budapest, Hungary, from Monday, June 24 to Friday, June 28. This Symposium provides an opportunity for the dissemination of recent advances in Information Theory in the form of long papers (40 minutes) and short papers (20 minutes). In addition, four plenary lectures will be given by invited, distinguished speakers, and a special invidet session in honor of A. Rényi will be organized by I. Csiszár. The Shannon Lecture, a regular feature of these Symposia, will be delivered this time by A. J. Viterbi on "Shannon Theory from a Communication Engineer's Perspective."

The plenary lectures of the symposium are as follows:

Körner: "A Generalized Shannon Theory for Zero Error"
Ornstein: "Universal Data Compression from Ergodic Theory Point of View"
M. Kunt: "HDTV"
R. Grahan: "Aspects of Randomness in Graphs and Hypergraphs"

The programme of the symposium contains the following sessions:

Shannon Theory
Rényi Invited Session
Error Correcting Coding
Error Correcting Constrained Codes
Constrained Codes
Codes and Algebraic Geometries
Convolutional Codes
Coded Modulation
Storage Channels
Estimation
Random Processes
Signal Processing
Distributed Information Processing
Neural Networks
Source Coding
Noiseless Source Coding
Quantization
Vector Quantization
Image Coding
Cryptography
Sequences
Data Networks
Multiple Acess and Data Networks
Multi-User Comunications
Communications Systems
Pattern Recognition
Detection
Optical Communications
Information Theory Applications

Informations on the symposium can be received from
Prof. L. Győrfy,
Technical University of Budapest
phone: 361 166 5824

■ TWO INTELSAT EARTH STATIONS AND A MICROWAVE LINK — A HUNGARIAN TENDER

The Hungarian Telecommunications Company and the ELEKTROIMPEX foreign trading company issued a tender inviting bids to deliver

- a gateway Atlantic Ocean earth station,
- a gateway Indian Ocean earth station,
- a digital microwave link.

Both new earth stations, to be installed in the vicinity of the Hungarian INTERSPUTNIK earth station already in operation, have to meet the requirements applicable for INTELSAT Standard A earth stations, i.e. G/T= =35 dB/K. The Atlantic Ocean station is intended to serve public voice and data traffic, and during the initial stage, 220 voice circuits to the USA and 110 voice circuits to Canada will be put into operation. Subsequently, additional modems will be operated to meet increasing traffic requirements. 2 Mbit/s IDR (Intermediate Digital Rate) carriers will be operated during the initial stage, with the possibility of conversion to IBS (International Business System) carriers. To increase channel capacity, traffic will probably be routed via multi-destinational DCME (Digital Circuit Multiplication Equipment). Further tender requirements called for engineering service circuit and monitor control equipment, and for the optional supply of an additional two-way TV channel.

The microwave link is intented to connect the earth stations with the town of Győr, branching there in the direction of Budapest and Vienna. The link will operate in the 6 GHz frequency range at a transmission speed of 140 Mbit/s over one operating and one stand-by channel, with the possibility of increasing the number of working channels at a later date. The tender also called for a service telephone channel, and for remote control of the repeater stations.

The complete project will be financed by the National Bank of Hungary by using a loan from the International Bank of Reconstruction and Development.

A pre-bid conference took place in February 1991 at the Hungarian Telecommunications Company for answering written and verbal questions of the bidders who were invited to inspect the site of the prospective earth stations near the village of Taliándörögd. Deadline of bid submission was the 15th of April, 1991 when the sealed bids were opened in the presence of the bidder's representatives at the office of ELEKTROIMPEX. The three items of the tender call will be separately evaluated by independent groups of experts, with following main aspects of evaluation:

- compliance with the tender specification,
- equipment price,
- fraction of Hungarian participation in the project.

Bids were submitted by following companies:
ALCATEL (France)
COMSAT (USA)
EB Nera (Norway)
Fujitsu (Japan)
GTE (USA)
Hispanet (Spain)
JVC (Japan)
MIK-CO (Hungary)
NEC (Japan)
OTC (Australia)
RSI (USA)
Scientific Atlanta.(USA)
STS (USA) ■

■ CEPT GOES EAST

The Conference of European Postal and Telecommunications Administrations (CEPT) held its XIVth ordinary session of Plenary in London during the period 26 September to 2 October last year, under the Chairmanship of Mr. Paul Salvidge of the Department of Trade and Industry UK.

The session celebrated an important landmark in its history by increasing its membership from 26 countries to 31. With the admission of Hungary, Poland, Czechoslovakia, Bulgaria and Romania and with the reunification of Germany, the population in CEPT member countries will increase by over 25%. This impressive expansion of CEPT will open up major opportunities for the development of communications throughout Europe.

During this Plenary Session the two Commissions changed their names to CEPT Posts and CEPT Telecoms respectively.

Discussion on the telecommunications side focused on the recent draft guidelines from the European Commission on the application of the competition rules to that sector, and on the ways in which CEPT Telecom might continue to adapt to the changing environment.

In addition, a Memorandum of Understanding to establish the European Radiocommunications Office was opened for signature and, by the close of the Plenary Assembly, had been signed by 16 Administrations. The Office, which will be based in Copenhagen, is being established to provide greater resources and to meet the demand for wider consultation on radio frequency management activities in Europe.

From October 1990, Greece succeeded the United Kingdom as the Managing Administration of CEPT. The UK had undertaken the task since October 1987. ■

■ MOTOROLA TOURS EASTERN EUROPEAN COUNTRIES

Motorola, a worldwide leader in the design and manufacture of electronics components and mobile radio communications equipment and services, is taking a broad range of some of the company's most current products to five Eastern European capital cities between May 9 and 23.

Motorola has a significant presence in Europe, with $2.3 billion sales in this region, accounting for 20% of total corporate revenue. Over the last 25 years, the company has established 12 manufacturing plants and 53 major locations in 15 European countries, employing 10,900 European nationals.

Founded in Illinois, USA, in 1928 to manufacture car battery eliminators, Motorola has grown into one of the world's largest manufacturers of electronic components and products. In 1990, worldwide sales reached $10.9 billion, up 13% on the previous year.

Interest in Motorola's products and technologies in Eastern European Countries dates back to before the opening up of these countries to free market forces. Now, with the easing of many of the COCOM restrictions which until very recently have affected high technology electronics in particular, it is possible for Motorola to show semiconductors, mobile two-way radio communications equipment for voice and data, pagers, trunking systems, computers, and mobile telephone systems, for a wide range of applications in both public and private sectors.

Motorola has built up a strong reputation for training and educatink its work-force, implementing a world-wide programme which requires all of its 105,000 employees to attend at least five days of work-related training every year. This ambitious programme is developed and managed by the Motorola University, based at the company's corporate headquarters in Schaumburg, outside Chicago in the United States. Motorola University will be presenting its capabilities to the audiences of the Eastern European Seminar series organized in the following capitals: Sofia, Bucharest, Budapest, Prague and Warsaw. ■

RELECTRONIC '91
8th SYMPOSIUM ON RELIABILITY IN ELECTRONICS
26—30 August, 1991—Budapest, Hungary

Organized by the Scientific Society for Telecommunication and the Optical, Acoustical and Film-technical Society

Sponsored by the Department for Technical Science of the Hungarian Academy of Sciences, and the IEEE Hungarian Section.

Chronology: The Symposia on Reliability in Electronics were held in 1964, 1968, 1973, 1977, 1982, 1985 and 1988 in Budapest.
Outstanding experts were attending from abroad and Hungary.

Future: The 8th Symposium on Reliability in Electronics will be held in Budapest from 26—30 August, 1991.

SCOPE: The Symposium is intended as a forum for presenting new results and developments, case studies and experiences,
in the following preferred aspects of **reliability, maintainability and availability:**

— Reliability Theory
— Service Quality
— Network Relability
— Human Factors
— Software Reliability
— Failure Phyisics of Components
— Production Yield

NOTES: The working language of the Symposium is English. A POSTER session will also be organized.
PANEL discussions will be held on topics of special interests.

Registration form and detailed information can be requested from the Secretariat of HTE, Budapest, Kossuth tér 6—8, H—1055

**THE ORGANIZING COMMITTEE
of RELECTRONIC '91**

Informations for authors

JOURNAL ON COMMUNICATIONS is published monthly, alternately in English and Hungarian. In each issue a significant topic is covered by selected comprehensive papers.

Other contributions may be included in the following sections:

- INDIVIDUAL PAPERS for contributions outside the focus of the issue,
- PRODUCTS-SERVICES for papers on manufactured devices, equipments and software products,
- BUSINESS-RESEARCH-EDUCATION for contributions dealing with economic relations, research and development trends and engineering education,
- NEWS-EVENTS for reports on events related to electronics and communications,
- VIEWS-OPINIONS for comments expressed by readers of the journal.

Manuscripts should be submitted in two copies to the Editor (see inside front cover). Papers should have a length of up to 30 double-spaced typewritten pages (counting each figure as one page). Each paper must include a 100—200 word abstract at the head of the manuscript. Papers should be accompanied by brief biographies and clear, glossy photographs of the authors.

Contributions for the PRODUCTS-SERVICES and BUSINESS-RESEARCH-EDUCATION sections should be limited to 16 doublespaced typewritten pages.

Original illustrations should be submitted along the manuscript. All line drawings should be prepared on a white background in black ink. Lettering on drawings should be large enough to be readily legible when the drawing is reduced to one- or two-column width. On figures capital lettering should be used. Photographs should be used sparingly. All photographs must be glossy prints. Figure captions should be typed on a separate sheet.

For contributions in the PRODUCTS-SERVICES section, a USD 110 page charge will be requested from the author's company or institution.