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IEEE

Microwave Theory and Techniques Society  
C.S. Italy, France and Germany MTT Chapters

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# 1992 International Workshop on Submillimeter Waves

22-24, 1992

Capitane del Popolo  
Orvieto, Italy

## Organized by:

Central and South Italy MTT Chapter  
France MTT Chapter  
Germany MTT Chapter  
University of Perugia  
C.S. Italy, Orvieto

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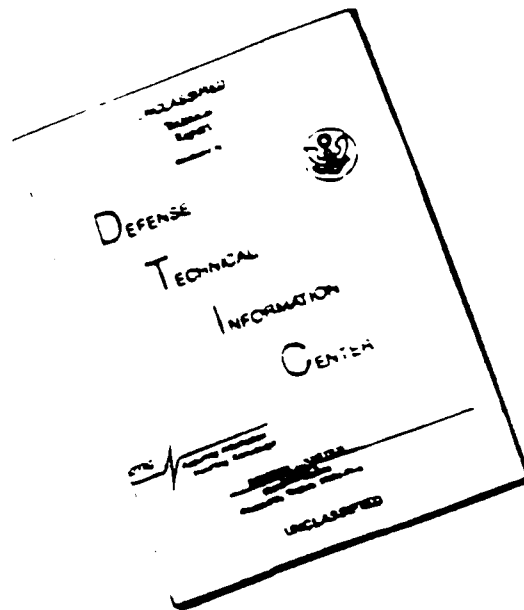
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IEEE  
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April 22-24, 1992

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ITELCO S.p.a., Orvieto

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## FOREWORD

Welcome to third edition of the International workshop on Millimeter Waves, being held at the *Palazzo del Capitano del Popolo* in Orvieto. After the encouraging success of the previous editions at the University of Rome "Tor Vergata" in 1986 and at the University of Perugia in 1989, it was decided to organize a new workshop as a cooperation between the Central and South Italy Chapter with the France and Germany Chapters of the IEEE Microwave Theory and Techniques Society.

The city of Orvieto has been chosen as the venue for this event not simply because it is located midway between Rome and Perugia, but mainly for its unique charm due to its suggestive medieval appearance and the rich artistic patrimony dating back to the Etruscan period. The Palazzo del Capitano del Popolo is no doubt an extraordinary combination between a historical medieval *palazzo* and a modern and efficient Congress Center.

ITELCO S.p.a. enthusiastically agreed to sponsor the workshop and has been a fundamental partner in the organization. The social program is also very attractive, including a very nice concert of two Canadian artists, T. Goudie and D. Lawton.

We would like to thank all the invited speakers who kindly agreed to participate in the workshop and the many others who have worked for its success. We are deeply grateful to Professor Carrara, one of the *fathers* of microwaves in Italy, who accepted our invitation to welcome the participants at the opening session.

Paolo Lampariello and Roberto Sorrentino

Workshop Co-Chairmen

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# **MILLIMETER WAVE APPLICATIONS**

## **AND TECHNOLOGY TRENDS**

**H. Meinel**

**Deutsche Aerospace • Germany**

1992

INTERNATIONAL WORKSHOP

on

MILLIMETER WAVES

Palazzo del Capitano del Popolo

Orvieto, Italy

MILLIMETER WAVE APPLICATIONS  
AND TECHNOLOGY TRENDS

Holger H. Meinel

Deutsche Aerospace AG  
Leopoldstrasse 175  
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Germany

## ABSTRACT

The advantages and new application areas being offered by millimeter-wave systems have been known for a long time. But only the advent of reliable and efficient solid-state sources as well as the progress in low-cost integration procedures, i.e. Hybrid- and Monolithic-Integration Techniques, respectively, has made millimeter-wave technology mature, opening a wide field of employment areas today. It is widely accepted, that the solution to traffic congestion, being a serious problem worldwide, does not lie in the construction of more roads, but in the application of advanced technologies to control the flow of vehicles. The efficiency of the already existing infrastructure will be increased, incorporating RTI (road traffic informatics) in the millimeter-wave range.

## APPLICATIONS

Three major application areas have to be distinguished in this field today. communications, radar, and automotive traffic control, the latter combining the earlier two. Apart from the unique propagation behaviour, millimeter-wave applications take advantage of the highly directive nature of the millimeter-wave beam, i.e. good angular resolution can be achieved with moderately small antennas, thus millimeter-wave systems are small in size and light in weight, compared to their microwave counterparts. The lack of frequencies in the microwave bands commonly used today, is an additional important reason.

## COMMUNICATIONS

Atmospheric propagation effects strongly influence considerations related to the application of millimeter-wave transmission. Normally such systems will be operated in the atmospheric windows around 35 and 94 GHz; for specific applications, the maximum absorption range around 60 GHz will be taken.

The employment of the 40 - 50 GHz band for EHF military satellite up-links has been very important for the development of millimeter-wave technology. This frequency band offers increased bandwidth over the UHF and SHF bands being in current use today and permits the construction of small terminals; thus covert operation for various military platforms is achieved. The SKYNET 4 series of satellites in the U.K. and the MILSTAR program in the US are such applications.

As new services emerge for LAN (local area network) and PCN (personal communication network) systems as well as for traffic control, the demand for a new generation of high data rate radio links has fostered the worldwide development of commercial millimeter-wave POINT-TO-POINT transmission link equipment.

VIDEO-CONFERENCE-SERVICES or SATSTREAM and MEGASTREAM are application examples for high data rate radio links, while remote telephone- or 'cable-TV'-subscriber connection is another; here with special emphasis on 'low cost and effort' infrastructure enhancement. The safety control of high speed trains with velocities up to 400 km/h, like the German ICE (Inter City Express) and the French TGV (Train à Grande Vitesse) as well as LCRN-services (local cellular radio network) are mobile millimeter-wave applications.

- POINT-TO-POINT transmission links, using portable equipment can be installed very rapidly on maximum hop lengths of up to about 10 km, the equipment, due to size and weight, can be mounted easily.

The main employment area for such millimeter-wave equipment is in connecting subscribers, for example in private data -, TV -, and telephone circuits. More than 1,000 systems have become operational since the late 1970's, mostly in Japan, but also in Europe.

Besides these point-to-point and area type connections, the millimeter-wave range has been used and is further investigated for mobile communication purposes.

- Local Cellular Radio Networks (LCRN), for wideband mobile services, are a major investigation effort, for instance in England.

The demand for mobile radio has been such, that severe spectrum congestion is occurring within the today allocated bands. Thus a region of radio spectrum has been sought, which could accommodate the future LCRN bands. The application of millimeter-waves could be the solution.

## RADAR

The widest spread employments of millimeter-waves are found in radar, having fostered the development of millimeter-wave technology during the last two decades. Seekers for terminal guided submunitions or battlefield and terrain mapping radars are military application examples, while the content-free measurement of distances or velocity in industry and traffic control are typical civilian employments.

The most important application in this area were seekers for terminal guided submunitions. Besides stringent performance and low cost constraints a high production volume was required; thus the world-wide seeker development has stimulated millimeter-wave development efforts and secured today's technology status. Such seekers, engineered to production status, are available from various manufacturers.

Millimeter-wave radar is advantageous for applications, where high resolution is a major requirement. Short range surveillance radars, like the Helicopter Obstacle Warning Radar in the lower frequency regime (40 - 60 GHz), is an example of this type. Wire like obstacles with diameters as small as 3 mm have been detected under flight conditions. Being a military necessity at first, helicopter obstacle warning today is an important civilian application, concerning the safety of airport commuter traffic.

In industry millimeter-wave radar can be employed in any situation, where movements, length or height have to be measured in a contactless manner. Doppler sensing is taken to gain information for automatic process control. An example of the latter is the top-dead-center measurement system for combustion engines, while the controlled glue application on windows in car manufacturing is an example of the earlier.

#### RECONNAISSANCE

Millimeter-wave technology has matured and operational systems are deployed in the military arena; thus millimeter-wave reconnaissance has become an urgent necessity today. Different systems, including millimeter-wave superheterodyne and DIFM receivers, have been built.

#### TRAFFIC CONTROL

Millimeter-wave techniques offer the means of sensing vehicles and communicating with them, while they have the advantage of being immune to changing weather conditions. Four major areas of applications have to be distinguished: Automatic Debiting Systems (ADS) and Road Traffic Informatics (RTI) are communication type applications, while true motion- and collision avoidance sensors are radar type approaches. Especially millimeter-wave systems have found an increasing interest for such purposes due to their specific advantages: to be small in size and light in weight. The lack of frequencies in the microwave bands commonly used today, is another very important reason. The frequency allocations for European road transport applications, as suggested by CEPT, clearly identifies the importance of millimeter-wave techniques in this field.

ADS TELEPASS is an ADS installed along the Autostrada Milan - Naples in Italy. Communication is achieved at 5.72 GHz between a roadside transceiver at the toll plaza and a transponder within the vehicle. Such an ADS approach will also be taken to reduce city traffic significantly, e.g. in Cambridge, England, as city entrance debiting is the second major ADS application. In the future such systems will be operated in the 60 GHz range.

RTI A major concern of RTI is vehicle-to-vehicle and vehicle-to-roadside communication. An example of this type is the DACAR system within the DRIVE programme. Communication occurs at 61 GHz over a taken prototype bandwidth of 500 MHz. DPSK (Differential Phase Shift Keying) with error detecting and correcting coding was applied. This leads to a bit error rate (BER) of  $5 \times 10^{-10}$  after demodulation, as multiple tests trials, being carried out in May/June '91 have shown. Further systems, like the improved German AVES (AEG Verkehrs Erfassung System) at 61 GHz are under test and development, respectively, for the same purpose.

TRUE MOTION SENSORS Future traffic control systems require true ground speed information for ABS, ASC or car navigation systems. Using Doppler systems at 24 and 61 GHz, respectively, it has been shown, that at least two independent sensors have to be employed, - looking forward and backward - in order to correct the systematic measurement error due to vehicle movements. Several sensors have been developed at different Institutes and Companies in cooperation with the major car manufacturers in Europe.

AUTOMOTIVE COLLISION AVOIDANCE RADAR The use of radar for automotive collision avoidance or intelligent cruise control is not new. Many companies, institutions and laboratories experimented with various millimeter-wave frequency pulse- and/or FM-CW radar systems in the past two decades. Having small size antennas to be fitted into the front of a car such radars are built for short range use only, up to 200 m maximum have to be and can be achieved easily, even under adverse weather conditions.

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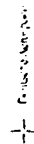
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# LPI - COMMUNICATION

Performance Data

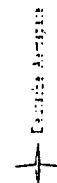


| MANUFACT. | FREQ. | OUTPUT POWER | MODULATION | DATATYPE   | HOPELENGTH | YEAR |
|-----------|-------|--------------|------------|------------|------------|------|
|           | GHz   | mW           |            | Modulation | km         |      |
| Hughes    | 33    | 100 (max)    | FM         | S          | 8          | 1975 |
| NEA COM   | 33    | 70           | NCFSK      | 20         |            |      |
| Philips   | 64    | 100          | FM         |            | 4          | 1975 |
| WJ        | 60    | 70           | FM         | BW 5 MHz   |            |      |
| NOSC      | 60    | 50           | FM         |            | 0.1        | 1980 |
| AEG       | 53    | 10           | FM         | BW 240 kHz | 1.5        | 1980 |
| TRW       | 70    | 10 (max)     | PCM/FM     | BW 150 kHz | 7.3        | 1979 |
| MARCONI   | 60    | 50           | AM         |            |            | 1955 |

LPI: Lo = Propensity of Intercept



Deutsche Aerospace



## SHORT-RANGE RADAR SYSTEMS

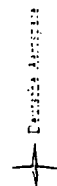
- 1. Short-range radar system (medium range)
- 2. Short-range radar system (short range)
- 3. Short-range radar system (very short range)

## COMMUNICATION SYSTEMS

- 1. Short-range communication
- 2. Short-range communication
- 3. Short-range communication (medium range)
- 4. Short-range communication (very short range)

## TRAFFIC CONTROL

- 1. Short-range traffic control
- 2. Short-range traffic control



## NOTES

NOTES: See page 100 and 101

1. Short-range radar system (medium range)

## BLOCK 1

1. Short-range radar system (medium range)

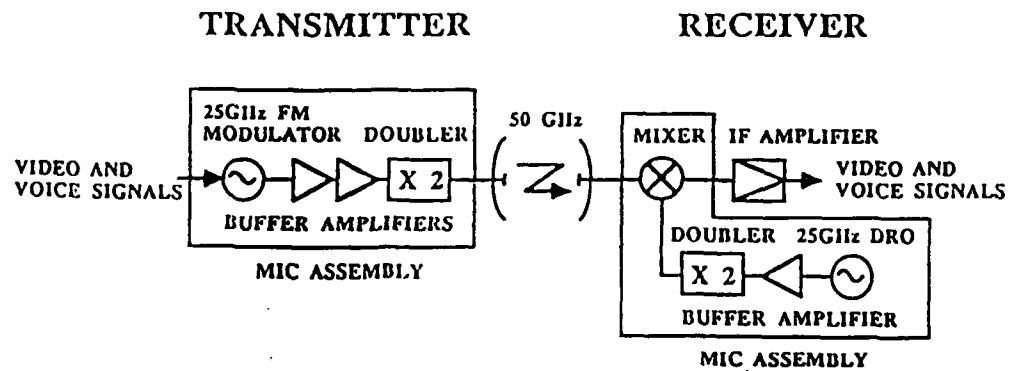
## BLOCK 2

1. Short-range radar system (medium range)

## CRITICAL COMPONENTS

1. Short-range radar system (medium range)

# BLOCK DIAGRAM OF A 50 GHz TRANSMITTER AND A RECEIVER



**MATSUSHITA**  
Electric Industrial Company, Ltd.

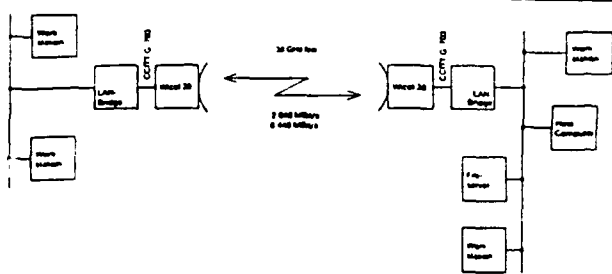
**MATSUSHITA**  
Electric Industrial Company, Ltd.

Major characteristics of the 50 GHz communication system

|                        |                           |
|------------------------|---------------------------|
| Frequency              | 50.46 GHz                 |
| Transmitting power     | 10 dBm                    |
| Noise figure           | 13 dB                     |
| Frequency stability    | $\pm 100$ ppm (-20-60 °C) |
| Modulation             | FM                        |
| Frequency deviation    | 16 MHz pp                 |
| Intermediate frequency | 960 MHz                   |
| IF bandwidth           | 27 MHz                    |
| Video signal bandwidth | 4 MHz                     |
| Audio signal bandwidth | 15 KHz                    |

**Wicol 38**  
38 GHz Wireless Link  
for Digital Data Transmission

**TELEFUNKEN**  
Sendertechnik



simplest system configuration using Wicol 38

critical data:

Application

wireless interconnection of data networks  
e.g. Ethernet, IEEE 802.3, Token Ring  
FDDI

Wicol 38 is equipped with an interface  
corresponding to CCITT G.703 and may  
thus be easily used in telecommunications  
area

Frequency range

in the range 37.5 GHz to 38.5 GHz

Frequency tuning

channel selection by means of incorpo-  
rated keyboard

Antenna beam width

approx. 2°

Bit rate

2.048 Mbit/s or 8.448 Mbit/s

Interface

corresponding to CCITT G.703

Hop length

high transmission reliability up to 4 km

Power supply

20 V to 50 V DC  
230/110 V AC optional

Power consumption

approx. 50 W

Temperature range

-20 °C to +50 °C for all ranges operable

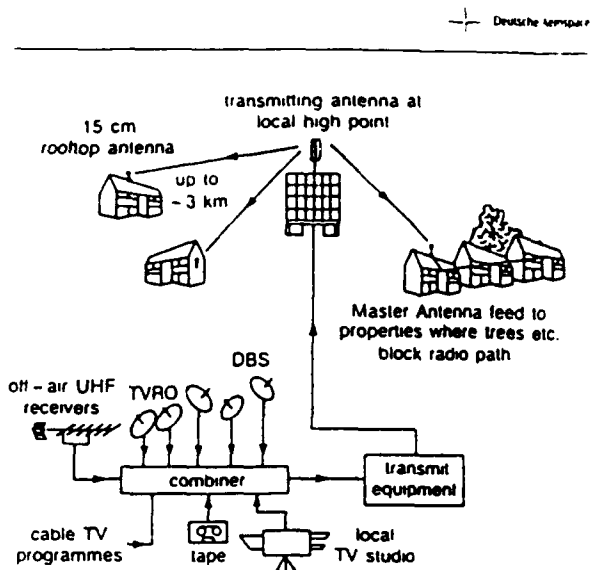
Dimensions

Width: approx. 290 mm  
Height: approx. 290 mm  
Depth: approx. 450 mm

Weight

approx. 18 kg

TELEFUNKEN Sendertechnik GmbH  
Postfach 100000, D-5000 Köln 21  
Telefon (0212) 3053-0 Telex 187819 Telefax (0212) 3053-119



**System Concept**

38 GHz MULTICHANNEL TV SYSTEM  
FROM BTRL, HARTLESHAM HEATH, ENGLAND

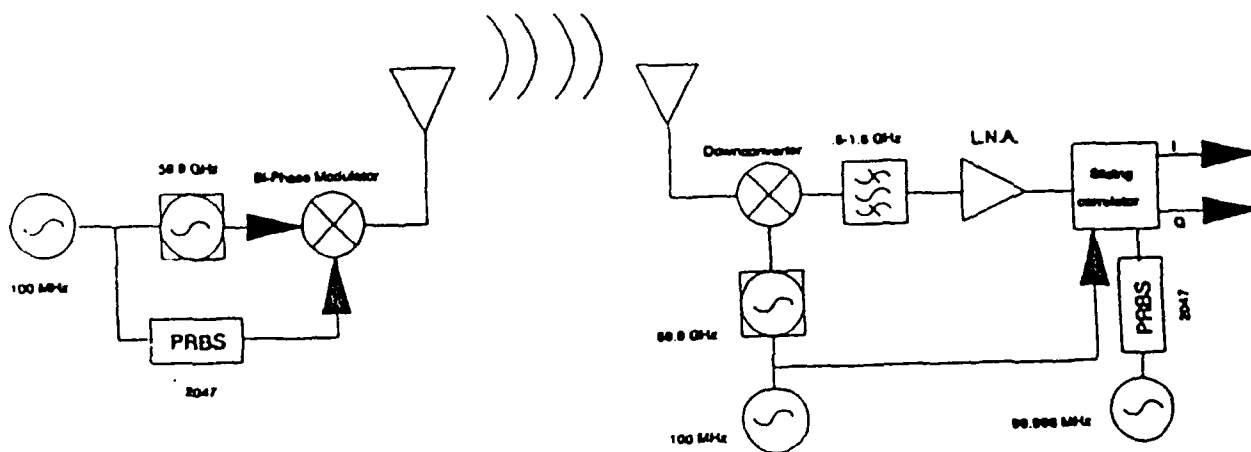


**Deutsche Aerospace**

| Manufact.  | Type     | Freq.<br>GHz | Output<br>power<br>mW | Modulation | Data Rate<br>kbits/s | Hoplength<br>km |
|------------|----------|--------------|-----------------------|------------|----------------------|-----------------|
| Alc.-Th.   | TM 440   | 40           | 50                    | FM         | BW:15MHz             | 15              |
| BTRL       |          | 29           | 120                   | FSK/PSK    | 8k/70k               | 10              |
| Telefunken | Wicol 38 | 38           |                       |            | 8,000                | 4               |
| NEC        | PL 50    | 50           | 15                    | FSK        | 1,544                | 2               |
| FUJITSU    | FC2160   | 50           | 10                    | FM/FSK     | 20,000               | 2               |
| MATSUSHITA |          | 50           | 10                    | FM         | BW:15MHz             |                 |
| SONY       |          | 50           | 20                    | FSK        | 30,000               |                 |

**POINT-TO-POINT Millimeter-Wave Communication systems**  
Performance Data

System 1, 200

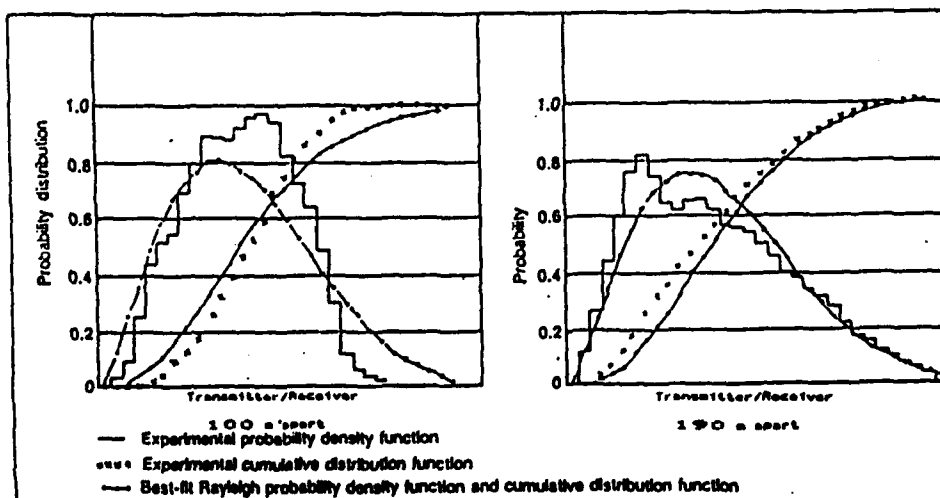


## Sounder Diagram

60 GHZ LBN PROPAGATION MEASUREMENTS  
FROM: UNIVERSITY OF BRISTOL, UK

Radio and Radar Systems Division

AEG



from: M.J. Thomas, G.L. Sigworth, R.G. Cole  
UNIVERSITY COLLEGE LONDON,  
London, England

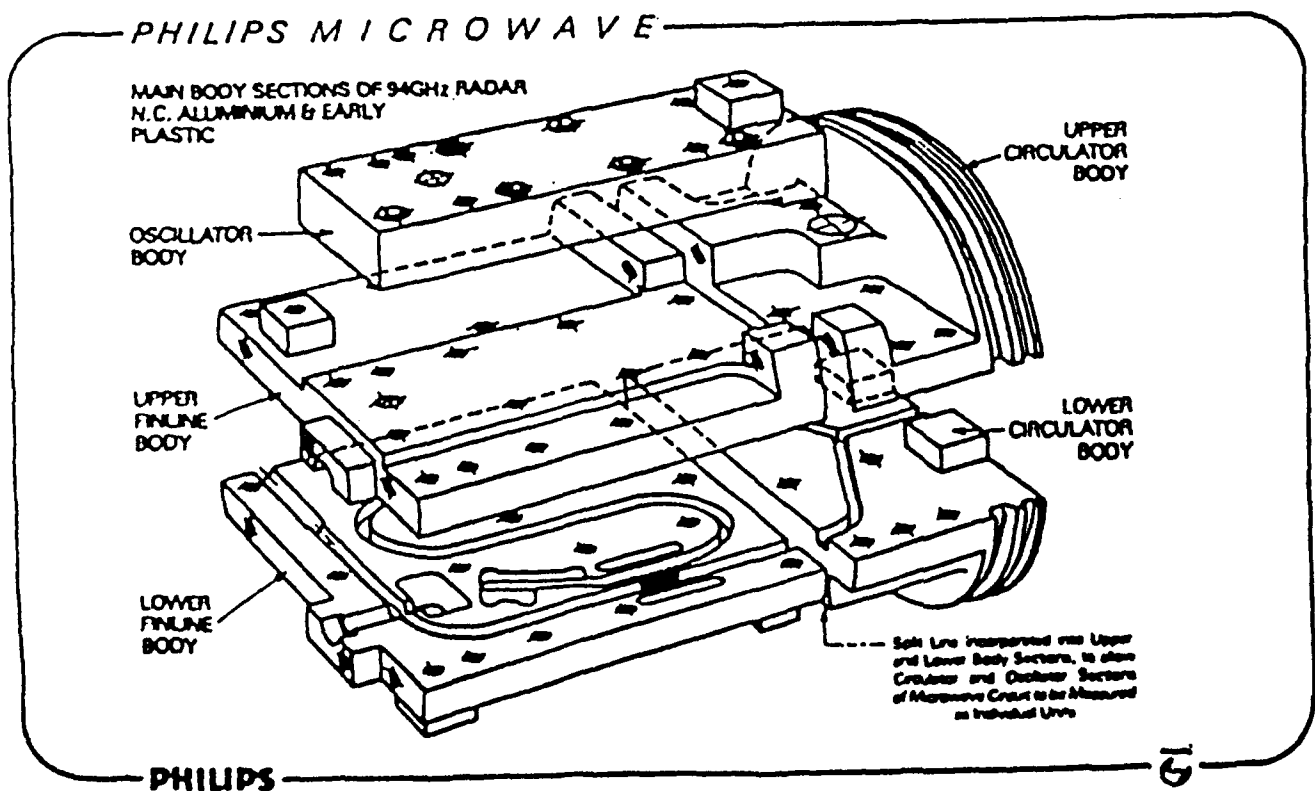
60 GHZ PROPAGATION IN AN URBAN ENVIRONMENT  
PROBABILITY DENSITY FUNCTION

| Manufact.          | Freq. | Output      | Modulation | Bandwidth | Runlength    |
|--------------------|-------|-------------|------------|-----------|--------------|
|                    | GHz   | Power<br>mW |            | MHz       | m            |
| Univ. Col.<br>Ldn. | 55    | 60          |            | 5         | 150          |
| Univ. Bristol      | 60    | 20          | PRBS       | 2         | L-O-S<br>200 |

**Experimental Millimeter-Wave System for Mobile Broadband Services (MBS)  
Performance Data**

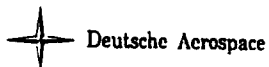
12

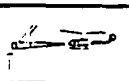
System Data



 Deutsche Aerospace

16



|                                                                                     |                                                        |            |
|-------------------------------------------------------------------------------------|--------------------------------------------------------|------------|
|  | <b>REQUIREMENTS FOR AN<br/>OBSTACLE WARNING SYSTEM</b> | <b>AEG</b> |
|-------------------------------------------------------------------------------------|--------------------------------------------------------|------------|

- **DETECTION OF CABLES AND WIRES WITH  
DIAMETERS  $\geq 3\text{mm}$**
- **RANGE OF CA. 500m**
- **ALL WEATHER CAPABILITY**
- **LIGHT WEIGHT AND SMALL VOLUME**
- **COOPERATION WITH EXISTING  
NAVIGATION AIDS**

## Philips Microwave 94 GHz seeker data

## References

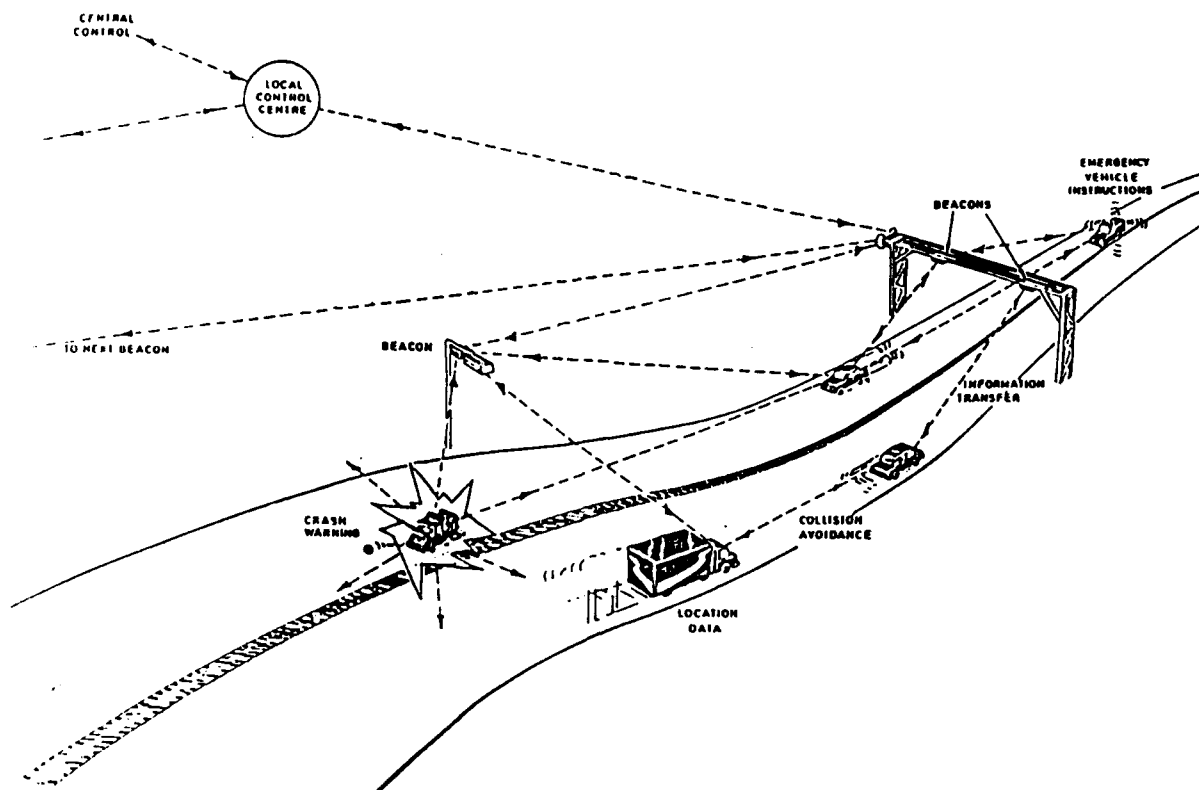


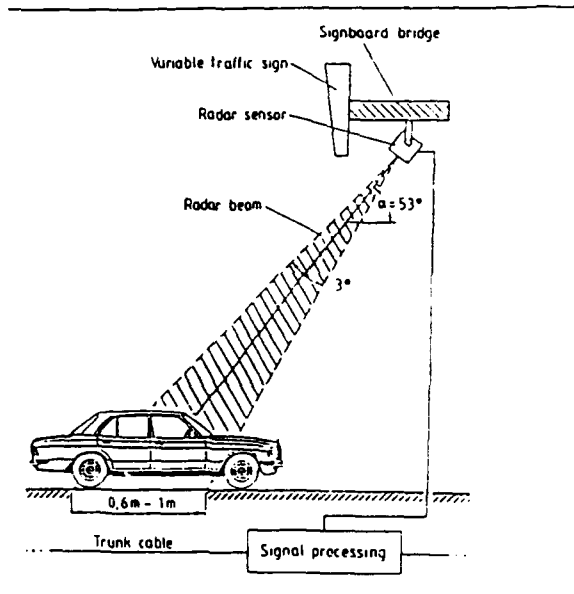
Figure 2.1: Operational system for RTI applications

18

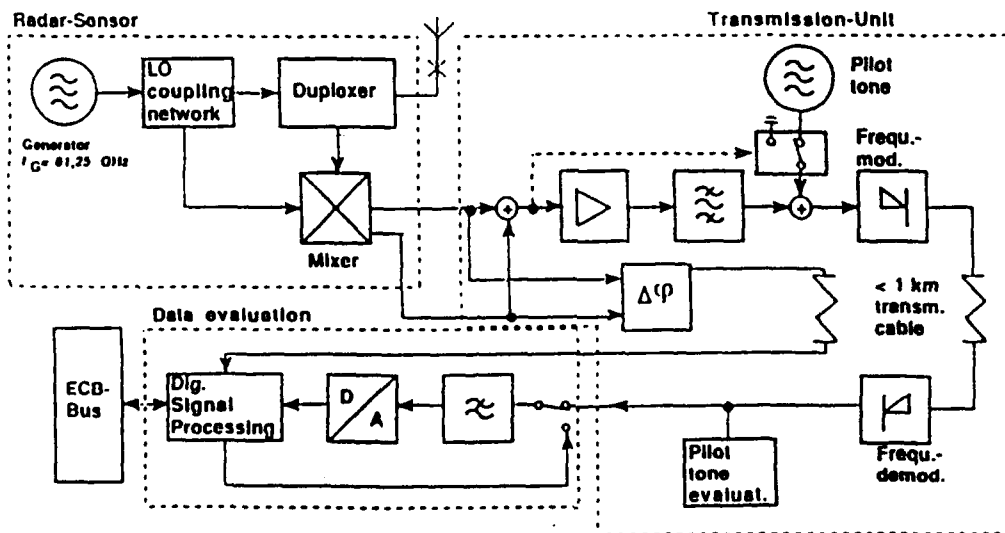


| Bands<br>GHz                | 5.795 - 5.805<br>5.805 - 5.815       | 63 - 64                             | 76 - 77                             |
|-----------------------------|--------------------------------------|-------------------------------------|-------------------------------------|
| Applications                | Automatic toll<br>debiting systems   | Transmissions<br>Road ↔ V<br>V ↔ V  | Anticollision radars                |
| Bandwidth                   | 2*10MHz<br>or 4 channels of 5<br>MHz | 1 GHz<br>Channels of 5 to 20<br>MHz | FM/CW : 100 MHz<br>Pulsed : 500 MHz |
| Allowed power               | 3 dBw                                | 3 - 16 dBw                          | 16 - 20 dBw                         |
| Recommended<br>antenna gain | 10 - 15 dB                           | 10 - 30 dB                          | 30 - 35 dB                          |
| Rate                        | 1 - 3 Mbits/sec                      | Few Mbits/sec                       |                                     |
| Modulation                  | FSK, PSK, ASK                        | FSK, PSK                            | FM/CW<br>Pulsed                     |

from: HESS, President of the CEPT frequency group - DRIVE Conference, 4 - 6/2/91



Application of AVES - Millimeterwave Radar Sensor



|                                    |            |
|------------------------------------|------------|
| Output power                       | 5 mW       |
| Frequency<br>(2nd harm. Gunn)      | 61.25 GHz  |
| Harmonic suppression<br>(H-irises) | > 25 dB    |
| Antenna beamwidth<br>azimuth       | 13 degrees |
| elevation                          | 3 degrees  |
| (asymmetric lens)                  |            |
| gain                               | 25 dB      |
| Mixer NF<br>(Ku-Band diodes)       | dB         |

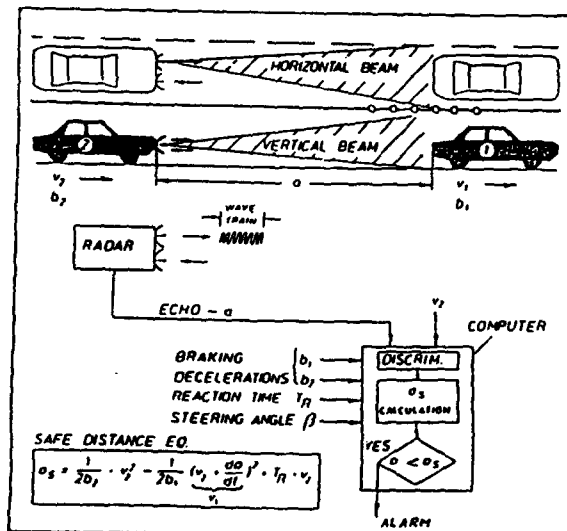
Telefunken Systemtechnik  
AVES sensor data

- 1) RANGE CONFIRMED.
- 2) OPERATES SATISFACTORILY IN PRESENCE OF MULTIPATH EFFECTS.
- 3) OPERATES IN PRESENCE OF OTHER VEHICLES.
- 4) OPERATES WHEN DIRECT PATH IS OBSCURED BY OTHER VEHICLES.
- 5) ERROR RATE BEFORE ERROR CORRECTION  $\approx 10^{-4}$ .
- 6) ERROR RATE AFTER ERROR CORRECTION SO LOW THAT IT IS DIFFICULT TO MEASURE (PROBABLY LESS THAN  $10^{-9}$ ).

DRIVE - DACAR - Program  
Road test results

VEG

Radio and Radar Systems Group.



5 GHz collision avoidance radar  
principle of operation

TOYOTA MOTOR CORPORATION, JAPAN

#### PERFORMANCE DATA

STAND ALONE 'SKIN TRACKING' UNIT

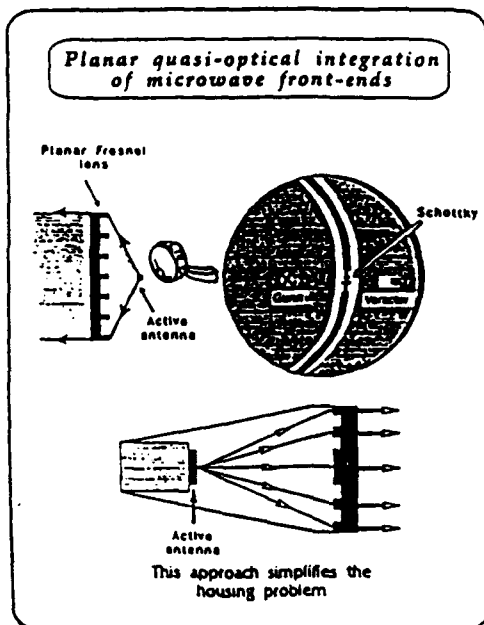
|                   |          |
|-------------------|----------|
| FREQUENCY         | 49.5 GHz |
| MODULATION        |          |
| FM-TRIANGULAR     | 75 MHz   |
| RATE              | 750 Hz   |
| OUTPUT POWER      | 30 mW    |
| MEASUREMENT RANGE | 3 to 60m |
| 150m/m RAIN       | to 50m   |
| ACCURACY          | 5cm      |

#### ANTENNA DESIGN

SEPARATE TRANSMIT AND RECEIVE ANTENNA  
2 DEGREES BEAM WITH  
CROSS POLARIZED  
POLARIZED 45 DEGREES TO ROADWAY

49.5 GHz AUTO MOTIVE RADAR

WORKSHOP ON: ADVANCED CAR ELECTRONICS AND FUTURE TRAFFIC CONTROL SYSTEMS RELATED TO MICROWAVES



Quelle: P.A. Rolland, C.H.S. - Université de Lille

Abstandswarnradar

Quasi-optischer Aufbau bei 60 und 86 GHz

|      | Collision Avoidance | Doppler Sensing      | Beacon Systems |                                |
|------|---------------------|----------------------|----------------|--------------------------------|
| 1970 | Bendix              | 16 GHz duplex CW     |                |                                |
| 1972 | RCA                 | 10 GHz FM-CW         | TI             | 55 GHz Spondermeter            |
| 1974 | AEG-TFK             | 35 GHz pulsed        |                |                                |
|      | Lucas Ltd.          | 32 GHz FM-CW         |                |                                |
| 1975 | SEL                 | 16.5 GHz FM-CW       |                |                                |
| 1976 | AEG-TFK             | 50 GHz pulsed        | Philips        | 10 GHz motion tracker          |
| 1977 | Bendix              | 36 GHz duplex CW     | AEG-TFK        | 35 GHz vsb-Radar               |
| 1978 | SEL                 | 35 GHz FM-CW         | Japan          | Skid control sensor            |
|      | Nissan              | 24 GHz pulsed/FM     |                |                                |
| 1980 | Toyota/Fujitsu      | 50 GHz FM-CW         |                | Vehicular Communication 60 GHz |
| 1982 | Nissan              | 60 GHz pulsed/FM     |                |                                |
| 1984 |                     |                      |                |                                |
| 1986 |                     |                      |                |                                |
| 1988 | Philips             | 94 GHz L/C Hybrid    | AEG            | 60 GHz AVES                    |
| 1990 | Uni Lille           | 94 GHz Quasi-optical |                | MCCS / SEL 60 GHz DACAR        |
|      |                     |                      |                | TST 60 GHz AVES-COM            |

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Millimeter-wave technology is mature today

- covert communication systems
- LAN radios (PCN, MBS)

- Terminal guidance systems
- collision avoidance sensors

have been built and tested successfully

Low cost techniques

for high rate production, hybrid and monolithic, are available and prototype production has been launched

Commercial applications

have gained momentum

- RTI-Road Traffic Informatics
- LAN radio

are today's driving forces

Conclusion

# Millimetre-wave commercial applications and technologies

As the microwave spectrum becomes more crowded the need for application of the millimetre spectrum becomes more acute. Holger Meland from Deutsche Aerospace, Munich has collected together some of the most likely applications of this part of the spectrum and the technologies that are an essential part of the successful application.

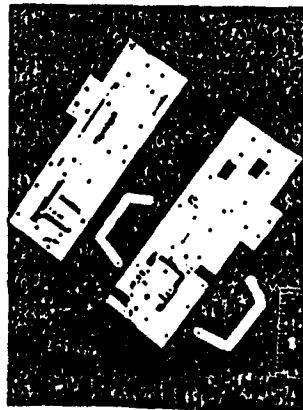


Figure 1: A 3D model of a millimetre-wave antenna module fabricated using M.A.COM's advanced integrated package technique. (Courtesy of M.A.COM)

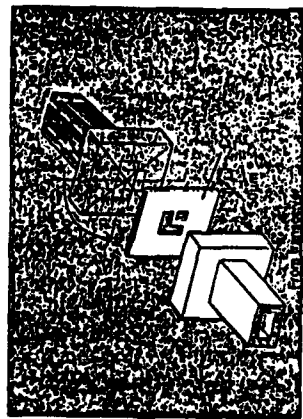


Figure 2: FACOM 2160 cross coupling filter (CCF) design. (From Doi et al [7]) Horizontal or vertical polarisation can be selected simply by rotating the antenna by 90°

Millimetre-wave technology is available and mature today, and it is being exploited in various specialist applications. In the commercial arena two major types of applications have to be distinguished: communications and traffic-control systems. The former being either point-to-point or LAN (local area network) systems, the latter incorporating radar and beacon (short hop link) systems [1].

Apart from the atmosphere's unique propagation behaviour [2], at mm-wave applications take advantage of the highly directive nature of the mm-wave beam, good angular resolution can be achieved with moderately small antennas, thus mm-wave systems are small in size and light in weight, compared to their microwave counterparts. The lack of frequencies in the microwave bands currently used today, especially for communication purposes, is an important reason to take advantage of mm-wave techniques.

Atmospheric propagation effects strongly influence considerations related to the application of mm-wave transmission links. Normally such systems will be operated in the atmospheric windows around 35 and

94GHz respectively. For specific applications and design reasons the maximum absorption range around 60GHz might be taken to ease the reuse of frequencies, for example 85GHz receiver noise figure.

Being the first to take advantage of the mm-wave approach the technology of early communication systems was mostly waveguide related or derived.

**Waveguide integration**  
The VIP (Via-Integrated Package) technique for integrating several different waveguide components into one compact, uniplanar, integrated assembly has been developed and used at M.A.COM of Burlington, MA, USA, for many years in the microwave range. It has been used as a cost-effective method for high-volume production, providing size and weight reduction without sacrificing performance. VIP technology uses waveguide as the primary transmission medium and computerized NC milling combined with laser welding for fabrication.

The approach was taken to develop a low-cost compact integrated mm-wave front-end for LPI (low probability of intercept) communication for the US Army [3]. This Ka-band

Beyond that, existing in-house component technology is heavily influencing this decision and thus the system design. The physical set-up for a particular application might differ significantly from company to company, as seen in the following.

Based on promising measurement results customized digital distribution radios have been developed worldwide. About 700 mm-wave point-to-point radio links are already installed in Japan alone [6].

Being either usable for digital data transmission up to 20Mb/s or for wide-band analogue signals such as colour TV, the FACOM 2160 50GHz system built by Fujitsu of Kawasaki, Japan [7] features 2 to 5km of hop-length, depending on circuit performance being required.

Typical configuration is with a parabolic antenna with 40dB gain, 10mW output power and a receiver noise figure of 17dB. In order to prevent interference problems, if two or more transmitters on the same frequency are used in the same neighbourhood, the FACOM 2160 features a unique, but simple solution: the Cross Coupling Filter (CCF), shown in figure 2. Horizontal or vertical polarization can be selected simply by rotating the antenna by 90°.

Conventional MIC technology on 0.2mm alumina (ferrite) substrate is used for the integrated 50GHz components to allow mass production of the RF-circuits. The single-balanced mixer was implemented using a 3dB branch hybrid and Si beam-lead Schottky-barrier diodes.

A good example for a transmission link making use of modern semicon-

ductor technology is the 50GHz system built by Matsushita Electric Ltd of Osaka, Japan [8]. In this example the output power is 10mW and the receiver noise figure is 13dB.

This system employs 3 Terminal devices only and takes advantage of hybrid MIC integration at 25GHz with double circuitry following. The approach will reduce production costs significantly. Figure 3 depicts the assembled 25GHz FM modulator, including a dielectric resonator.

An example of a European point-to-point transmission link is the 20GHz BTL SYSTEM. Early work at British Telecom Laboratories (BTL) in Marleham, UK, concentrated on developing a 70Mb/s system for transmission of data and full-band video traffic. Output power here is 150mW, receiver noise figure is 8dB and antenna gain is 41dB.

Philips Research at Redhill, England, who developed the RF components of this 29GHz system, has chosen inline technology for the realization. It has emerged as a front runner for frequencies of about 25-100GHz [5]. Inline techniques achieve relatively low loss and the conductor patterns are on circuit boards, which can be made reproducible using photolithographic techniques.

The RF front end contains the integrated inline modules, to which the Gunn diode local oscillator and the transmitter sources are mounted. The transmitter unit, includes a balanced phase modulator, a voltage controlled attenuator, and a monitor coupler for checking power and frequency. The receiver unit shown in figure 4, contains a low noise bal-

anced mixer with integrated IF pre-amplifier, an attenuator to adjust the LO level to the mixer, and a monitor coupler for oscillator power and frequency checking. The entire RF assembly is desiccated. 70 transceiver units of this type have been built.

Hybrid mm-wave technology is mature today. However, MIC technology will be incorporated into such systems as soon as it becomes available. After several years of operation, the frequency of the BTL system was reallocated to 41GHz and the entire front end will now be redesigned incorporating MICs [10].

## Local Cellular

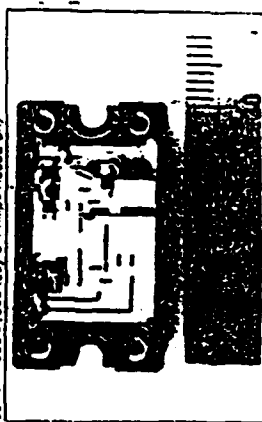
Demand for communication services that can be provided without the need for cable connections is growing rapidly as potential users become aware of services such as cordless telephones and mobile cellular radio.

The 453MHz and 900MHz bands have been extensively used but two groups of researchers in England, namely H J Thorsas et al [11] and J P McGeehan et al [12] from the Universities of London and Bristol, respectively, have investigated the 55-60GHz range for LCRN applications. Figure 5 shows the power profile of 60GHz transmission measurements carried out in Bristol.

Based on these initial measurements it appears possible to establish LCRN's local cellular radio networks in the mm-wave range. Thus the commercial exploitation of this greatly underutilised region of the spectrum is anticipated.

So far both groups have used

Figure 3: 50GHz MATSUHITA system Assembled 25GHz FM modulator (from Ogasawa et al [8])  
Figure 4: 29GHz BTL communication system Integrated receiver module (Courtesy of Philips Research)



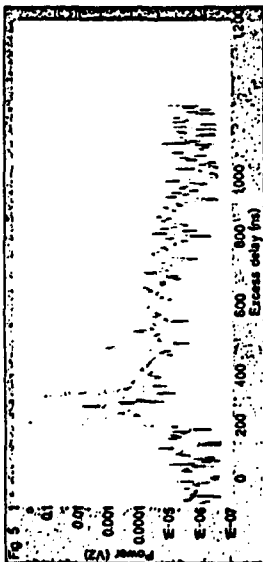


Figure 5 60GHz transmission experiment's Power profile (From McGeerhan [17])

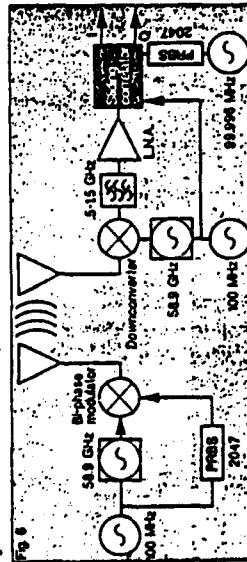


Figure 6 60GHz transmission experiment's Schematic diagram of the equipment (From McGeerhan et al [13])

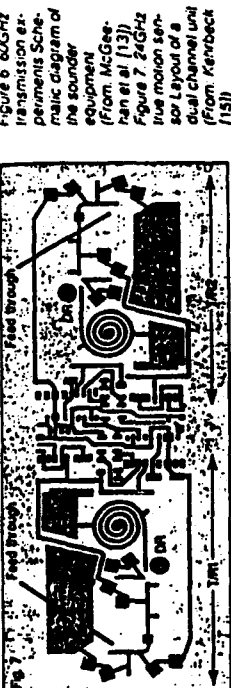


Figure 7 24GHz sensor layout of a dual channel Doppler sensor (From Karmbeck [15])

Table 1 CEPT frequency allocation for European road transport applications

| Frequency band | Service                           | Power  | Modulation    | Bandwidth | Channel spacing | Frequency reuse |
|----------------|-----------------------------------|--------|---------------|-----------|-----------------|-----------------|
| 24-26 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 26-28 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 28-30 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 30-32 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 32-34 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 34-36 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 36-38 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 38-40 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 40-42 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 42-44 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 44-46 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 46-48 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 48-50 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 50-52 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 52-54 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 54-56 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 56-58 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |
| 58-60 GHz      | Automatic toll collecting systems | 100 mW | FSK, PSK, ASK | 10 MHz    | 10 MHz          | 10 MHz          |

Figure 7 24GHz sensor layout of a dual channel Doppler sensor (From Karmbeck [15])

commercially available 60GHz equipment (Standard Via-guide type), see figure 6.

### Traffic control

In Europe, road transportation and traffic congestion are major problems. Ever increasing demand has adversely affected traffic safety, efficiency and economy. For the future, the protection of the environment is a special challenge. Pan-European projects like PROMETHEUS or DRIVE are supporting solutions to control and assist the traffic flow, using Road Traffic Information (RTI), such as route guidance, road pricing, or incident detection, for example. Thus mm-wave systems have found an increasing interest [12]. Doppler radar systems for true motion control and traffic monitoring as well as FM-CW or pulse-radar for collision avoidance purposes, and beacon systems for road pricing and advanced road infrastructures are examples. Table 1 shows the frequency allocations for European road transport applications [14], as suggested by the CEPT. This clearly identifies the importance of mm-wave techniques in this field.

Not taking into account the very early research in this field, hybrid and monolithic mm-wave integrated circuits have been incorporated mostly. Microstrip and thin-film technologies are the two most important transmission media being employed in traffic control systems. However, quasi-optical technology, i.e. the optical integration of planar MMICs, is being studied as an option.

Future traffic control systems require true ground speed information for ABS, ASC or car navigation systems, using microwave Doppler systems for this purpose.

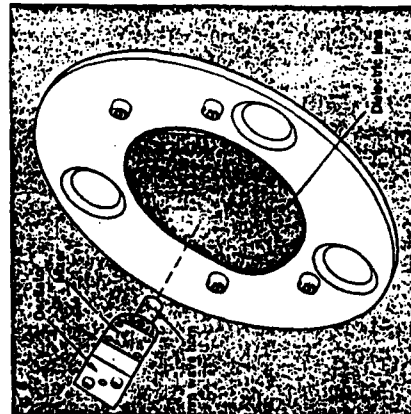


Figure 8 61.5GHz AVEs System Schematic setup of the mm-wave front-end (From Sammer et al [16])

self-biased single-ended mixer employs a Si beam-lead Schottky-barrier diode on a sili-substrate. The dimensions of the complete micro-wave front-end are 120x60x15mm.

The sensor module fulfills all requirements for car integration. Due to ongoing developments in high frequency satellite broadcasting systems (12GHz) the cost of the hybrid version can be expected to reduce correspondingly. Furthermore, the RF front-end could be a MMIC.

### 61.5GHz AVEs system

61.5GHz vehicle sensors based on the Doppler principle have been applied successfully for traffic jam avoidance or warning. The AVEs system (AEG Verkehrs Erfassung System) obtains specific traffic information from appropriate Doppler signal processing. Built by Telefunken Systemtechnik of Urm, Germany [16], the chosen frequency of operation, 61.5GHz offers several advantages:

- the band is available for this application, ISM - band (Industrial, Scientific and Medical) in Europe
- due to the high frequency small antenna dimensions allow good focusing, individual lanes are illuminated only
- short range application, thus not hampered by high atmospheric absorption
- economic reuse of frequencies

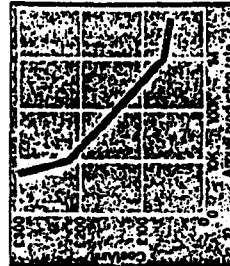


Figure 10 61.5GHz radar seeker demonstrator Cost/loss curve employing polymer injection technique (From Birns [26])

The sensor is relatively small, approximately 150x150x150mm, and can be mounted on existing signposts and bridges. Thus, it is not necessary to install the system underground, as is done with inductive loops, making such a sensor especially suited for mobile installations, e.g. on road construction sites.

Output power is 5mW and antenna beamwidth is 3x13°. Hybrid integration was implemented combining line and waveguide to reduce cost, as is the employment of a 2nd harmonic type Gunn oscillator and the implementation of Ku-band devices in the single balanced mixer. The schematic setup of the mm-wave front-end is presented in figure 8, while the actual components are in figure 9.

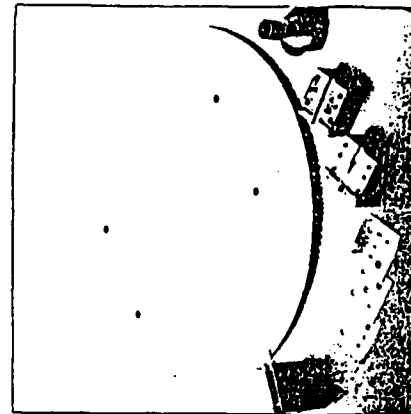


Figure 9 61.5GHz AVEs System Realized millimetre wave components (Courtesy of Telefunken Systemtechnik)

Vehicle speed, vehicle length, and the distance between vehicles is obtained by the AVEs sensor, which has been tested successfully on German highways, outperforming a competing infrared system. The system has been approved by the German Bundesanstalt für Strassenwesen and a production contract for the first 1,000 units has been placed.

A future application of a further developed AVEs sensor can be seen in vehicle-to-roadside communication. Together with a smart card reader such a 60GHz transceiver (hybrid or monolithic integrated) offers the possibility to establish a link between individual vehicles and a roadside traffic management network [13]. Concerning the traffic status ahead, drivers can be provided with real-time information.

### Automotive radar

The use of radar for automotive collision avoidance or intelligent cruise control is not new. Many companies, institutions and laboratories experimented with various mm-wave pulse and/or FM-CW radar systems in the past two decades summarized in table 2.

Having small antennas lined into the front of a car, such radars are built for short range use only, up to 200m, achieved even under adverse weather conditions.

Present technology demands that

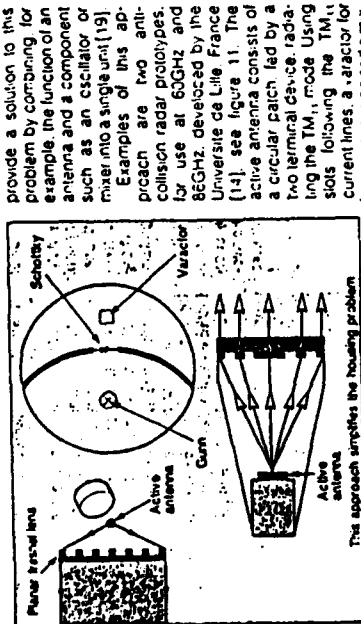


Figure 11. Quasi-optical integration of radar front-end principle

the antenna be manufactured either as a cost-effective waveguide feed structure, the most efficient millimetre-wave cavity waveguide, or as a planar waveguide. All the remaining front-end components can be conveniently manufactured in a single hybrid assembly using thin-film technology. The overall performance of such a unit is determined by the conducting housing, forming the surrounding waveguide. Plasmas have generated much interest as a low-cost solution for mass production. Polymer injection moulding is one possibility and exhibits a considerable background of process information [13]. The mass production of compact Gunn diodes with micro surface features (a more stringent requirement compared to planar wave radar) has been solved using a metal-plated polymer injection moulding process pioneered by Philips. The performance of a 30-GHz radar-seeker demonstrator realized by Philips Microwave of Stockport, England, was an output power of 35 mW, a sweep range of 500 km and better than 1% linearity. The mass is less than 500 g.

Component values and layout design of this first demonstrator unit are not optimum for radar use, as it was necessary to have electrical access to each component for experimental purposes. However, the feasibility of this design approach was proven. Thin-film black technology with thin and microstrip circuits inserted was used for the realization of the RF front-end. The housing was manufactured using metalized polymer injection moulding. The projected cost is less than £10 UK with a production rate of 400 per hour, figure 12.

providing a millimetre-wave radar product ideally suited to the needs of the automotive industry.

The operating frequency was chosen due to military requirements, but has been changed to 76 GHz, the CEPT allocation.

### Quasi-optical

As mm wave systems progress to higher frequencies and components, the engineer may be forced to combine simply components and to combine

provide a solution to this problem by combining for example, the function of an antenna and a component such as an oscillator or mixer into a single unit [19].

Examples of this approach are two anti-collision radar prototypes, for use at 60 GHz and 86 GHz, developed by the Université de Lille, France [14], see figure 11. The active antenna consists of a circular patch fed by a two terminal device radiating the TM<sub>11</sub> mode. Using slots following the TM<sub>11</sub> current lines, a reactor for frequency tuning and a mixer diode for rectifying can be incorporated as well.

### SiMIMWC

The application of molecular beam epitaxy (MBE) and X-ray lithography for the fabrication of monolithic millimetre-wave devices on high-resistivity silicon has been progressing and has been proven [22]. Discrete IMPATT and Schottky diodes fabricated from this material have shown excellent performance around 100 GHz. Hybrid integrated transmitters using the SiMIMWC (Silicon Monolithic Microwave Integrated Circuit) approach have delivered a maximum CW output power of 200 mW at 73 GHz. A completely monolithically integrated transmitting element using a planar IMPATT diode in a coplanar

structure was realized recently, demonstrating the feasibility. Without additional heat sinking, the output of this unit was only 0.05 mW, so more work will be needed. The integration of a coplanar Schottky-barrier diode with a microstrip antenna yields a simple monolithic detector receiver chip. Developed by the Daimler-Benz Research Institute in Ulm, Germany, it is shown in figure 13. The Schottky-barrier diode is terminated by a quarter-wave microstrip line two networks for biasing and for decoupling of the received signal being connected. The Schottky diode was biased with 270 mV. A receiver sensitivity of 65 pV/√Hz was obtained at 94 GHz.

### Conclusion

This overview shows that millimetre-wave technology to meet system requirements is available today. The near advent of monolithic integration techniques for production purposes has already changed the commercial arena and major MMIC suppliers from the US, for example, have already found partners in the European automotive industry, so the scientific discussion of such topics has become very sensitive.

### Acknowledgements

Thanks are due to Dr. Peter Tabor, CEO of Philips Microwave, for his support and to Dr. J. H. J. van der Linde, Philips Research Laboratories, for his support and to Dr. J. H. J. van der Linde, Philips Research Laboratories, for his support and to Dr. J. H. J. van der Linde, Philips Research Laboratories, for his support.

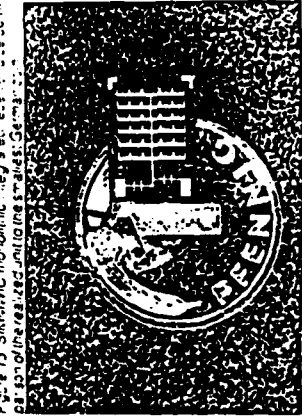


Figure 13. SiMIMWC monolithic millimetre-wave receiver

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**A NEW PASSIVE MICROWAVE SCANNER FOR  
AIRBORNE MEASUREMENTS OF MARITIM OIL  
POLLUTIONS**

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# **A New Passive Microwave Linescanner for Airborne Measurements of Maritime Oil Pollutions**

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## **Abstract**

A new passive microwave linescanner has been developed that will be used for airborne operational measurements of maritime oil pollutions. The system consists of two offset rotating parabolic mirrors and two radiometer sets, where each set contains three radiometers at 18.7, 36.5 and 89 GHz center frequency. A further 89 GHz radiometer is used for the measurement of the average radiometric sky temperature. The system is continuously calibrated by the use of a "hot load" at ambient temperature and a peltier cooled "cold load". A computer system allows online data reduction for the estimation of the amount of oil on the sea surface.

## **I. Introduction**

Pollution of the sea has increased up to levels which can result in severe risks for the marine ecosystem. High loads of oil have been observed in many coastal waters. In addition to the procurement of "oil combating ships", this situation has led to the demand for airborne surveying methods for providing helpful data for oil spill clean-up operations following accidental discharges, and to monitor the legal aspects of permanent contributions to maritime oil pollutions.

In Europe, aircraft equipped with remote sensing instruments have been provided in various countries in the period 1983-1988 for a regular survey of coastal waters and, in view of a sensor system of second generation, for the purpose of basic research. Since 1986, for example, two Dornier Do28 aircrafts equipped with an IR/UV-Linescanner, a Side Looking Airborne Radar (SLAR) and a scanning 35GHz-Radiometer have been flown in the German-responsibility areas of the North Sea and the Baltic Sea, aiming at a continuous surveillance of maritime pollution and an improved guidance of oil combating operations. In 1983, 1985 and 1988, under the sponsorship of the European Community, a series of international airborne exercises took place in the North Sea (Archimedes Campaigns [1-3]). Different institutions

operated optical and active microwave sensors in parallel to a multitude of microwave radiometers (linescanners and profilers). The general aim of these exercises was to establish specifications for an improved operational sensor package for quantitative registration of oil pollution at sea under nearly all-weather conditions.

In 1989, the German Ministry of Traffic decided, to realize a new maritime surveillance system on board a modern Do228 aircraft [4]. In addition to improved sensors utilized up to this time in the first generation aircraft, the new system will include newly developed sensors, a Laser Fluorosensor and a multiple frequency microwave radiometer which allow a more thorough analysis of oil spills. This includes a quantification of the spilled volume over a range covering small discharges up to accidental events, and the classification of oil types.

## **II. The New Radiometer**

### **A. Requirements**

The Archimedes exercises offered a unique chance to study radiometer outputs in the 5GHz to 90GHz range for different types of oil and oilwater emulsions of varying thickness (up to several millimeter) during a variety of realistic environmental conditions. A series of findings resulted which, together with some operational restraints are now forming the basis for the development of the new radiometer. In detail, basic requirements for a radiometer of second generation are:

- simultaneous measurements at several frequencies to avoid ambiguity problems during the determination of layer thickness and, above all, the oil volume; an error of volume estimation smaller than 50% is tolerated,
- selection of frequency bands with reference to possible interferences due to man-made noise and registration of oil layer thickness in the 50µm to 3mm range,
- simultaneous registration of the sky radiation at least at the highest frequency band which is the most sensitive one with respect to variation of weather conditions,

- a scan angle as large as possible to realize wide scan widths,
- a spatial resolution as high as possible under the proviso of the maximum tolerable dimensions for the radiometer package (length:88cm, width:50cm),
- consideration of typical flight altitudes of 300m to 1000m and aircraft velocities near 70m/s.

#### B. Frequency Ranges

All requirements, listed above, have been considered during the development of the new radiometer. The construction of the linescanner was achieved with a compact configuration that realizes simultaneous imaging at 18.7, 36.5 and 89GHz. The frequency bands selected are reserved for passive measurements. During the Archimedes campaigns, measurements at 17GHz regularly resulted in the best volume estimations for thicker layers. A frequency around 35GHz seems to be the best compromise with respect to geometrical resolution and bad weather capability. 90GHz images are highly resolved and comparable in quality to IR images (Fig. 1). A sensitivity down to a layer thickness of 50 $\mu$ m exists. Oil-in-water emulsions and white caps (false alarm) can be determined qualitatively well in this frequency range.

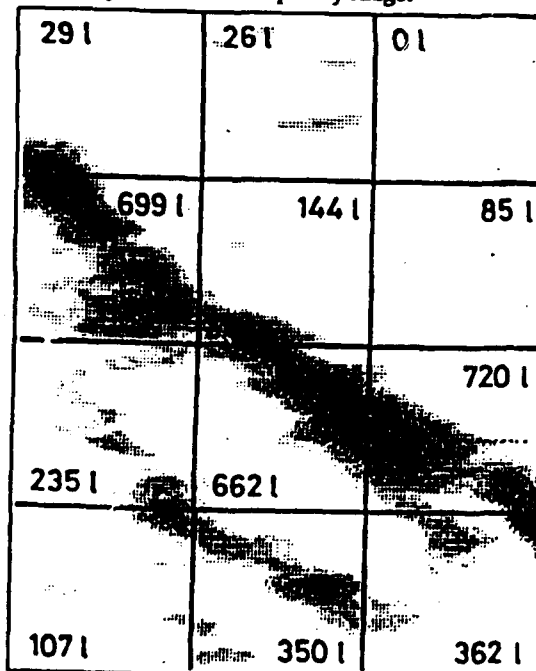


Fig. 1. Example for the representation of passive microwave oil spill measurements. Oil slick taken at 90GHz during night time together with volume estimations (liter); altitude 4000 ft, scan width 440 m, maximum variation of brightness temperature 21 K.

In general, for optimum data evaluation, the sky radiation which influences the linescanner output should also be constantly mapped with a separate system in all frequency bands. But for simplification and cost reduction only one further 89GHz radiometer with a fixed sector-horn antenna has been designed for the sky radiation measurement. Its output delivers a mean-value of the relevant sky radiation which is the optimum basis for numerical estimation of the corresponding values at 36.5 and 18.7 GHz. The errors resulting from that procedure are expected to be smaller than 50%.

#### C. The Scanning Principle

Scanning happens with the aid of two parabolic mirrors installed in a continuously rotating cylinder (Fig.2). Beyond both front sides of the cylinder, identical three-channel receivers are placed which are coupled to the parabolic mirrors via plane reflectors and through holes at the front sides. This scanning principle should be particularly advantageous with respect to operational use. It avoids oscillating movements of the main reflectors, rotational switches and guarantees redundancy in the case of a failure of receiver channels. Further it supports a simple method for continuous calibration. A swath width of 476m is achieved for an altitude of 308m (1000ft).

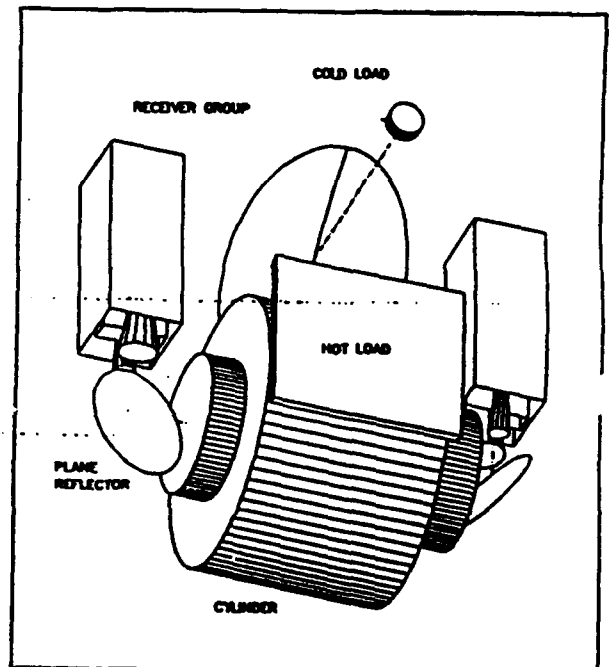


Fig. 2. Perspective representation of the MWR configuration developed. Two parabolic reflectors in a rotating cylinder and two receiver groups realize two scanlines per revolution. Thermal hot/cold sources are used for a continuous calibration.

#### D. The Calibration Principle

The calibration of the radiometer output signals is done by using a hot load at ambient temperature and a peltier-cooled cold load. The hot load consists of a sheet of absorbing material that is mounted on the inner wall of the radiometer main rack. As the effective area of the cold load is much smaller, a further mirror is used for focussing the beams of the rotating antennas onto the cold load. During each revolution of the cylinder a complete series of hot/cold measurements for all frequency channels is accomplished.

A cross sectional view of the cold load is shown in Fig. 3. The absorber is cooled by a 3-stage Peltier element that has a power consumption of 45W, this results in a temperature difference of 51K with reference to the case temperature. The core of the cooled absorber is made of aluminum in order to get a nearly constant temperature distribution up to the tips of the pyramids. The surface is covered with a layer of absorbing material. The reflecting inner wall of the load case has a conical shape resulting in an increased effective area of the cold load. This is important for the 18GHz channels which have the largest beam waist that is also somewhat off axis. The space between the inner wall of the case and the cold load is filled with small balls of expanded polystyrene in order to prevent convection of the surrounding air and condensation of water. The overall efficiency of the cold load has been determined with an external cold load consisting of a large sheet of absorbing material cooled with liquid nitrogen.

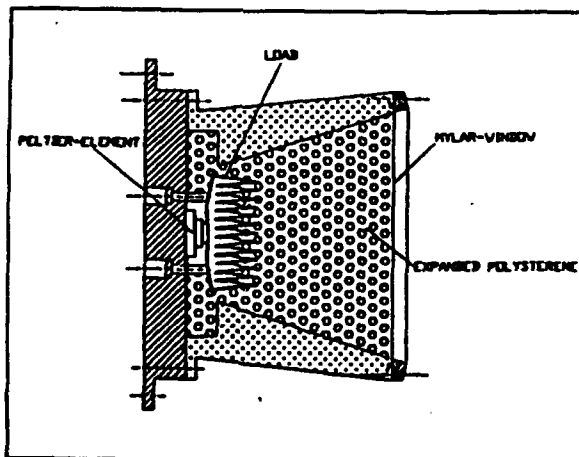


Fig. 3. Principal configuration of the cold load.

The sky-radiometer is calibrated by the same principle. However here we are using waveguide ferrite switches to switch the radiometer input to the loads. The cold load consists of a waveguide absorber that is also cooled by a 3-stage Peltier element.

#### E. Receiver Front- and Back-End Characteristics

To realize optimum radiometric resolution total power and heterodyne principles are applied to all channels. Balanced mixers are used throughout. They show the lowest noise figures which are, presently, obtainable with commercial devices at room temperature. The receiver housings shown in Fig.2 only contain the single front-end modules which can be simply removed like a drawer for better access during repair. The adjustment of a single radiometer channel for calibration happens by the use of a variable offset amplifier which is placed in a separate back-end housing. Low pass filtering and A/D conversion follow after low frequency amplification.

The whole radiometer system is completely computer controlled. Not only calibration is done automatically but also a system configuration on the basis of a priority list (after switching on the radiometer). Further, during the measurements, a numerical correction of the antenna beam squint angles which result from the displacement of the feedhorns is realized. Digital processing, in addition, allows a continuous observation of the operational conditions of the system by monitoring and storing data such as the physical temperatures of the housings and calibration loads, the noise figures of the front-ends, and the gain and offset voltages of the variable amplifiers. Finally, in the case of oil spill measurements, layer-thickness and volume is directly processed from the received signals.

Different spatial resolutions are obtained for the single frequency bands by optimum utilization of the antenna apertures especially at 89GHz. However an identical resolution which would be important for an automatic unambiguous estimation of the volume, could be achieved numerically.

In table 1, the most important radiometer characteristics are listed.

#### III. First Airborne Measurements

After integration and installation into the aircraft, the new system has been flown for the first time in November 1991 in the Oberpfaffenhofen area. A special quick-look system had been developed for image representation, to qualify the radiometer. The first results showed that the concept of the new system works very well and that the resolutions meet the specified values (Fig. 4). On the other hand, a deeper analysis revealed some spurious interferences, among other things, due to local-oscillator radiation in the 18.7 and 36.5GHz ranges. These problems arise from the absence of unilines at the receiver inputs which normally results in higher sensitivities for well adjusted devices. In view of the first measurements above oilpolluted sea that are scheduled for April 1992, additional filtering in the 18.7GHz channels will be a short term solution. It is planned, however, to replace the het-

Tab. 1. Microwave Radiometer Characteristics

| Tab. 1. Microwave Radiometer Characteristics |                |                                     |                |             |
|----------------------------------------------|----------------|-------------------------------------|----------------|-------------|
| Antenna                                      |                |                                     |                |             |
| Sensor :                                     |                | linescanner                         | sky-radiometer |             |
| type :                                       |                | rotating offset parabolic reflector | sector horn    |             |
| angular resolution :                         |                | 0.8° / 2.0° / 4.0°                  | 8° • 80°       |             |
| scan-width utilized :                        |                | 76°                                 | ./.            |             |
| scan-frequency :                             |                | 10 revolutions/sec = 20 lines/sec   | ./.            |             |
| Receiver front-ends                          |                |                                     |                |             |
| channel                                      | type           | center freq.                        | i.f.           | noise temp. |
| 1, 1'                                        | heterodyne DSB | 89.0 GHz                            | 1-3 GHz        | 590 K       |
| 2, 2'                                        | heterodyne DSB | 36.5 GHz                            | .1-5 GHz       | 565 K       |
| 3, 3'                                        | heterodyne SSB | 18.7 GHz                            | .6-8 GHz       | 550 K       |
| sky                                          | heterodyne DSB | 89.0 GHz                            | .1-1 GHz       | 590 K       |

erodyne principle by direct receiver principle, at an appropriate time, by using HEMT amplifiers in the 18.7 and 36.5GHz ranges.

#### Acknowledgement

We wish to express our thanks to our industrial partners "Krupp MaK Maschinenbau" and "Wissenschaftlich-technische Beratungen, Dr. Hoffmann". Development of the Microwave Radiometer is financed by grants from the "Bundesministerium für Technologie", FRG. We thank O. Krause from "DLR Projektträgergesellschaft Umweltschutztechnik", for his continued support and encouragement.

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- [1] Gillot, R.H., Toselli, F., "The Archimedes 1 Experiment", Commission of the European Communities, Series: Environment and Quality of Life. EUR 10216 EN, 228 pp. Luxembourg 1985.
- [2] Gillot, R.H., "The Archimedes 2 Experiment", Commission of the European Communities, Series: Environment and Quality of Life. EUR 11249 EN, 238 pp. Luxembourg 1987.
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Fig. 4. First airborne measurements with the new radiometer. 89GHz (left) and 36.5GHz (right) quick-look representation of a section with different types of vegetation or soils and water surfaces (black colour, storage lake). Demonstration of resolution capability and brightness temperature range awaited for polluted sea surfaces. Spurious interferences by external electronics. Altitude 3000ft.

# **MMIC TECHNOLOGY FOR COMMUNICATION SYSTEMS**

**M. Aikawa**

**NTT • Japan**

## MMIC Technology for Communication Systems

MMICs are potentially very suited to high frequency bands (microwave band to millimeter-wave band), due to

- (1) mass productivity with high repeatability and excellent accuracy in the fabrication process,

and

- (2) low cost potentiality.

April 22, 1992

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1-2356 Yokosuka-shi, Kanagawa, 239-03 Japan

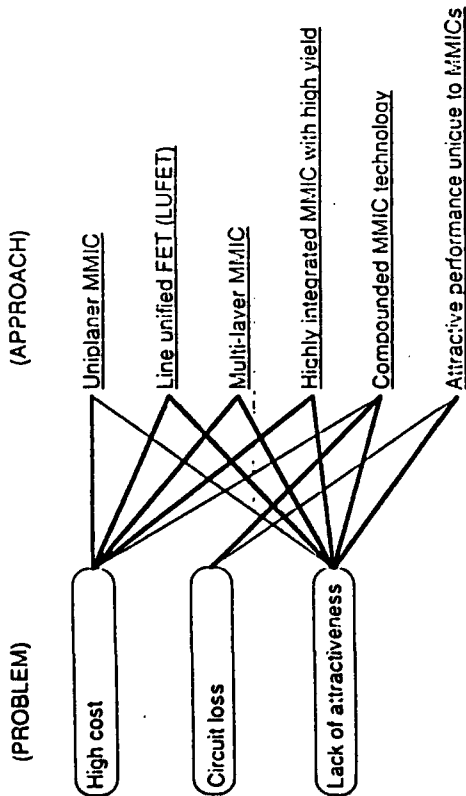
## OUTLINE

1. Problems in Applying MMICs to Communication Systems
2. MMIC Technology for Low Cost
3. Technology for Overcoming Disadvantages and Increasing Attractiveness of MMICs
4. MMICs Applications in Radio Communication Systems
  - Digital Microwave Radio
  - Subscriber Radio
5. Conclusion

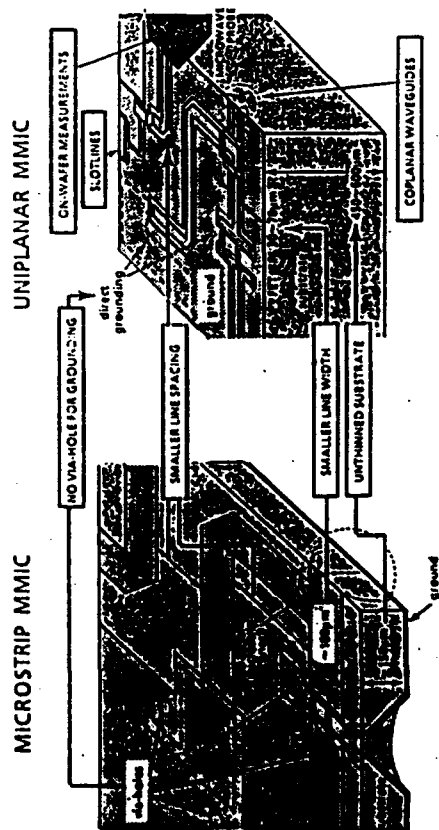
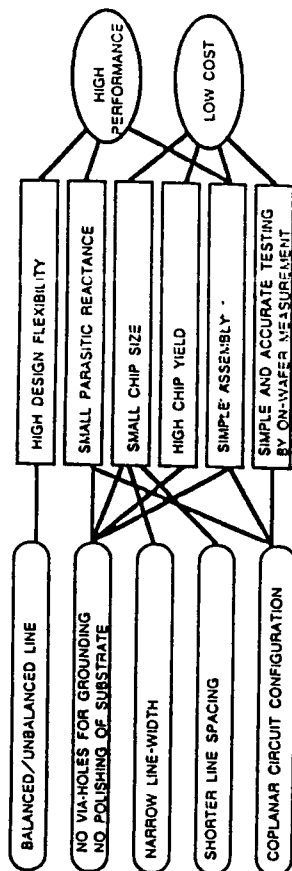
## TECHNICAL PROBLEMS IN APPLYING MMICs TO COMMUNICATION SYSTEMS

- (1) High cost at present  
because of the small market size and specialized needs,
- (2) Circuit loss  
because of the small chip size  
⇔ high Q resonator, filter, power efficiency, NF, etc.
- (3) Lack of MMIC attractiveness for systems to which MMICs are applied

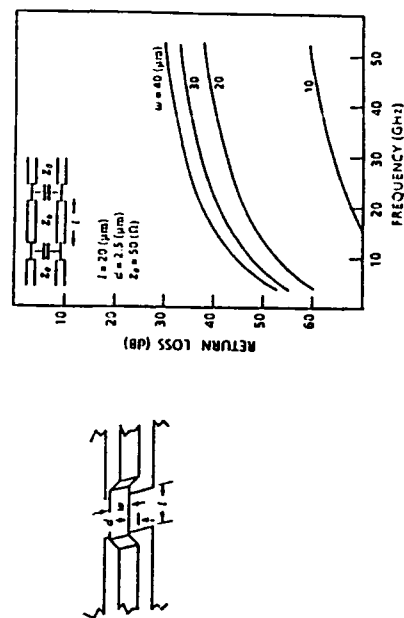
# TECHNICAL APPROACHES TO OVERCOME THE PROBLEMS



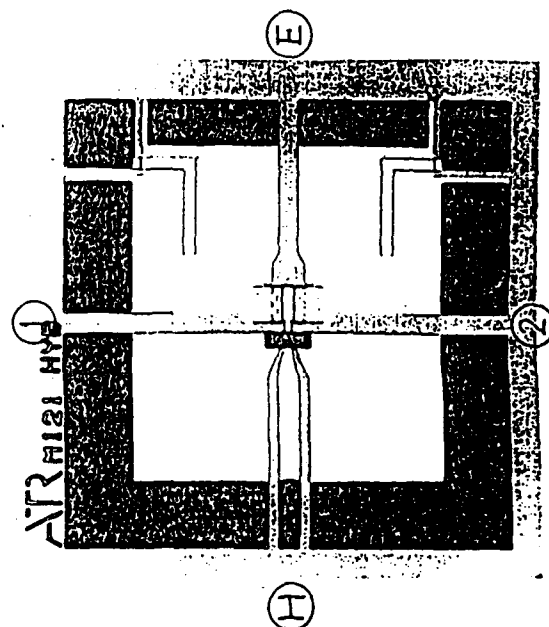
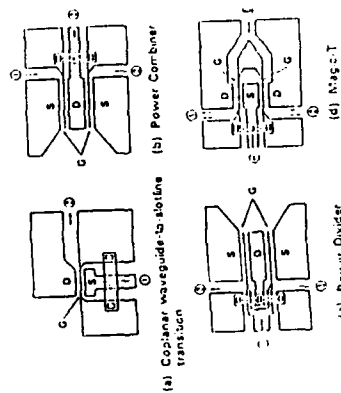
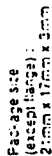
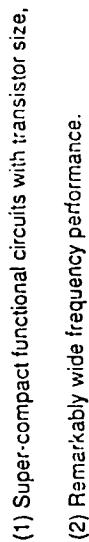
## ADVANTAGES OF UNIPLANAR MMICs



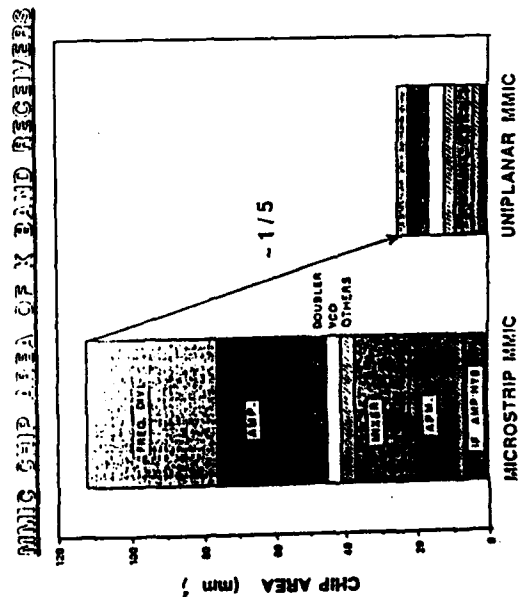
## Characteristics of Air-Bridge



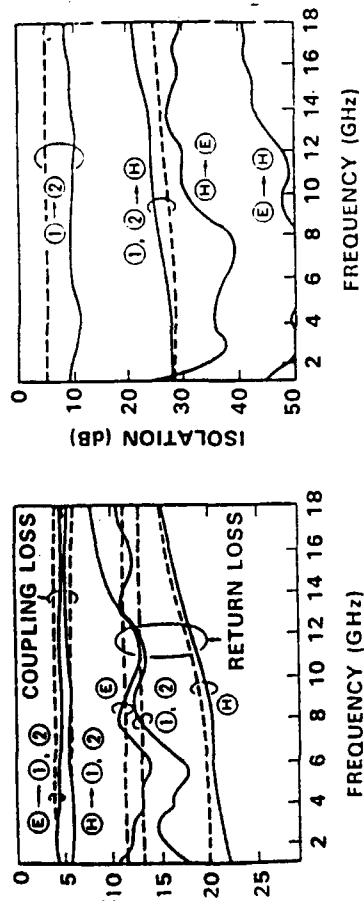
LINE-UNIFIED FET (LUFET)



Photograph of Magic T LUFET



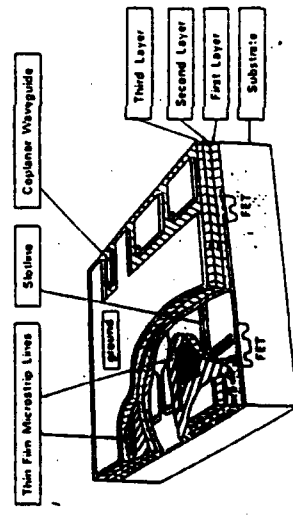
## Line-Unified FET MMIC Magic-T Performance



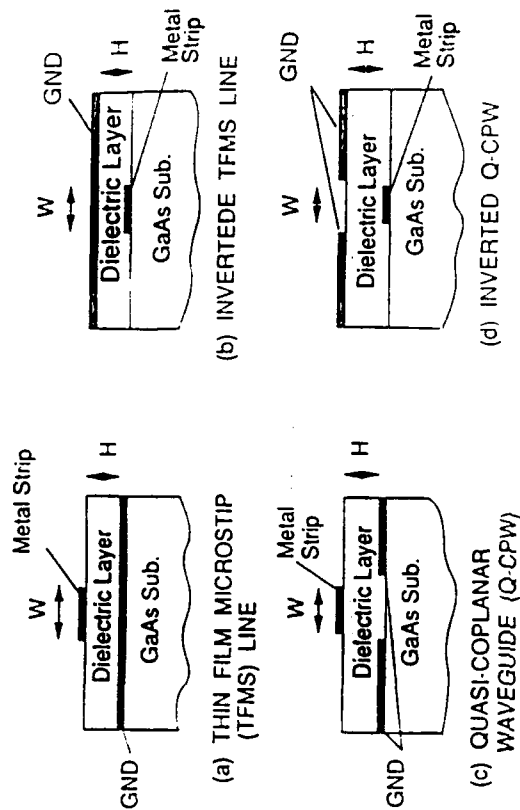
## MULTI-LAYER MMIC

Purposes are as follows:

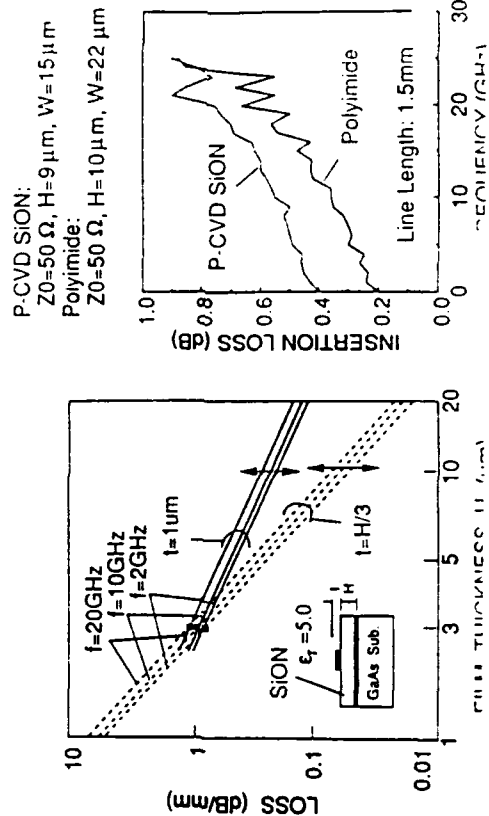
- (1) Substantially reducing MMIC chip size,
- (2) Making high design flexibility for advanced MMICs possible,
- (3) Achieving highly integrated MMICs.



## MINIATURE TRANSMISSION LINES FABRICATED USING A THIN DIELECTRIC FILM

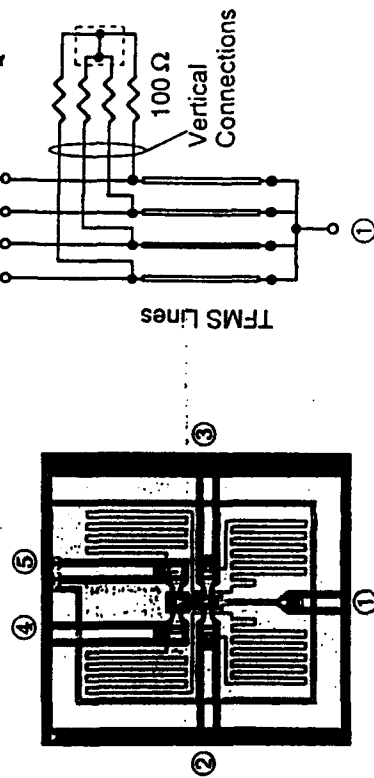


## LOSS CHARACTERISTICS OF TFMS LINE

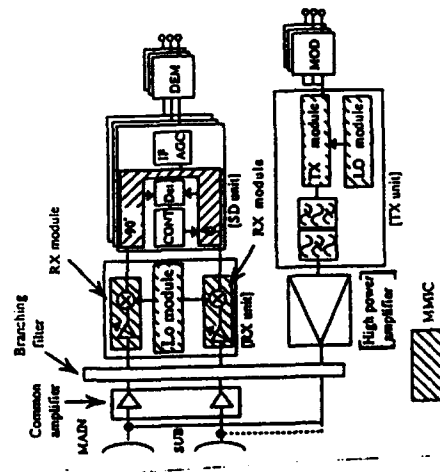


# A 12GHZ, 4-PORT WILKINSON DIVIDER MMIC USING VERTICAL CONNECTIONS

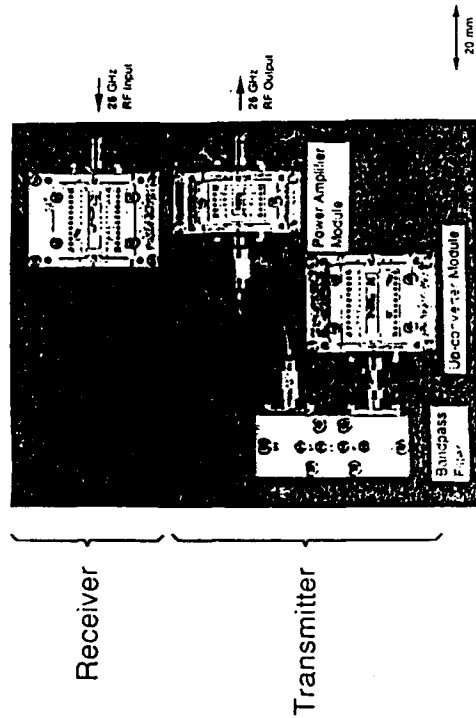
SIZE: 1.2 x 1.0 mm<sup>2</sup>



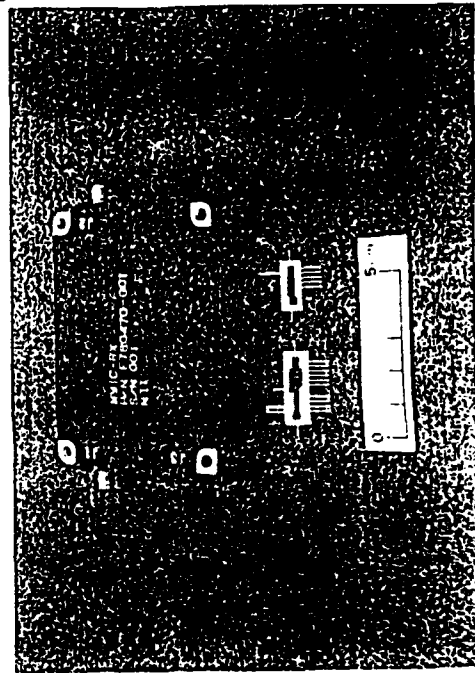
## 4/5/6 GHz-Band MMIC Transmitter & Receiver for 16QAM Digital Microwave System



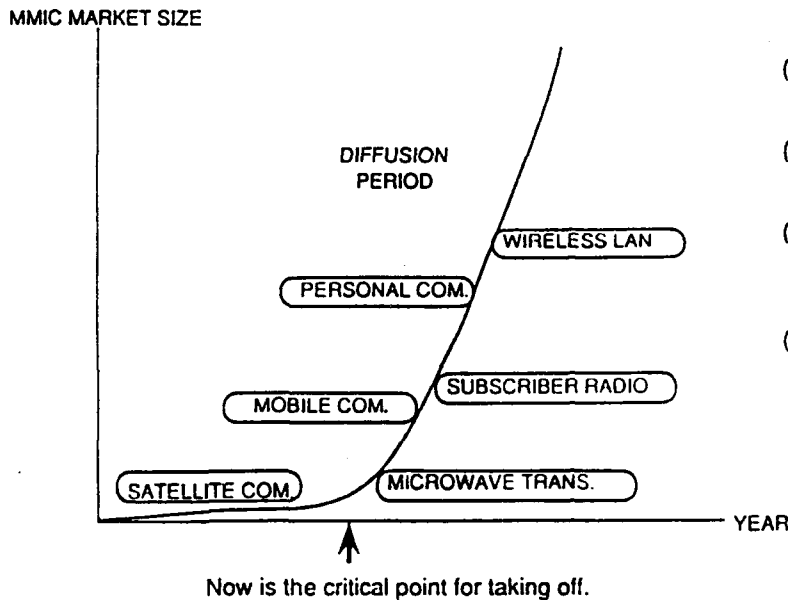
## 26 GHz-Band Full MMIC T/R Breadboard Model



## 30GHz-BAND MMIC RECEIVER FOR ETS-VI



## Prospects for MMIC Applications in the Communication Field



- (1) MMICs are now at the critical point for taking off.
- (2) The initial application was satellite communication systems.
- (3) Presently, digital microwave transmission systems are boosting the popularity of MMICs.
- (4) The principal accelerators will be mobile communications, subscriber radio, wireless LANs etc..

## CONCLUSION

1. Uniplanar MMICs and LUFET MMICs as well as highly-integrated MMICs are remarkably effective for the cost problem, and multi-layer MMICs are also very promising for the next generation.
2. In addition, the uniplanar-, LUFET- and multi-layer MMICs are very suited to high-frequency-band applications.
3. Performances / functions unique to MMICs and Compounded MMIC technology should be studied and developed more eagerly in order to overcome the disadvantages of MMICs, and to promote wider, more active MMIC applications in the communication field.

**NEW INTEGRATED SUBSYSTEMS FOR MILLIMETER -  
WAVE APPLICATIONS**

**R. Funck**

**Dassault Electronique • France**

# NEW INTEGRATED SUBSYSTEMS FOR MILLIMETER-WAVE APPLICATIONS

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## INTRODUCTION

The technology of millimeter wave circuits has been developing for many years and is now available for the realization of complex subassemblies to be used in various applications. The classical technologies are waveguide technology, finline or E-plane circuits, microstrip circuits, and are often used jointly in the same subsystem, depending on the performance and cost objectives.

Historically, the first millimeter wave applications have been for military (radars, sensors, seekers) and space (radiometers, communications) purposes. They have allowed, after the use of waveguide components, the development of planar and quasi-planar hybrid integration technologies (microstrip, finline or E-plane circuits). We can observe that, today, the development of new integrated subassemblies using these basic technologies is still continuing.

The emergence of new applications, especially civilian, and the search for cost reductions, have favoured the appearance of new principles of integration, among which the most promising are quasi-optical techniques.

Beside the development of hybrid circuits, much work is ongoing in the field of monolithic millimeter wave circuits development. Two types of technologies are concerned: Microwave Monolithic Integrated Circuits on GaAs (MMICs) and Silicon Monolithic Millimeter Wave Integrated Circuits (SiMMICs), silicon having very interesting properties at such high frequencies.

This paper will first describe the new developments of millimeter wave subsystems for specific applications. After that, it will be focussing on more general or multi-applications developments.

## RADARS AND SENSORS

The integrated subassemblies developed for these applications generally use a combination of waveguide, E-plane and microstrip circuits.

The finline or E-plane technique was one of the most promising hybrid integration techniques and research was being carried out worldwide [1], as this technology had emerged as a front-runner. The finline structure, which is a planar printed circuit enclosed in a waveguide like split-block housing, has relatively low loss, and the conductor patterns are on circuit boards, which can be made reproducible using conventional photolithographic techniques. Entire systems were designed and realized using finline technology. We will describe some recent developments made at Telefunken System Technik of Ulm, Germany [2,3,4].

The block diagram of a polarization agile 94 GHz monopulse radar front-end for interrupted FM-CW operation with 600 MHz bandwidth [2] is given in fig. 1. The transmitter part of the radar front-end is formed by a 500 mW pulsed IMPATT oscillator, which is injection synchronized by a bias tuned CW Gunn oscillator. A high isolation SPST switch with 60 dB isolation and 1.6 dB insertion loss is inserted between Gunn and IMPATT oscillators in order to ensure maximum receiver sensitivity during pulse intervals. Transmit-receive duplexing is carried out by E-plane circulators, integrated in common housings with finline balanced receiver mixers and IF amplifiers. The technology used for the design of the front-end is a combination of E-plane and integrated waveguide millimeter block technology, enabling straight forward design and manufacturing of integrated subsystems in the frequency range 30-150 GHz. The front-end overall design and its separation into modules suitable for easy production and testing is shown in fig. 2. The global front-end performances are: 270 mW output power, 4 dB receiver conversion gain.

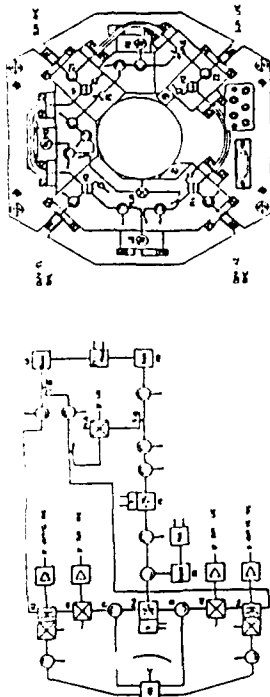


Fig. 1: Block diagram of a 94 GHz FM-CW radar front-end with polarization agility [2]

The integrated millimeter block technology has been used for the design of an active passive sensor front-end in the 90 GHz range [4]. This combined radar/radiometer sensor, the block-diagram of which is given in fig. 3 and its photograph in fig. 4, has been developed for a low cost application. The Gunn oscillator provides transmitting power (30 mW) as well as local oscillator power for the mixer. The received signal is converted down to video frequency which can, after being amplified, be separated by filtering into radar and radiometer channels. The entire RF unit is very compact and measures only 20 mm x 20 mm x 23 mm, and has a mass of 16 g. A number of complete subsystems have been successfully tested at accelerations reaching 1600 g.

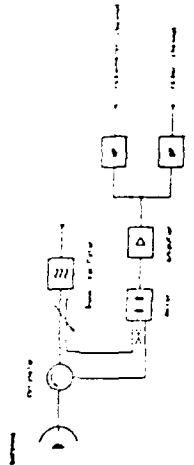


Fig. 3: Block diagram of a 90 GHz active passive sensor front-end [4]

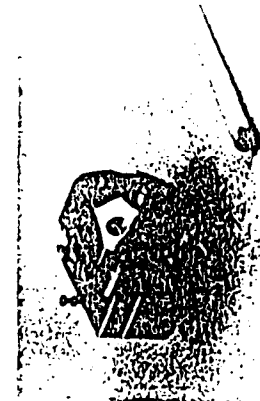


Fig. 4: Active passive sensor [4]

These sensors were originally developed for military purposes. Today they are under consideration for automotive radar applications, due to their potential low cost.

Microstrip technology has been used by GEC, U.K., for the production of millimeter wave circuits for more than 10 years. Its development progressed upward in frequency, with 35 GHz components

being developed in the mid 1970's and 54 GHz components in the 1980's. Microstrip components are now available to operate up to 220 GHz [5]. This technology uses hard substrates such as single crystal quartz or glass ceramics, which allows the insertion of ferrite pucks for iso-circulators and provide compact rugged assemblies.

The fig 5 gives an example of a fully integrated 54 GHz FM-CW transceiver using thin microstrip technology [6], which has been designed specifically for an anti-armour sensor requirement. The transceiver is realized in a volume of 1 cubic inch. A single quartz microstrip circuit contains all the millimeter wave components, except the VCO, which is in a waveguide cavity. The transmit power is approximately +8 dBm, and the system overall noise figure is less than 16 dB.

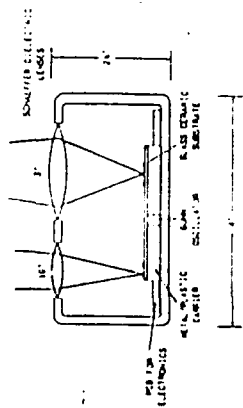
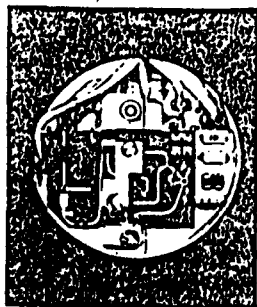


Fig 5. 54 GHz FM-CW microstrip transceiver [6]

Fig 6. Proposed structure of a 54 GHz microstrip unit [7]

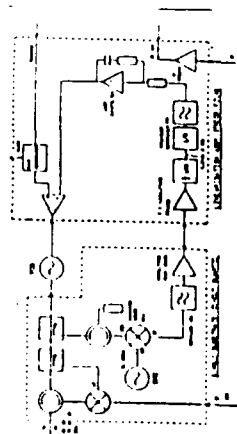


Fig 7. Block diagram of the cruise control unit [7]

The same technology has been used to design a millimeter wave system for cruise control on automobiles [7]. This system is operating at 77 GHz and comprises an FM-CW radar realized in thin film microstrip technology and a switched beam quasi optical antenna. Active cruise control enables the driver to engage cruise control when travelling at a 'driver determined' safe distance behind a vehicle. As the selected vehicle speeds up or slows down, your vehicle tracks this change by accelerating or decelerating. A proposed structure of the control unit is given in fig 6. The system (fig 7) is a basic FM-CW radar based on a varactor tuned waveguide Gunn oscillator using the radial disk geometry with extraction of the second harmonic (32 mW output power, 1 GHz tuning range, -85 dBc/Hz phase noise at 100 kHz off carrier). The received signal is mixed with the oscillator sampled with a 10 dB coupler to derive LO power. A further sampling coupler is used in a FM feedback loop which is used to linearize the ramped output of the oscillator; the 77 GHz output signal is downconverted in the 850-1150 MHz range with a subharmonic mixer driven by a 14 GHz DRO and applied to a 10 MHz frequency discriminator through a 1/100 frequency divider. All the microwave components are located on a single

thin film microstrip circuit. A quasi optical concept is used for the antenna: patch antennas on the microstrip circuit are used to illuminate a dielectric lens. By illuminating the lens from different patches, a beam steering effect can be achieved. The dielectric lens forms an isometric soal can be produced as part of the housing.

Microstrip technology has also been used for a millimetric sensor for stand-off weapons realized by Marconi Defence Systems U.K. [8]. The received signal is fed from the antenna to the receiver via a quasi-optical feed. This is capable of providing the necessary R.F. input to the receiver so that co- and cross-polar sum, elevation and azimuth error channel data may be simply extracted. The R.F. receiver uses quartz microstrip techniques which enables the integration of all the passive microwave components (mixers, isolators, couplers, etc.) on a single circuit. The L.O. uses a Gunn diode in a waveguide cavity (10 to 15 mW of output power). The transmitter uses an IMPATT in a waveguide cavity (300 mW) which is injection locked to a Gunn diode VCO (20 to 30 mW) swept in a linear fashion. The linearity of the sweep is controlled by the use of a closed loop system which maintains the frequency ramp at a given rate.

A very interesting concept has been described by the RSRE, U.K. [9]. This paper describes a staring, multibeam, local-plane array technology and radar development which has evolved from research into lens-fed microwave and millimeter wave receivers with integral antennas. The key components are a high permittivity dielectric lens and a local-plane array of receivers in intimate contact with the lens (Fig 8). The lens generates a multiplicity of simultaneous narrow beams, each with its own direction of look, covering between them the total field of view. Each beam is focussed to a unique position on the lens local-plane. Combining this with an array of receivers in the local-plane creates an array of high gain receivers which collectively stare across a wide field of view, with no moving parts and without separate beam forming circuitry. Each receiver is built round a crossed-dipole antenna, in which one dipole receives the signal and the other the local oscillator. This provides a convenient means to drive a quad diode mixer (Fig 9). Receiver arrays of dipole coupled mixers replicated on a single high permittivity substrate have been realized: a 35 GHz 4x4 microstrip silicon array and 5x5 hybrid arrays on thin-film have been realized. Typical noise figure of receivers of this type were measured to be 8.5 dB including 2.0 dB IF noise. A 35 GHz heterodyne radar demonstrator has been built, incorporating an array of nine 5x5 thin-film receivers. As can be seen, this is a good example of a system combining different technological concepts: microstrip circuits on silicon and quasi-optical techniques.

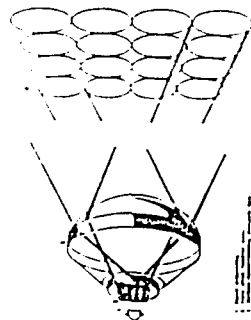


Fig 8. Staring array receiver [9]

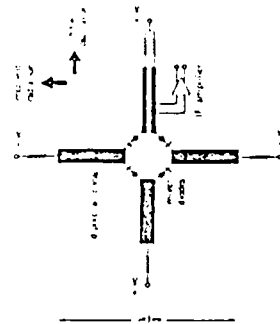


Fig 9. Quad diode mixer [9]

## COMMUNICATIONS

The main applications are for in-door communications or communications with mobiles at 60 GHz, and for satellite communications. The technologies which are used are on the same type as previously described.

T.S.T. has designed a 2 W solid state transmitter for short range data communication at 60 GHz [10] composed of 3 stages of amplification. The first uses a reflection type Gunn diode amplifier (20 mW), the second an injection locked Gunn diode oscillator (150 mW) and the third an injection locked

GaAs IMPATT oscillator (2 W). The block-diagram of this amplifier is given in fig. 10. The breadboard has been realized in waveguide technology.

The work done by Siemens Telecommunications in Italy [11], concerns commercial inter-orbit and inter-satellite communication applications. In this sense, advanced studies are in progress in order to evaluate a possible experimental mission of a 59.64 GHz communication payload. Their work demonstrates that resistive mixers present a reliable and performing solution for mm-wave low noise mixers. The mixer which has been realized uses a cross-bar configuration which allows a clear separation of the various LO sideband harmonics, as well as image recovery. Tests have been made at 40 GHz with a 4 GHz IF. Insertion loss is about 3.2 dB in a 1 GHz bandwidth (fig. 11), showing clearly the performance improvement due to image frequency recovery. It is expected that at 65 GHz, the conversion loss would be 3.7 dB.

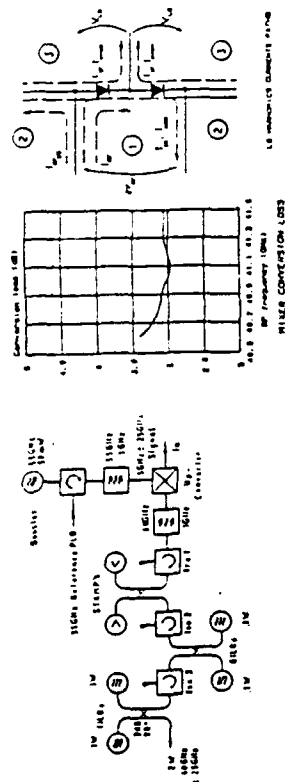


Fig. 10: Block diagram of the 2W transmitter [10]

The European AIMS/ESPRIT programme, led by Thomson-CSF, France, addresses components up to subsystems level for applications between 20 and 30 GHz. System pulls are from short-hop land links and ultra-small aperture satellite terminals (USAT). A range of generic MMICs are being developed, covering switches, VCOs, solid-state amplifiers. Apart from the T/R switch, which makes use of MESFETs, all other circuits are based on the use of heterojunction transistors: HEMT, P-HEMT, HBT. These activities will be briefly described in a next paragraph. Within this project, Thomson-CSF/TCM is developing a 27.5-30 GHz synthesizer. The fig. 12 and 13 represent respectively the microwave part of the synthesizer and the 30 GHz MMIC amplifier chip. The main performances are: 10 dBm output power, less than -70 dBc/Hz phase noise at 100 kHz off carrier, and 27 MHz frequency steps. The subassembly consists in cascaded buffer, frequency divider, buffer, VCO, power divider, frequency multiplier, amplifier. At this step, the functionality has been proven with conventional thin-film hybrid technology to interconnect individual MMIC chips. Later on, compact versions will be designed together with comparing improvements gained from advanced MMIC technology.



Fig. 12: 30 GHz synthesizer for AIMS

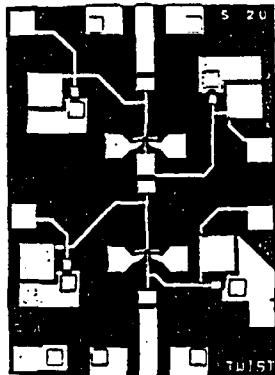


Fig. 13: 30 GHz MMIC Amplifier [19]

## AUTOMOTIVE APPLICATIONS

Many programs related to automotive applications are now under study in Europe. In particular, there are European projects like DRIVE and EUREKA. The main applications are: automatic debiting systems, Doppler speedometers, vehicle-vehicle and vehicle-road communications, obstacle detection. These programs could lead to large microwave applications if two conditions are satisfied: low cost technology compatible with mass production, low size and low weight design to fulfill the implementation requirements.

The search for low cost and low volume at millimeter wave frequencies may force the engineer to simplify components and to combine functions whenever possible. The use of MMICs could be appropriate, but these components don't solve the problem of circuit integration.

Rather than scaling low frequency techniques to shorter wavelengths, a completely new approach is often needed. Quasi-optical components provide a solution to this problem.

An example of realization of a radar sensor for automobile cruise control, directly derived from radar developments, has been described above [7].

T.S.T., Germany, has developed a very simple radar sensor for traffic data acquisition system named AVES [12]. This system obtains specific traffic information from appropriate Doppler frequency. The chosen frequency of operation, 61.5 GHz, offers several advantages: the band is available for this application in Europe (ISM-band, Industrial, Scientific and Medical). Due to the high frequency, small antenna dimensions allow good focussing, short range application and economic reuse of frequencies due to high atmospheric absorption. The sensor is relatively small (150x150x150 mm<sup>3</sup>), and can be mounted on existing sign posts and bridges. The schematic set up of the mm-wave front end is presented in figure 14. Hybrid integration was implemented, combining inline and waveguide to reduce cost: this unit is composed of a cavity second-harmonic type Gunn oscillator, a waveguide filter and a inline mixer. Vehicle speed, vehicle length and the distance between vehicles is obtained by the AVES sensor. A future application of this type of sensor can be seen in vehicle-roadside communications.

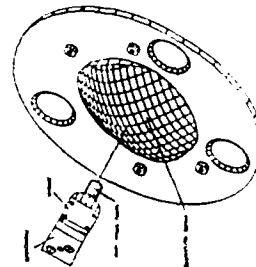


Fig. 14: The AVES system [12]

A first example of the quasi-optical approach is given by the design of ALDER from the RSRE [3]. The University of Lille, France, presents an other circuit approach based on a quasi-optical planar integration [13] which suppresses the physical connections between functions, the transient of energy being achieved either by electromagnetic radiation or by electromagnetic coupling. This allows the realization of low cost active antennas which can be coupled to Fresnel lenses to fulfill the various directivity requirements. The concept of this antenna is given in fig. 15. The transmitter is composed of a circular patch fed by two terminal devices (Gunn or IMPATT diode) in a  $TM_{11}$  radiating mode. The insertion location of the diode is chosen to optimize its matching. Electronic frequency tuning is provided by a varactor which is implemented in parallel inside the circular patch cavity, the value of the coupling depending on the distance from the disc center. The receiver is realized by the means of a Schottky diode mixer integrated on the patch. The principle is to isolate a strip following the  $TM_{11}$  mode current lines, the receiver diode being connected in serie on this strip. The concept has been demonstrated at X-band. Some modifications have been made for mm-wave applications, in order to increase the size of the antenna to allow the use of packaged Gunn or IMPATT diodes. This is achieved by coupling a  $TM_{11}$  circular patch antenna to a  $TM_{12}$  annular patch antenna by means of a circular slot antenna as illustrated in fig. 16. In addition, simple electronic beam forming can be obtained by proper perturbation of the electromagnetic fields : scanning of the beam can be achieved at  $\pm 10^\circ$ . Two anticolision radar prototypes have been developed : a pulsed 60 GHz Gunn radar which gave excellent results for a 150 m range, and a 86 GHz FH-CW IMPATT radar which exhibited lower sensitivity due to the IMPATT diode noise.

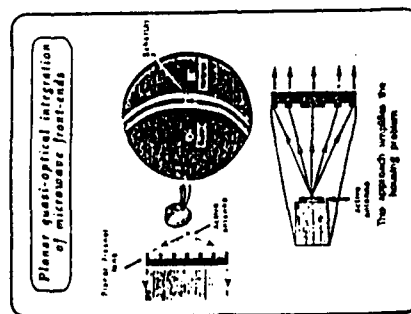


Fig. 15 : [13]

## QUASI-OPTICAL

Quasi-optical techniques can also be used as an elegant way to improve the power handling by coupling many devices distributed over a surface several wavelengths in dimensions [14]. This technique is referred to as quasi-optical array or grid technique. This technique can be applied to oscillators, multipliers, phase shifters and mixers. The fig. 17 illustrates the principle of a grid oscillator.

Theoretical modeling of grid power combiners is being made by the LEST, France [15]. Quasi-optical techniques are also well adapted to the realization of filters. The LEST is studying planar grating-reflector antennas [16] on a theoretical point of view. Dichroic filtering structures are being realized for radar applications by MATRA ESPACE, France, [17]. Such structures are transparent or reflective depending on the wavelength of the incident electromagnetic wave. A dichroic

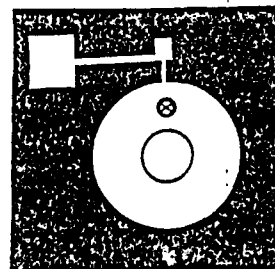


Fig. 16 : Integration of the emitting diode [13]

triplexer using resonant grid structures etched on a fused silica substrate has been realized demonstrating very low loss : insertion loss at 90 and 157 GHz is 0.2 and 0.3 dB, reflexion loss at 183 GHz is 0.2 dB. A different approach is used by the Institut National de Télécommunications and the Observatoire de Meudon, France, to be used for Météosat satellite [18]. The dichroic triplexer are realized by using perforated screens. The performances obtained are : 0.54 dB insertion loss at 183 GHz, 0.05 and 0.11 dB reflexion loss at 150 GHz and 110 GHz respectively.

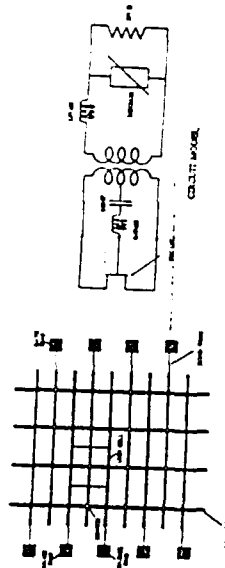


Fig. 17 : Planar grid oscillator

## MONOLITHIC INTEGRATION

Two technologies are used for the realization of monolithic integrated circuits and subsystems : Microwave Monolithic Integrated Circuits on GaAs (MMICs) and Silicon Monolithic Millimeter Wave Integrated Circuits (SiMMWICs).

### MMICs

Three ESPRIT programs are concerned with MMICs developments for mm-wave applications. AIMS addresses components and subsystems for applications between 20 and 30 GHz. The fig. 13 shows a 27-40 GHz amplifier from Thomson, which is used in the synthesizer described above. The gain is 15 dB, noise figure 14dB and size 1.5 mm<sup>2</sup>. This amplifier uses a 0.25  $\mu$ m pseudomorphic technology [19]. The same laboratory has designed a 60 GHz LNA having a 12 dB gain and a 5 dB noise figure.

MONOFAST is focussed at 44 GHz. The overall aim of this project is to integrate unfamiliar technologies and new design tools into a high proven methodology for high frequency MMICs. An MMIC LNA at 44 GHz based on sub-micron MESFET and coplanar waveguide technology is being designed by Alcatel Espace, France [20] : a gain of 15 dB and a noise figure less than 4 dB have been simulated.

GIANTS is concerned with the development of InGaAs based transistors where one of the aims is to demonstrate a low noise MMIC amplifier at 60 GHz.

Philips Microwave Limel, France, has developed several mm-wave MMICs using a pseudomorphic HEMT technology [21, 22]. A push-push oscillator using a 0.5  $\mu$ m HEMT has been realized demonstrating a -3 dBm output power and a phase noise of -75 dBc/Hz at 1 MHz off carrier at 52 GHz. An image rejection mixer (fig. 18) designed in the 52-60 GHz frequency range and using a 0.25  $\mu$ m HEMT has a typical conversion loss of 14 dB and an image rejection higher than 30 dB for a 2-4 GHz IF frequency range.

In Germany, Telefunken System Technik is also engaged in an important mm-wave MMICs development program [23, 24, 25]. Using the integration of MESFET devices ( $I_{max} = 90$  GHz) and Schottky diodes ( $FI = 2300$  GHz) on the same chip, mixers working at 35 GHz and 60 GHz have been realized. The 35 GHz mixer, integrating a 2-stage IF amplifier, has a 6 dB conversion gain with a 9 dB DSB noise figure. The 60 GHz receiver chip has a 2-stage IF amplifier : the conversion gain of this chip is 12.5 dB with an associated DSB noise figure of 9.5 dB. The fig. 19 represents the 60 GHz receiver chip with a chip size of 3x3.5 mm<sup>2</sup>. The mixer itself consists of a Lange-coupler feeding two Schottky

diodes and LO and signal band stop filters at the IF output. The diodes have a finger width of 0.3  $\mu\text{m}$ . The low noise IF amp further employs two MESFETs with a gate length of 0.75  $\mu\text{m}$ . Further developments have been made using a PHEMT technology which has proven to be superior to all other available monolithic technologies for both advanced low noise and power circuits in the mm-wave frequency range. Quarter-micron HEMTs with a  $f_{\text{max}}$  of 200 GHz have been developed and fabricated. Using these devices, a 2-stage low noise amplifier has been fabricated having a gain of 5 dB at 53 GHz.

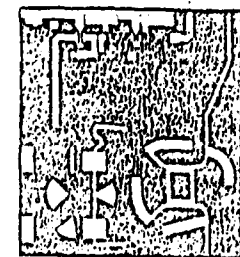


Fig. 18. 50 GHz receiver chip (100x)

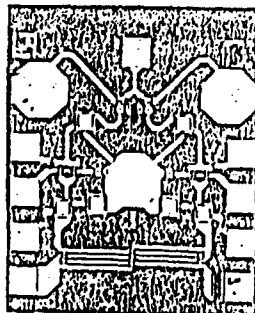


Fig. 19. 50 GHz transmitter chip (100x)

## SIMULICS

The attention of transmission lines based on high resistivity silicon (1000 Ohm-cm) is dominated by conductor losses above 40 GHz. On the other hand, silicon substrates allow the integration of two terminal devices such as IMPATT diodes, Schottky and FET diodes, PIN diodes. Therefore, silicon is also a good substrate for the design of monolithic integrated circuits in the millimeter-wave range.

Monolithically integrated 60 GHz transmitters have been fabricated by Daimler-Benz Research, Germany [10]. The substrate is a high resistivity silicon (1000 Ohm-cm) in a radiating slot antenna structure on a high resistivity silicon substrate (Fig. 20). This substrate is realized using the so-called "air" process. For simplicity, the resonator and the antenna are merged into one structure as a resonator. Thus no additional antenna structure is needed, the chip area can be drastically reduced and coupling losses between antenna and resonator are avoided. Additionally there is no need for vias and backside metalization. The IMPATT diode is integrated in the center of the slot, which has a length of one substrate wavelength. For biasing the diode, two small slots are provided a distance of one wavelength apart from the ends of the slot line resonator. For beam focusing in a horn antenna, an additional diaphragm is provided (up to now, 2.6 mW CW and 50 mW for short pulses with 50 ns pulses, duty cycle 0.5%) (have been observed at 60 GHz). The spectral purity is satisfactory, spurious noise is typically -30 dBc/Hz at 500 kHz offset.

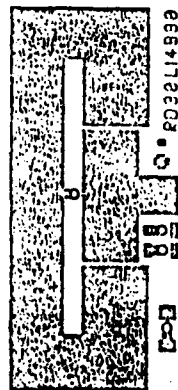


Fig. 20. 50 GHz transmitter chip (100x)

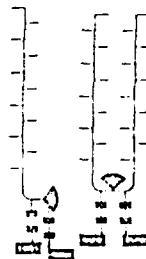


Fig. 21. Schematic diagram of a 50 GHz transmitter

Planar millimeter-wave silicon detector circuits have been realized by the same laboratory [27]. Schottky barrier diodes have been monolithically integrated with planar antenna structures on high resistivity silicon. The Schottky barrier diodes were fabricated by MBE: low series resistance (65 ohms), ideality factor of less than 1.1 and cut-off frequency up to 1 THz have been achieved. The Fig. 21 gives the layout of two types of receiver which have been realized. With the two antenna structures, detection sensitivities of 40.8 mV/(mW/cm<sup>2</sup>) without bias, and of 90 mV/(mW/cm<sup>2</sup>) with 260 mV bias, have been achieved at 4 GHz.

## CONCLUSION

In reviewing the progress made in the last years in millimeter-wave substrates, we have seen that the search for low cost and large volume quantities for commercial applications such as communications, target control, collision avoidance, etc., has favored the emergence of new principles of integration, in parallel with the development of millimeter-wave circuits.

Millimeter-wave applications and technologies, which have been initiated and pushed by military needs first, are now available for new commercial applications.

## ACKNOWLEDGEMENTS

Thanks to colleagues at Alcatel Espace, Daimler-Benz Research, GEOS-Patent, Marconi Defence System, Siemens Telecommunication, Tealunken System Technik, Thomson-CSF, Université de Lille, for supplying data and information and materials.

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**NEW MILLIMETER - WAVE DEVICES REALIZATION  
AND PERFORMANCE**

**D. Pons**

**Thomson - CSF LCR • France**

**FIELD - EFFECT TRANSISTORS FOR THE MILLIMETER -  
WAVE RANGE: PHYSICAL ANALYSIS, MODELING AND  
EXPECTED PERFORMANCE**

**G. Salmer**

**Université de Lille • France**

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**FIELD EFFECT TRANSISTORS  
FOR THE MILLIMETER WAVE RANGE**

**Physical Analysis - Modelling  
and Expected Performance**

**G. SALMER**

**I.E.M.N. - U.M.R. CNRS 9929**

**Département Hyperfréquences  
et Semiconducteurs**

**Université des Sciences et Techniques de Lille  
Villeneuve d'Ascq - France.**

**With the help of**

- E. CONSTANT**
- Y. CROSNIER**
- A. CAPPY**
- R. FAUQUEMBERGUE**
- P. BOURREL**

**C.H.S.**

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**OUTLINE**

**• PHYSICAL DEVICE MODELS**

**• LOW NOISE DEVICES**

- Noise simulation**
- MESFETs**
- MODFETs on GaAs Substrate**
- MODFETs on InP Substrate**
- Discussion**

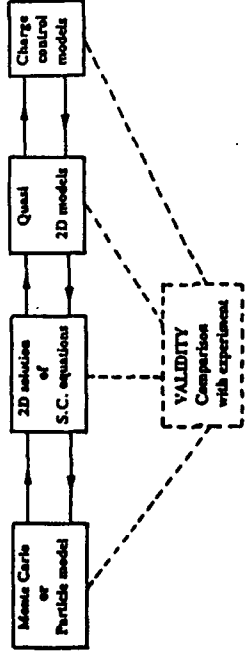
**• POWER DEVICES**

- Power requirements**
- Breakdown phenomena**
  - Power FETs.**

**C.H.S.**

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## PHYSICAL MODELLING OF F.E.T.: METHODOLOGY



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## MONTE CARLO OR PARTICLE MODELS

### Principle

The stochastic motion of the carriers is studied simultaneously:

- in the 3D  $k$  space: for each time step  $\Delta t$ , are considered
  - the effect of the electric field  $\cdot \vec{A}k = \frac{q}{\hbar} \vec{E} \Delta t$
  - the effect of scattering mechanisms.
- in the 2D  $r$  space: that gives the two dimensional description of
  - electric fields
  - carriers concentrations
  - carrier velocities

from which can be deduced -external currents and voltages  
-equivalent circuits

Remark: other possibilities: Direct solution of the Boltzmann transport equation.

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## MONTE CARLO OR PARTICLE MODELS

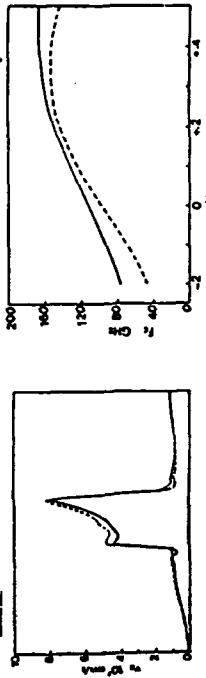
- Advantages:
  - S.C. Equations automatically solved
  - All physical effects can be included
  - transport properties
  - Determination of all device properties.
- Requirements:
  - Accurate description of energy band diagram
  - Knowledge of various scattering mechanisms
- Problems:
  - Bad knowledge of several effects: screening, ionization
  - Transport properties unknown for several ternary or quaternary compounds
  - Introduction of quantum effects
- Drawback:
  - Very long computational time
  - Powerful computers
  - Difficult to study dynamic regimes.

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## MONTE CARLO SIMULATION OF A AlInAs/GaInAs MODFET:

PARAMETRICAL STUDY: INFLUENCE OF AlInAs LAYER CHARACTERISTICS

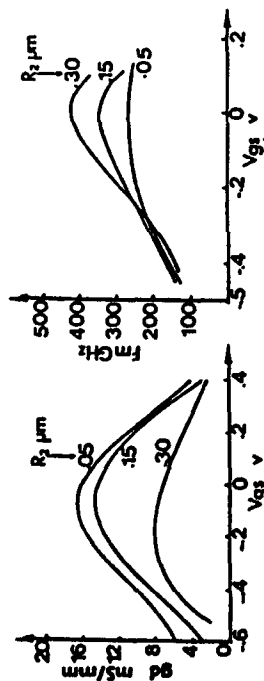


Longitudinal velocity  
AlInAs layer:  $N_D = 4 \cdot 10^{18}$   $a = 75$  Å  
 $N_D = 2 \cdot 10^{18}$   $a = 125$  Å

from Bourrel and Fauquembergue, 1991

C.H.S.

# INFLUENCE OF GATE TO DRAIN CAP LAYER SPACING $R_2$



Output Conductance Maximum oscillation frequency  
for  $L_g = 0.3 \mu$   $R_1 = 0.05 \mu$

Following Shauki and Salmer (1990)

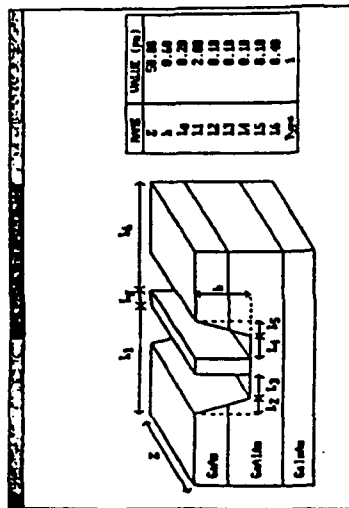
## QUASI TWO DIMENSIONAL MODELS: ONE DIMENSIONAL SOLUTION OF SEMICONDUCTOR EQUATIONS

- **Main assumptions:**
  - Equipotential lines perpendicular to the source to drain axis  
→ one dimensional treatment
  - Depletion regions are fully depleted.
- **Main advantages:**
  - Computational effort is strongly reduced
  - Main physical effects can be account for : overshoot, surface, buffer injection, ionization...
  - Parametrical studies are easy
  - Large signal behaviour and noise properties can be deduced.
- **Problems:**
  - Difficulty to treat very complicated structures or two dimensional effects
  - Validity for very short gate devices.

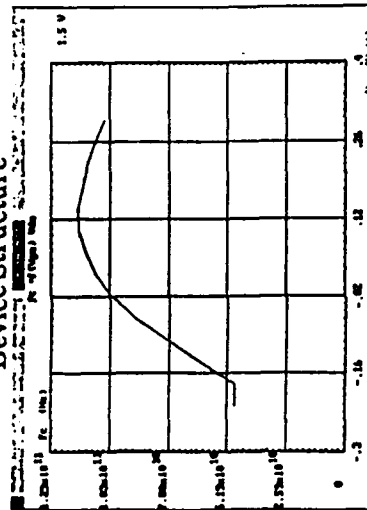
C.H.S.

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## EXAMPLE OF QUASI 2D MODEL HELENA SOFTWARE



### Device Structure



Current gain cut-off frequencies predicted

H. Happy, A. Cappy, G. Dambrine, 1991.

C.H.S.

*iem*

## TWO DIMENSIONAL SOLUTIONS OF SEMICONDUCTOR EQUATIONS

### - Principle:

Two dimensional solution of:

- Equations deduced from B.T.E.: continuity, momentum relaxation, energy relaxation.
- Poisson equation
- By taking into account
- Physical effects and material properties : trapping in the substrate, surface potential, ionization effects
- Parameter values deduced from Monte Carlo simulation:  $\tau$ ,  $\mu$

### - Remarks:

- Various kinds of models corresponding to various degrees of assumptions: single electron gas, drift diffusion models.
- Possibility to account for quantum effects by partially coupling with Schrödinger equations
- Finite difference methods are generally used.

C.H.S.

*iem*

## TWO DIMENSIONAL SOLUTIONS OF SEMICONDUCTOR EQUATIONS

### - Advantages:

- The most complete equations allow to account for non steady state transport phenomena and other physical effects (trapping, surface, ionization...)
- Dynamic characteristics, transient response and microwave performance can be deduced
- Parametrical studies can be made

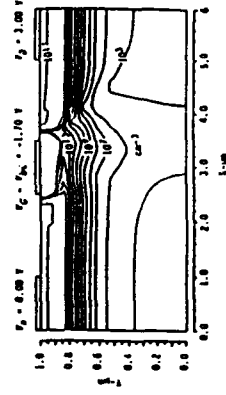
### - Problems:

- Validity of semiconductor equations or simplified assumptions (drift diffusion for submicrometer gates)
- Inclusion of quantum effects
- Treatment of - Schottky diodes
- Highly doped regions: mesh dimensions
- Computation time remains long.

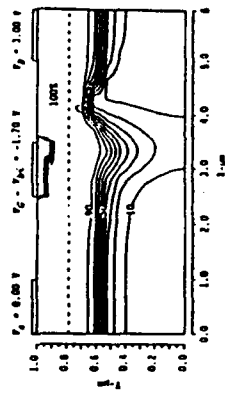
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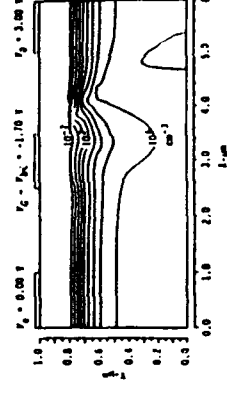
## 2D SOLUTIONS OF SEMICONDUCTOR EQUATIONS - TYPICAL RESULTS: MESFET SIMULATION



Electron concentration



D.L.T. occupancy



Hole Density

Effect of substrate electron injection on acceptor deep trap level occupancy.

From Ghione et al, 1990.

C.H.S.

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**- Principle:**

- One dimensional solution of equilibrium equations (Schrödinger or current continuity equations) following the transversal direction
- Analytical formulations of drain current such as very simplified one:  $J_D \approx q n_d \langle v \rangle$

**Interat:**

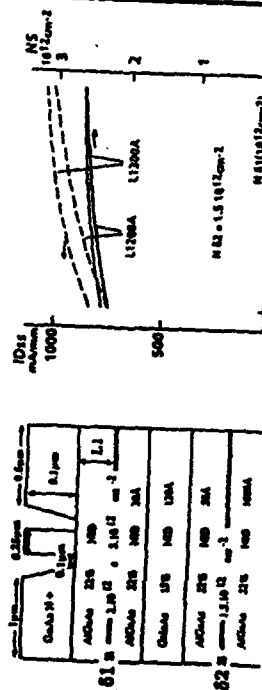
- **Usable for first order optimization**
- **Allows to point out interesting physical dependences**
- **Needs very small computation times.**

### Problems:

- **Validity of simplified assumptions : unrealistic values of saturation velocity**
- **Unable to treat complicated structures.**

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## APPLICATIONS OF SIMPLIFIED CHARGE CONTROL MODEL: COMPARISONS WITH QUASI 2D MODEL



**Considered structure**

From Crandall et al, 1992  
ESPRIT 1993 AIMS

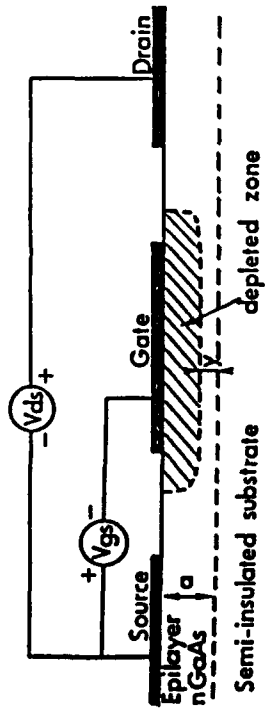
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## LOW NOISE DEVICES

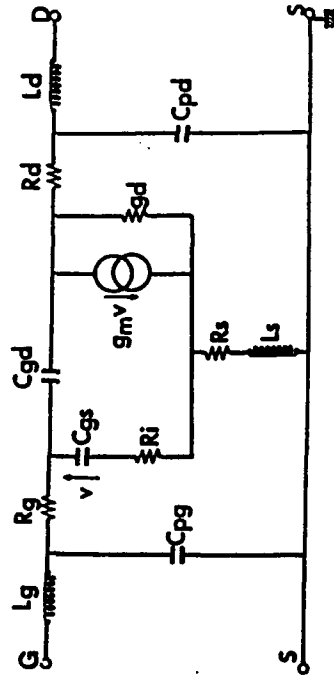
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GaAs MESFET : Basic principle.



Electrical equivalent circuit.



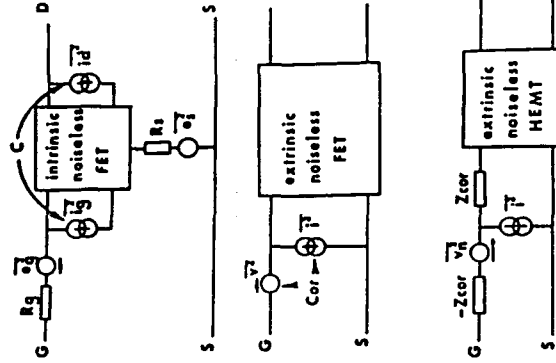
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NOISE CALCULATIONS FOR F.E.T.

Noise origins :

- Carriers velocity fluctuations in the channel that induce - drain current fluctuations  $<i_d> > 2$
- gate current fluctuations  $<i_g> > 2$  that are correlated
- Noise in the access resistances  $R_s$  and  $R_g$ .

Circuit transformations for the noise determination



C.H.S.

*10mm*

## ANALYTICAL FORMULATIONS OF NOISE FIGURE following Cappy

- When neglecting gate noise  $\langle i_g \rangle^2$ :

- $F_{min} = 1 + 2 \sqrt{P} \frac{f}{f_c} \sqrt{g_m (R_s + R_d)}$

with  $P = \frac{\langle i_d \rangle^2}{4kTg_m \Delta f}$  close to "Fukui" formula

- According to Delagebeaudeuf:  $P \approx \frac{I_{d0}}{E_c L_g g_m}$

$$F_{min} = 1 + 2 \sqrt{\frac{I_{d0}}{E_c L_g}} \times \frac{f}{f_c} \sqrt{R_s + R_d}$$

• Consequences: For improving noise figure

- $R_s$  and  $R_d$  must be reduced
- $\frac{\langle v \rangle}{f_c} \frac{1}{2\pi L_g}$  must be as high as possible.

• Remarks:

- 1) The intrinsic noise figure Fint (for  $R_s + R_d = 0$ ) remains always close to 1!!!
- 2) For comparing devices of similar technologies ( $L_g, R_s, R_d$ ), the quality factor can be the ratio:

$$r = \frac{I_{d0}}{f_c^2}$$

- When taking into account gate noise, more complicated formulae with Fint given by

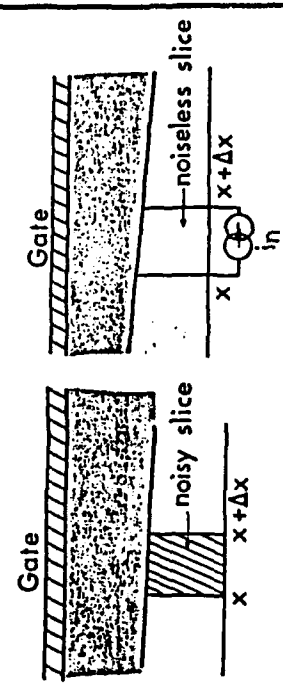
$$F_{int} = 1 + 2 \frac{f}{f_c} \sqrt{PR(1 - C^2)}$$

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*10mm*

## CALCULATIONS OF GATE AND DRAIN NOISE SOURCES

- 1) In a quasi 2D scheme, representation of a noisy slice by a noisy source in proportionnal to  $n(x) D(x) \Delta f$



- 2) Use of the Impedance Field Method

(Van der Ziel, Schockley, Nougier, Cappy, Wang, Ghione)

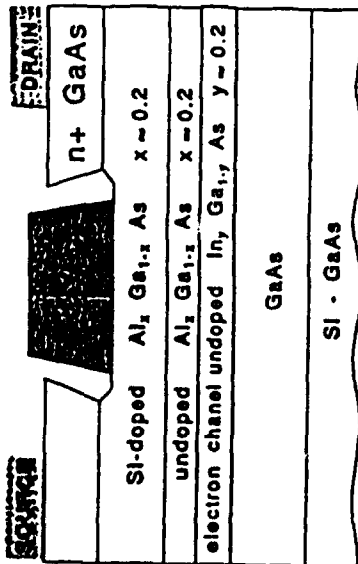
- Calculation of transfer small signal impedances  $Z(\omega, x)$  between noise sources in the channel and drain and gate voltages.
- Summation over the whole length of the channel.

Remarks:

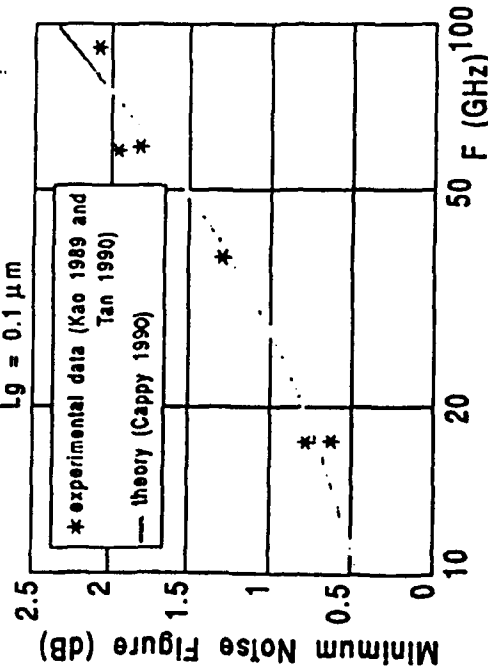
- Interest of taking into account the distributed character of the channel (Cappy et al, 1990)
- Possibility of using a complete two dimensional treatment (Ghione 1990).

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## Noise Performance of High Electron Mobility Transistors (HEMT)



Pseudomorphic AlGaAs-InGaAs-GaAs HEMTs

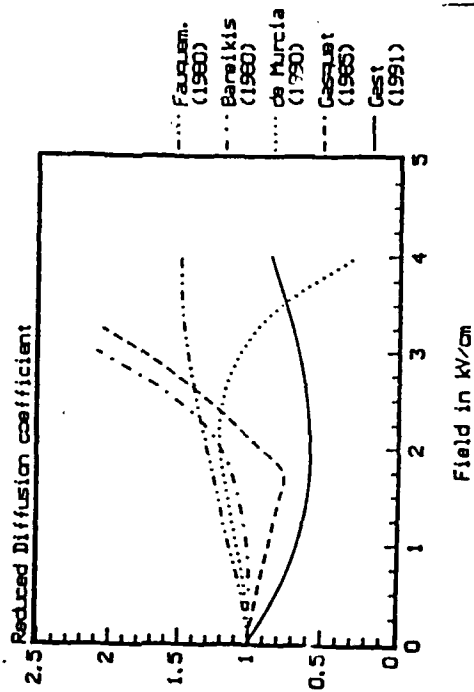


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## NOISE CALCULATIONS : PROBLEMS

- Physical basis : uncorrelation between the noise sources in the channel?  
Does it remain valid for very short gate length devices ( $L_g \approx 500 \text{ \AA}$ )?
- Values of diffusion coefficient are not well known : for instance in AlGaAs/GaAs heterojunction.



From Zimmerman et al, 1991

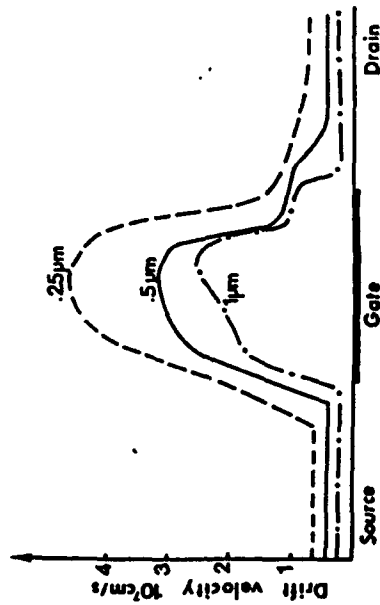
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## LOW NOISE MESFET's

-Improvement of noise figure

- Intrinsic  $f_c = \frac{g_m}{2\pi C_{gs}} \propto \frac{1}{L_g}$  must be increased
- Main possibility. A decrease of gate length  $L_g$  that allows also an increase of  $\langle v \rangle$



Typical variations of drift velocities in MESFET's for various gate lengths (Carnoz et al)

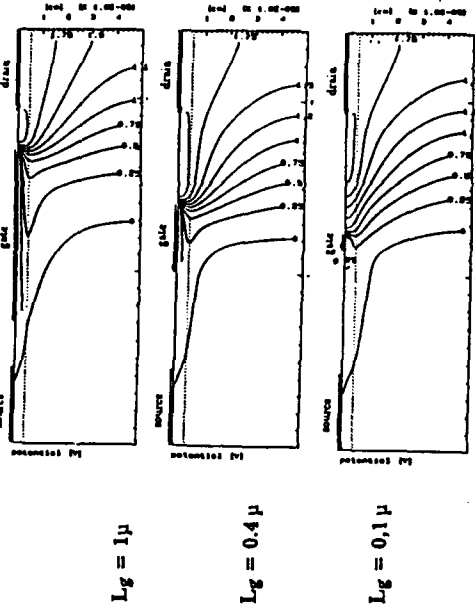
But if the aspect ratio  $L_g/a$  is not high enough, short channel effects may cause a degradation of performances :

When  $L_g$  decreases  $g_m$  may decrease } Gain  $\nearrow$   
 $g_d$  increases  
 $f_c$  is not improved

G.H.S.

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## GaAs MESFET INFLUENCE OF THE ASPECT RATIO



Equipotential contours

| $L_g$<br>$\mu m$ | $L_g/a$ | $g_m$<br>mS/mm | $g_d$<br>mS/mm |
|------------------|---------|----------------|----------------|
| 1                | 19      | 320            | 25             |
| 0.4              | 7.6     | 420            | 45             |
| 0.1              | 1.9     | 380            | 60             |

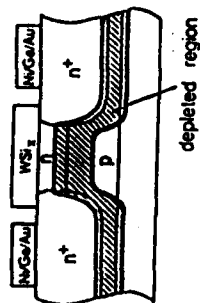
Simulations performed by Makowikz and Brockerhoff.  
 (Europ. Trans. on Tel.N° 4 19 90).

G.H.S.

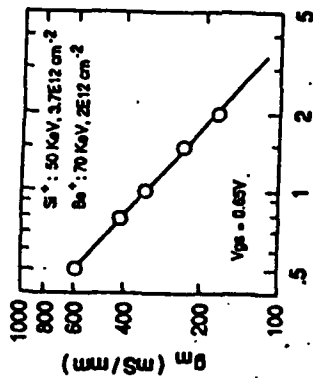
Siemens

## SUPPRESSION OF SHORT CHANNEL EFFECTS

- Use of a partially depleted layer:



Schematic cross sectional view of the MESFET.



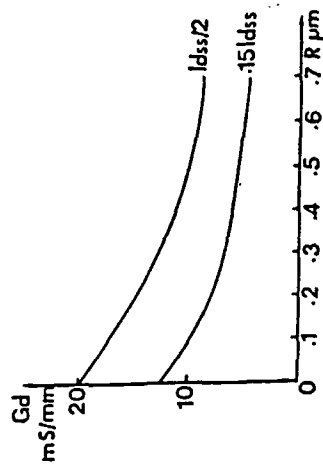
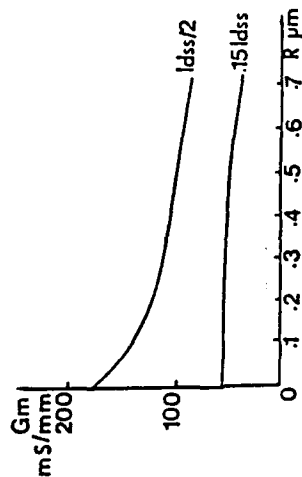
Gate length dependence of  $g_m$ .

From Noda et al (IEEE Trans. 1991).

C.H.S.

C.H.S.

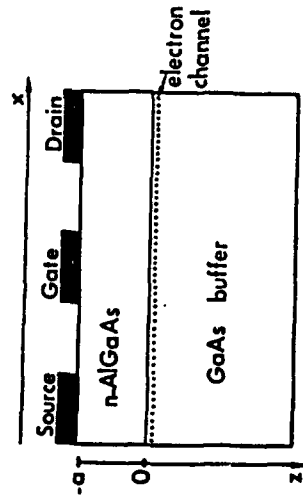
## RECESSED GATE GaAs MESFET



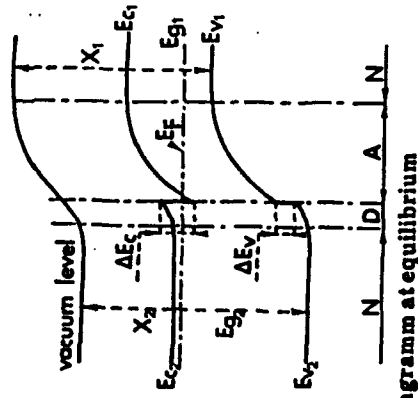
Transconductance and output conductance dependences upon the distance  $R$  for a  $0.3 \mu\text{m}$  gate MESFET at  $I_{dss}/2$  and  $0.15 I_{dss}$

Heliodore et al IEEE Trans. 1988.

# CONVENTIONAL $\text{AlGaAs}/\text{GaAs}$ MODFET's

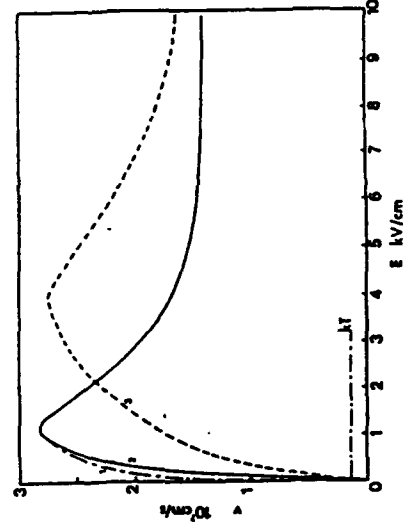


- Device Structure



- Band diagram at equilibrium

CHS Lille



Theoretical velocity field characteristics at 77° K

*iemn*

## VERY SHORT GATE LENGTHS MESFET's

Other possibilities for improving the performance:

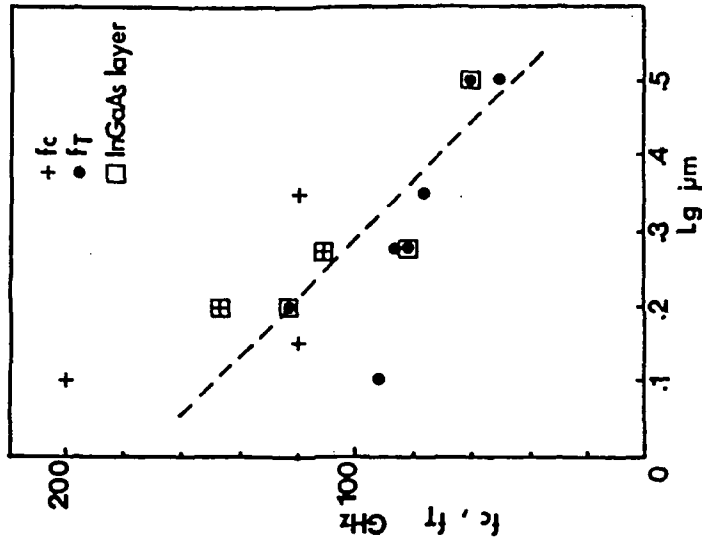
- Unsymmetrical recessed gate structures  $\frac{g_m}{g_d}$  can be improved by a factor of two
- Use of highly doped (ion implanted) InGaAs unmatched layers (Feng et al)
- Use of very highly doped (4.10<sup>18</sup>) thin GaAs layers (150 Å) in order to maintain a good aspect ratio even for small gate length.

but carriers mobility decreases strongly down to 1 000 cm<sup>2</sup>/Vsec.

C.H.S.

*iemn*

## MESFET's CAPABILITIES: Recent results

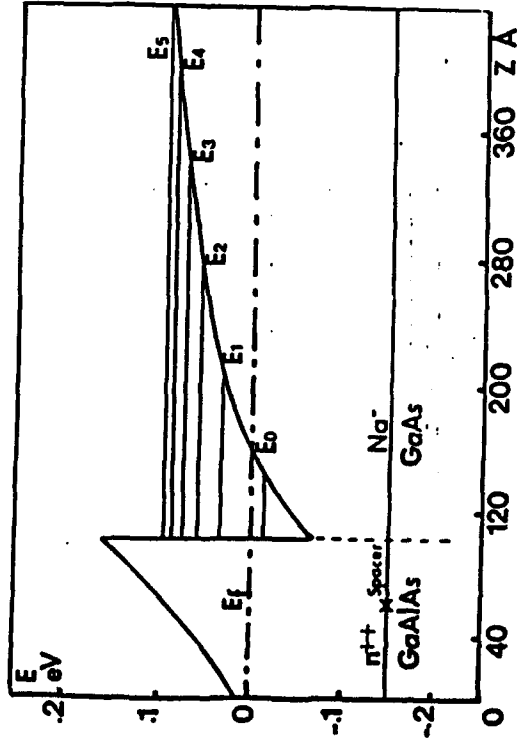


Current gain cut-off frequency  
versus gate length.

- 1991 -

C.H.S.

## SUBBAND STRUCTURE IN THE QUANTUM WELL OF A CONVENTIONAL MODFET



$$E_k = E_{nz} + \frac{\hbar^2}{2m^*} (k_x^2 + k_y^2)$$

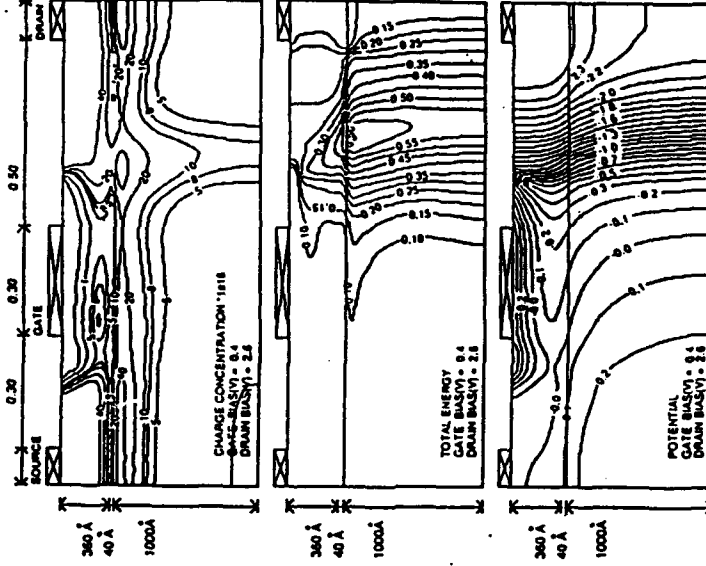
- If space charge reaction is neglected, and well triangular shape is assumed:

$$E_{nz} = \left( \frac{\hbar}{2m} \right)^{1/2} \left( \frac{3}{2} \pi q E_s \right)^{2/3} \left( n + \frac{3}{4} \right)^{2/3}$$

Note that  $E_s$  is proportional to  $n_s$ .

CHS Lille

## PHYSICAL BEHAVIOUR OF CONVENTIONAL MODFET'S

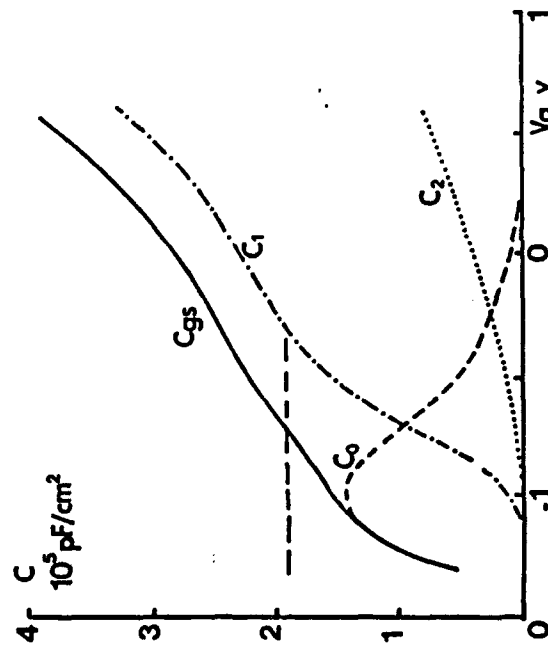


- Equiconcentration. Equipotential and equipotential contours for a 0.3 μm gate MODFET by taking into account surface effects

(from Shawki and Salmer)

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# PARTICULARITIES OF CHARGE CONTROL LAW IN CONVENTIONAL MODFET's

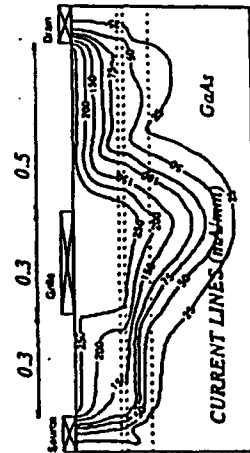


Variations of total capacitance  $C_g$  as function of  $V_g$

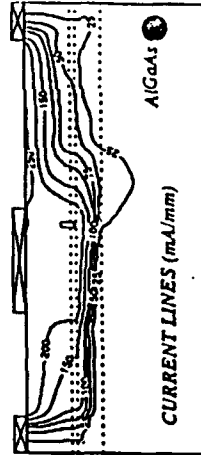
Contributions of electrons  
 $C_0$  in the well  
 $C_1$  free in AlGaAs  
 $C_2$  trapped in AlGaAs

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## USE OF AN AlGaAs BUFFER LAYER



Conventional MODFET



AlGaAs Buffer Layer

Current lines in submicron gate MODFET's

Following Shawki and Salmer (1990)

*tem*

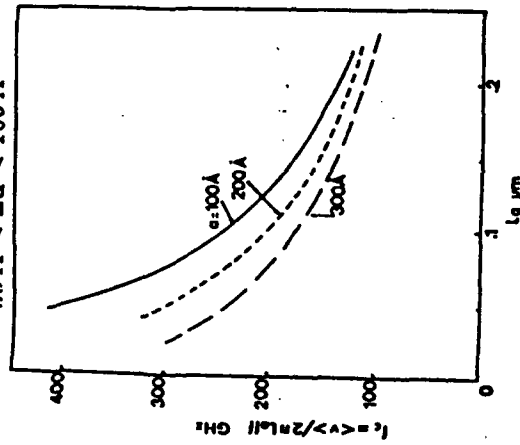
## SHORT CHANNEL EFFECTS IN CONVENTIONAL HEMTS

Following Han and Ferry (1990), it is possible to define an effective gate length

$$L_{eff} = L_g + 2(a + \Delta d)$$

a AlGaAs layer thickness  
 $\Delta d$  equivalent distance of the 2DEG.

$$50 \text{ \AA} < \Delta d < 100 \text{ \AA}$$



First order evaluation of the influence of aspect ratio on intrinsic current gain frequency.

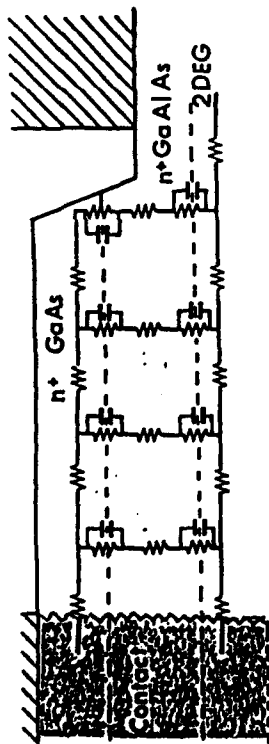
C.H.S.

## CONVENTIONAL MODFET'S

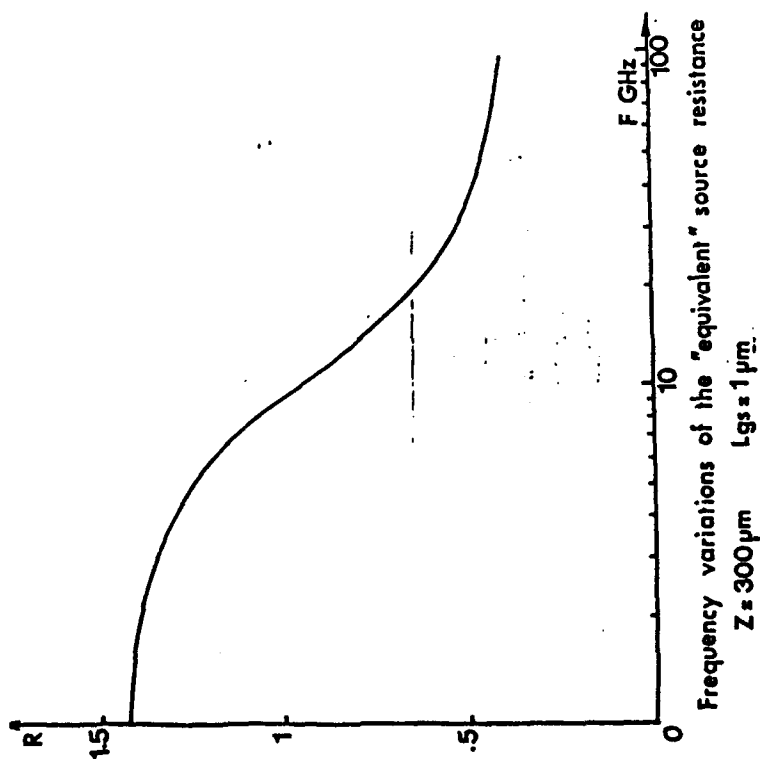
### Main limitations

- Total carrier concentration in the channel :  
 $n_s < 1.2 \cdot 10^{12} / \text{cm}^2$   
 $I_{DSS} < 250 \text{ mA/mm}$
- Quantum well height :  $\Delta E_c < 0.25 \text{ eV}$
- Carrier injection into the buffer layer  
 $\Rightarrow$  High values of  $g_d$ .
- Threshold voltage variations.
- Parasitic effects under low temperature conditions.

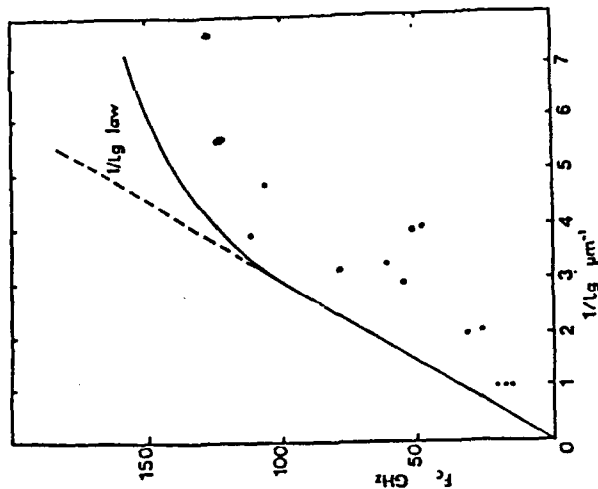
CHS Lille



Equivalent scheme of the source access zone.



## IMPORTANCE OF THE ASPECT RATIO: $L_g/a$



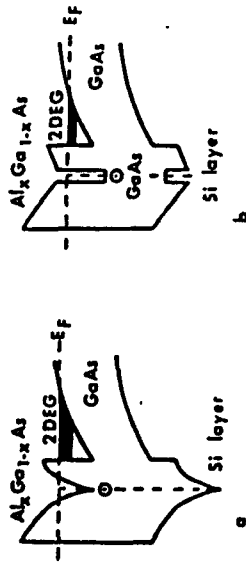
Theoretical intrinsic current gain cut-off frequencies ( $f_c = g_m/2\pi C_{gs}$ ) dependence upon gate length  $L_g$  for a conventional MODFET, the AlGaAs layer thickness a remaining constant ( $a = 400 \text{ \AA}$ )

Following Fauquembergue calculations.

- Experimental extrinsic current gain cut-off frequencies.

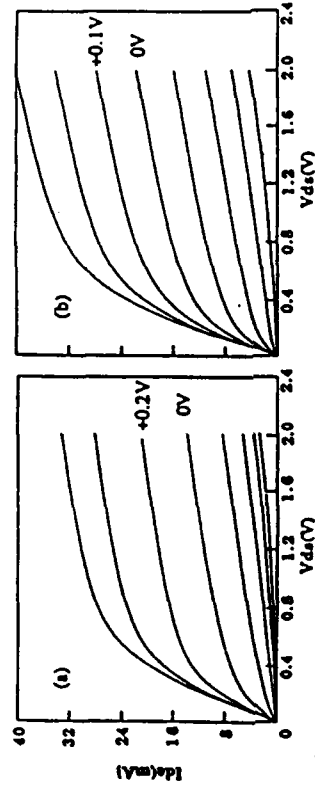
CHS Lille

## PLANAR DOPED HETEROSTRUCTURES



- Higher carrier concentration:  $n_s \approx 1.5 \times 10^{12}$
  - Improvement of the ratio  $L_g/a$  and then  $g_m/g_d$  (gain).
  - Improvement of the breakdown voltage.
  - Suppression of parasitic effects due to DX Center.
- But: Parasitic access resistances may increase.

## PLANAR DOPED MODFET: Improvement of the transconductance



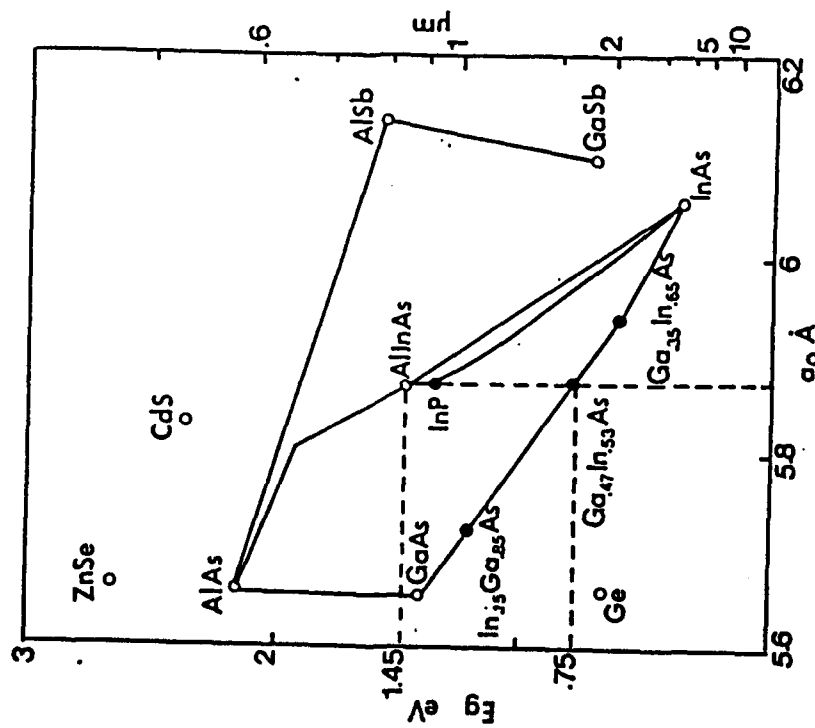
Bulk doped MODFET

Planar doped MODFET

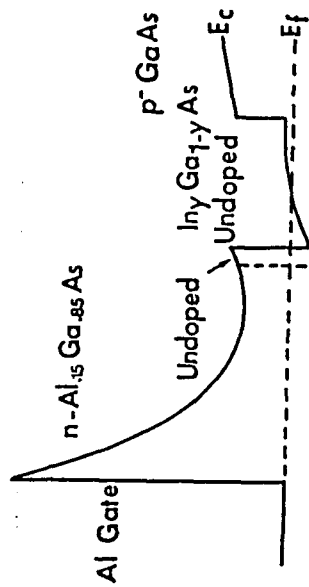
Jin et al (1991)

**C.H.S.**

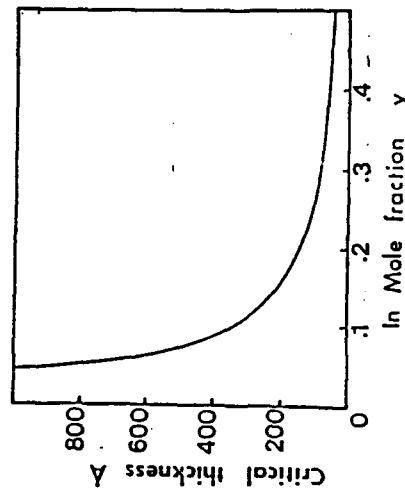
# PSEUDOMORPHIC AlGaAs/InGaAs MODFET's



Energy band gap versus lattice constant.

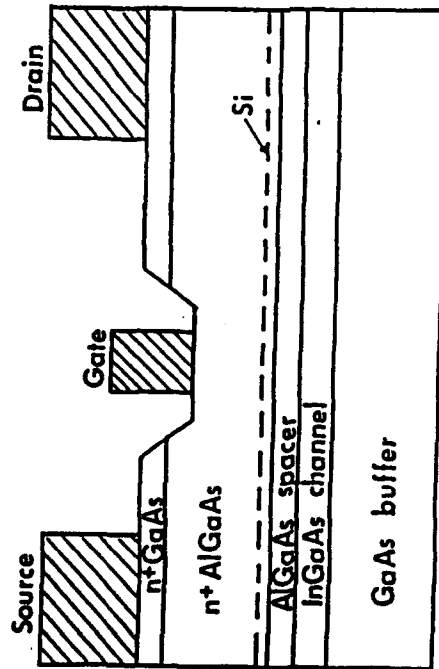


• Energy band diagram



• Critical thickness of the well as function of In mole fraction

## PSEUDOMORPHIC AlGaAs/InGaAs MODFET'S



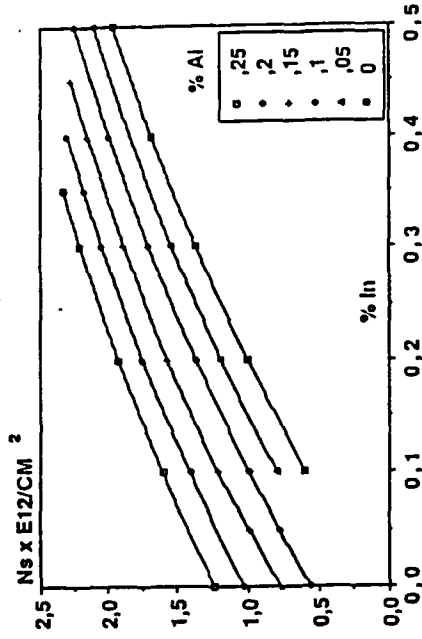
### Advantages :

- More favourable transport properties ?
- Better confinement.
- Higher carrier concentration values :

$n_s \sim 2.5 \cdot 10^{12}/\text{cm}^2$

CHS Lille

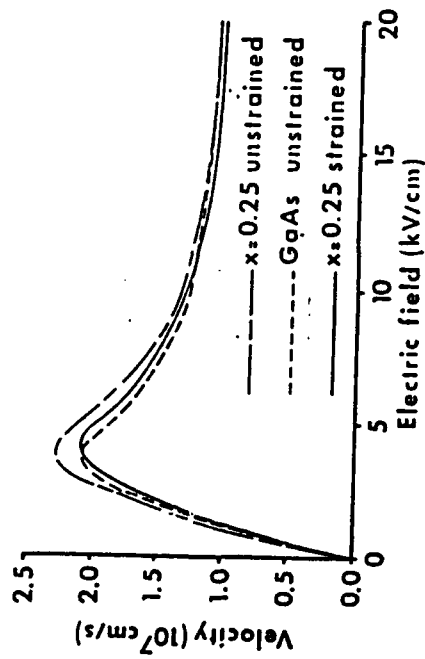
## PSEUDOMORPHIC AlGaAs/InGaAs MODFET's



Carrier concentration dependence upon Al and In contents :  
Self consistent solution of Schrödinger equation

From Cappy et al (1989)

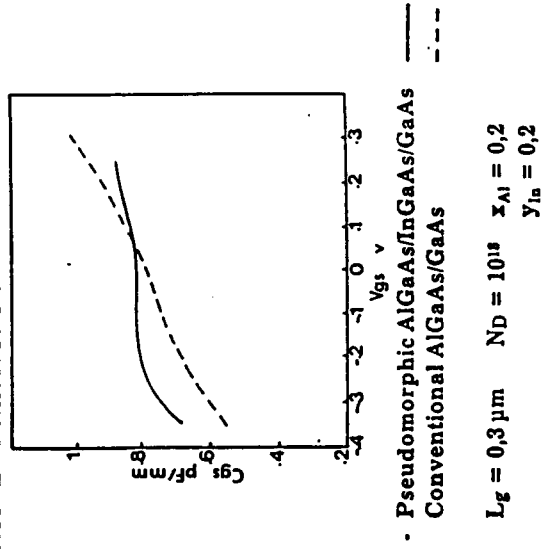
# PSEUDOMORPHIC MATERIALS Influence of strain



## Velocity field characteristics in In<sub>x</sub>Ga<sub>1-x</sub>As layers

From Thobel et al (1989)

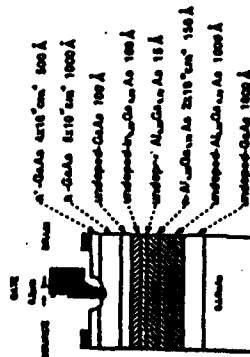
## COMPARATIVE VARIATIONS OF INTRINSIC CAPACITANCE C<sub>gs</sub>



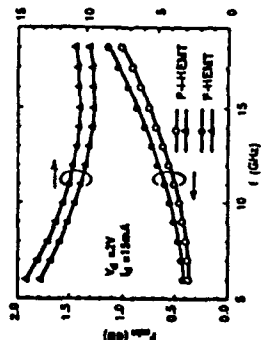
$L_g = 0.3 \mu m$   $N_D = 10^{18}$   $x_{Al} = 0.2$   
 $y_{In} = 0.2$

CHS Lille

# PSEUDOMORPHIC INVERTED MODFET



### Schematic cross-section of the device

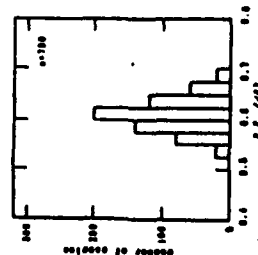


**Comparative performance  
Pacude Inverted HEMT  
Pseudomorphic HEMT**  
From Ohmuro - OKI - 1991.

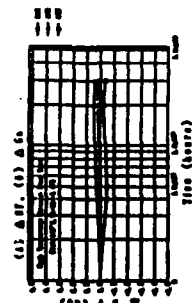
**C.H.G.**

- р. чбан

# LOW NOISE PSEUDOMORPHIC MODFET's EXAMPLE OF REPRODUCIBILITY AND FIABILITY



### Distribution of noise figure at 12 GHz



**NF as function of test time.**

From Tokyo-EDM 91  
C.H.S.

# PSEUDOMORPHIC AlGaAs/InGaAs MODFET'S

● **Various structures:**

- n channel - uniform doping
- n channel - planar doped
- n multiple channel
- n short periode superlattice channel
- n channel with variable In content
- inverted device
- MIS Like FET
- p channel

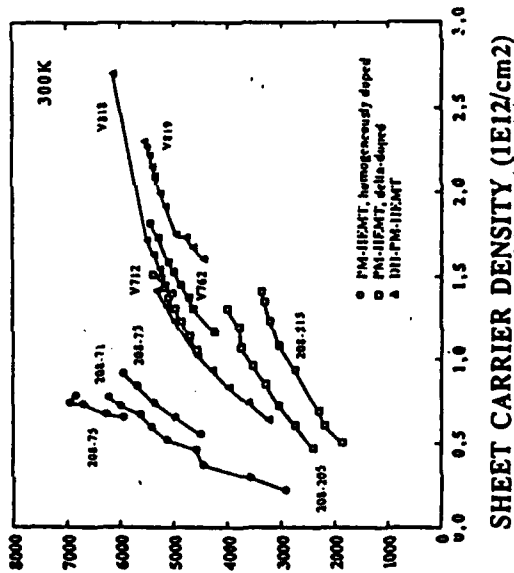
- Among the best performance:

- Tan et al (TRW)

**C.H.S.**

*item*

# CARRIER MOBILITY VARIATIONS IN AlGaAs/InGaAs PSEUDOMORPHIC STRUCTURES

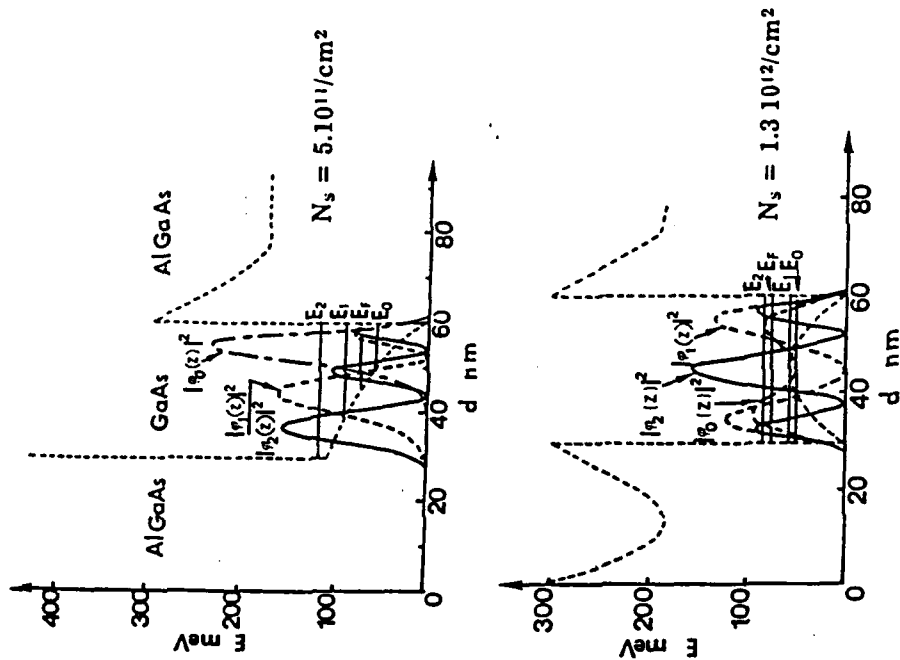


Gated Van der Pauw Hall Effect Measurements

From Pons - Thomson CSF.

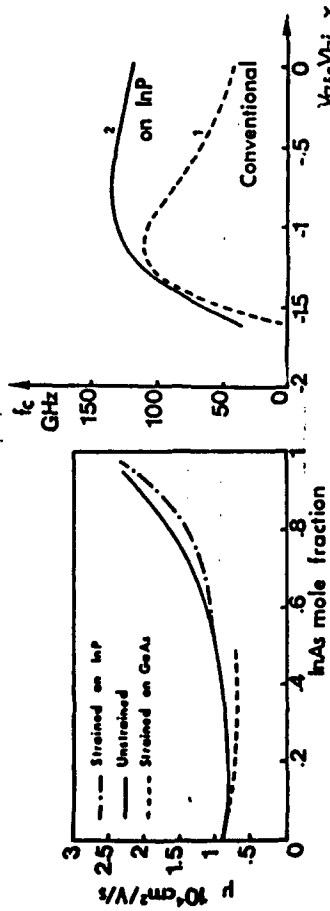
**C.H.S.**

## INFLUENCE OF QUANTUM EFFECTS



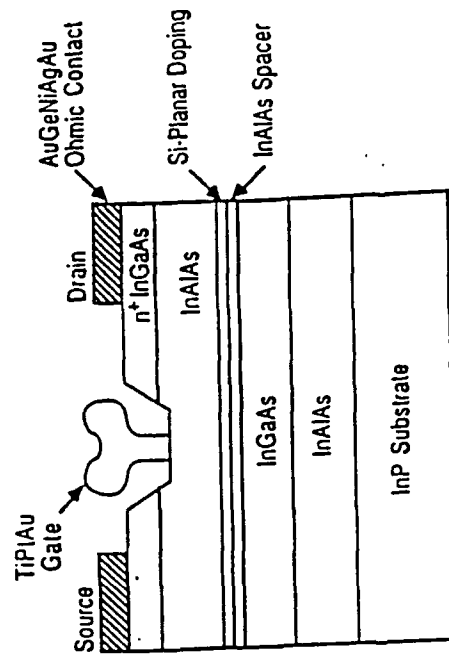
Conduction band edge and squared electron wave functions in a modulation doped double heterojunction (following Tomizawa, oct. 89)

## ADVANTAGES OF InAlAs/InGaAs on InP MODFETs



Improvements of carrier mobility and cut-off frequency

## MODFET on InP SUBSTRATE



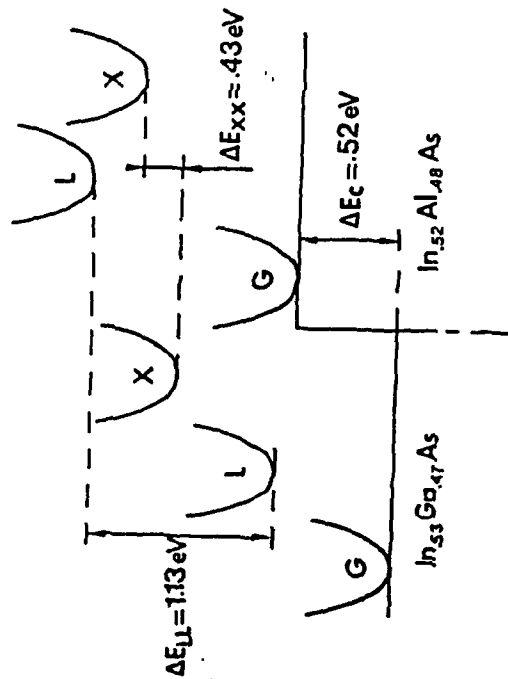
Schematic cross section of a

In<sub>0.52</sub>Al<sub>0.48</sub>As/In<sub>0.53</sub>Ga<sub>0.47</sub>As MODFET

Potential interest

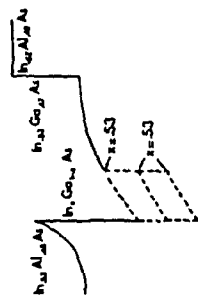
- Better velocity field characteristics
- Higher conduction based discontinuity and then better confinement
- Higher carrier concentration in the channel.

# MULTIVALLEY TRANSFER ACROSS THE HETEROJUNCTION

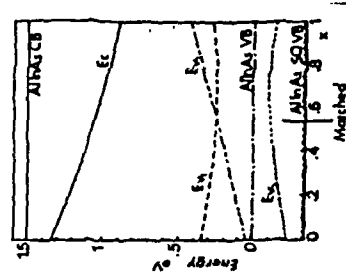


*item*

## INTEREST OF InGaAs ON InP PSEUDOMORPHIC STRUCTURES

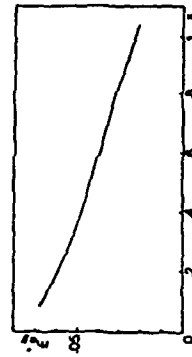


- Conduction Band Structure



-  $\text{Ga}_{1-x}\text{In}_x\text{As}/\text{Al}_{.48}\text{In}_{.52}\text{As}$   
Band offsets.

- Electron effective mass in  $\text{Ga}_{1-x}\text{In}_x\text{As}$  strained on InP substrate

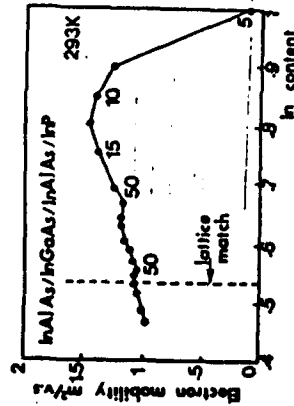


Following Priester (1990).

C.H.S.

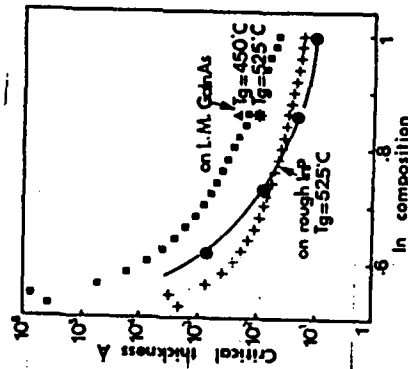
iemn

# $\text{In}_{0.52}\text{Al}_{0.48}\text{As}/\text{In}_{1-x}\text{Ga}_x\text{As}$ on $\text{InP}$ PSEUDOMORPHIC STRUCTURES



Electron mobility  
as function of  $\text{In}$  content  
Tacano et al (1991)

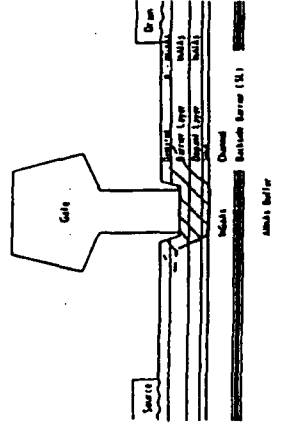
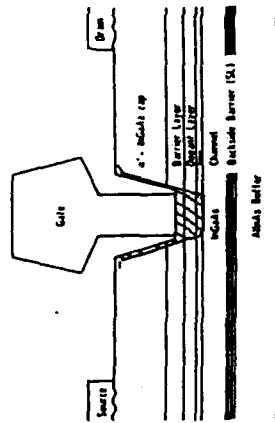
Critical layer thickness  
Victorovitch et al (1991)



C.H.S.

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# INFLUENCE OF CAP LAYER CHARACTERISTICS



Trade-off between :

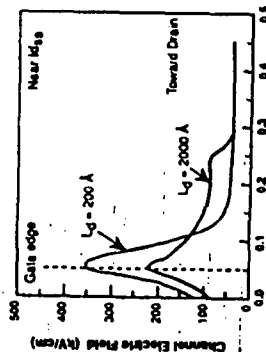
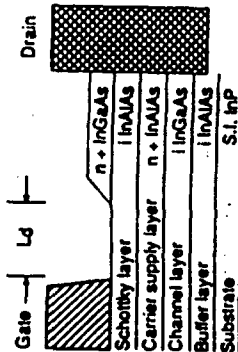
- Improvement of parasitic resistance :  
Thick-Doped.
- Improvement of output conductance :  
Thin and depleted cap layer

From Dickman (1991).

C.H.S.

# IMPROVEMENT OF THE OUTPUT CONDUCTANCE AND DECREASE OF KINK EFFECT IN InGaAs/InAlAs MODFET (Zhan et al) 1991.

Main cause: Ionization in the channel



Schematic cross-section of the device

Electric field in the channel deduced from simulation

# HIGH PERFORMANCE InGaAs on InP MODFET

0.1  $\mu$ m T gate Pseudomorphic (In content = 60 %)

Tan et al IEDM 91

|                             |                 |
|-----------------------------|-----------------|
| 200 Å 4E18 cm <sup>-3</sup> | InGaAs (x=0.60) |
| 200 Å undoped               | InAlAs (x=0.52) |
| 250 Å undoped               | InGaAs (x=0.60) |
| AlInAs Buffer (x=0.52)      |                 |
| Superlattice Buffer         |                 |
| S.I. InP Substrate          |                 |

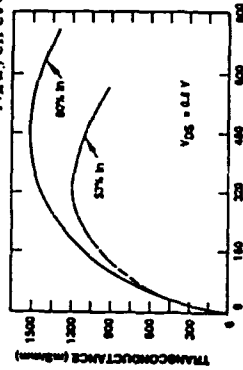
$f_T = 200$  GHz  
 $f_c = 260$  GHz

$N_F = 1.3$  dB  
 $G_a = 8.2$  dB  
 $G_s = 7.3$  dB at 141 GHz

Device cross section

650 Å Selfaligned gate Pseudomorphic (x = 80 %)

Nguyen et al. IEEE El. Dev. March 92



Best current gain  
cut-off frequency  
 $f_T = 305$  GHz

Transconductance values:  
Lattice matched versus  
pseudomorphic devices.

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## MODFET's ON InP SUBSTRATE: NOVEL STRUCTURES

• InAlAs/InAs: Eugster et al - Dec. 91

| Structure   | Layer  | Thickness (nm) | Material |
|-------------|--------|----------------|----------|
| InAlAs/InAs | InAlAs | 10             | InAlAs   |
|             | InAs   | 10             | InAs     |
|             | InAlAs | 10             | InAlAs   |
|             | InAs   | 10             | InAs     |
| InAlAs/InAs | InAlAs | 10             | InAlAs   |
|             | InAs   | 10             | InAs     |
|             | InAlAs | 10             | InAlAs   |
|             | InAs   | 10             | InAs     |
| InAlAs/InAs | InAlAs | 10             | InAlAs   |
|             | InAs   | 10             | InAs     |
|             | InAlAs | 10             | InAlAs   |
|             | InAs   | 10             | InAs     |

Remarks: Very thin InAs layer for avoiding dislocations.

Small mobility: 5 400 cm<sup>2</sup>/Vsec

Reasons are not clear:

- Strongly quantized channel?
- Dislocations?
- Roughness of interfaces.

$L_g = 2 \mu\text{m}$   $g_m = 170 \text{ mS/mm}$

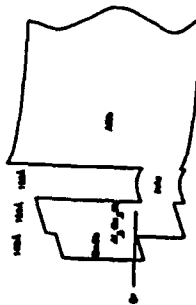
• InAs/AlSb - Werking et al (1992) - Yo et al (1991)

- Very promising properties

$n_s = 3.8 \cdot 10^{17} \text{ cm}^{-3}$   
 $\mu = 21\,000 \text{ cm}^2/\text{Vsec}$

For  $L_g = 2 \mu\text{m}$   
 $g_m = 473 \text{ mS/mm}$

- Very difficult technologies



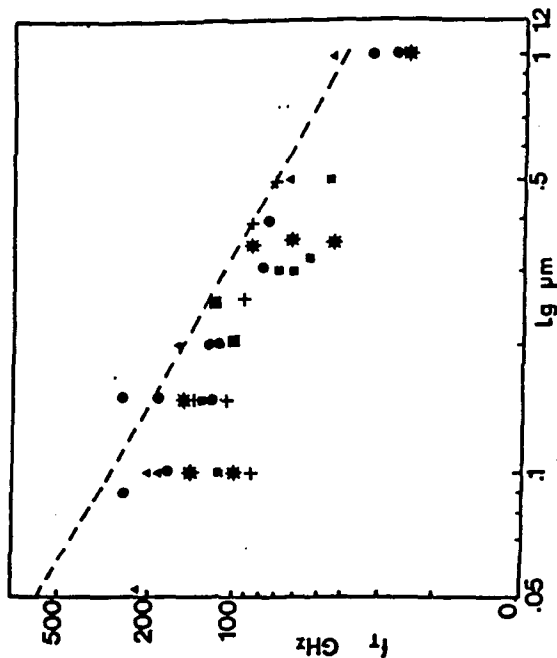
Energy band diagram

C.H.S.

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## FIELD EFFECT TRANSISTORS

State of the art (March 1992)

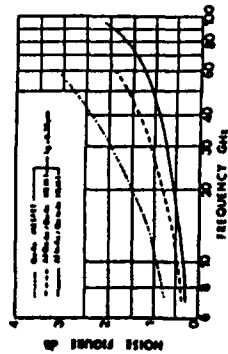


- + GaAs and InGaAs MESFET
- Strained InAlAs/InGaAs MODFET
- Lattice Matched InGaAs on InP MODFET
- \* Pseudomorphic AlGaAs/InGaAs MODFET
- AlGaAs/GaAs MODFET

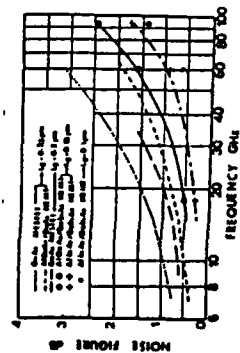
C.H.S.

*iemn*

## LOW NOISE FIELD EFFECTS TRANSISTORS



Comparison between FET's  
for the same gate length (0.25 μm)  
from DUK (MITT 88)

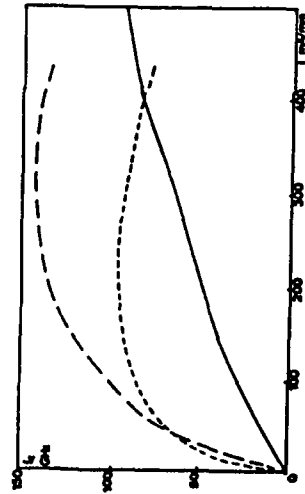


Noise figure versus frequency  
Among the best results  
published (March 92).

[C.H.S.]

*iemn*

## COMPARATIVE BEHAVIOUR OF 0.2 μm GATE FET'S



GaAs MESFET - very well optimised (from FONS)  
- not optimised  
- Pseudomorphic AlGaAs/GaAs MODFET

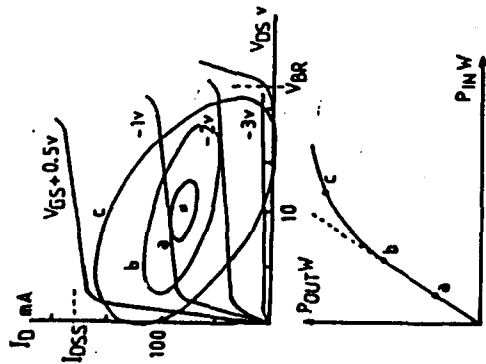
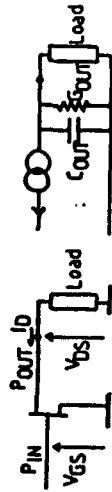
[C.H.S.]

## POWER FET

[C.H.S.]

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## POWER FET LIMITATIONS



-  $P_{out} \# 1/8 I_{DSS} V_{BR} \cos \phi$

### Requirements:

- $I_{DSS}$  } as high as possible.
- $V_{BR}$  }
- Output impedance less reactive as possible
- Gain linearity: variation of MAG with  $V_{GS}$ .

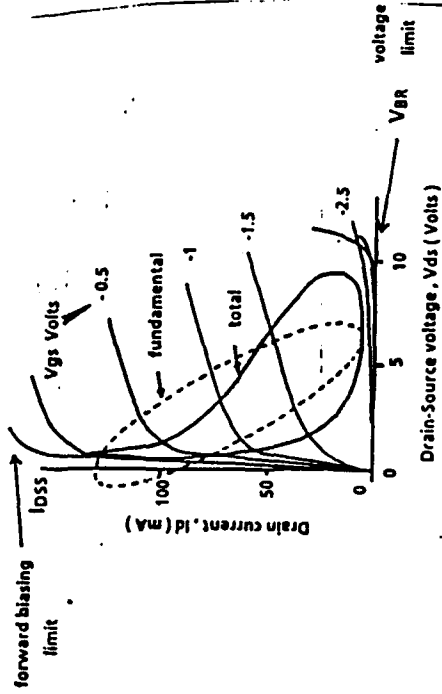
C.H.S.

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## LARGE SIGNAL SIMULATIONS OF POWER FETs

It is necessary but very difficult to take into account at the same time:

- Physical device behaviour (hot electron effects, surface effects) and physical limitations (breakdown)
- Harmonic components of drain voltages and currents



from Halkias (1988).

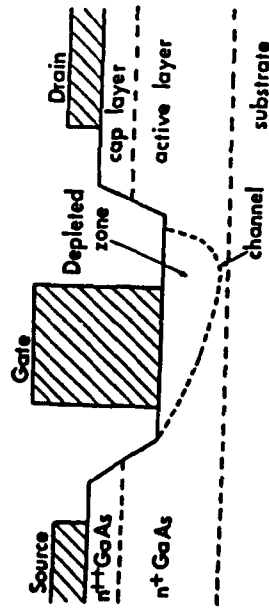
- Parasitic elements
- Device configuration: association of a large number of gates (100 at 100 GHz?)
- Thermal considerations (dc and pulse).

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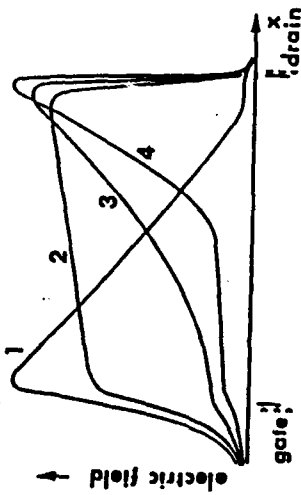
## BREAKDOWN PHENOMENA: PHYSICAL KNOWLEDGE

### - Example of MESFET



### - Physical causes

- Impact ionization in the channel:



1 at pinch-off gate;  
2-3-4:  $I_{DS}$  increasing From Wroblewski et al (1983)

- In the gate depletion region
  - impact ionization (at pinch-off)
  - gate current (thermionic or tunneling)

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## SIMULATIONS OF BREAKDOWN PHENOMENA

### - Simplified models

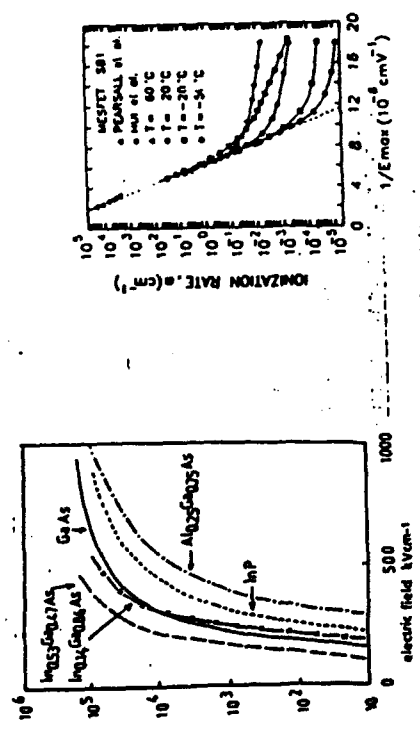
- Quasi 2D models: (Wroblewski 1983)
  - Channel breakdown: good estimation
  - Optimization of MESFET (recess structure)
  - Problems of gate breakdown, calculation.
- 2D Solution of Poisson equation (Wroblewski - Zaitlin, Fellon).
  - Calculation of multiplication factor.
  - Evaluation of electric field at the gate edges  
→ Tunneling currents.
  - Problems:
    - knowledge of ionization rates and accurate predictions of tunneling currents.
    - space charge reaction is not included: valid only for small  $I_{ds}$ .
- 2D models (Mizuta, Wada, Lefebvre, Chan, Barton, Ashworth, ...)

- Introduction of ionization effects or tunneling currents.
- Problems:
  - drift diffusion models
  - Ionization effects and tunneling currents are not introduced together.

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# IONIZATION RATES IN BINARY AND TERNARY III-V MATERIALS



Ionization coefficients data for InGaAs

Accurate determination

of  $\alpha$  in GaAs by Canali et al (1991)

C.H.S.

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# BREAKDOWN VOLTAGE: MAIN DEPENDANCES

- Material properties:
  - Ionization rates in bulk material
  - Problems: Ionization rates in Very thin channels ??
  - Strained materials: Improvement of the breakdown voltage due to strained:  $m^*h$  (Eisenbeiser - 1991)
- Epilayer characteristics:
  - Doping level of the last layer under the gate is the most important.
  - Trade off with high  $I_{dss}$
  - Use of Low-High doping profile
  - Very thin top layer (100 to 200 Å). Influence of tunneling currents.
- Semiinsulated substrate: Hole capture.
- Recess zone configuration:
  - Double recess gate process allow to improve breakdown voltage: from 8 to 16 V in InAlAs/InGaAs/InP HEMT (Boss - 1991).
  - Problems of surface states under dynamic conditions (Huang - 1991).

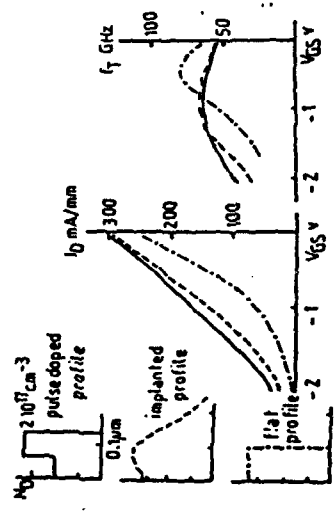
C.H.S.



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## IMPROVEMENTS OF GaAs MESFET

- Pulse doped active layer :  
 . It allows to improve breakdown voltage and linearity : for instance Chu obtained 23 dB between  $IP_3$  and  $P_{0.1\text{ dB}}$   
 . Such improvement was predicted by Heliodore



- Example of result : 220 mW output power at 35 GHz (RCA)  
 $0.5\text{ }\mu\text{m} \times 600\text{ }\mu\text{m}$  gate.

## - Highly doped InGaAs channels (Feng and Wang)

$L_g = 0.25\text{ }\mu\text{m}$   
 96 mW at 44 GHz      0.6 W/mm in  
 121 mW at 60 GHz      both cases

## - Pulse doped InGaAs channel : Kim

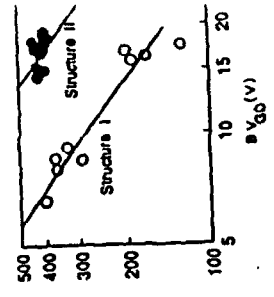
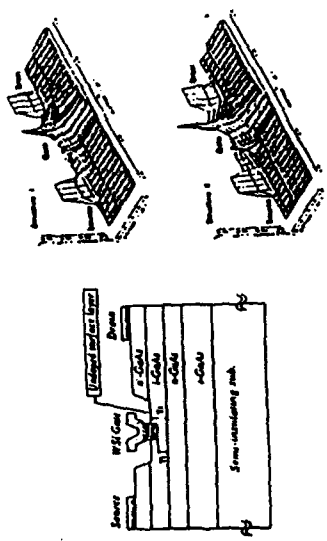
$L_g = 0.25\text{ }\mu\text{m}$   $I_{\text{DSS}} = 700\text{ mA/mm}$   
 $P_0 = 50\text{ mW} - 0.6\text{ W/mm}$  at 60 GHz.

## - Other improvements : AlGaAs buffer - via holes.

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## IMPROVEMENTS OF BREAKDOWN VOLTAGE AND PERFORMANCE BY USING A LOW-HIGH DOPING PROFILE (Takahashi - 1991).



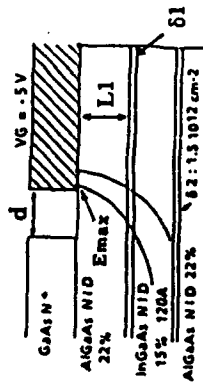
Device Structure.  
 Electric field distribution  
 Structure I :  $T_1 = 300\text{ }\text{\AA}$

$I_{\text{MAX}} = f(\text{Breakdown V})$   
 Structure II :  $T_1 = 0$ .

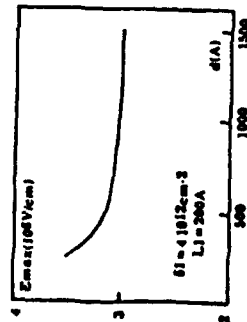
C.H.S.

# SENSITIVITY OF THE MAXIMUM ELECTRIC FIELD AT THE GATE CORNER OF A PLANAR DOPED PM HFET TO THE STRUCTURE PARAMETERS.

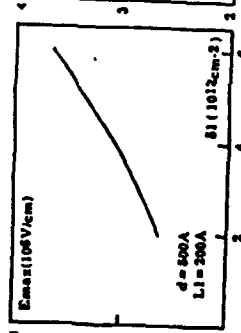
extracted from the 2D electric field map under pinch-off conditions



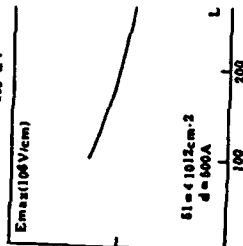
Influence of the recess spacing  $d$



Influence of the top planar doping  $\delta I$



Influence of the top NID AlGaAs layer thickness  $L_1$



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## POWER FIELD EFFECT TRANSISTORS

### • MESFET:

- GaAs
- InGaAs on GaAs substrate.

### • MODFET:

- Conventional AlGaAs/GaAs
- Multichannel AlGaAs/GaAs
- Single channel AlGaAs/InGaAs
- Multichannel AlGaAs/InGaAs/GaAs
- InAlAs/InGaAs on InP substrate..

### • HFET or D.M.T.:

- AlGaAs/InGaAs ( $n^+$  doped) strained or not on GaAs.
- InAlAs/InGaAs ( $n^+$  doped) on InP.
- InAlAs/InP ( $n^+$ ) on InP substrate.

### • MISFET:

- InGaAs  $n^+$  doped ) on InP
- InP  $n^+$  doped
- LT GaAs insulating layer.

C.H.S.

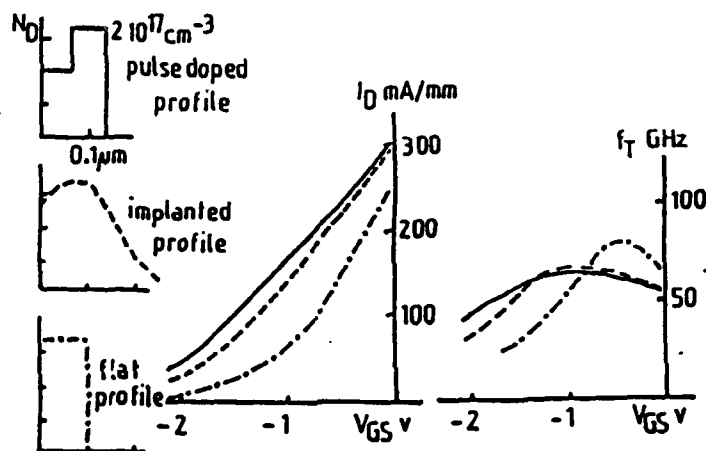
*iemn*

## IMPROVEMENTS OF GaAs MESFET

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. It allows to improve breakdown voltage and linearity : for instance Chu obtained 23 dB between IP3 and P0.1 dB

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96 mW at 44 GHz

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both cases

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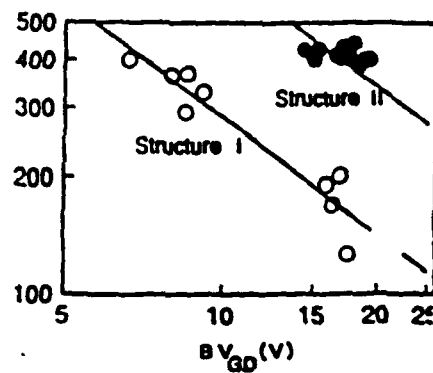
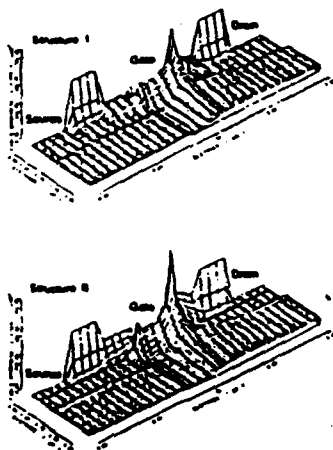
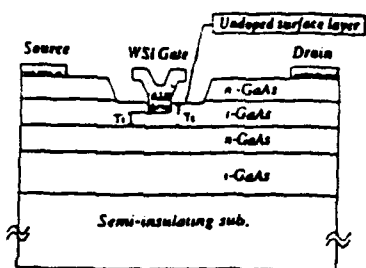
$P_0 = 50 \text{ mW} - 0.6 \text{ W/mm}$  at 60 GHz.

### - Other improvements : AlGaAs buffer - via holes.

**C.H.S.**

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Device Structure.

Electric field distribution  
Structure I :  $T_i = 300\text{\AA}$

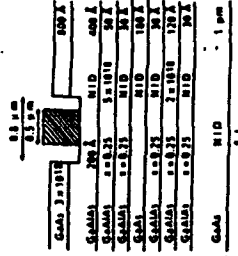
$I_{MAX} = f(\text{Breakdown } V)$   
Structure : II  $T_i = 0$ .

**C.H.S.**

*diem*

## CONVENTIONAL AlGaAs/GaAs HEMT

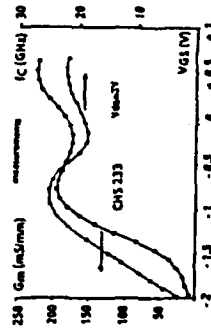
- Single channel
- Sheet carrier density smaller than  $1.2 \times 10^{12} \text{ cm}^{-2}$ : influence of DX centers for  $x \approx 0.2$
- $I_{\text{DSS}}$  is too small:  $< 300 \text{ mA/mm}$ .
- Multichannel devices
- $n_s$  and  $I_{\text{DSS}}$  may be higher
- $n_s = 5.3 \times 10^{12} \text{ cm}^{-2}$
- $I_{\text{DSS}} = 1.100 \text{ mA/mm}$  for a 3 channels.



- Improvement of linearity

- Problem of  $R_s$

- Example of performance: Fujitsu
- $0.5 \times 2400 \mu\text{m}$   $I_{\text{DSS}} = 530 \text{ mA/mm}$
- $P_0 = 1.2 \text{ W}$  at 30 GHz.



From Temcamani

C.H.S.

## PSEUDOMORPHIC HFET LAYER STRUCTURES

### HEMT

|                                                                |
|----------------------------------------------------------------|
| $4 \times 10^{18} \text{ GaAs}$                                |
| $1 \times 10^{17} \text{ Al}_{0.30}\text{Ga}_{0.70}\text{As}$  |
| $5 \times 10^{12} \text{ Si planar doping}$                    |
| undoped $\text{Al}_{0.30}\text{Ga}_{0.70}\text{As}$ spacer     |
| undoped $\text{In}_{0.22}\text{Ga}_{0.78}\text{As}$ well 125 Å |
| $\text{Al}_{0.30} \text{Ga}_{0.70}\text{As/GaAs}$ superlattice |
| GaAs buffer                                                    |
| Semi-insulating GaAs                                           |

$$n_s = 2.5 \times 10^{12}$$

$$\mu = 6300$$

### DCH MODFET

|                                                                |
|----------------------------------------------------------------|
| $4 \times 10^{18} \text{ GaAs}$                                |
| $1 \times 10^{17} \text{ Al}_{0.25}\text{Ga}_{0.75}\text{As}$  |
| $5 \times 10^{12} \text{ Si planar doping}$                    |
| undoped $\text{Al}_{0.25}\text{Ga}_{0.75}\text{As}$ spacer     |
| undoped $\text{In}_{0.22}\text{Ga}_{0.78}\text{As}$ well 63 Å  |
| $1.1 \times 10^{12} \text{ Si planar doping}$                  |
| undoped $\text{In}_{0.22}\text{Ga}_{0.78}\text{As}$ well 63 Å  |
| $\text{Al}_{0.25} \text{Ga}_{0.75}\text{As/GaAs}$ superlattice |
| GaAs buffer                                                    |
| Semi-insulating GaAs                                           |

$$n_s = 4 \times 10^{12}$$

$$\mu = 3100$$

### DCHFET

|                                                                            |
|----------------------------------------------------------------------------|
| $4 \times 10^{18} \text{ GaAs}$                                            |
| $1 \times 10^{17} \text{ Al}_{0.25}\text{Ga}_{0.75}\text{As}$              |
| $2.6 \times 10^{18} \text{ In}_{0.22}\text{Ga}_{0.78}\text{As}$ well 125 Å |
| $\text{Al}_{0.25} \text{Ga}_{0.75}\text{As/GaAs}$ superlattice             |
| GaAs buffer                                                                |
| Semi-insulating GaAs                                                       |

$$n_s = 4.9 \times 10^{12}$$

$$\mu = 1900$$

### DH HEMT

|                                                                    |
|--------------------------------------------------------------------|
| $4 \times 10^{18} \text{ GaAs}$                                    |
| $1 \times 10^{17} \text{ Al}_{0.25}\text{Ga}_{0.75}\text{As}$      |
| $4 \times 10^{12} \text{ Si planar doping}$                        |
| undoped $\text{Al}_{0.25}\text{Ga}_{0.75}\text{As}$ spacer         |
| undoped $\text{In}_{0.22}\text{Ga}_{0.78}\text{As}$ well 125 Å     |
| undoped $\text{Al}_{0.25}\text{Ga}_{0.75}\text{As}$ spacer         |
| $1 \times 10^{18} \text{ Al}_{0.25}\text{Ga}_{0.75}\text{As}$ 50 Å |
| $\text{Al}_{0.25} \text{Ga}_{0.75}\text{As/GaAs}$ superlattice     |
| GaAs buffer                                                        |
| Semi-insulating GaAs                                               |

$$n_s = 3.4 \times 10^{12}$$

$$\mu = 5200$$

From Lester et al (1989)

## PSEUDOMORPHIC HFET STRUCTURES

### Summary of D.C. characteristics and 35 GHz Power Performance

| DEVICE                    | HEMT | DCHFET | DHEMT | DCH<br>MODFET |
|---------------------------|------|--------|-------|---------------|
| Current Density<br>mA/mm  | 455  | 470    | 630   | 600           |
| Transconductance<br>mS/mm | 625  | 745    | 790   | 747           |
| Breakdown<br>Voltage v    | 9.5  | 7.2    | 9.5   | 9.5           |
| Efficiency %              | 40   | 35.5   | 43    | 47            |
| Maximum Power<br>W/mm     | 0.75 | 0.77   | 0.97  | 0.94          |

Following Lester et al - DRC 1989

*iemn*

## AlGaAs/InGaAs PSEUDOMORPHIC HFET

### DOPED CHANNEL OR NOT?

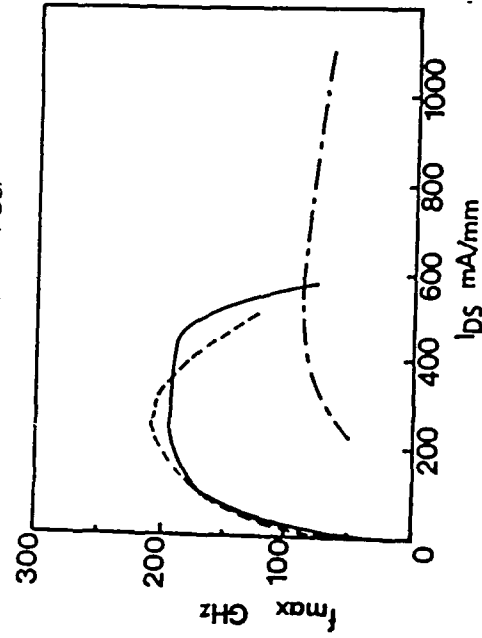
- Best result at 60 GHz : 1 W/mm -  $L_g = 0.25 \mu\text{m}$   
Homogeneously doped channel  $f_M = 150 \text{ GHz}$ .

- Work of Dickmann et al (1991)

#### \* Comparison

- Conventional HEMT - - - - -
- Centrally pulse doped channel HFET - - - - -
- Backside pulse doped channel HFET - - - - -

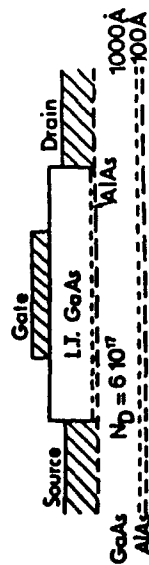
\* Experimental results:  $f_M = f(I_{DS})$ .



C.H.S.

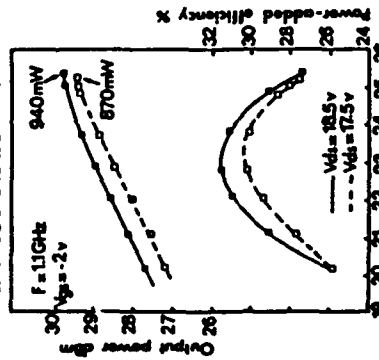
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## HIGH POWER DENSITY GaAs MISFET WITH A LT GROWN AS THE INSULATOR



Buffer layer GaAs

### Device Structure.

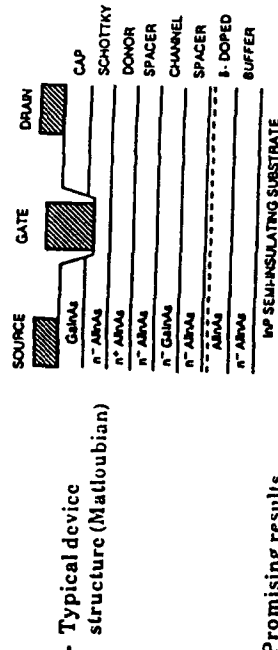


Power performance at 1,1 GHz  
from Chen (1991).

hem

## HEMT ON InP SUBSTRATE FOR POWER APPLICATIONS

- $\text{AlInAs}/\text{InGaAs}/\text{InAlAs}$  HEMT



### Promising results

- Matloubian et al:  $L_g = 0.22 \mu\text{m}$   
 $V_g = 9 \text{ V}$   $I_{\text{DSS}} = 770 \text{ mA/mm}$   
 $P_0 = 0.96 \text{ W/mm at } 12 \text{ GHz}$   $\eta = 40 \%$
- Kao et al:  $L_g = 0.12 \mu\text{m}$ . Pseudomorphic  
 $V_H = 4.8 \text{ V}$   $I_{\text{DSS}} = 600 \text{ mA/mm}$   
 $P_0 = 0.35 \text{ W/mm at } 60 \text{ GHz}$   
 $P_0 = 0.2 \text{ W/mm at } 94 \text{ GHz}$   $\sim 100 \text{ mW}$   
(as compared to 67 mW for InGaAs/GaAs HEMT)
- Kuang et al: DHEMT  
 $I_{\text{DSS}} = 1.5 \text{ A/mm}$

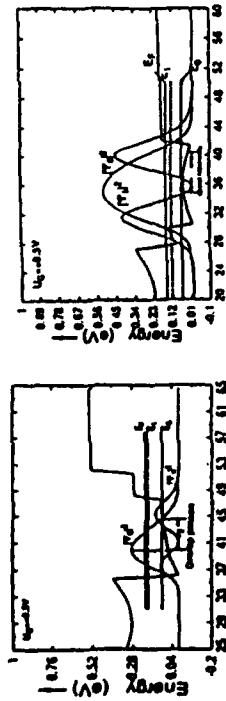
But problems of reliability.

C.H.S.

10mm

# SCHRODINGER EQUATION SOLUTION → COMPARISON BETWEEN

Centrally pulse doped HFET and Backside pulse doped HFET



Backside pulse doped is better  
But transfer to the E1 level occurs earlier.

From Dickman et al (1991)

C.H.S.

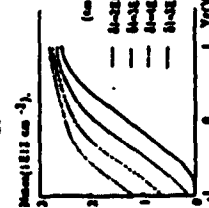
SPMUT 603 ALMS

ICM - CUIS UNIVERSITE DE LILLE

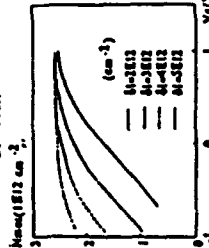
## 1 D SCHRODINGER-POISSON SIMULATION OF THE ELECTRON GAS CONTROL NS = f(VG) IN PLANAR DOPED PSEUDOMORPHIC HFETS.

| gate $V_g$ |                    | $L_1$              |       |
|------------|--------------------|--------------------|-------|
| 61 %       | 20% 100 1000 10000 | 20% 100 1000 10000 | 10000 |
| 62 %       | 20% 100 1000 10000 | 20% 100 1000 10000 | 10000 |

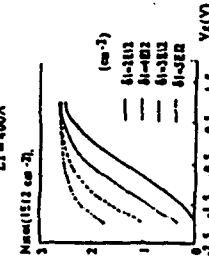
$L_1 = 200\text{\AA}$



$L_1 = 500\text{\AA}$

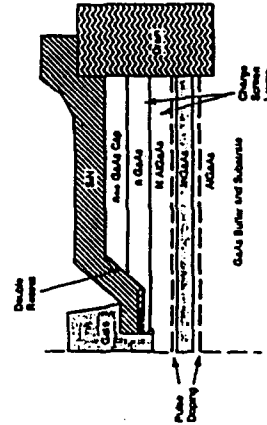


$L_1 = 400\text{\AA}$

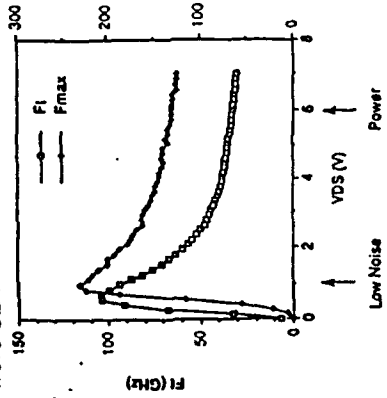


10mm

# EXAMPLE OF HIGH BREAKDOWN PSEUDOMORPHIC HEMT



Schematic cross section  
of the gate drain region.



Current gain cut-off frequency  
and maximum gain cut-off frequency  
versus drain voltage:  
increase of transit time.

From Huang (1991).

C.H.S.

*iemn*

### OTHER POWER FET DEVICES ON InP SUBSTRATE FOR MILLIMETER WAVE RANGE

#### - HFET :

- InAlAs/n + InGaAs      Del Alamo
  - Interesting results but trade-off between  $V_g$  (max 19 V) and  $IDSS$  (max 450 mH/mm)
  - No power results available.
- InAlAs/n + InP      Del Alamo - Aina
  - Promising results in term of  $g_m$  and  $IDSS$  for long devices.
  - No power results.

#### - MISFET :

- InP :
  - Very high breakdown voltage
  - High drain current density
  - Small gain
  - Problems of stability.

#### Very promising results :

- 1.6 W/mm at 18 GHz. Bredenbender
- 1.8 W/mm at 30 GHz. Saunier et al.

- InGaAs :
  - High drain current density
  - Small gain.

Encouraging results : Johnson - 1991.  
 $L_g = 1 \mu\text{m}$   $P_0 = 1.07 \text{ W/mm}$  at 12 GHz

**C.H.S.**

*iemn*

### CONCLUSION

#### - PHYSICAL MODELS

- They are able to furnish very interesting informations on device behaviour ;
- They can be used systematically for device optimization ;
- They cannot give always accurate predictions due to the lack of knowledge in basic physics.

#### - LOW NOISE DEVICES :

- Pseudomorphic AlGaAs/InGaAs MODFET can now constitute the basis of a mature technology ;
- InGaAs on InP MODFET is the most promising but problems of reliability must be solved ;
- Noise figure smaller than 1 dB at 94 GHz can be hoped in near future ;
- Temperature dependance must be studied.

#### - POWER DEVICES

- The situation is not clear, but for next future, MODFETs on GaAs remain preferable ;
- Studies must be done on several kinds of devices. New device configurations must be envisaged for adding properly output powers of devices.

**C.H.S.**

# **TWO - TERMINAL MILLIMETER - WAVE DEVICES**

**J. Freyer**

**University of München • Germany**

## Two terminal millimetre-wave devices

J. Freyer, Technical University of Munich

- Basic principles of power generation
- Gunn element (Transferred electron device)
- IMPATT diode
- MITATT and TUNNETT diode
- PIN-avalanche diode
- BARITT diode
- Double-barrier and QWITT diode
- Device fabrication and packaging
- Resonators
- Application and state-of-the-art performance

$$Z_D + Z_L = 0 \quad \text{Kirchhoff's law}$$

$$R_D + R_L = 0$$

$$X_D + X_L = 0$$

$$P_{rf} = 1/2 |I_{rf}|^2 R_L = 1/2 |I_{rf}|^2 |R_D|$$

(available output power)

$$\text{series resistance } R_s$$

$$P_{rf} = 1/2 |I_{rf}|^2 (|R_D| + R_s)$$

## Basic equations

$$\frac{\partial n}{\partial t} = \frac{1}{q} \frac{\partial J_n}{\partial x} + G$$

$$\frac{\partial p}{\partial t} = -\frac{1}{q} \frac{\partial J_p}{\partial x} + G$$

$$J_n = q v_n n$$

$$J_p = q v_p p$$

$$\frac{\partial E}{\partial x} = q(N_D - N_A + p - n)$$

$$n : \text{electron density}$$

$$p : \text{hole density}$$

$$J_n : \text{electron current density}$$

$$J_p : \text{hole current density}$$

$$G : \text{generation}$$

$$q : \text{electronic charge}$$

$$t : \text{time}$$

continuity equations

current density equations

Poisson equation

$$x : \text{space coordinate}$$

$$v_n : \text{electron drift velocity}$$

$$v_p : \text{hole drift velocity}$$

$$\epsilon : \text{dielectric constant}$$

$$E : \text{electric field}$$

$$N_D : \text{donor density}$$

$$N_A : \text{acceptor density}$$

## n-type semiconductor

$$\begin{aligned} \frac{\partial E}{\partial x} &= -\frac{q}{\epsilon} (n - N_D) & \text{Poisson equation} \\ I_c &= A q n v(E) & \text{convection current} \\ I_{\text{tot}} &= I_c + A \epsilon \frac{\partial E}{\partial t} & \text{total current} \\ I_{\text{tot}} &= A \left\{ \epsilon \frac{\partial E}{\partial t} + q (N_D - n) v(E) \right\} \end{aligned}$$

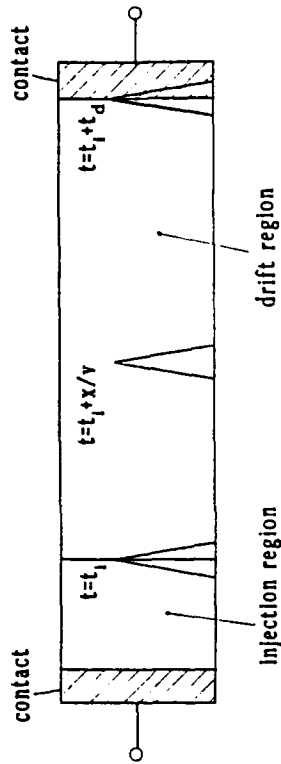
## small signal assumption

$$\begin{aligned} E &= E_0 + E_1 \exp(j\omega t) \\ I_{\text{tot}} &= I_{\text{tot}0} + I_{\text{tot}1} \exp(j\omega t) \\ E_1(x) &= \frac{I_{\text{tot}1}}{j \epsilon A v_0 (\omega / v_0 - j q N_D / \epsilon v_0 \partial v / \partial E)} \cdot \\ &\quad \{ 1 - \exp(-j \omega x / v_0) \exp(-q N_D / \epsilon v_0 \partial v / \partial E) \} \end{aligned}$$

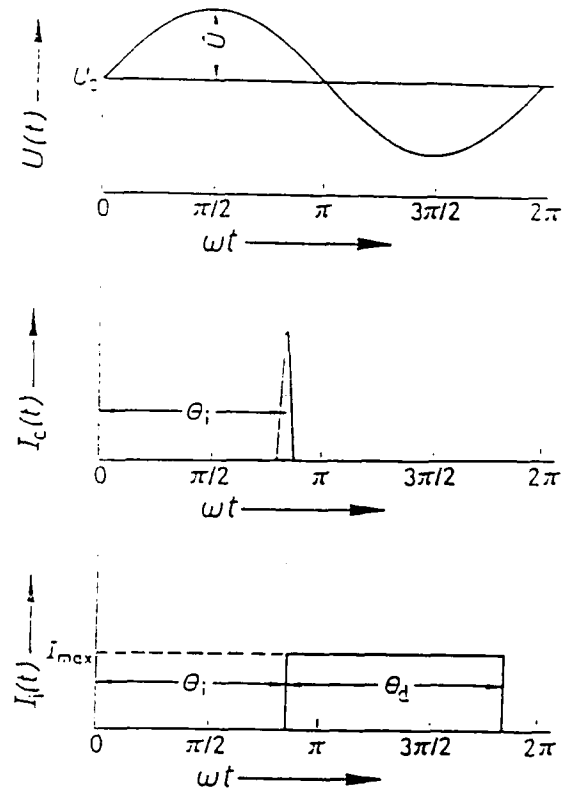
$$\frac{\partial v}{\partial E} > 0 \quad \text{damping}$$

$$\frac{\partial v}{\partial E} = 0 \quad \text{no attenuation}$$

$$\frac{\partial v}{\partial E} < 0 \quad \text{growing up}$$



Schematic presentation of a transit-time device

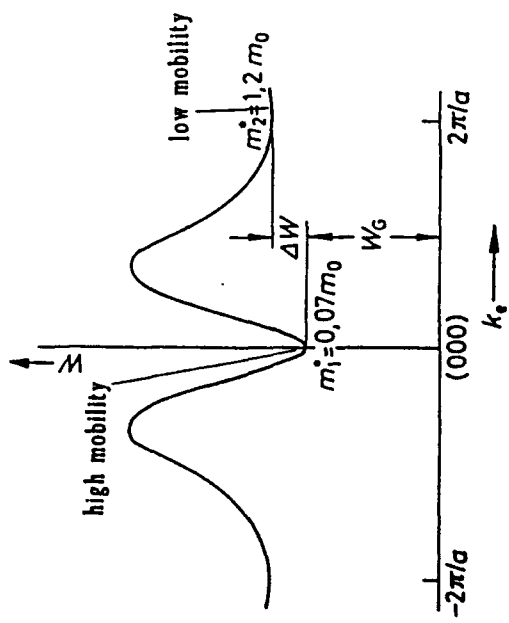


Voltage, injected convection current and induced current of a transit-time device

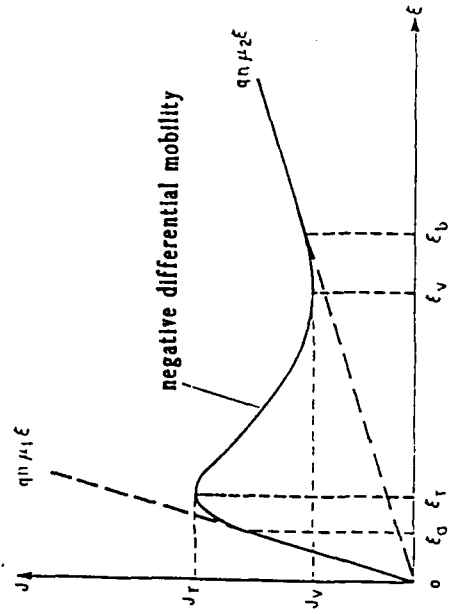
# Transit-time device

|                                                                                       |                   |  |
|---------------------------------------------------------------------------------------|-------------------|--|
| • injection region                                                                    |                   |  |
| • drift region                                                                        |                   |  |
| $I_{rf} = 2 I_0 \sin(\Theta_d/2)/(\Theta_d/2)$                                        | Fourier component |  |
| $I_0 = I_{\max} \Theta_d/(2\pi)$                                                      | dc-current        |  |
| $P_{rf} = U_{rf} I_0 \frac{\sin(\Theta_d/2)}{\Theta_d/2} \sin(\Theta_i + \Theta_d/2)$ |                   |  |
| $\Theta_i = \omega t_i$                                                               | injection angle   |  |
| $\Theta_d = \omega t_d$                                                               | drift angle       |  |

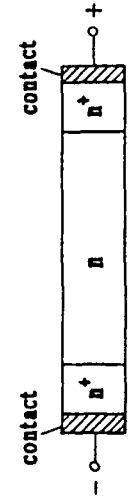
generation of output power :  $\pi < \Theta_i + \Theta_d/2 < 2\pi$



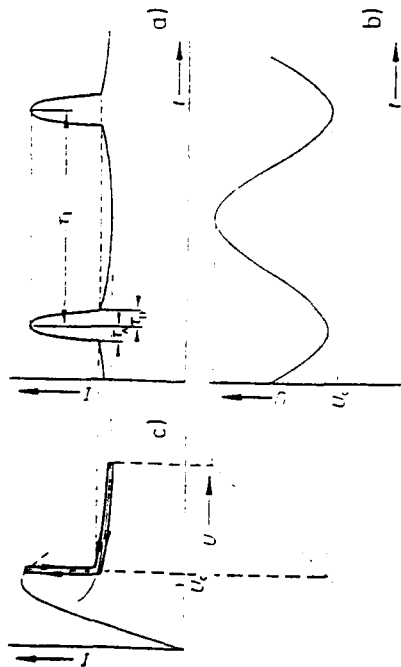
Energy band structure of GaAs



Current density versus electric field of a two-valley semiconductor



Schematic structure of a Gunn element



Current and voltage of a Gunn element in transit-time mode

### Limitation of output power of Gunn elements

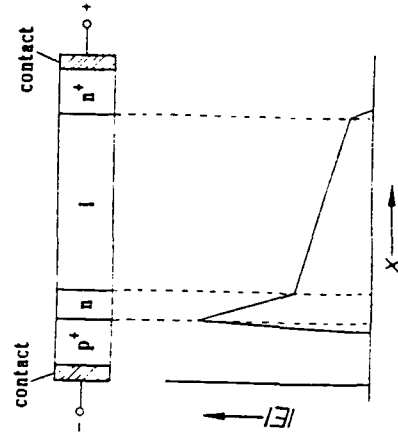
- The negative mobility vanishes if the oscillation periode reaches the scattering time
- The  $n^+$  cathode injects "cold" electrons which must gain energy before they can transfer. A "dead" zone results which occupies a large fraction of the total n-zone

### Improvement

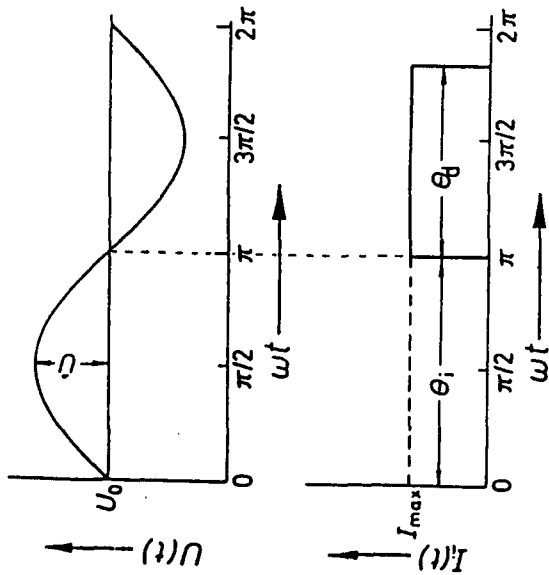
- Injection of "hot" electrons
- GaAlAs/GaAs launcher
- planar doped barriers
- camel diodes
- tunnel injection
- reverse biased Schottky barriers

### Noise of Gunn elements

- primary noise
- diffusion (statistical velocity fluctuations of drifting carriers)
- modulation (low frequency fluctuations of dc-current, temperature)
- up-conversion via the device impedance



Structure and field distribution of a Gunn element



Rf-voltage and induced current of a Read IMPATT diode

## Negative resistance of IMPATT diode

$$G = \alpha v_s (J_n + J_p)$$

$$\frac{\partial J_n}{\partial t} = -v_s \frac{\partial J_n}{\partial x} + \alpha v_s (J_n + J_p)$$

$$\frac{\partial J_p}{\partial t} = v_s \frac{\partial J_p}{\partial x} + \alpha v_s (J_n + J_p)$$

$$J_c = J_n + J_p$$

$$\frac{v_s}{2} \frac{dJ_c}{dt} = J_c \left( \int_0^L \alpha dx - 1 \right) + J_s \quad \text{Read equation}$$

$$v_s = I_s / v_s$$

$J_s$  : saturation current density

$$R_D = \frac{1}{\omega C_d} \frac{(1 - \cos \theta_d) / \theta_d}{(1 - \omega^2 / \omega_s^2)}$$

$$\omega_s = \sqrt{2\alpha v_s J_0 / \epsilon} \quad J_0 : \text{dc current density, } \alpha' = \alpha / dE$$

## Limitation of output power of IMPATT diodes

- $I_s/w$  increases with increasing doping concentration
- The fraction of the "dead" space increases with decreasing avalanche zone width
- For high frequencies the generation of carriers is reduced because of the finite intrinsic avalanche response time
- With increasing electric field more carriers are generated by tunnel effect
- The saturation drift velocity reduces with increasing electric field ( $\omega_s$  is reduced)
- The ionization rates saturate with increasing electric ( $\omega_s$  is reduced)
- The thermal conductivity limits the maximum input power  
Si : 1.45 W/cmK, InP : 0.66 W/cmK, GaAs : 0.44 W/cmK

## IMPATT diode

$$\theta_i = \pi$$

$$Prf = -U_{rf} I_0 \frac{(1 - \cos \theta_d)}{\theta_d} (1 - I_s / w)$$

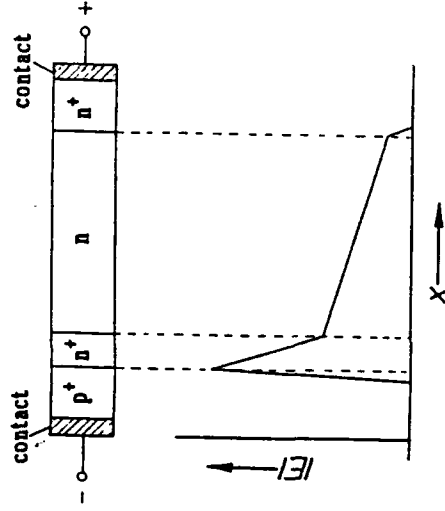
$l_a$  : length of the avalanche zone  
 $w$  : total width

$$\theta_{dopt} = 3\pi/4$$

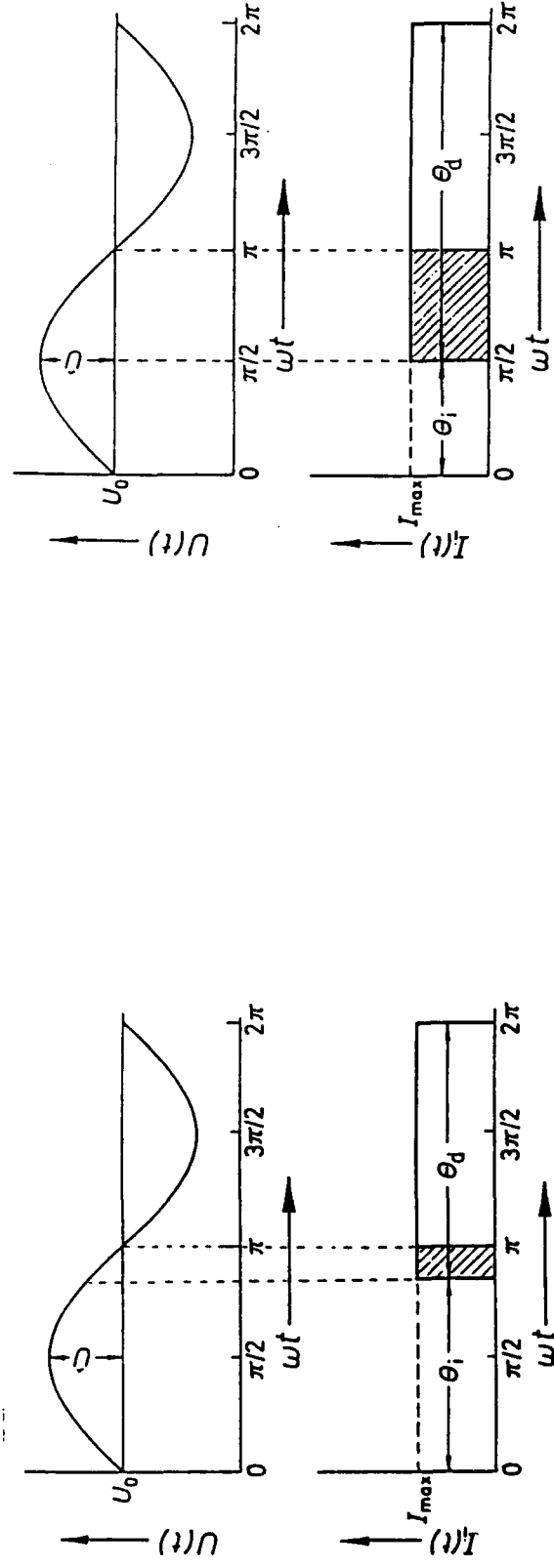
$$\text{optimum drift length : } l_d = \frac{\theta_d v_s}{2\pi f} = \frac{3}{8} \frac{v_s}{f}$$

## Noise of IMPATT diodes

- The IMPATT mode exhibits an intrinsically high noise level
- The noise power is mainly generated by the statistical nature of the avalanche process
- The noise behaviour is dependent on
  - semiconductor material
  - diode doping profile
  - bias conditions
  - rf-signal levels
  - parametric interactions
- Generally GaAs IMPATT diode are less noisy than Si devices

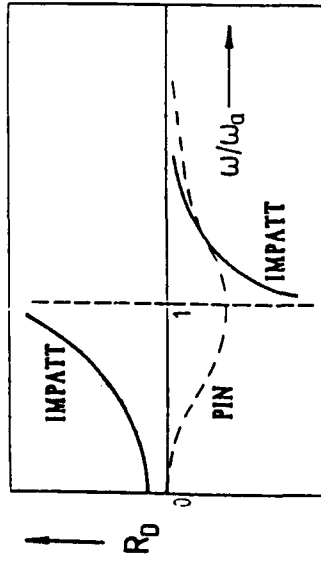


## Structure and field distribution of a TUNNETT diode



Rf-voltage and induced current of a

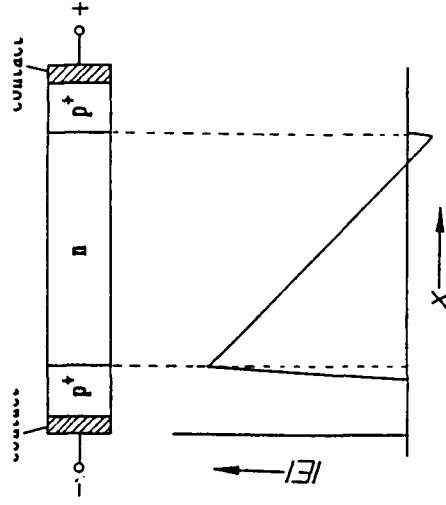
TUNNETT diode



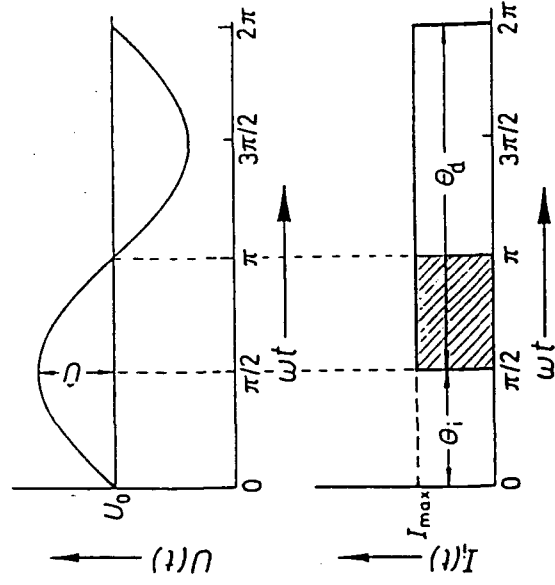
Real part of IMPATT and PIN-avalanche diode impedance versus frequency

### PIN-avalanche diode

- Negative resistance not only above avalanche frequency but also below
- Broadband negative resistance
- Advantage for pulsed diodes : relatively large negative resistance at  $\omega = \omega_a$
- The reactance is zero at  $\omega = \omega_a$ . No large capacitive currents which limit output power



Structure and field distribution of a BARITT diode



Rf-voltage and induced current of a BARITT diode

# TUNNETT diode

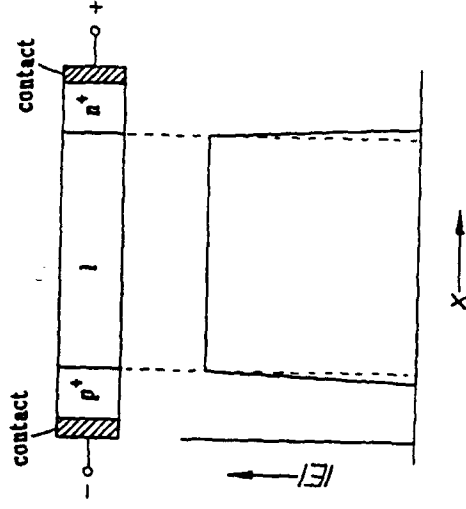
$$\theta_l = \pi/2$$

$$P_{rf} = U_{rf} I_0 \frac{\sin \theta_d}{\theta_d} (1 - I_a/w)$$

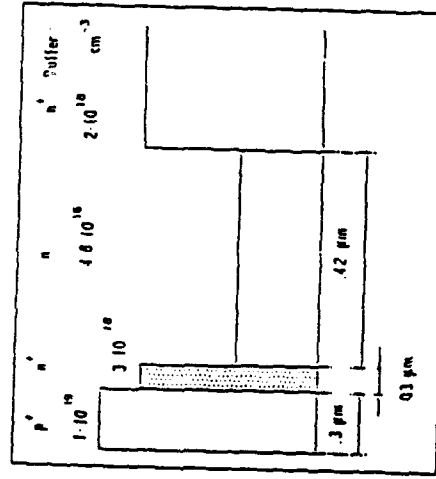
$I_a$  : length of the generation zone  
 $w$  : total width

$$\theta_{d_{opt}} = 3\pi/2$$

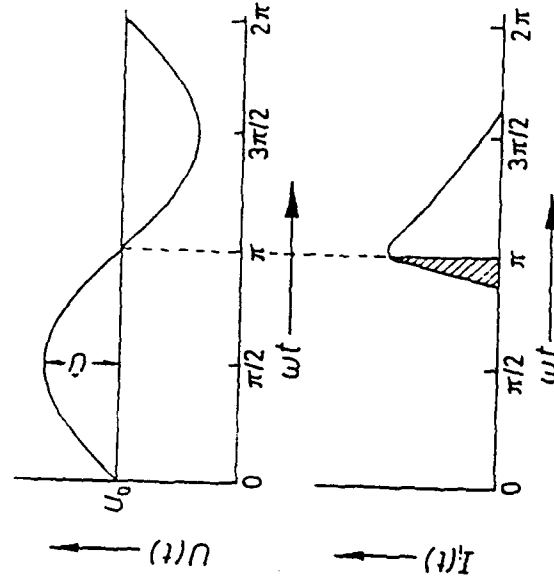
$$\text{optimum drift length : } l_d = \frac{\theta_d v_d}{2\pi f} = \frac{3}{4} \frac{v_d}{f}$$



Structure and field distribution of a PIN-avalanche diode



Doping profile of a W-band TUNNETT diode



Rf-voltage and induced current of a

## BARITT diode

$$\theta_1 = \pi/2$$

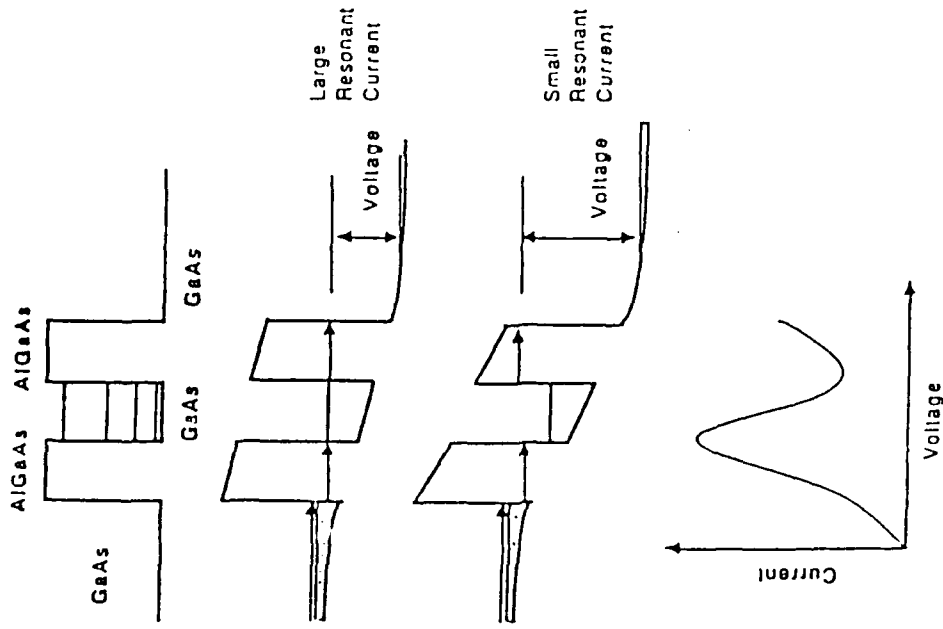
$$P_{rf} = U_{rf} I_0 \frac{\sin \theta_d}{\theta_d}$$

$$\theta_{dopt} = 3\pi/2$$

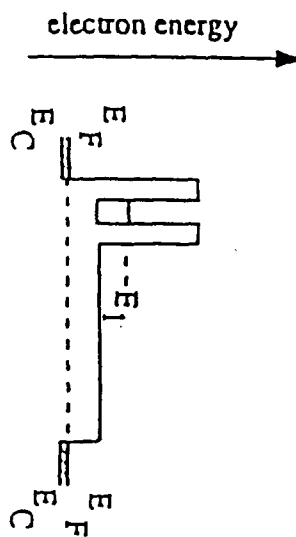
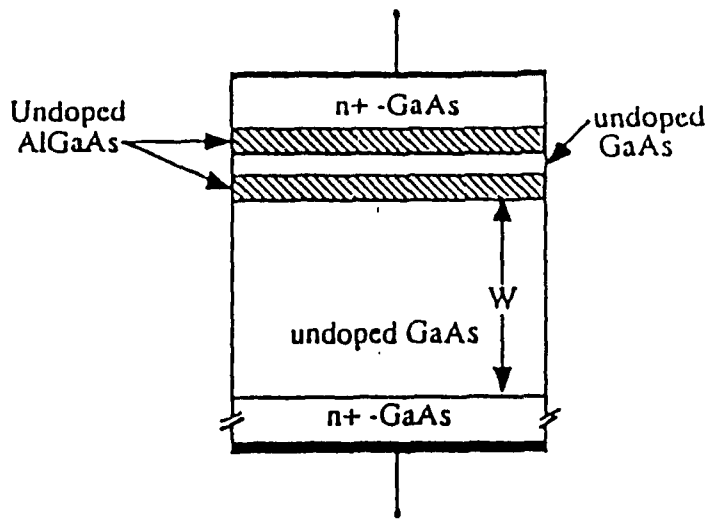
$$\text{optimum length : } l_d = \frac{\theta_d v_A}{2\pi f} = \frac{3}{4} \frac{v_A}{f}$$

## Noise of BARITT diodes

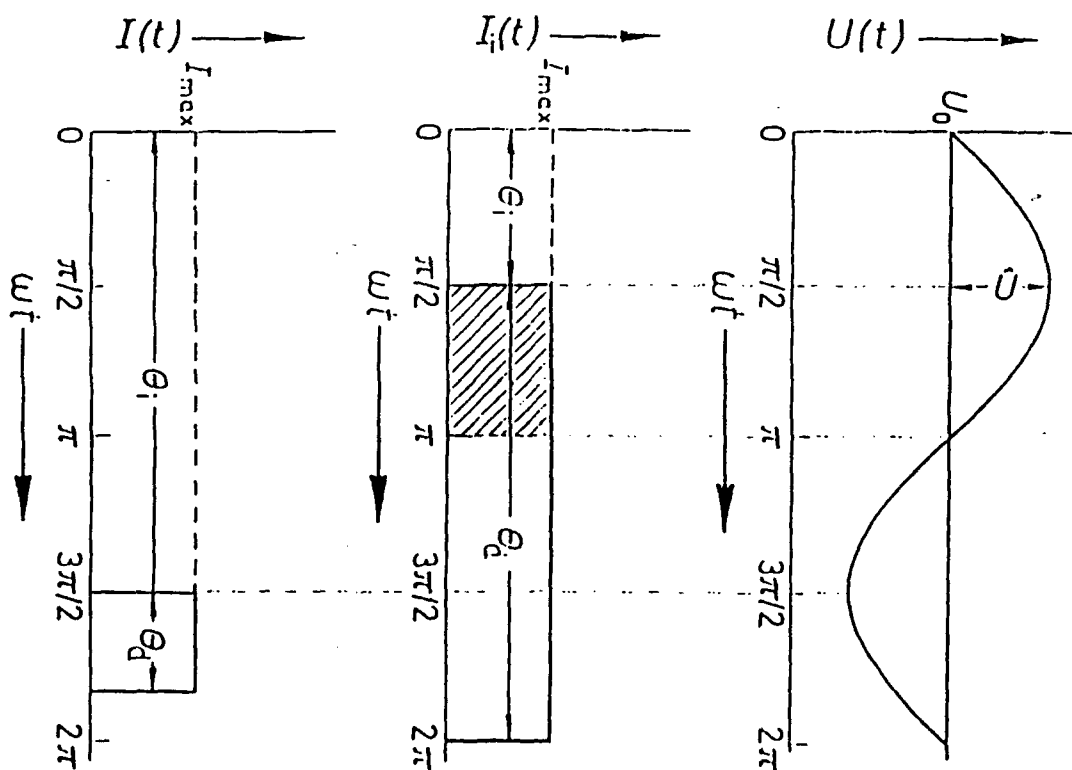
- diffusion noise of drifting carriers
- shot noise due to the injection mechanism
- low frequency noise (up-conversion)
- Phase noise as low as for Gunn elements
- Application as self-oscillating mixer  
60 GHz : minimum detectable signal  $\approx -160$  dBm  
conversion gain  $\approx 30$  dB  
(15 to 20 dB better than for Gunn elements)



Energy band structure of a quantum-well resonant tunneling device

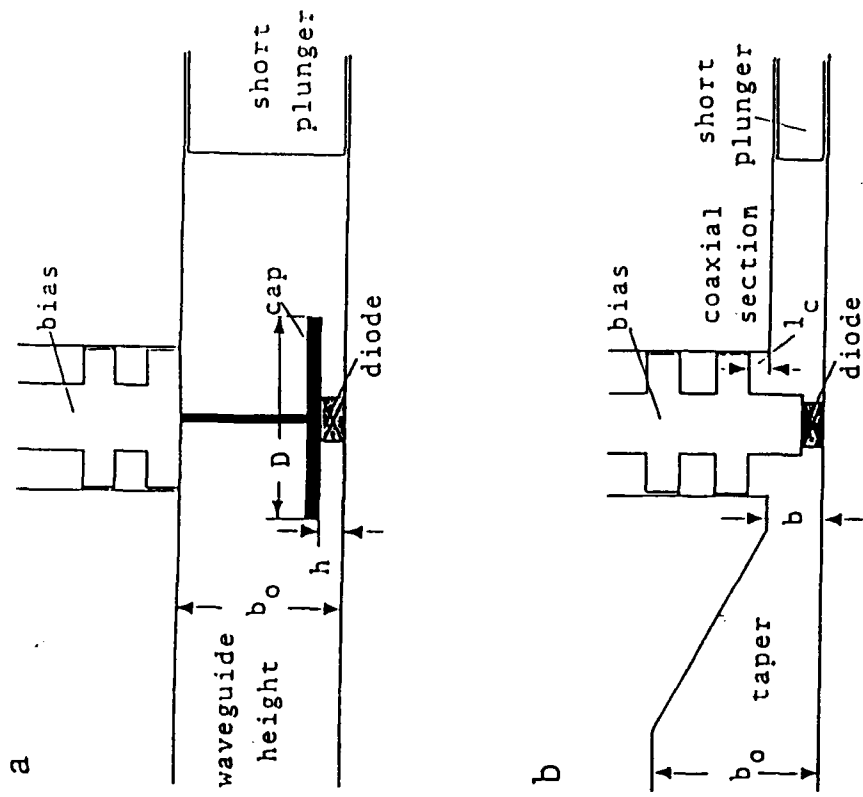
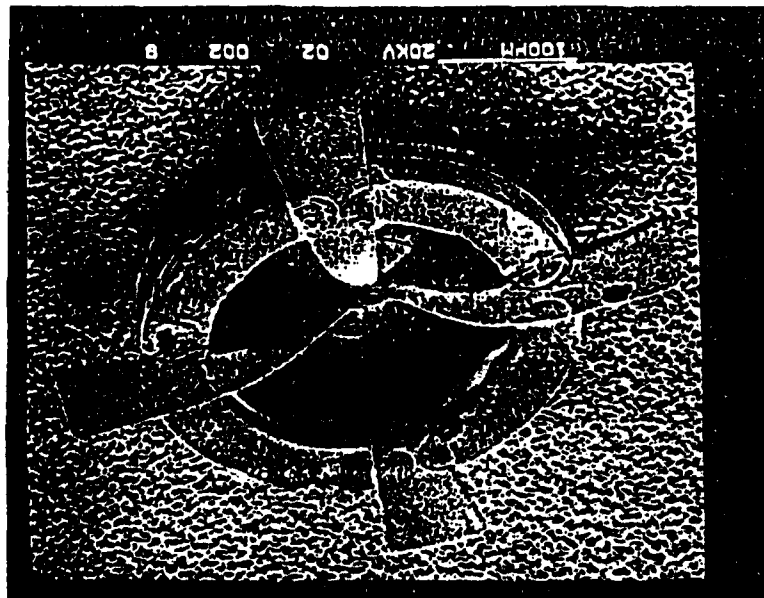


Structure and energy band diagram of a GaAs/GaAlAs QWITT diode



Rf-voltage and induced current of a QWITT diode  
 $(\Theta_1 = \pi/2 \text{ and } \Theta_1 = 3\pi/2)$

# Encapsulated W-band GaAs IMPATT diode



Full and reduced waveguide height resonator

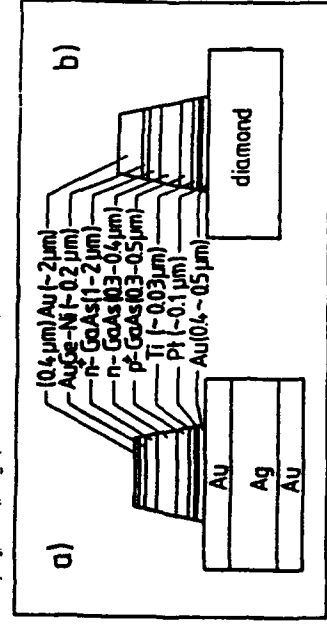
## Device fabrication

Si, GaAs, InP for mm-wave devices

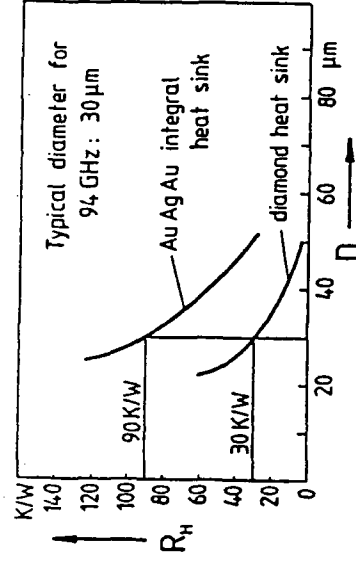
- Molecular beam epitaxy (important for elevated frequencies)
- Photorealist technology
- Etching processes (wet, dry)
- Metallization (evaporation, sputtering, plating)

## Reduction of losses

- low ohmic contacts
  - substrate thickness : 1 - 3  $\mu\text{m}$
  - low heat flow resistance (diamond)
  - reduction of parasitic capacitance and inductance
  - matching to the resonator impedance
- 94 GHz :  $C < 0.07 \text{ pF}$ ,  $L < 15 \text{ pH}$   
quartz ring height  $< 75 \mu\text{m}$

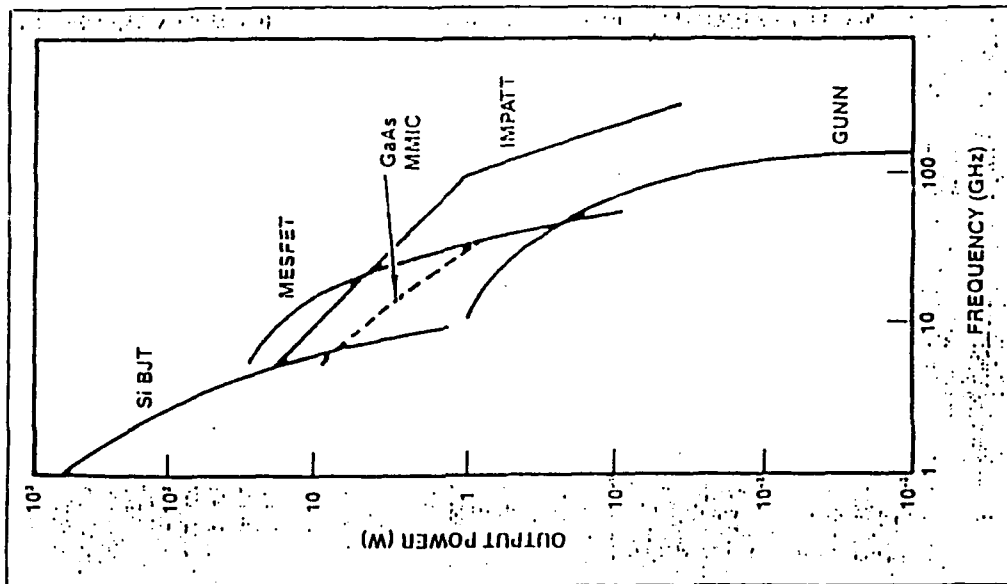
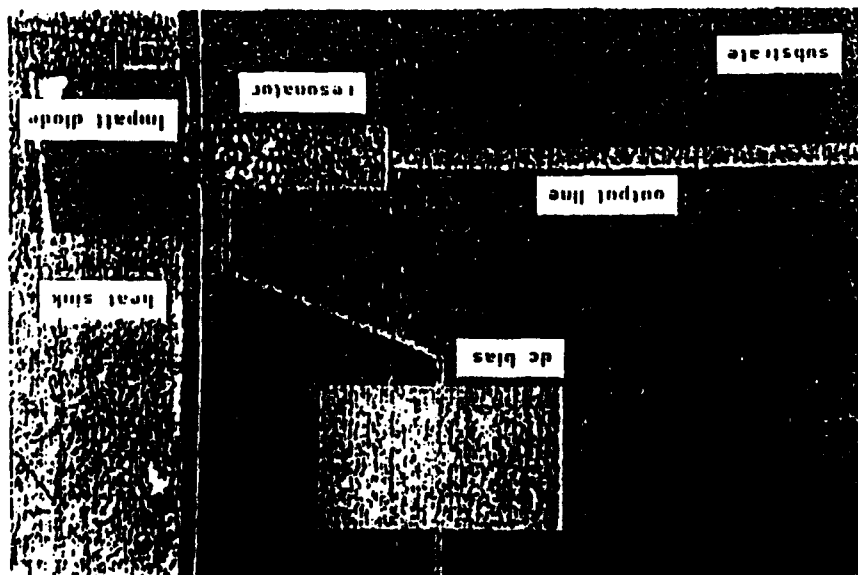


GaAs IMPATT diode with Au/Ag/Au and diamond heat sink

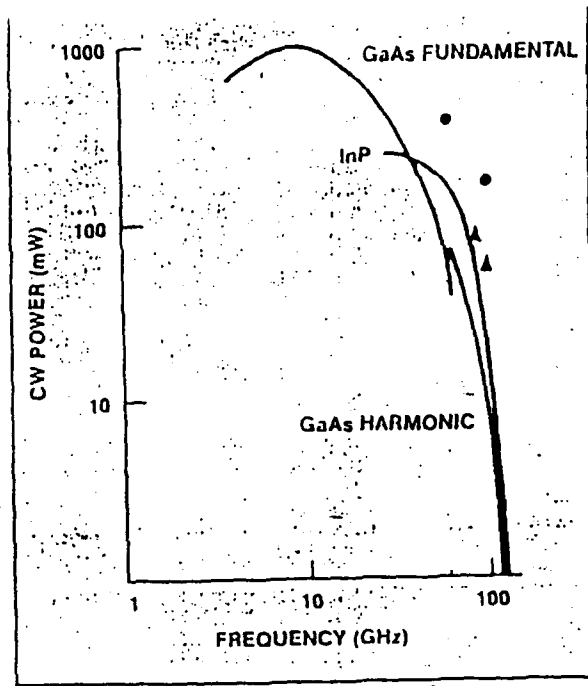


Heat flow resistance of 94 GHz IMPATT diodes versus device diameter

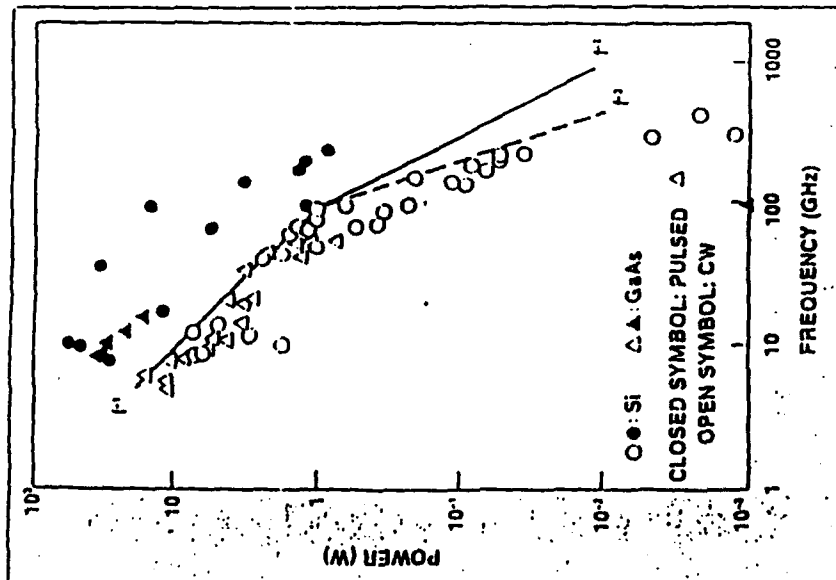
Photograph of a  $\lambda/4$  strip-line resonator on seminsulating GaAs substrate



State-of-the-art performance of solid-state power devices (after Shih and Kuno)



Cw-output power of Gunn elements  
 (▲ GaAlAs/GaAs launcher  
 ● InP current limiting cathode)



State-of-the-art performance  
 of Si and GaAs IMPATT diodes  
 After Shin and Kuno)

## Maximum efficiency

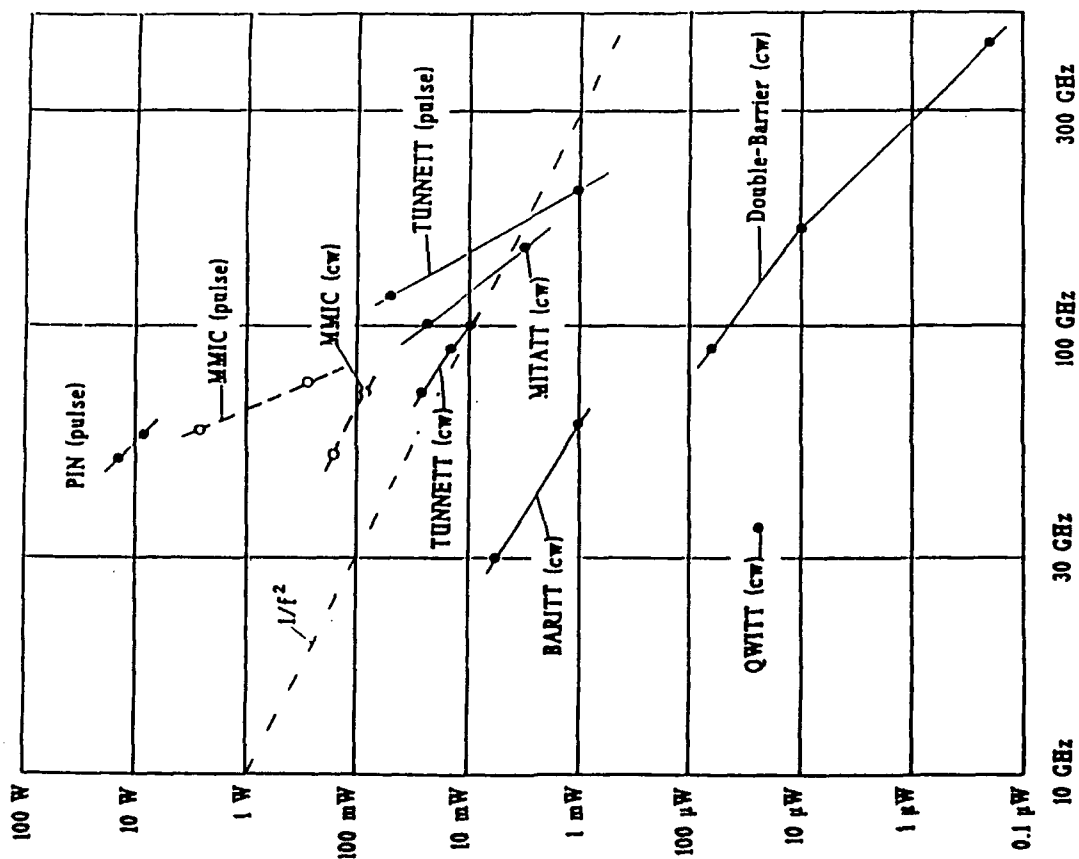
**Impatt**    20 %     $f < 40 \text{ GHz}$     for GaAs devices  
               10 %     $f > 40 \text{ GHz}$   
               10 - 12 %    for silicon devices

**Gunn**    2 - 5 %    for GaAs devices  
              5.7 % at 94 GHz    for InP devices  
              (12 % at 60 GHz, pulsed, for InP

**TUNNETT**    2 %    at 94 GHz

**BARRITT**    0.4 %    at 60 GHz

**MITATT**    0.5 %    at 150 GHz



Cw and pulsed output power of different mm-wave devices

## Noise measure

$$M = \frac{F - 1}{1 - 1/G} = \frac{\overline{u_n^2}/B}{4kT_0|R_D|}$$

**F** : Noise figure

**G** : amplification

$\overline{u_n^2}$  : primary noise voltage source

**B** : bandwidth

**k** : Boltzmann constant

**T<sub>0</sub>** : room temperature

**R<sub>D</sub>** : negative diode resistance

$$M = \frac{P}{kT_0 B} \frac{\Delta f_{rms}^2 Q_{ex}^2}{f_0^2}$$

94 GHz :  $M_{TUNNETT} \approx 18$  dB

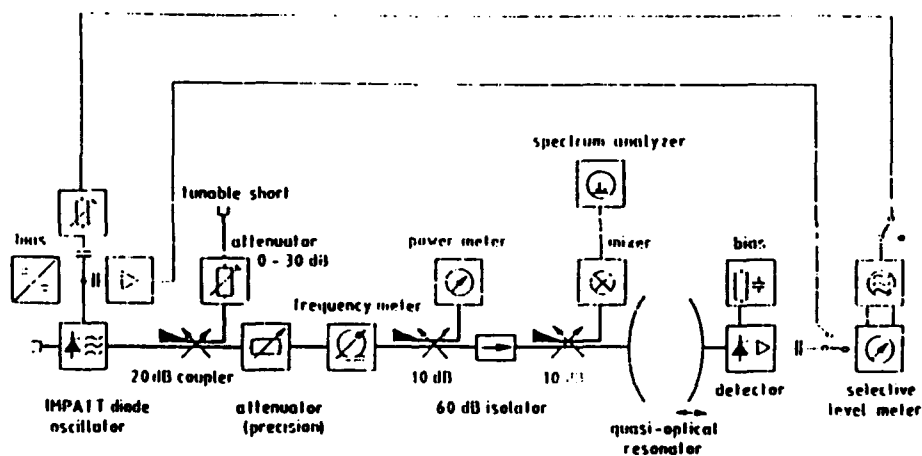
$M_{IMPATT} \approx 21$  dB

**P** : rf power

$\Delta f_{rms}$  : rms frequency deviation

$f_0$  : oscillation frequency

$Q_{ex}$  : external quality factor



Noise measurement set-up

**TECHNOLOGY DEVELOPMENTS TOWARDS NANOMETRIC  
STRUCTURES**

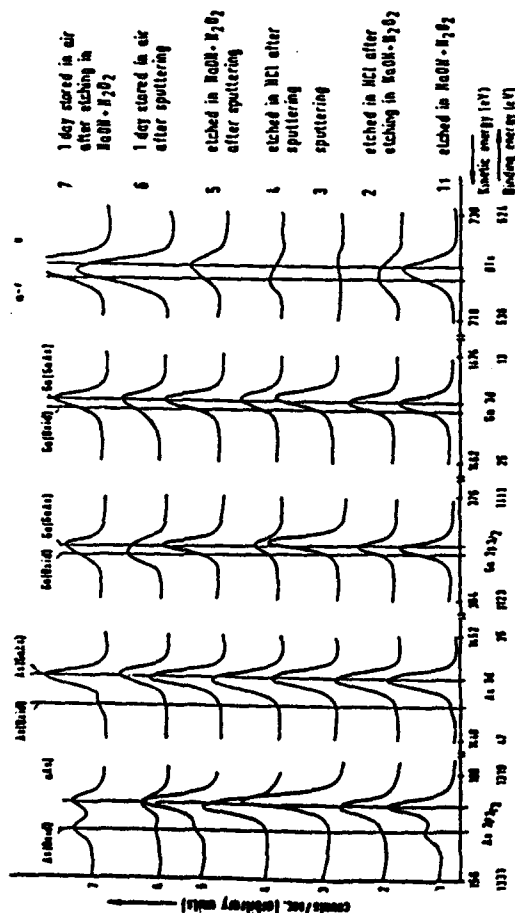
**H. L. Hartnagel**

**University of Darmstadt • Germany**

# Technology Developments towards Nanometric Structures

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61 Darmstadt  
Germany

- ) Basic Considerations towards Nanometric Devices
- ) Selective Epitaxy
- ) E-beam Lithography and soft Electrolytic Processes:  
Anodic, Etching and Cathodic metal deposition.
- , Local-Structure Deposition by point-like Electrode Current scanned by Piezo-Actuators.
- ) Evaluation of Nanometric Surfaces
- ) Outlook



XPS-spectra of GaAs(100) surfaces after different surface treatments  
The As 2p<sub>3/2</sub>, As 3d, Ga 2p<sub>3/2</sub>, Ga 3d and O 1s core level peaks are shown

## Basic Considerations towards Nanometric Devices

### - Dimensional Reduction

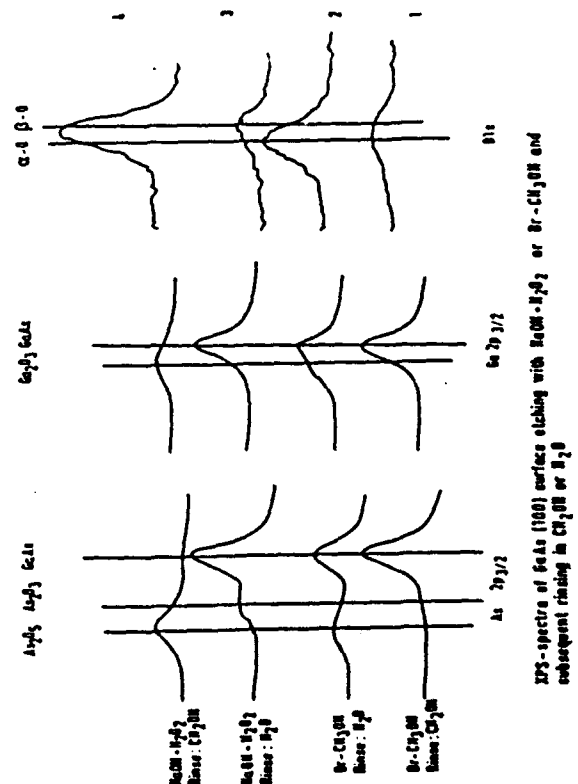
- a) hot-electron transport  $\longrightarrow$  High Speeds
- b) high packing densities for highly complex signal processing
- c) Impedance optimization for capacitances based on optimal space-charge layers for mm waves:

$$|Z_c| = \frac{1}{\omega C_j}$$

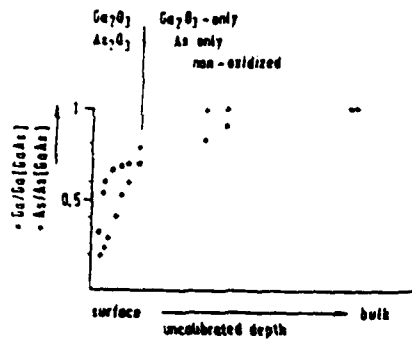
- d) minimization of heat loss for current crowding along emitter periphery of power transistors for mm waves

### - Increase of Surface to Volume Ratio

- a) no damage acceptable leading to traps
- b) space-charge layer minimization to obtain carrier-density independent, controllable conditions

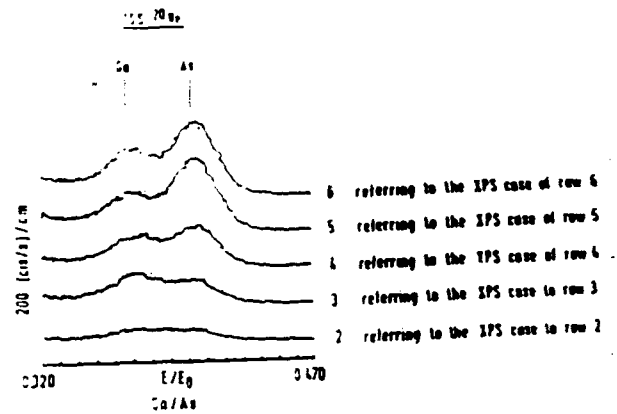


XPS-spectra of GaAs(100) surfaces etching with H2O2-H2O2 or Br-CH3OH and subsequent rinsing in CH3OH or H2O

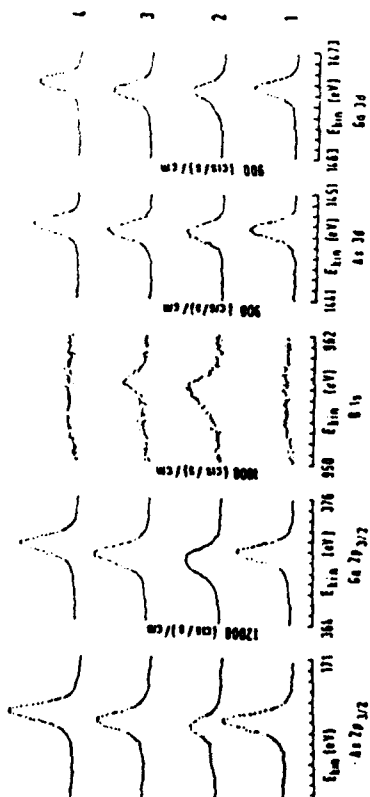


Depth profile of Ga and As by detailed  $^{70}\text{Re}$ -ISS analysis  
 $\text{Ga}/\text{GaAs}$ : concentration of Ga in GaAs-compound  
 $\text{As}/\text{GaAs}$ : concentration of As in GaAs-compound  
 $\text{Ga}$ : concentration of Ga after the given etch step  
 $\text{As}$ : concentration of As after the given etch step

## ISS spectra of GaAs

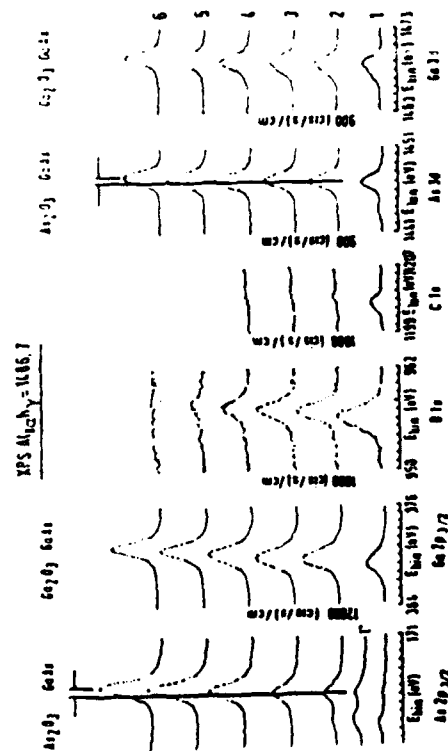


$^{70}\text{Re}$ -ISS spectra



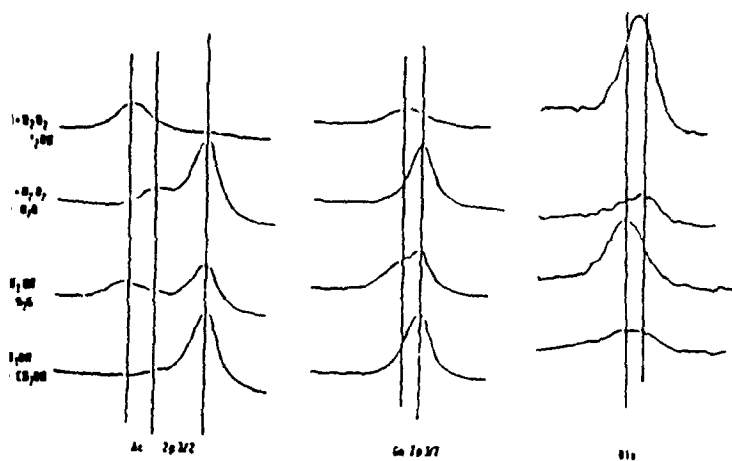
XPS-spectra

row 1 identical with Fig. 3 row 6  
 row 2 after reoxidation in air for 30 min  
 row 3 after short ionetching  
 row 4 after heavy Ar etching similar to that before spectrum given by Fig. 3 row 6



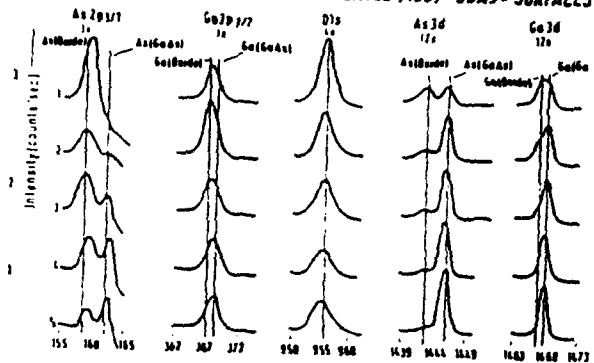
XPS-spectra

row 1 original GaAs-sample as inserted into the vacuum system  
 row 2 to 5 subsequent etching with  $\text{Ar}^+$  of different intensities and duration  
 row 6 last etch step when  $\text{Ar}^+$  was used ( $\text{Ar}^+$  is much heavier than  $\text{Ne}^+$ )



XPS-spectra of GaAs(100) surface etching with NaOH + H<sub>2</sub>O<sub>2</sub> or Br-Cl<sub>3</sub>OH and subsequent rinsing in CH<sub>3</sub>OH or H<sub>2</sub>O

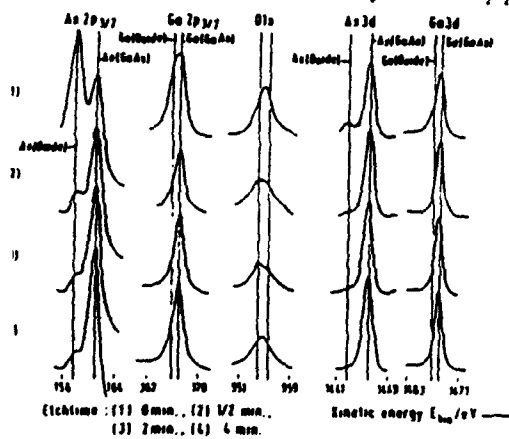
# XPS-SPECTRA OF DIFFERENTLY TREATED (100)-GaAs- SURFACES



The treatments based on: (1) H<sub>2</sub>SO<sub>4</sub>-H<sub>2</sub>O<sub>2</sub>,  
(2) only organic solvent cleaning,  
(3) NaOH-H<sub>2</sub>O<sub>2</sub>, (4) NaOH, (5) H<sub>2</sub>O<sub>2</sub>

Kinetic energy E<sub>kin</sub>/eV  
Binding energy E<sub>b</sub>=1686.6 eV (Al K<sub>α</sub>)

## XPS - SPECTRA OF SAMPLES TREATED BY NH<sub>4</sub>OH (25%):H<sub>2</sub>O<sub>2</sub>(30%):H<sub>2</sub>O (2:1:300)



Etchtime: (1) 0 min., (2) 1/2 min.,  
(3) 2 min., (4) 4 min.

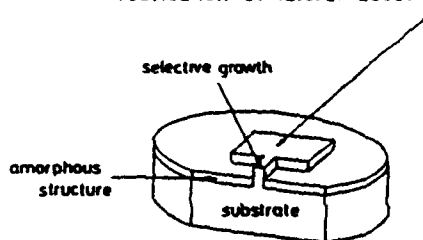
Kinetic energy E<sub>kin</sub>/eV

## Selective Epitaxy

Starting with high-temperature capable amorphous mask

Growth required with good quality on side walls

Same growth rate as with volume case to allow fabrication of lateral access structures



## E-beam Lithography and Soft Electrolytic Processes:

Anodic Etching and  
Cathodic Metal Deposition

- 1) nanometric structurization by electron-beam illumination
- 2) gas-plasma etching through E-beam resist pattern
  - a) steep structures possible
  - b) but ion bombardment minimized only by protective grids or plasma potential control
- 3) anodic wet etching through resist pattern
  - a) side-steepness by combined adjustment of bias and opt. illumination
  - b) passage through various epitaxial and doping layers by optical hole generation as required for anodic process

## Ion-Beam Lithography

$H^+$ , 60 KeV into PMMA  
spread:  $0.1 \mu m$   
no proximity effect  
no backscattering on Si

-15-

## E-Beam Lithography

pos. resist

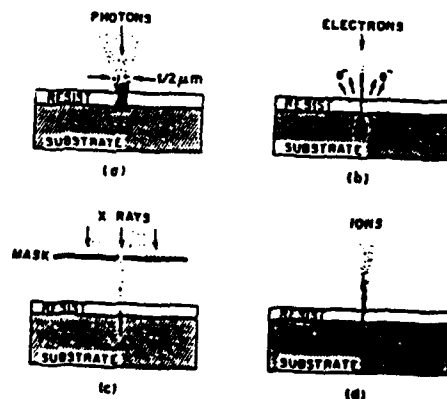
PMMA polymethyl metacrylate  
PBS polybutane -1 sulfone  
resolution:  $1 \mu m$

neg. resist:

COP Polyglycidyl methacrylate -  
co-ethyl acrylate  
resolution due to swelling:  $1 \mu m$

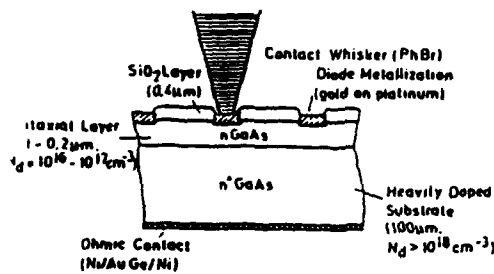
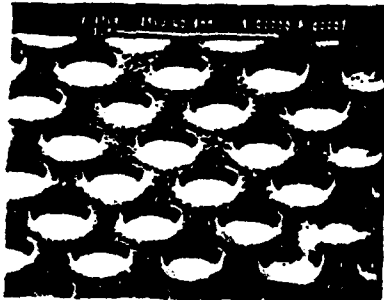
limitation:

electron scattering, 20 keV  
( $\lambda \sim 1 \text{\AA}$ ); proximity effect



Types of advanced lithographic methods. (a) Optical lithography. (b) Electron beam lithography. (c) X-ray lithography. (d) Ion beam lithography.

# Hottky-Barrier-Dioden für Sub-mm-Wellen-Mischung

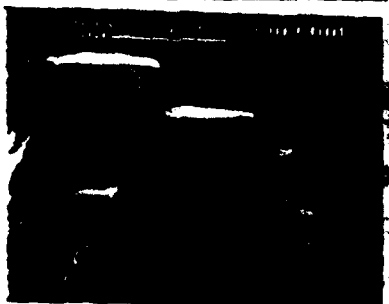
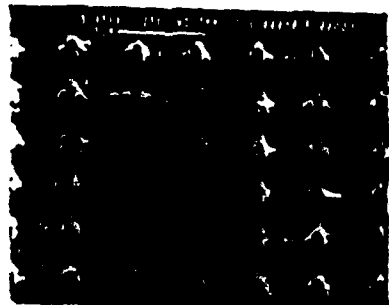
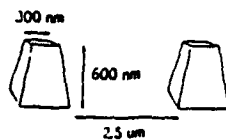


Schematischer Diodenquerschnitt

Etched structures (1):

(2)

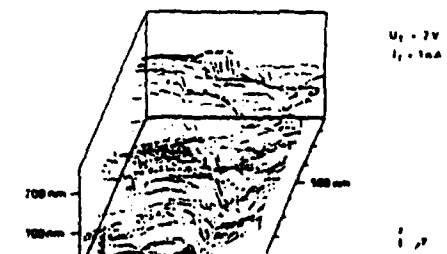
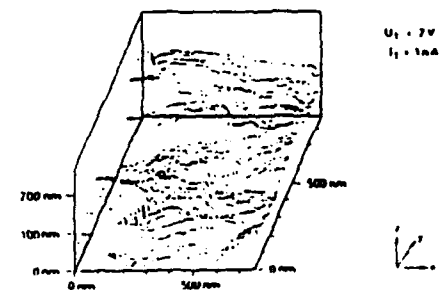
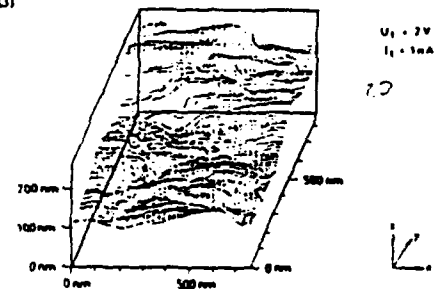
of GaAs towers



## Continuation: Electrolytic Processes

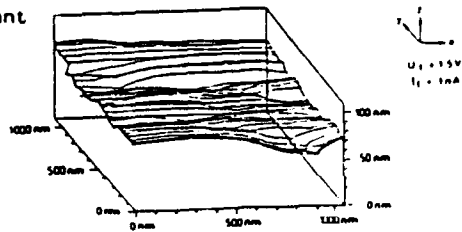
- 3)
  - c) Selection of electrolyte: reaction-limited solution instead of diffusion limited one
  - d) Equal solution rates for each of the components of the semiconductor
  - e) pulsed-bias
- 4) cathodic metal deposition through resist pattern
  - a) resulting metal thickness proportional to local current density
  - b) therefore illumination to increase current flow over layers of reduced conductivity
  - c) sometimes side-wall metallization avoided by intermediate insulating epi layer and deposition in darkness

Changes of surface structure  
in air after treatment b)  
measured by STM  
(15min between each  
measurement)

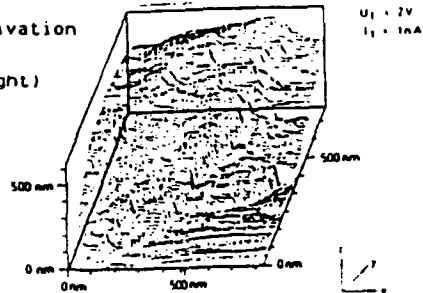


## STM-Studies of GaAs(100) Surfaces after different etch treatments

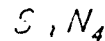
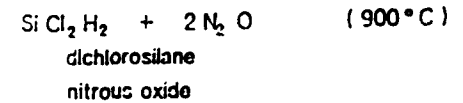
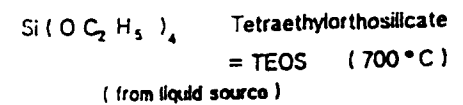
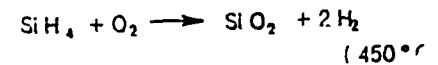
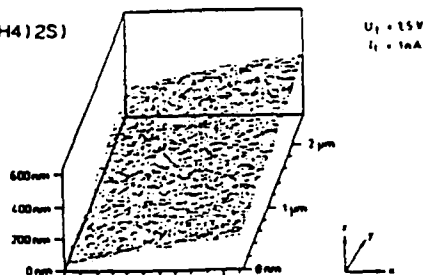
alkaline etchant  
140H:H2O2:H2O  
1:1:300)



photochemical passivation  
then 30min  
Cl:H2O 1:1, UV-light)



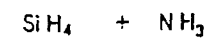
-passivation  
then 10min in (NH4)2S)



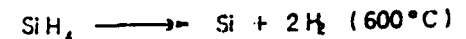
LPCVD : 750°C

PCVD : 300°C

( final chip passivation against  
scratch protection, moisture  
barrier, sodium barrier )



*Poly Si* ( larger stability than Al )



## Chemical Vapour Deposition

(e.g.  $SiO_2$  to insulate multilevel metallisation)

1) atmospheric -pressure CVD

2) low -pressure CVD

3) plasma -assisted CVD

- Hot -wall reactor

- Low -Temperature CVD for Si submicron techn.)

(reduced thermal broadening of dopants  
by diffusion)

energy -enhanced CVD

(possible:  $SiO_2$  at 150 Å/min for 50°C

by UV light ; in

" Chemical Vapour Deposition "

1984, ed.: M.D. Robinson et al

Electrochem. Soc., Pennington, USA)

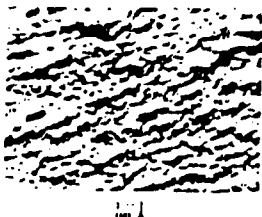
## Cont.: Electrolytic Processes

4)

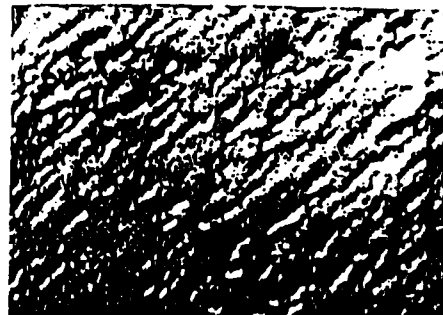
d) metal sequences by several  
cathodic steps as required  
for  
- ohmic contacts  
- Schottky contacts  
- good adhesion  
- high stability (diffusion  
barrier)

e) selective deposition on  
semiconducting islands in  
seminsulating surface in  
darkness

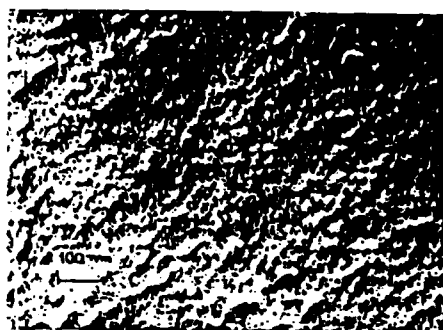
f) pulsed bias



DOUBLE-REPLICA TEM PHOTOGRAPH OF  
 $H_2SO_4:H_2O_2$  ETCHED GaAs SURFACE



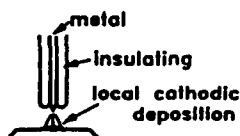
DOUBLE-REPLICA TEM PHOTOGRAPH OF Al-FILM SURFACE



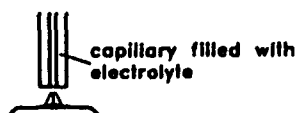
Local Structure Deposition by  
Point-like Electrode Current  
scanned by Piezo-Actuators

1) Point Current Source by

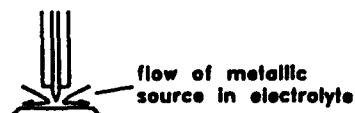
- a) needle with insulation except on tip



- b) fine capillary in conducting electrolyte



2) Pulsed Bias for optimization of results



- 3) Position control by piezo actuators similar to Scanning Tunnelling Microscope.

- 4) Combination with local etching under point current source

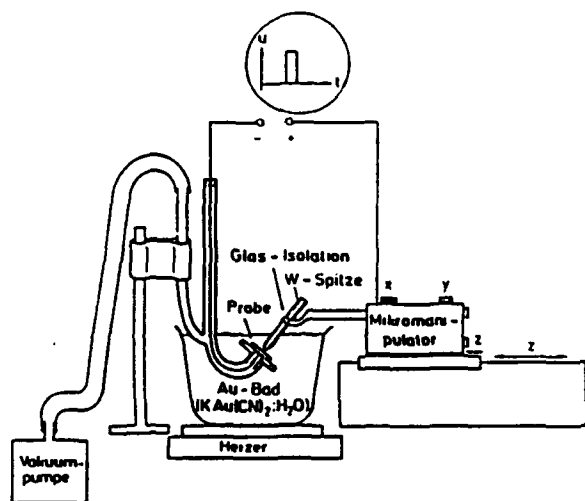


- 5) Surface preparation by focussed-electron (SEM) or focussed-ion illumination for local etching with point or line metallizations.

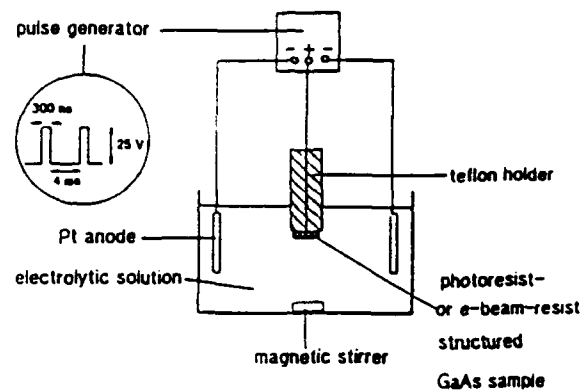
## Direct writing of submicron metallisation patterns on GaAs

Development of a new process for writing submicron metallisation patterns with a tungsten-tip on n-GaAs in an electrolytic Au-bath

Experimental set-up

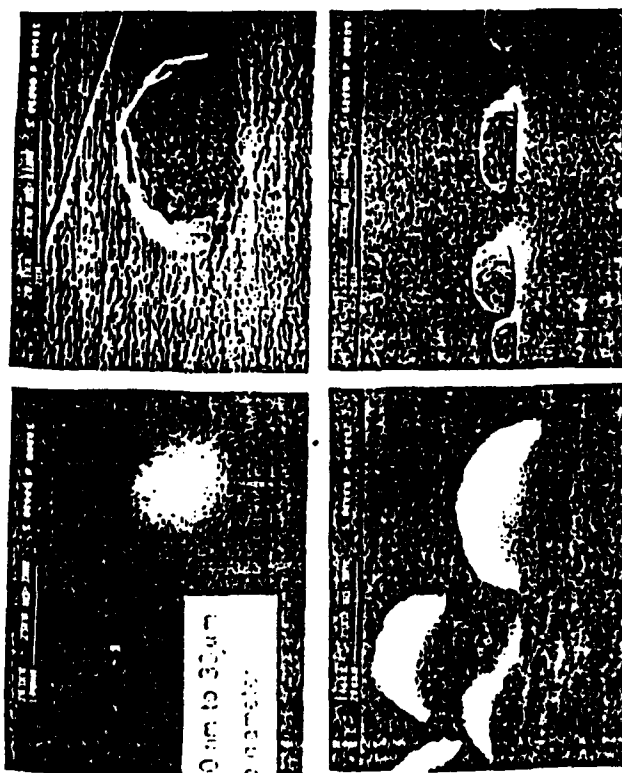


experimental setup:



Pt-solution temperature: 40°C

etch rate: 4 nm/pulse



### 3 Results

-dots:

diameter from 100 nm to 30 μm  
height  $\leq$  half the diameter

## Evaluation of Nanometric Surfaces

- 1) STM and its derivatives in particular scanning tunnelling luminescence
- 2) Electron-beam evaluation
  - a) SEM
  - b) Scanning beam cathodoluminescence
  - c) TEM, particularly at angle or transverse; possibly with double replica techniques.
- 3) Focussed-ion techniques  
incl. Ion Scattering Spectroscopy

- Soft structurization
- Local epitaxy
- relevant structure characterization
- will give
- mm-wave devices and IC's
- high packing densities for complex signal processing with short interconnection delays
- extension to  $\mu\text{m}$  waves
- broad-band approach

# **OPTICAL CONTROL OF MILLIMETER - WAVE CIRCUITS**

**T. Itoh**

**UCLA • USA**

# PROGRESS OF A TUNABLE ACTIVE BANDPASS FILTER

Professor Tatsuo Itoh

Electrical Engineering Department

University of California, Los Angeles

U.S.A.

## Introduction

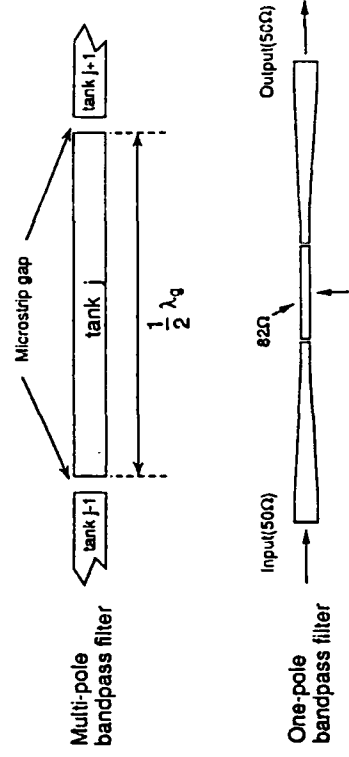
- Filter in MMIC form is difficult to realize because of the low Q value for MMIC element.  
→ **Active Filter**
- Direct optical control of microwave devices has been an area of growing interest.  
→ **Tunable Active Bandpass Filter Using MESFETs**
- Use of both negative resistance and reactance of an active device has design potential in microwave active circuit.  
→ **Tunable Active Bandpass Filter Using one MESFET**

## AGENDA

- Introduction
- Passive Bandpass Filter
- Active Bandpass Filter
  - coupled negative resistance method
- Tunable Active Bandpass Filter Using Varactor Diode
- Tunable Active Bandpass Filter Using MESFET as Varactor
  - optical tuning as well as electrical tuning
- Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor
  - different ways of using MESFETs as varactors
  - passive mode, active mode
- Conclusion

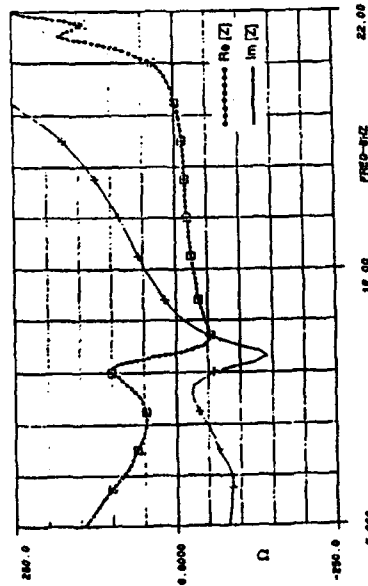
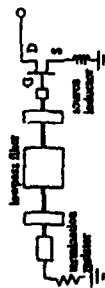
## Microstrip-line Bandpass Filter

- Passive circuit
  - loss is large, Q value is low
- Half-wavelength resonator
  - determine the frequency of each pole
- End-coupled bandpass filter
  - Chebyshev response if lossless



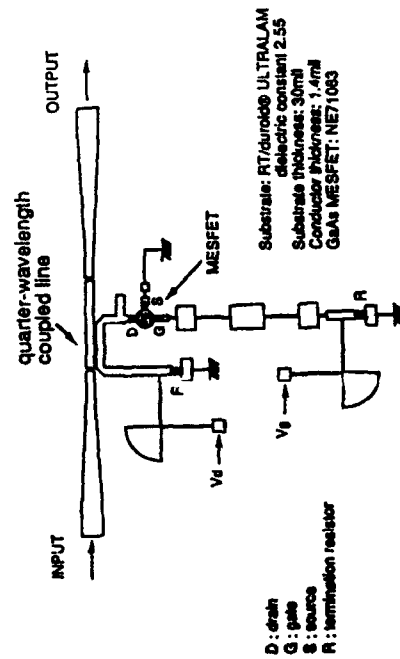
## Active Bandpass Filter

Simulation Result of Negative Resistance Using NE71083



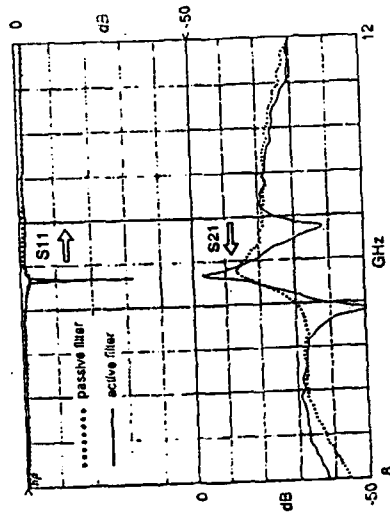
## Active Bandpass Filter

- Coupled negative resistance – compensate the loss of resonant tank, increase the Q value
- Fixed center frequency of passband



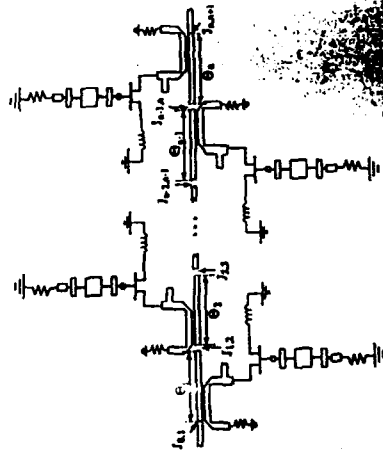
D : drain  
G : gate  
S : source  
R : termination resistor

- Comparison of passive filter and active filter
- The Q value is increased by coupled negative resistance method



## Active Bandpass Filter

- Multi-pole configuration
- Coupled negative resistance in each tank circuit
- Chebyshev response

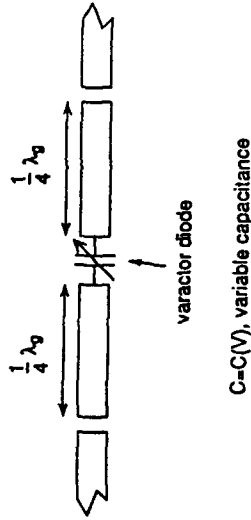


## Tunable Active Bandpass Filter Using Varactor Diode

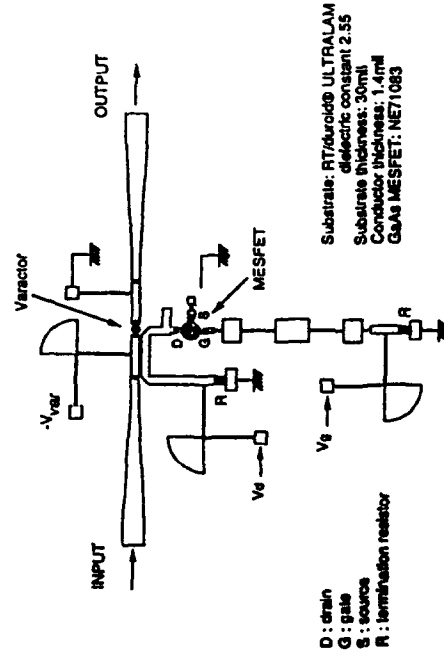
The function of varactor diode :

change the effective electrical length of half-wavelength resonant tank, thus change the center frequency of the passband

Two quarter-wavelength sections are connected by the varactor diode

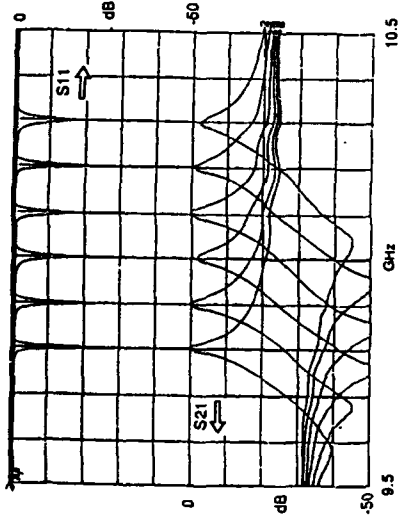


## Tunable Active Bandpass Filter Using Varactor Diode



## Tunable Active Bandpass Filter Using Varactor Diode

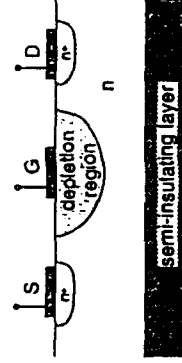
- 500 MHz tuning range
- MESFET bias voltage changes to compensate the different Q value of the tank



IEEE Trans. on MTT, MTT-38, December 1990

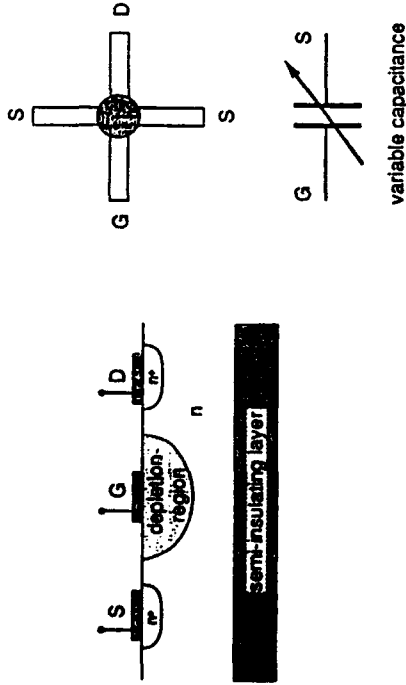
## Tunable Active Bandpass Filter Using

- Gate-to-source capacitance is a function of bias v
- used to replace the varactor diode



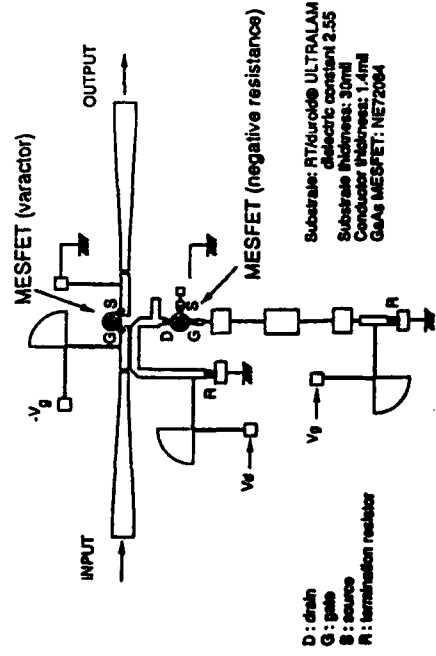
## Tunable Active Bandpass Filter Using MESFET as Varactor

- Gate-to-source capacitance is a function of bias voltage
- used to replace the varactor diode



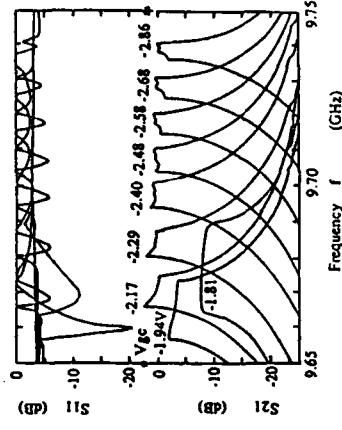
## Tunable Active Bandpass Filter Using MESFET as Varactor

- Both devices in this circuit are MESFETs – advantage



## Tunable Active Bandpass Filter Using MESFET as Varactor

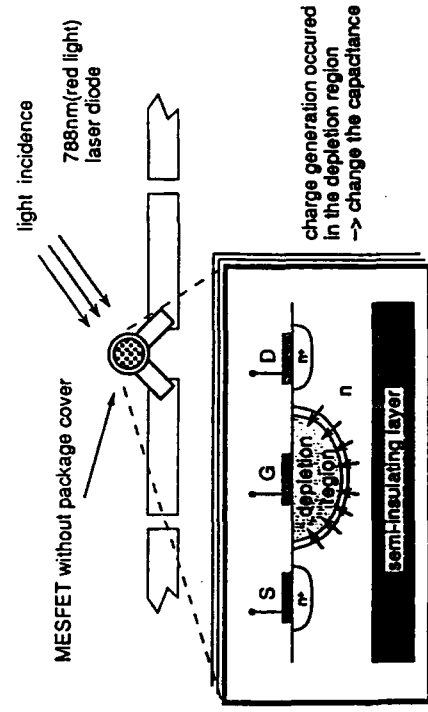
- electrical-tuning result
- 75 MHz tuning range without compensation of Q value



IEEE Microwave and Guided Wave Letters, May 1991

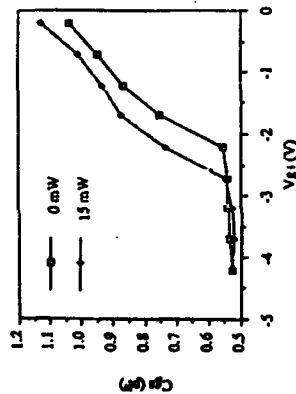
## Tunable Active Bandpass Filter Using MESFET as Varactor

- Optical control of solid state device



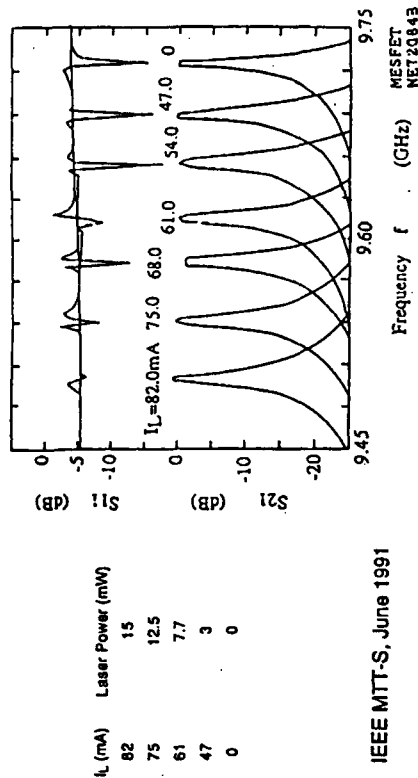
## Tunable Active Bandpass Filter Using MESFET as Varactor

$C_{gs}$  of the MESFET as a function of  $V_{gs}$  at two different levels of light illumination:  $P_L = 0$  mW and  $P_L = 15$  mW ( $P_L$  is the output power of a semiconductor laser diode, Sharp LT021MCO)



## Tunable Active Bandpass Filter Using MESFET as Varactor

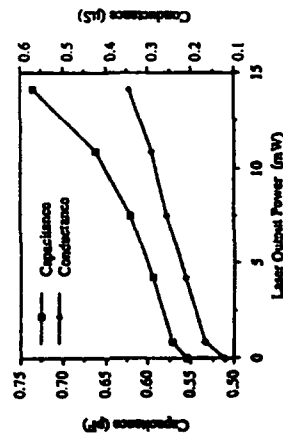
- optical-tuning result
- 225 MHz tuning range
- MESFET (negative resistance) bias voltage changes to compensate the different Q value of the tank



IEEE MTT-S, June 1991

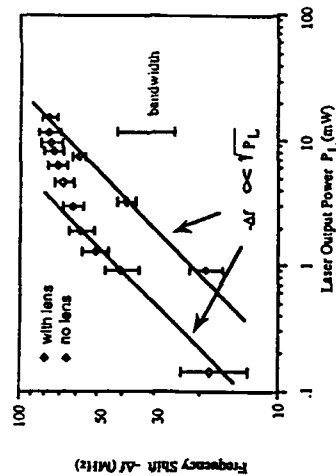
## Tunable Active Bandpass Filter Using MESFET as Varactor

$C_{gs}$  and  $G_{gs}$  as a function of the output power from a semiconductor laser diode (Sharp LT021MCO). Bias voltage ( $V_{gs}$ ) = -2.21 V



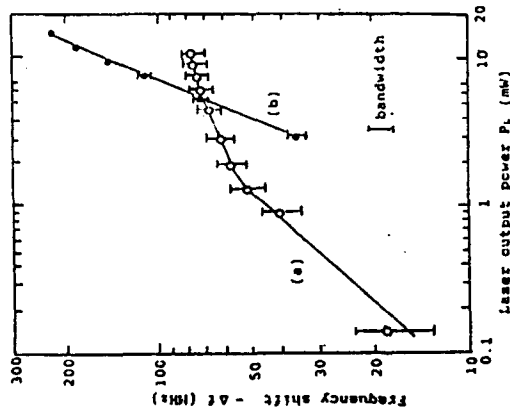
## Tunable Active Bandpass Filter Using MESFET as Varactor

- Optical Tuning
- frequency shift v.s. laser power



### Tunable Active Bandpass Filter Using MESFET as Varactor

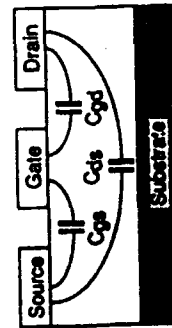
- Optical Tuning  
- frequency shift v.s. laser power



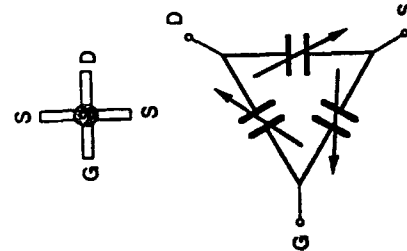
- (a) constant bias voltages
- (b) compensation case  
(varying bias voltages)

### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor

- MESFET in three-terminal application

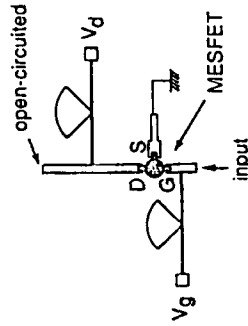


- Three capacitances are involved



### Using Three-terminal MESFET Varactor in the Passive Mode

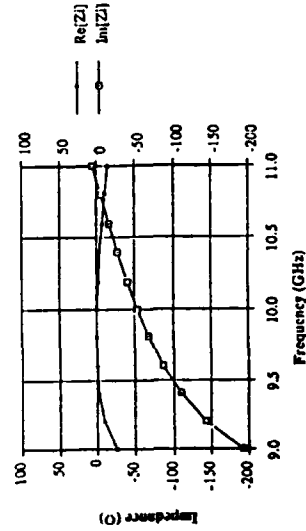
- Reactance-tuning circuit



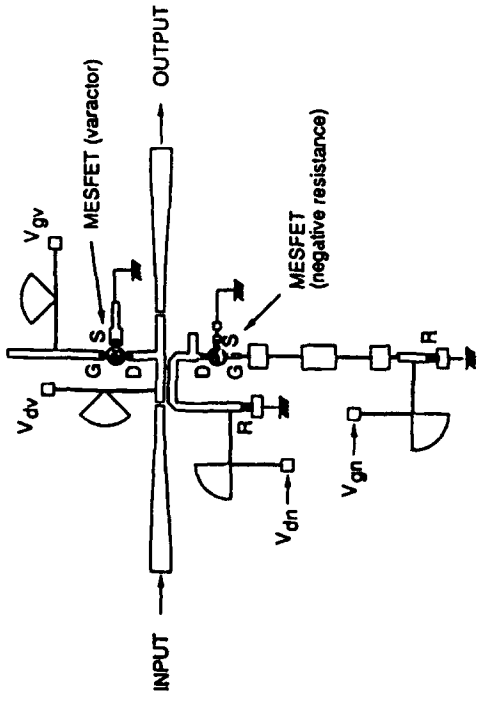
- The MESFET is biased in the passive mode
- The impedance at the input port is nearly reactance
- If resistance presents at the input port - Q value is affected

### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

- Simulation result of reactance-tuning circuit

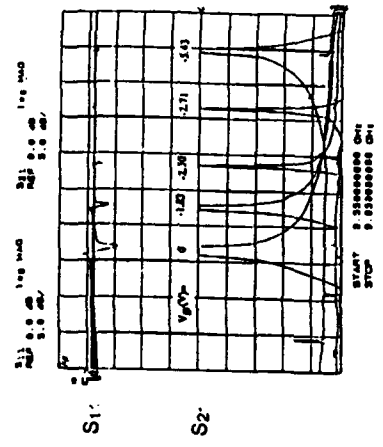


# Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

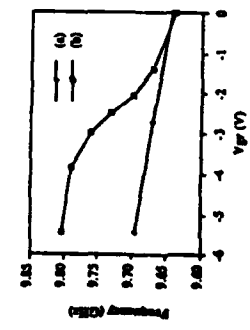
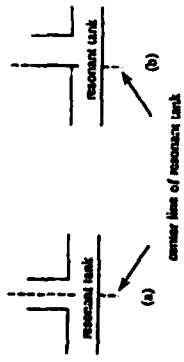


# Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

- electrical-tuning result
- no degradation occurs at  $V_{gr}=0$
- 170 MHz tuning range



# Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode



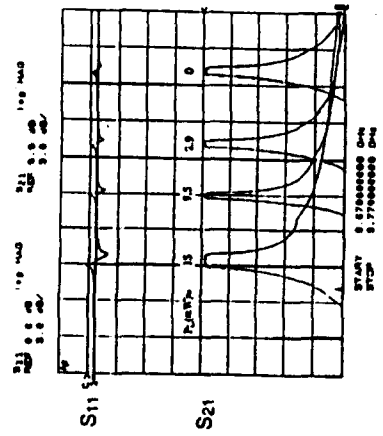
- The position to add the reactance – (b) is chosen

# Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

- electrical-tuning result
- $V_{gr}$  increases  
 →  $C_{gs}$  and  $C_{gd}$  increase  
 → electrical length increases  
 → frequency decreases

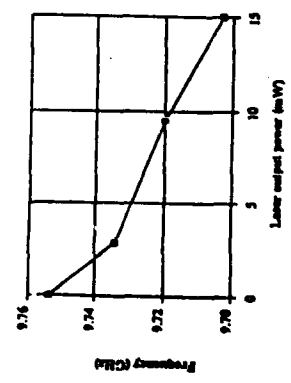
1115 700000 1000000

- optical-tuning result
- no degradation occurs
- 50 MHz tuning range

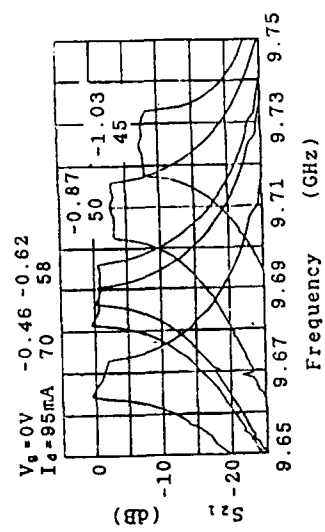


### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

- optical-tuning result
- Laser output power increases  
→  $C_{gs}$  and  $C_{gd}$  increase  
→ electrical length increases  
→ frequency decreases

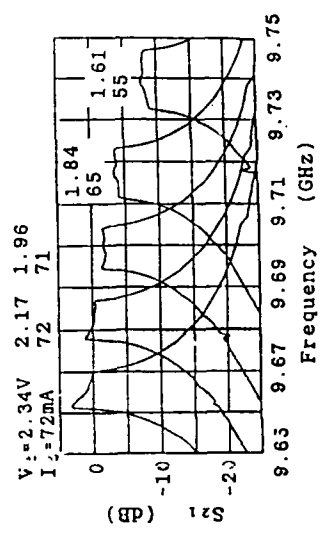


- Preliminary result from the filter using two-terminal MESFET varactor
- Passband shifts when  $V_g$  of the negative resistance MESFET varies from -1.03V to 0V
- $V_d$  of the negative resistance MESFET is kept constant (-2.13V)
- $V_{gc}$  of the MESFET varactor is kept constant (-2.19V)



### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- Preliminary result from the filter using two-terminal MESFET varactor
- Passband shifts when  $V_d$  of the negative resistance MESFET varies from 1.61V to 2.34V
- $V_g$  of the negative resistance MESFET is kept constant (-0.46V)
- $V_{gc}$  of the MESFET varactor is kept constant (-2.20V)



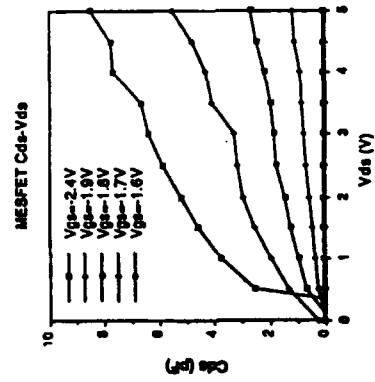
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

Some observations from the preliminary result

1. Tuning range of  $V_d$ -tuning is larger than tuning range of  $V_g$  tuning.
2. The tuning direction of  $V_g$ -tuning can be explained by the depletion-layer capacitance.  
 $V_g$  increases  $\rightarrow$  capacitance increases  $\rightarrow$  electrical length increases  $\rightarrow$  frequency decreases
3. The tuning direction of  $V_d$ -tuning cannot be explained by the depletion-layer capacitance.  
 $V_d$  increases  $\rightarrow$  capacitance decreases  $\rightarrow$  electrical length decreases  $\rightarrow$  frequency increases  $\rightarrow$  not agree with experimental result?

### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- Active-channel capacitance is important when the MESFET varactor is operated in the active mode.
- $C_{d3}$  is dominant in  $V_d$ -tuning
- Measured  $C_{d3}$ - $V_{ds}$  relation:



above  $V_{gs} = -1.6V$   
the channel is open  
and the conductance is  
too large  
 $\rightarrow$  can't measure  $C_{d3}$

1 MHz, HP4280A C-V Meter

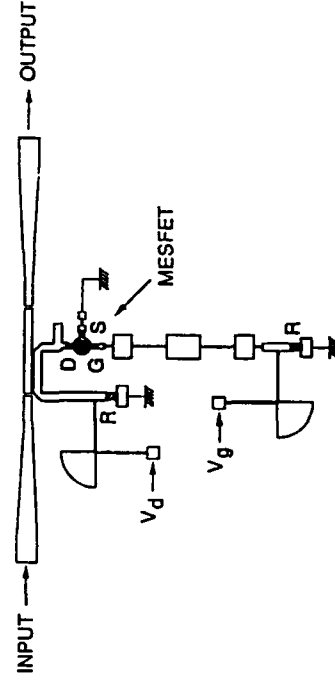
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

Try to design a tunable active bandpass filter using only one MESFET

- Capacitance variation is larger  $\rightarrow$  Tuning range is wider
- If  $V_{gs} = -1.6V$  is selected, capacitance variation is optimum
- To keep the insertion loss of active filter at 0dB, the separation of quarter-wavelength coupled line is adjusted from 18 mil to 8 mil
- Then, the gate voltage of the negative resistance needed to keep the insertion loss at 0dB is changed from -0.46V to -1.6V

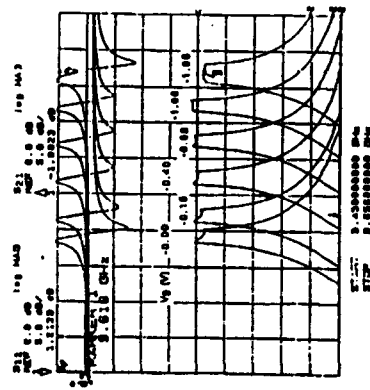
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor In the Active Mode

- Two MESFETs are replaced by only one MESFET
- The MESFET is biased in the active mode



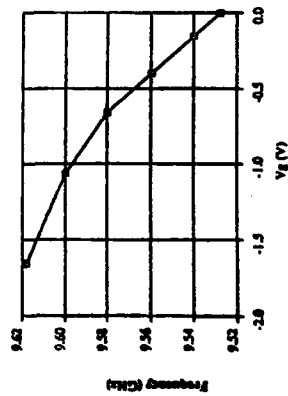
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of  $V_g$ -tuning
- 90 MHz tuning range – smaller than  $V_d$  tuning



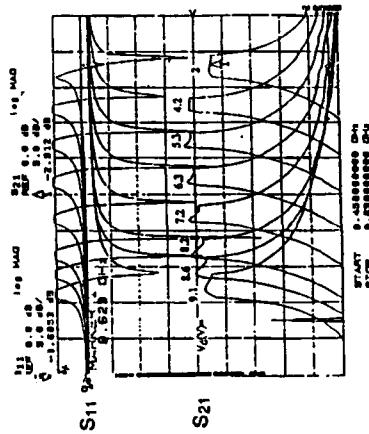
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of  $V_g$ -tuning
- $V_g$  increases
  - $C_{gs}$  and  $C_{gd}$  increase
  - electrical length increases
  - frequency decreases



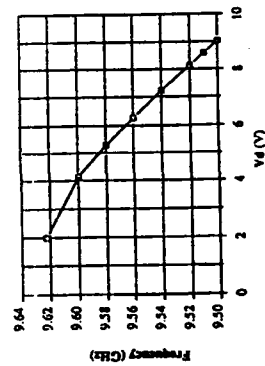
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of  $V_d$ -tuning
- 120 MHz tuning range



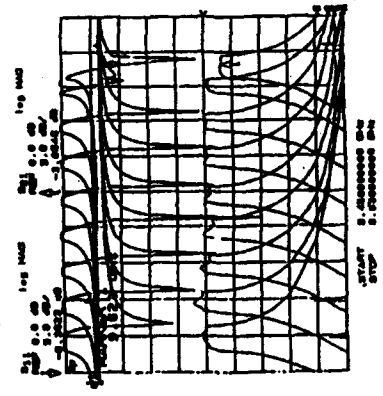
### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of  $V_d$ -tuning
- $V_d$  increases
  - $C_{gd}$  decreases, but  $C_{ds}$  increases and dominates
  - electrical length increases
  - frequency decreases



### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning by both  $V_d$  and  $V_g$
- 150 MHz tuning range



### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning by both  $V_d$  and  $V_g$

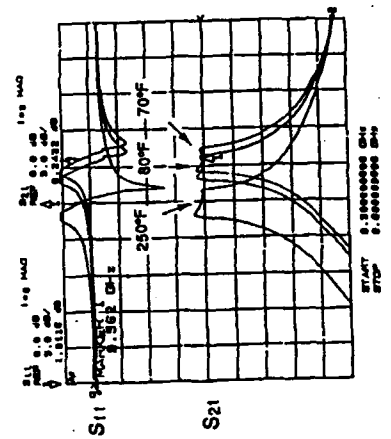
| Frequency (MHz) | $V_d$ (V) | $V_g$ (V) |
|-----------------|-----------|-----------|
| 2.620           | 1.85      | -1.38     |
| 2.610           | 2.27      | -1.38     |
| 2.599           | 2.99      | -1.67     |
| 2.570           | 3.64      | -1.67     |
| 2.520           | 4.64      | -1.64     |
| 2.520           | 2.54      | -1.64     |
| 2.510           | 3.21      | -1.55     |
| 2.490           | 3.23      | 0         |
| 2.470           | 2.63      | 0         |

### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- Optical tuning range is very small ( $<10$  MHz)
- The optimum gate bias for optical tuning is above -2V  
e.g. -2.35V for the three-terminal MESFET varactor in passive mode
- To keep the insertion loss of the filter at 0dB, a much higher gate bias is needed.  
e.g. -1.6V in  $V_d$ -tuning  
-1.6V to 0V in  $V_g$  tuning
- If the optimum bias condition for the optical tuning is the same as the bias condition needed for negative resistance, optical tuning becomes practical.

### Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor

- Temperature Stability
- 2 MHz shift from 70°F to 80°F



## Conclusion

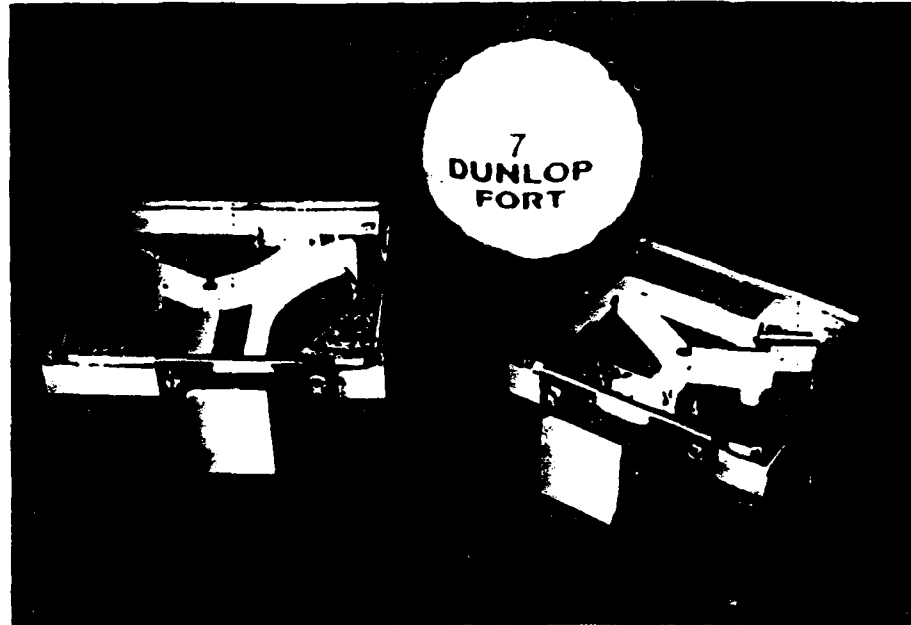
- Coupled Negative Resistance –  
increase the Q value for MMIC application
- Varactor Diode –  
tune the center frequency of the passband
- MESFET Varactor
  1. In addition to the electronic tuning, optical tuning is realized.
  2. Same type of device as the active device generating negative resistance.
  3. Operation in three-terminal mode is investigated.
  4. Show the possibility of using one active device to generate negative resistance and tune the frequency.

# **RECENT DEVELOPMENTS OF NRD - GUIDE TECHNOLOGY**

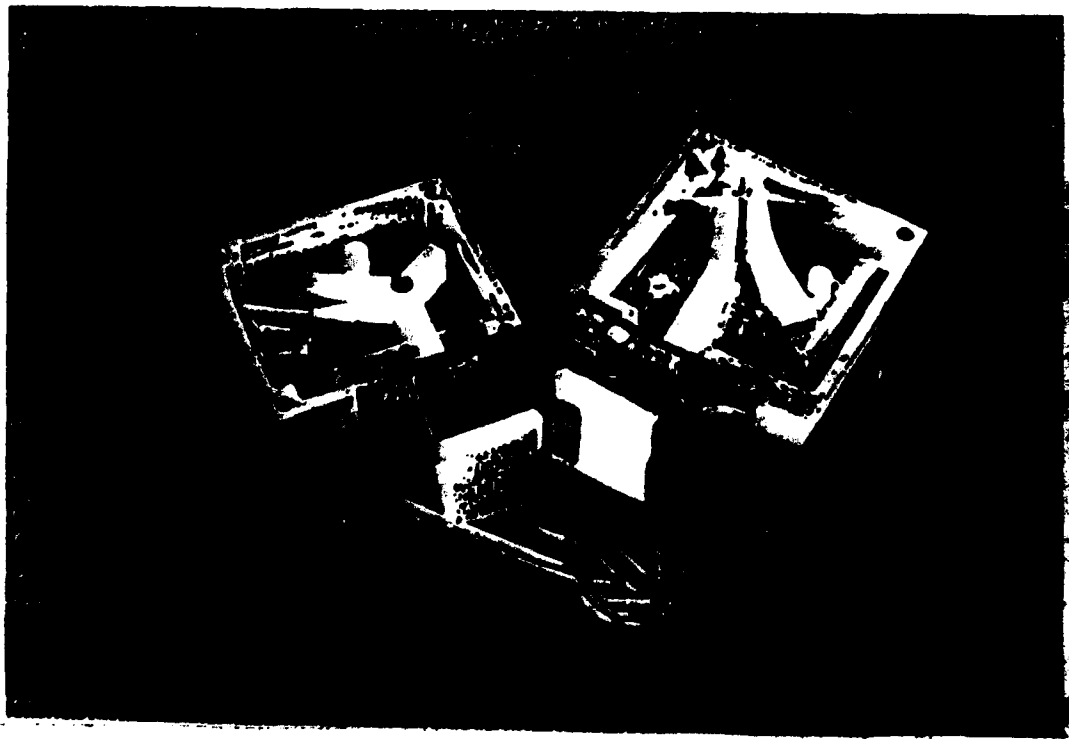
**T. Yoneyama**

**Tohoku University • Japan**

3



2



# Recent Development in NRD-Guide Technology

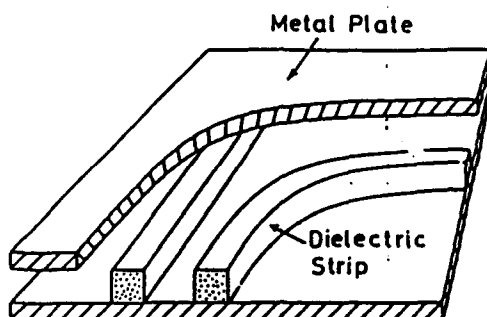
Tsukasa Yoneyama

Research Institute of Electrical Communication

Tohoku University

Katahira 2-1-1, Aoba-ku

Sendai, 980 Japan



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### I. Introduction

### II. NRD-Guide Integrated Circuits

#### 2.1 RF Front Ends at 35 GHz

#### 2.2 Gunn Oscillator at 60 GHz

### III. Proposal of Planar Antennas Fed by Leaky NRD-Guides

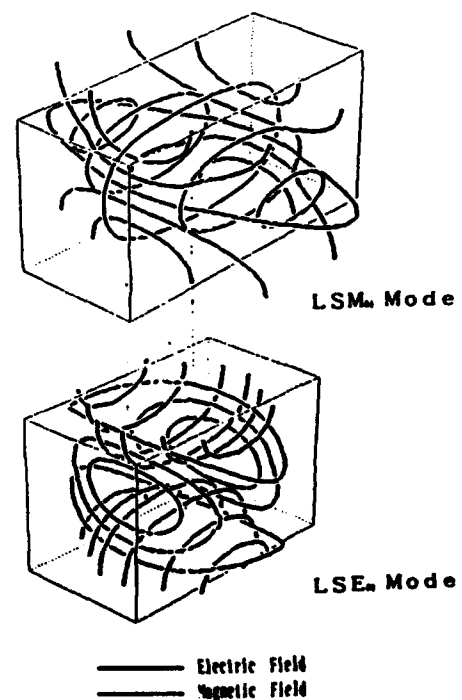
#### 3.1 Review of Uniform Leaky NRD-Guides

#### 3.2 Periodic Leaky NRD-Guides

#### 3.3 Planar Antennas Fed by Leaky NRD-Guides

#### 3.4 Applications

### IV. Conclusions



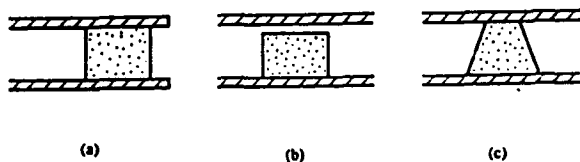


Fig.1 Classical types of leaky wave NRD-guides

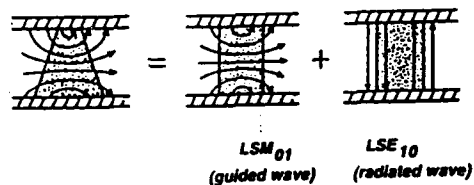


Fig.2 Operation principle of leaky wave NRD-guide

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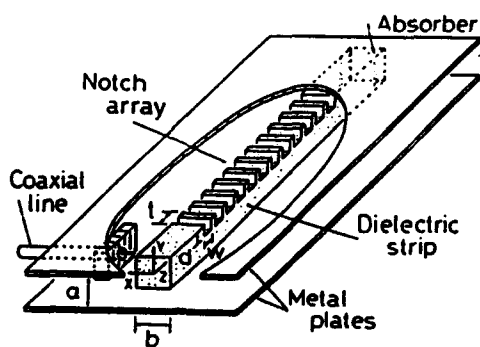


Fig.4 Leaky wave NRD-guide having notches

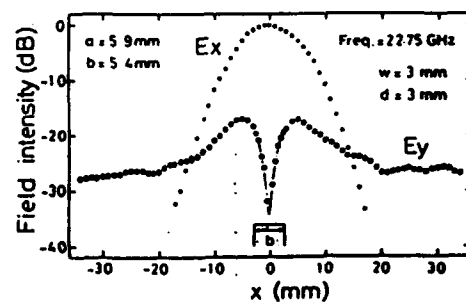


Fig.5 Field distributions in the cross sectional plane of leaky wave NRD-guide having notches

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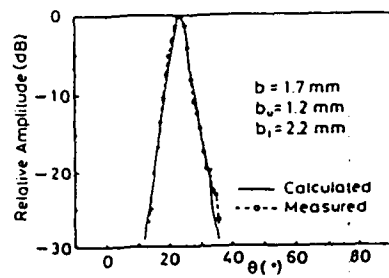
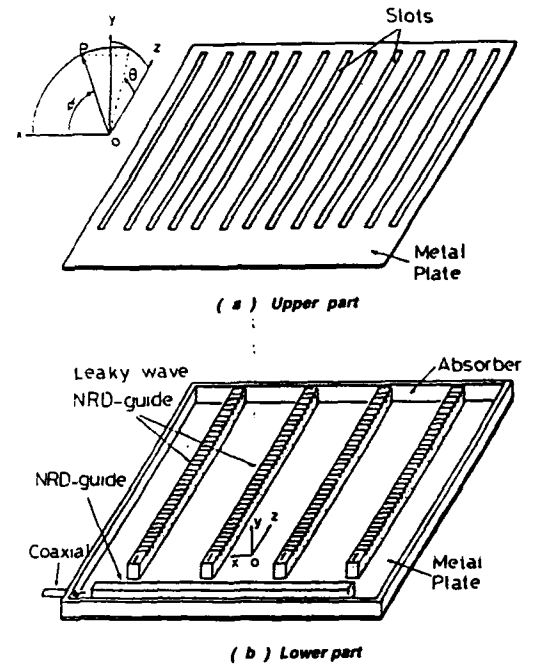
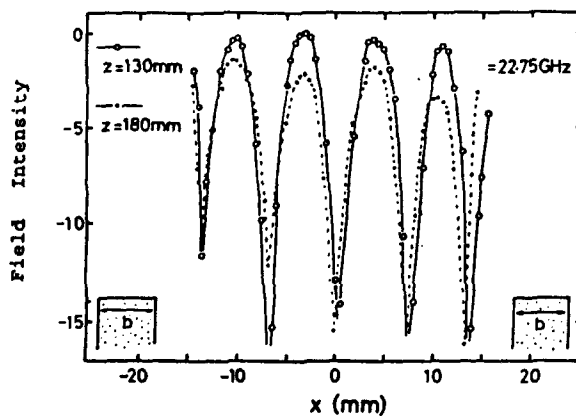


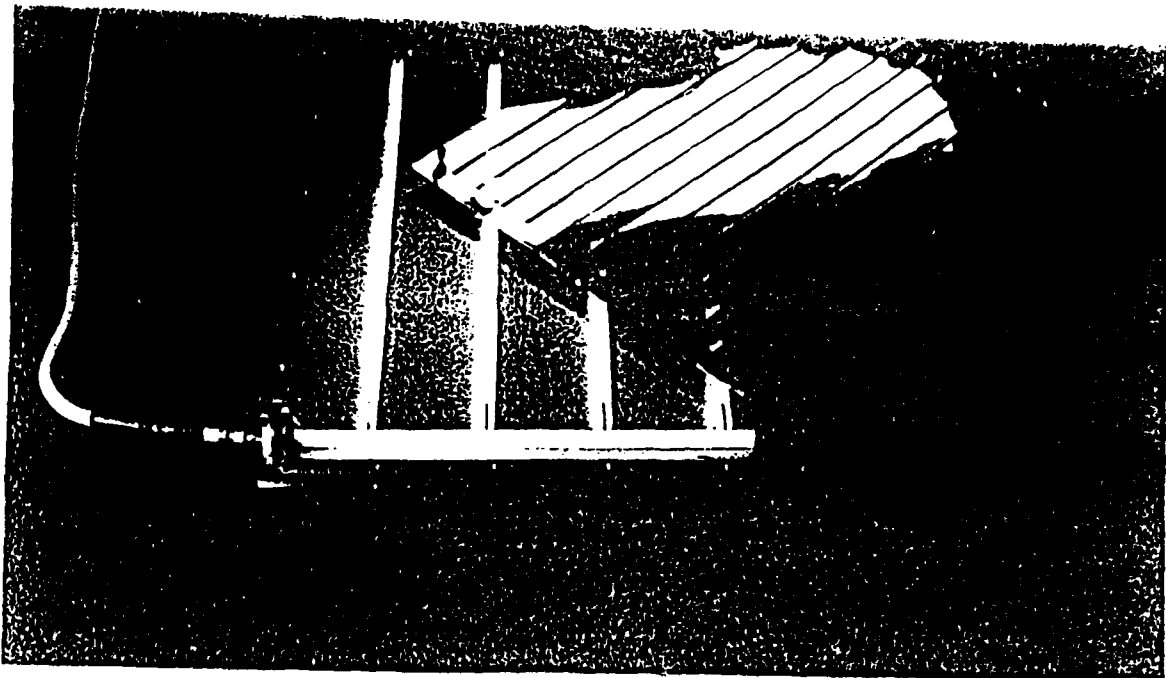
Fig.3 Radiation pattern of leaky wave NRD-guide having trapezoid dielectric strip



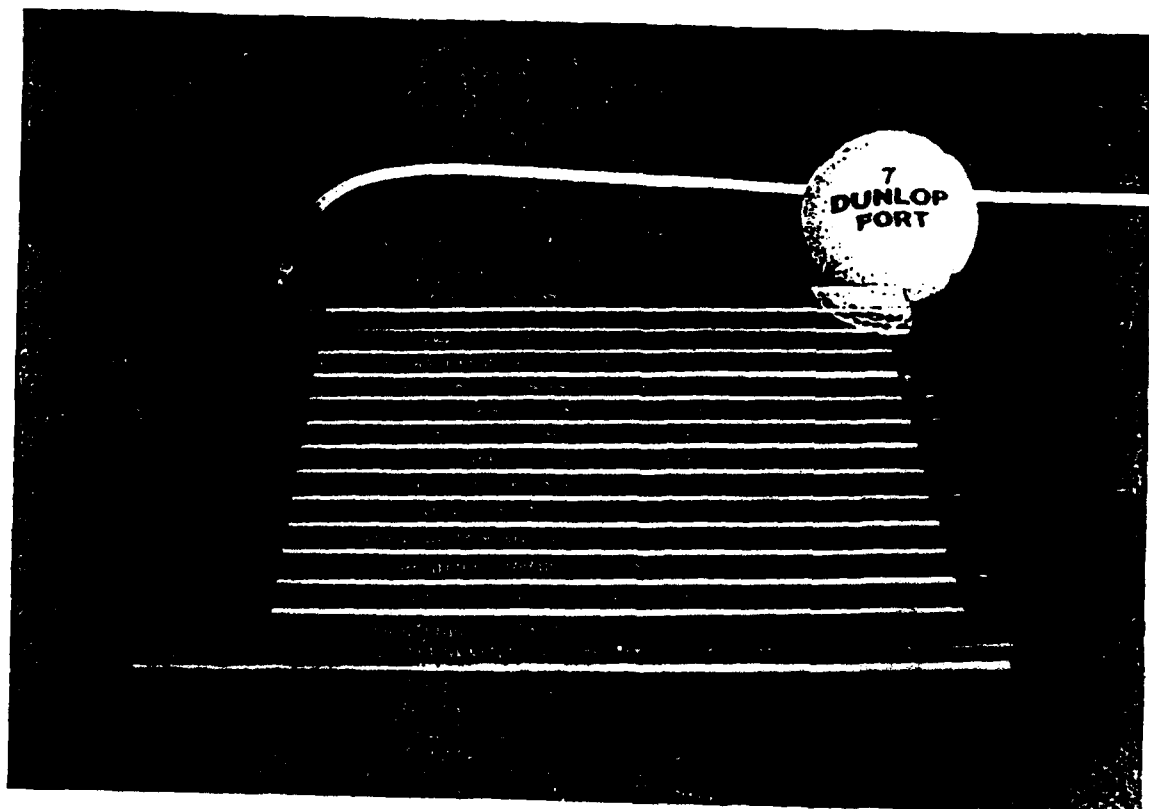
Structure of leaky wave NRD-guide fed planar antenna



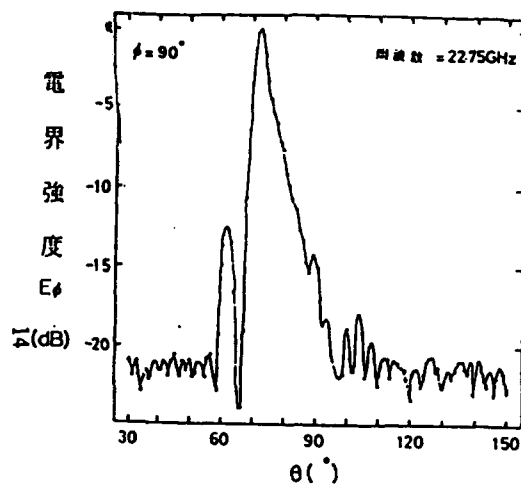
Distribution of vertical component of E-field in the planar antenna



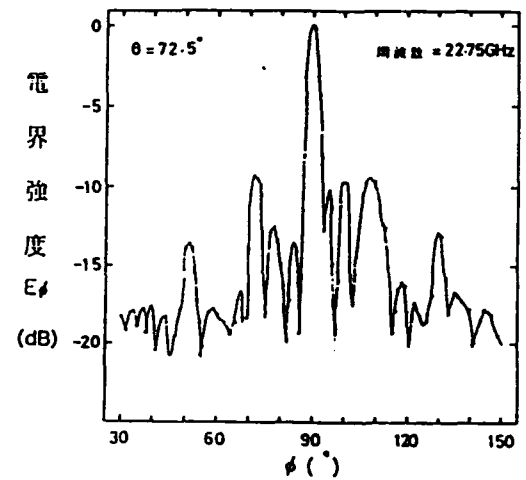
12



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(a)  $\theta$  方向特性



(b)  $\phi$  方向特性

図7 漏れ波NRDガイド給電平面アンテナの放射指向性

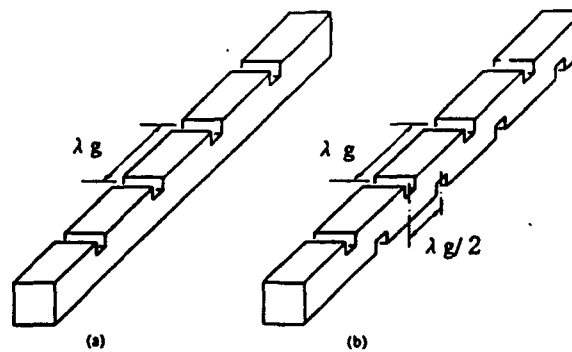


Fig.6 Structure of broadside leaky wave NRD-guide

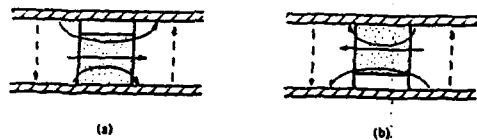
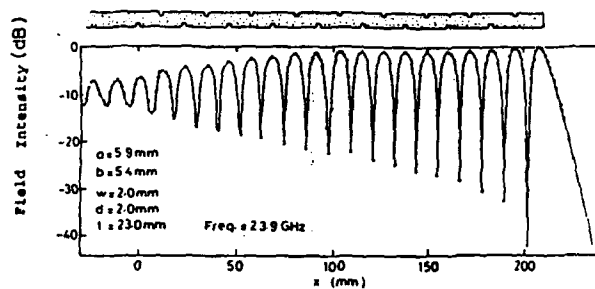


Fig.7 Field distributions of broadside leaky wave NRD- guide at notches separated by half a wavelength (solid line : guided wave, dotted line : radiated wave)



Standing wave pattern along broadside leaky wave NRD- guide

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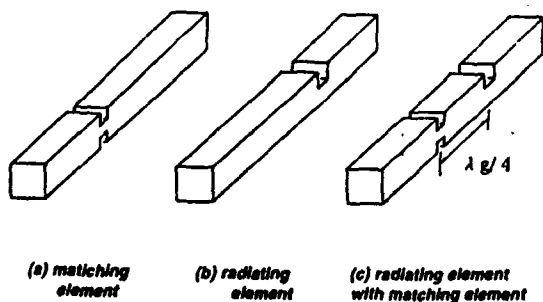


Fig.9 Matching element and radiating element

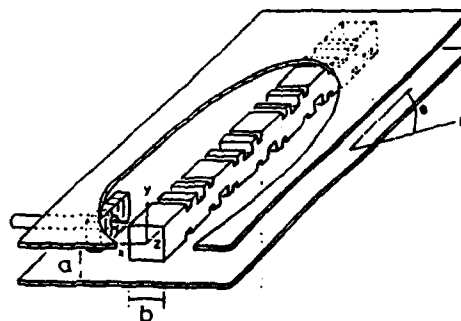


Fig.10 Structure of broadside leaky wave NRD-guide

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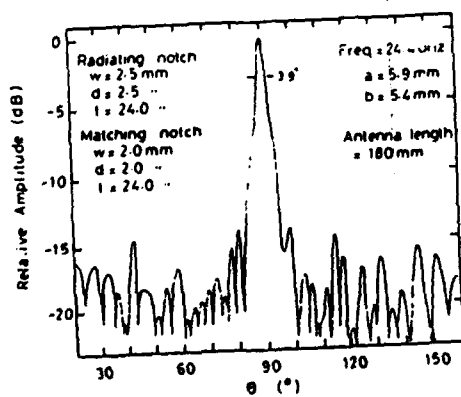
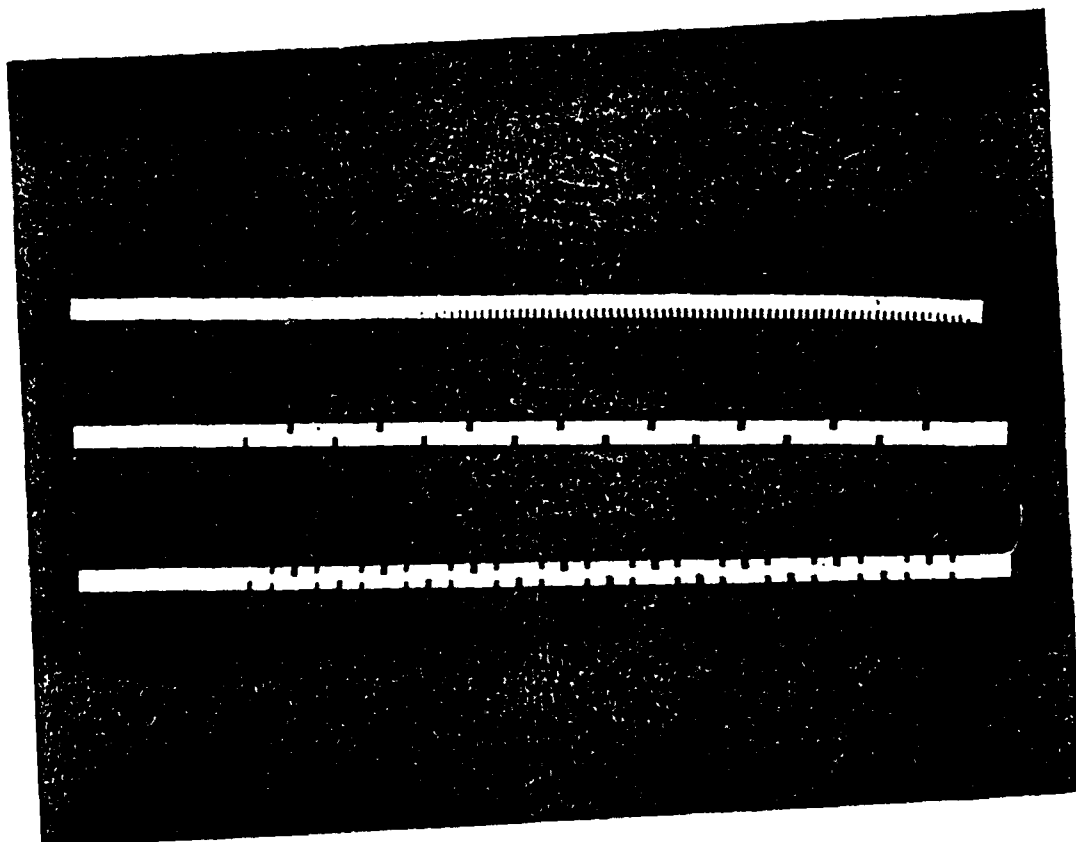


Fig.10 Radiation pattern of broadside leaky wave NRD-guide

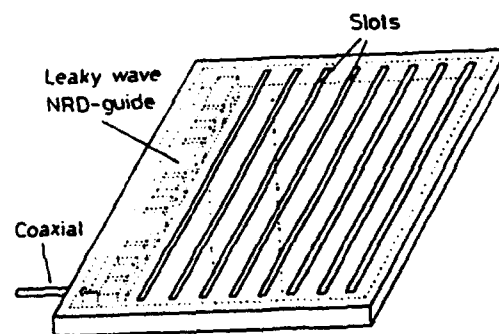
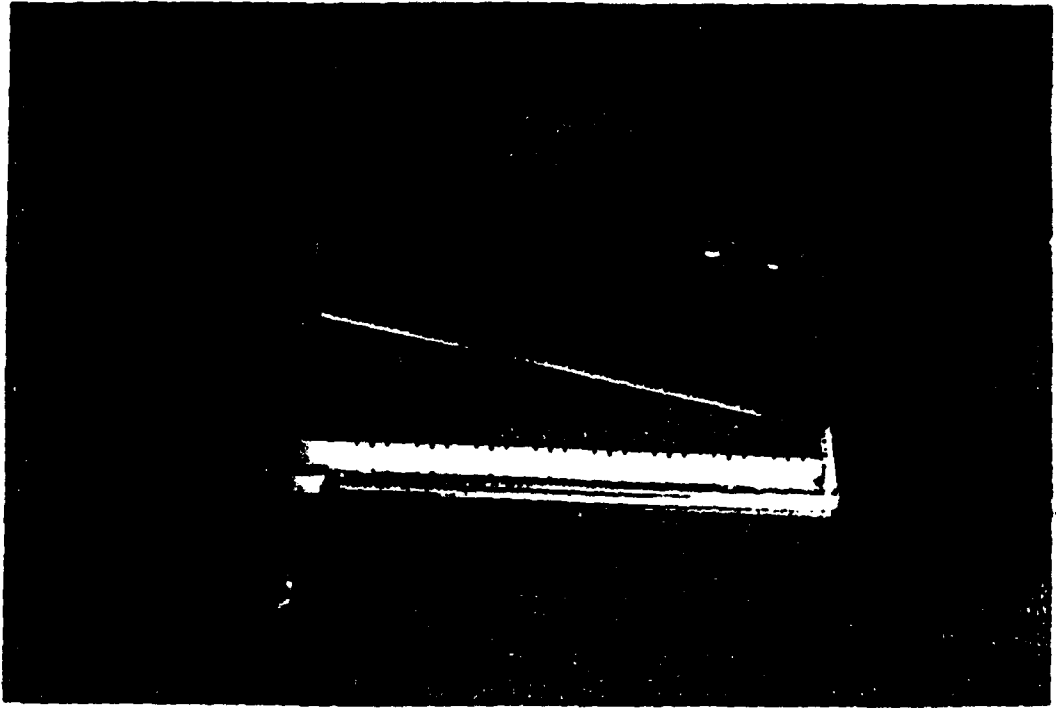
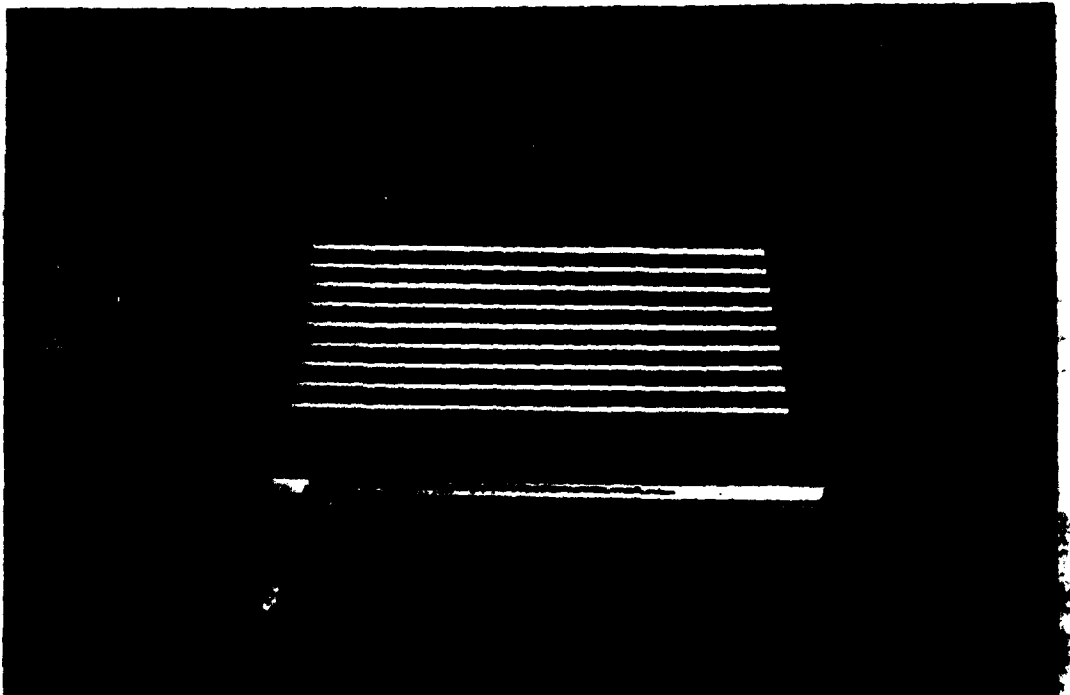


Fig.11 Planar antenna fed by leaky wave NRD-guide

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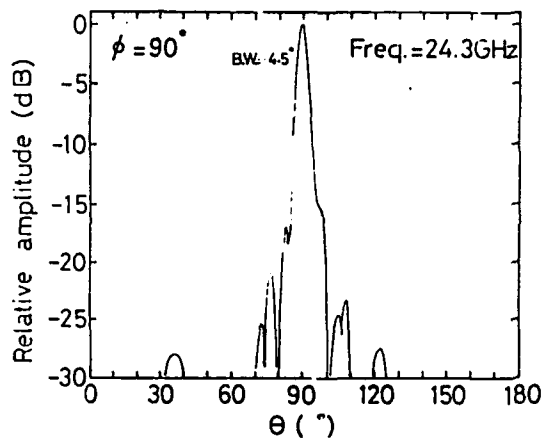


Fig.12-(a) Radiation pattern of broadside leaky wave NRD-guide fed planar antenna (230x150x7 in mm) ( in the vertical plane parallel to the slot )

25

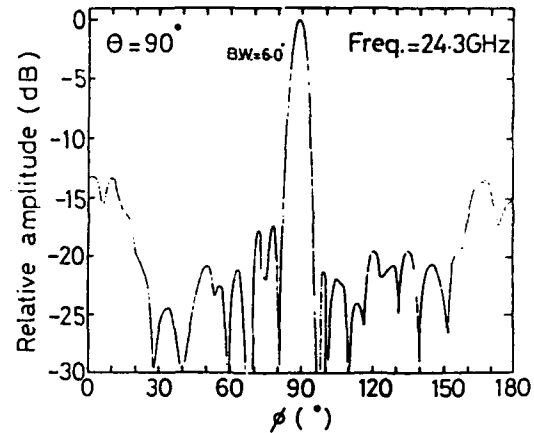
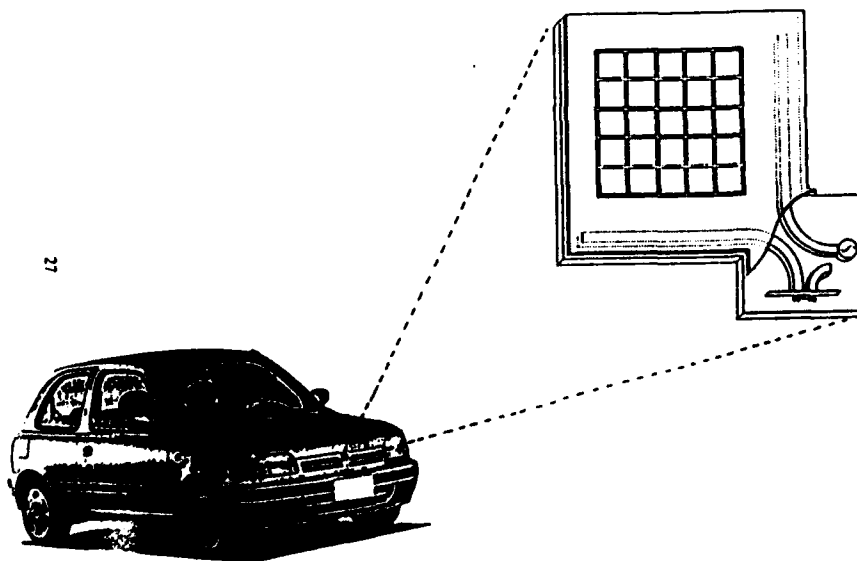


Fig.12-(b) Radiation pattern of broadside leaky wave NRD-guide fed planar antenna (230x150x7 in mm) ( in the vertical plane perpendicular to the slot )

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**ADVANCED DESIGN TECHNIQUES  
FOR LINEAR AND NON LINEAR MMIC'S INTO THE  
MILLIMETER - WAVE REGION**

**R. H. Jansen**

**Jansen Microwave • Germany**

# ADVANCED DESIGN TECHNIQUES FOR LINEAR AND NONLINEAR MMICS INTO THE MM-WAVE REGION

Rolf H. Jansen, Jansen Microwave, Ratingen, Germany

## SUMMARY

In the past few years, the field of high-speed and microwave monolithic integrated circuits (MMICs) has seen an increasing range of nonmilitary applications. Particularly large growth rates and competitive development prevails in the emerging markets of mobile and satellite communications as well as of automotive electronic equipment up to high mm-wave frequencies. Commercial pressure is now driving the demand for faster development and lower cost manufacturing of GaAs MESFET-based and HEMT-based MMICs. From a design point of view, this is reflected in the following trends:

- The requirement for high packing density circuits, i.e. MMICs with minimized chip size, and for multifunction circuits integrating several functional blocks on a single chip while maintaining reasonably small size and thus acceptable yield. This is paralleled by the creation of miniaturized lumped components working well into the mm-wave region.
- The tendency towards lower noise figures and higher frequencies through the gradual transition from MESFET to HEMT technologies and in view of mm-wave applications like collision-avoidance radars. Several MMIC foundry services are now in the stage of moving to higher frequency limits, i.e. from about 20 GHz presently to 40...60 GHz in the near future.
- The demand for better CAD capable of addressing the problems associated with high packing density design and with high mm-wave frequencies in process-related form. The goal here is the reduction of development time and the full exploitation of the degrees of freedom offered by the MMIC medium for linear and nonlinear circuit design.

Under contract with GEC-Marconi Materials Technology, UK, formerly Plessey Research Caswell, the author and his group have worked over a number of years on such as the above outlined issues. Progress achieved is visualized by representative examples. A brief overview is given over advances in CAD techniques directly related to the MMIC process technology. This includes third generation look-up table concepts for single and dual metallization structures as well as 3D electromagnetic full-wave simulation with high speed and for high geometrical complexity. A unified approach

has been made available here for microstrip type MMICs and for coplanar waveguide type (CPW) circuits. For specific, repeatedly used structures like via holes, approximate field-theory based models have been developed for fast interactive CAD. Further, linear and nonlinear distributed FET/HEMT modelling in terms of device layout geometry is described, including automated parameter extraction. CAD based on this approach has near interactive speed and has primarily been developed for frequency domain design involving larger devices and for analogue MMICs operating at high mm-wave frequencies.

## REFERENCES related to this workshop contribution

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- /2/ Jansen, R.H.: Recent advances in the full-wave analysis of transmission lines for application in MIC and MMIC design, Invited Paper, Proc. SBMO Intern. Microwave Symposium, Rio de Janeiro, July 1987, 467-475.
- /3/ Jansen, R.H. and Wertgen, W.: A 3D field-theoretical simulation tool for the CAD of mm-wave MMICs, Alta Frequenza, Italy, vol. LVII, 1987, 203-216.
- /4/ Jansen, R.H. et al: A comprehensive CAD approach to the design of monolithic microwave integrated circuits, Invited Paper, IEEE Trans., MTT-36, (1988), 208-219.
- /5/ Wiemer, L. and Jansen, R.H.: Reciprocity-related definition of strip characteristic impedances for multiconductor hybrid-mode transmission lines. Microwave and Optical Technology Letters, Wiley Interscience, vol. 1, (1988), 22-25.
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**CAD OF MILLIMETER - WAVE PASSIVE COMPONENTS  
IN SUSPENDED MICROSTRIP TECHNOLOGY**

**V. Fouad Hanna**

**CNET • France**

WHY?

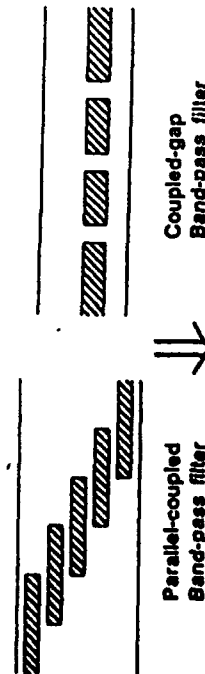
|                       | Stripline<br>(microstrip) | Suspended Stripline<br>(suspended microstrip) |
|-----------------------|---------------------------|-----------------------------------------------|
| Attenuation<br>losses | High                      | 75% Less than<br>microstrip                   |
| Dispersion            | Moderate                  | Little                                        |
| CAD<br>Facilities     | Rich                      | Poor                                          |

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WHY?



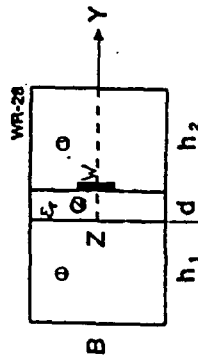
An Asymmetric Suspended Stripline

Fig. 3

## PLAN

- \* INTRODUCTION.
- \* EVALUATION OF EIGENMODES IN SUSPENDED STRIPLINE (SSL) TECHNOLOGY.
- \* CHARACTERISATION OF UNIAXIAL SSL DISCONTINUITIES.
- \* REALISATION OF PASSIVE MICROVE COMPONENTS IN SSL TECHNOLOGY : Directional Coupler-Resonator-Filter.
- \* CHARACTERISATION OF MULTIAXIAL SSL DISCONTINUITIES : Bend , Stub , Bent-Stub.

## SPECTRAL DOMAIN APPROACH for a SSL



$$E_z(x, y, z, t) = \sum_{\alpha_n} E_z(\alpha_n, y) \cdot e^{j\alpha_n x} \cdot e^{j(\omega t - \beta z)} \quad (1)$$

where  $\alpha_n = n \frac{\pi}{B}$

$$\begin{bmatrix} E_x(\alpha_n, d) \\ E_z(\alpha_n, d) \end{bmatrix} = \begin{bmatrix} G_{11}(\alpha_n, \beta, \omega) & G_{12}(\alpha_n, \beta, \omega) \\ G_{21}(\alpha_n, \beta, \omega) & G_{22}(\alpha_n, \beta, \omega) \end{bmatrix} \begin{bmatrix} J_x(\alpha_n, d) \\ J_z(\alpha_n, d) \end{bmatrix} \quad (2)$$

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CAD OF MILLIMETER-WAVE PASSIVE COMPONENTS  
IN SUSPENDED MICROSTRIP TECHNOLOGY

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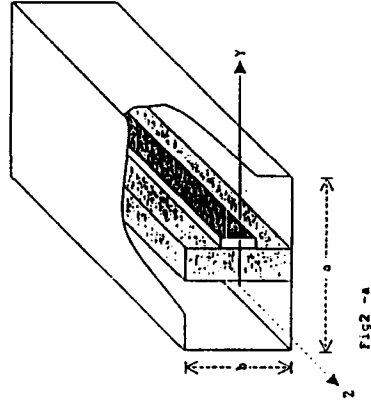


Fig 2-a

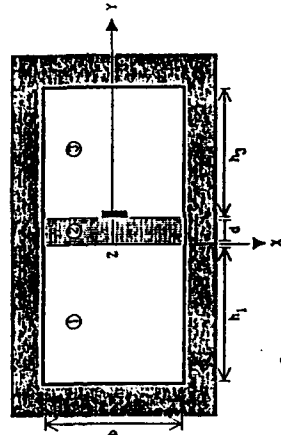


Fig 2-b

SUSPENDED AND INVERTED MICROSTRIP LINES

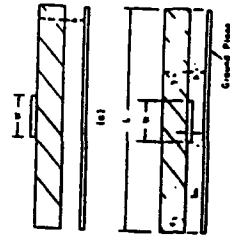


Fig 1

Longitudinal (1-a) and cross-sectional (1-b) view of suspended stripline.

- Reduced attenuation (detached ground plane)
- Millimeter applications --- Shielding
- Field configurations TEM (closer than microstrip)
- Circuit layout on both sides of the substrate.

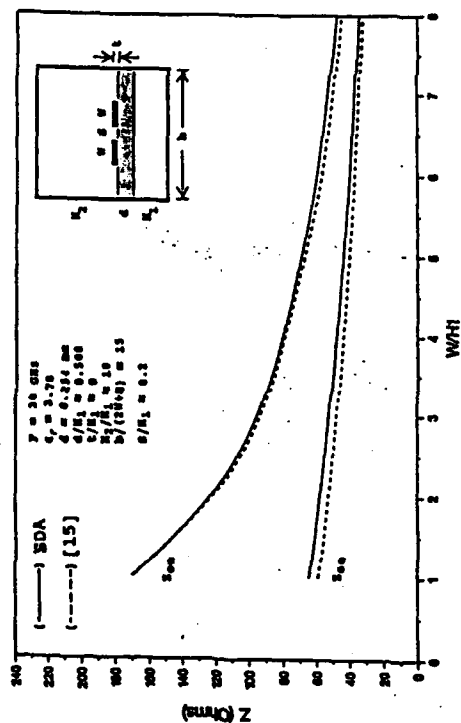


Fig 6 Variation of the characteristic impedance for the even and odd modes as a function of  $W/H$  for coupled SSL's: (—) our analysis, (---) reference [15].

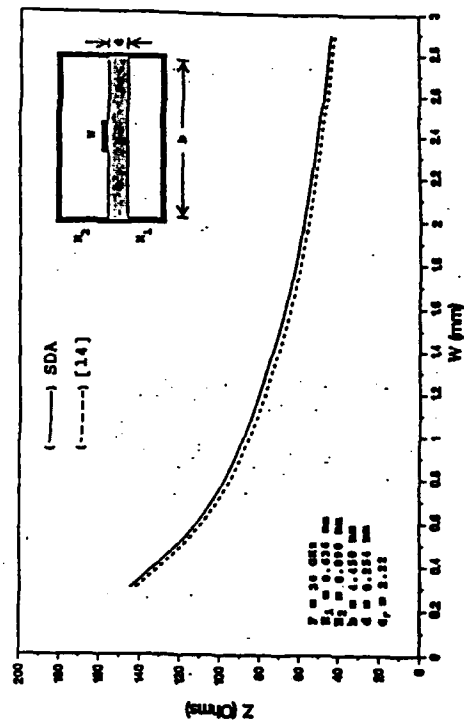


Fig 7 Variation of the characteristic impedance as a function of  $W$  for an SSL: (—) our analysis, (---) reference [14].

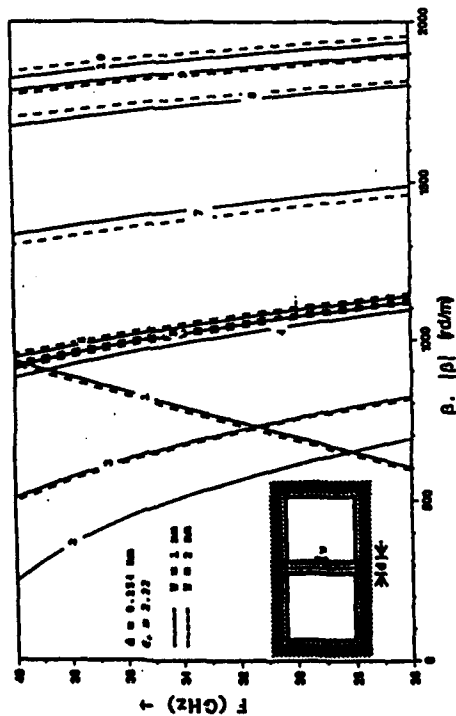


Fig 4 Dispersion characteristics of symmetric SSL's ( $\epsilon_r = 2.22$ ).

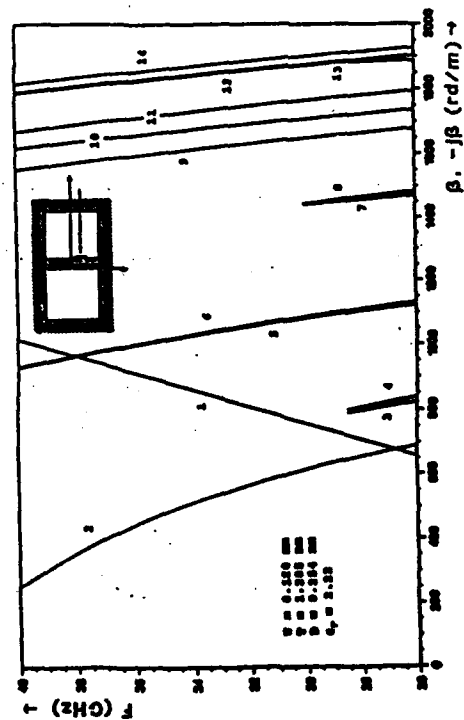


Fig 5 Dispersion characteristics of asymmetric SSL ( $\epsilon_r = 2.22$ ).

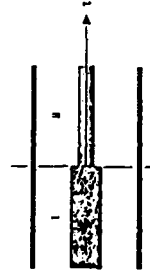
# Method of GALERKIN

$$\begin{aligned}
 J_s(x, d) &= \sum_{n=1}^N a_n \cdot J_{s,n}(x, d) \Rightarrow J_s(a, d) = \sum_{n=1}^N a_n \cdot J_{s,n}(a, d) \\
 J_s(x, d) &= \sum_{n=1}^N b_n \cdot J_{s,n}(x, d) \Rightarrow J_s(a, d) = \sum_{n=1}^N b_n \cdot J_{s,n}(a, d)
 \end{aligned}
 \tag{3}$$

## Relationship of Complementarity

$$\begin{aligned}
 \int_{-\infty}^{\infty} E_s(x, d) \cdot J_{s,n}'(x, d) \cdot dx &= 0 \Rightarrow (R+S) \text{ homogeneous linear equations} \\
 \int_{-\infty}^{\infty} E_s(x, d) \cdot J_{s,n}'(x, d) \cdot dx &= 0 \Downarrow \\
 \beta^2 > 0 &\leftarrow \text{Det}(\beta, \text{ geometry}) = 0 \rightarrow \beta^2 < 0 \\
 \text{propagating modes} &\qquad \qquad \qquad \text{evanescent modes}
 \end{aligned}
 \tag{5}$$

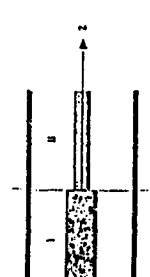
|          | $T(x)$              | $J(a_s)$                                            |
|----------|---------------------|-----------------------------------------------------|
| $J_{21}$ | <div> </div>        | $\frac{2}{a_s b} \sin a_s \frac{b}{2}$              |
| $J_{22}$ | <div> </div>        | $\frac{a_s b}{2b} J_2\left(a_s \frac{b}{2}\right)$  |
| $J_{21}$ | <div> </div>        | $-\frac{a_s b}{2b} J_2\left(a_s \frac{b}{2}\right)$ |
| $J_{22}$ | <div> </div>        | $-\frac{a_s b}{2b} J_2\left(a_s \frac{b}{2}\right)$ |
| $T-1.1$  | $a_s = (2n-1)\pi/b$ |                                                     |

$$\left\{ \begin{array}{l} \vec{E}_T' = \sum_{p=1}^P (A_p' + B_p') \cdot \vec{e}_{p,T} \\ \vec{H}_T' = \sum_{p=1}^P (A_p' - B_p') \cdot \vec{h}_{p,T} \end{array} \right\} \quad (6)$$


$$\left\{ \begin{array}{l} \vec{E}_T'' = \sum_{q=1}^Q (A_q'' + B_q'') \cdot \vec{e}_{q,T} \\ \vec{H}_T'' = \sum_{q=1}^Q (B_q'' - A_q'') \cdot \vec{h}_{q,T} \end{array} \right\} \quad (7)$$

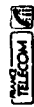
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Discontinuity Boundary Conditions

$$\left. \begin{array}{l} \text{on } \vec{E}_T' = 0 \\ S_1 \quad \vec{H}_T'' + \vec{J}_T' \times \vec{u} = 0 \end{array} \right\} \quad (8)$$


$$\left. \begin{array}{l} \text{on } S_{11} - S_1 \\ \vec{E}_T' = \vec{E}_T'' \\ \vec{H}_T' = \vec{H}_T'' \end{array} \right\} \quad (9)$$

$\Rightarrow$  Set of  $2(P+Q)$  linear equations  $\Rightarrow 2(P+Q)$  Modal Amplitudes

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# Generalised Discontinuity Scattering Matrix

$$[B] = [S] \cdot [A] \quad (10)$$

$$[B] = \begin{bmatrix} [B_P'] \\ [B_Q''] \end{bmatrix} \quad [A] = \begin{bmatrix} [A_P'] \\ [A_Q''] \end{bmatrix}$$

$$[S] = \begin{bmatrix} [S_{11}] & [S_{12}] \\ [S_{21}] & [S_{22}] \end{bmatrix}$$

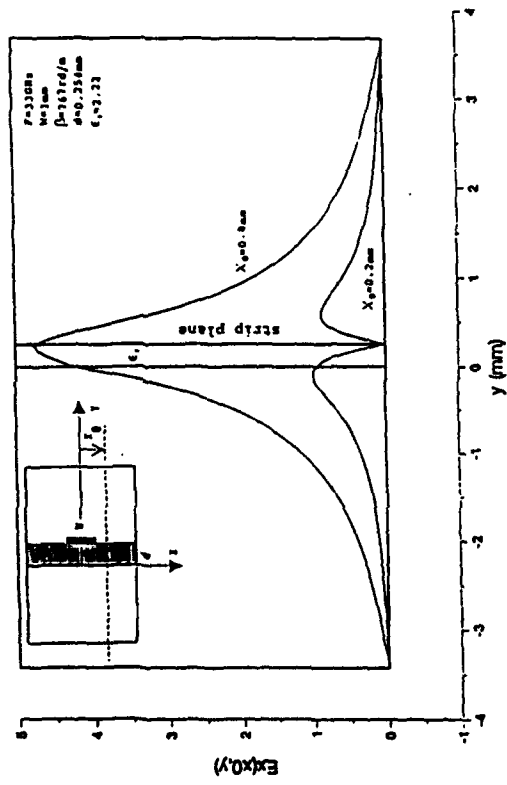


Fig 8 Variation of the x component of the electric field as a function of y (mm) in an SSL.

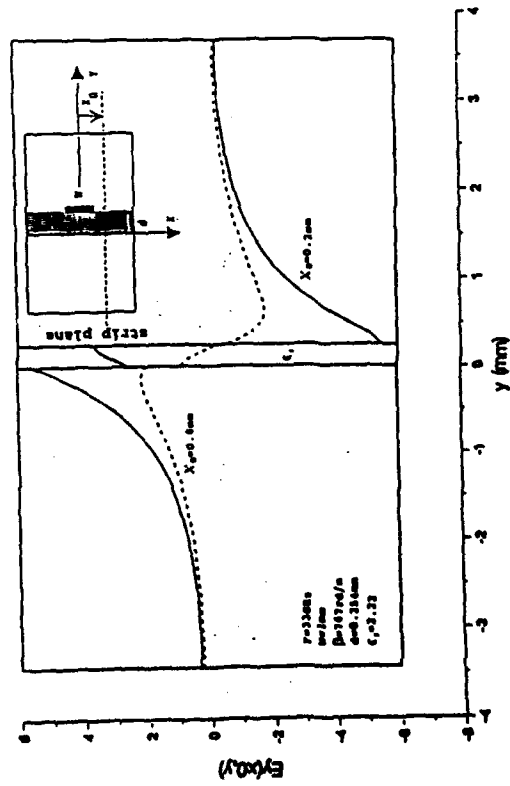


Fig 9 Variation of the y component of the electric field as a function of y (mm) in an SSL.

a step discontinuity in an SSL

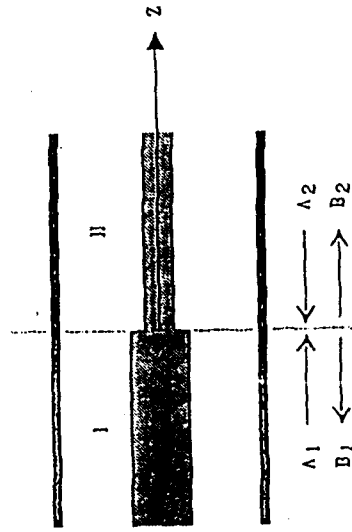
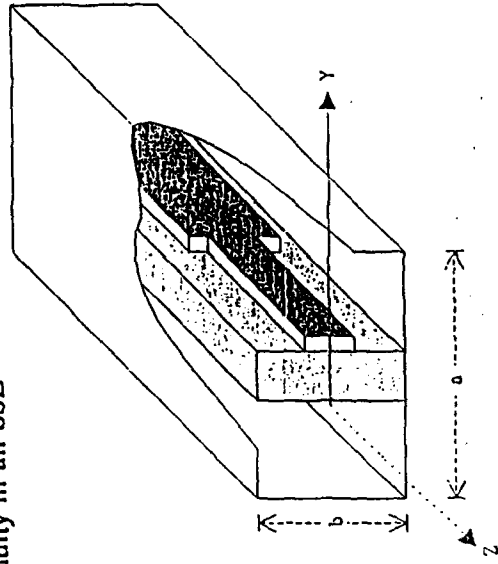


Fig 10

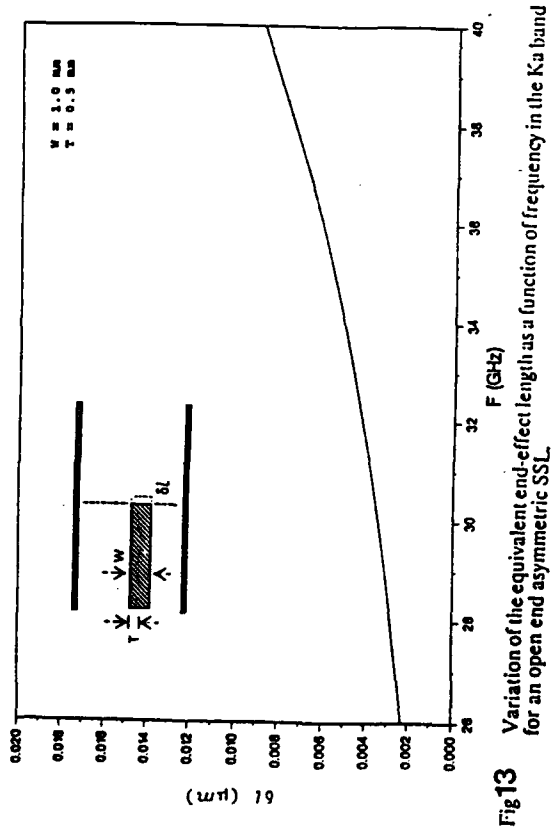


Fig 13 Variation of the equivalent end-effect length as a function of frequency in the Ka band for an open end asymmetric SSL.

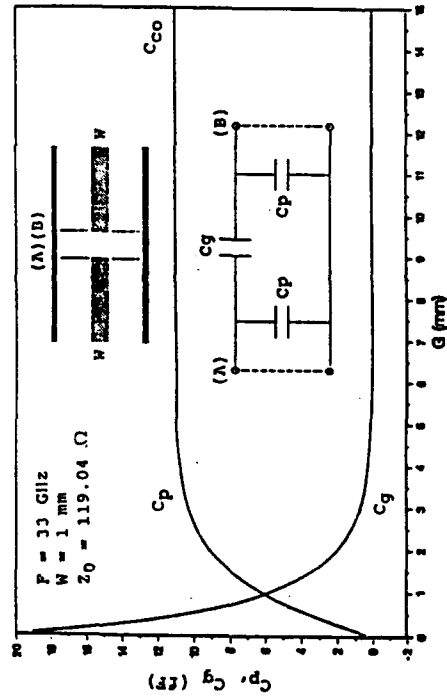


Fig 14 Variation of the equivalent capacitances as a function of the Gap's width in the equivalent circuit for an SSL Gap discontinuity.

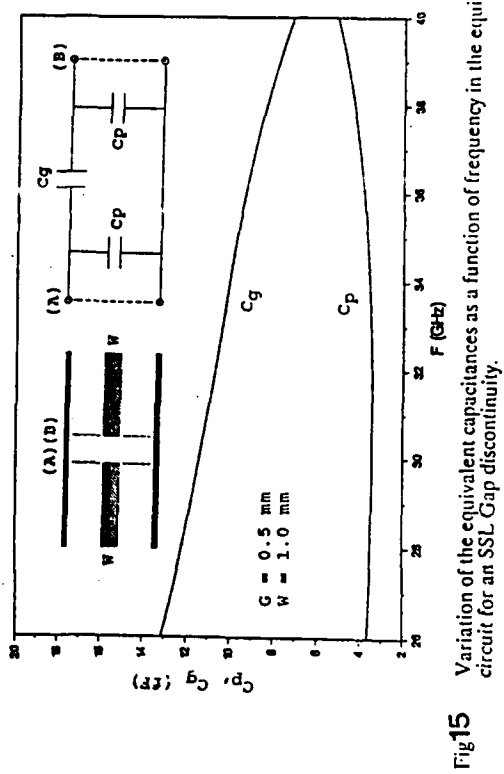


Fig 15 Variation of the equivalent capacitances as a function of frequency in the equivalent circuit for an SSL Gap discontinuity.

Fig 16 Theoretical and experimental variation of the coupling coefficient of the directional coupler as a function of frequency in the Ka band.

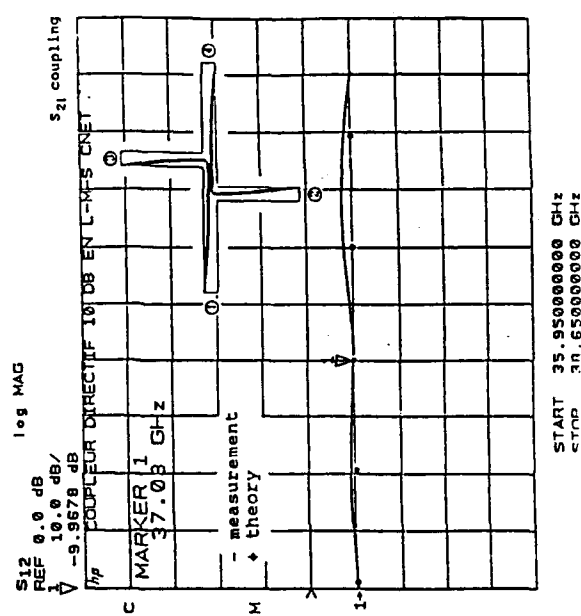


Fig 16 Theoretical and experimental variation of the coupling coefficient of the directional coupler as a function of frequency in the Ka band.

# Role of the Spectral Domain Approach

se constants.

amplitudes of eigenmode field components.

pling coefficients } directly in the  
er flow } spectral domain;

$$P_{p,q} = \frac{1}{2} \oint_S (\vec{E}' \wedge \vec{H}'') \cdot d\vec{S} \quad (11)$$

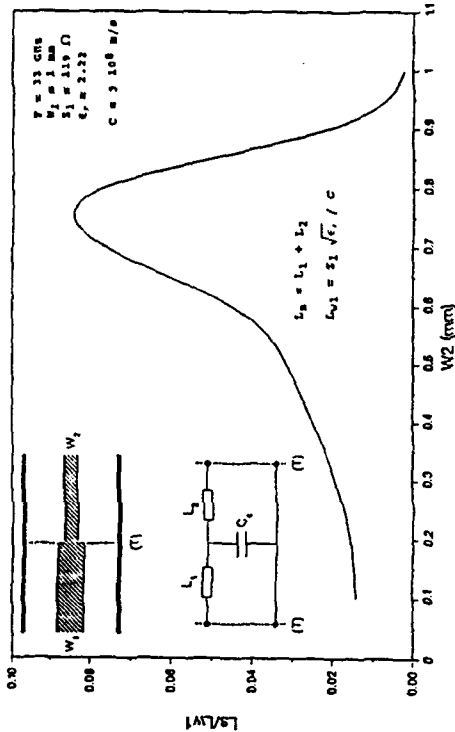


Fig11 Variation of the equivalent normalized inductance as a function of the strip width in the equivalent circuit for an SSL step discontinuity.

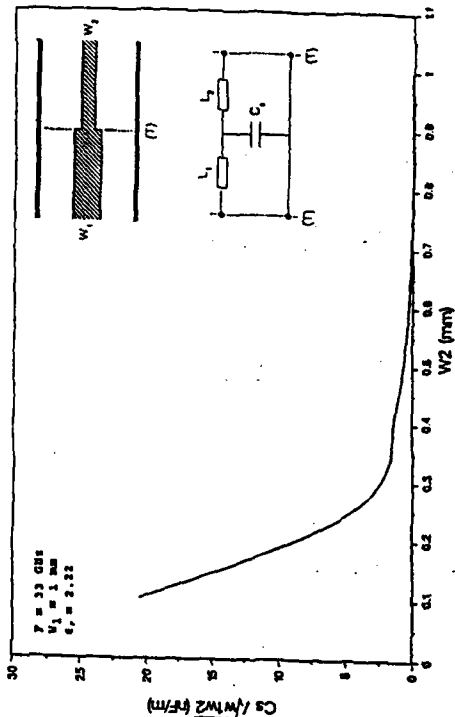


Fig12 Variation of the equivalent normalized capacitance as a function of the strip width in the equivalent circuit for an SSL step discontinuity.

INTEGRAL EQUATIONS TECHNIQUES.

- METHOD OF MOMENT.
- MATCHED LOAD SIMULATION.

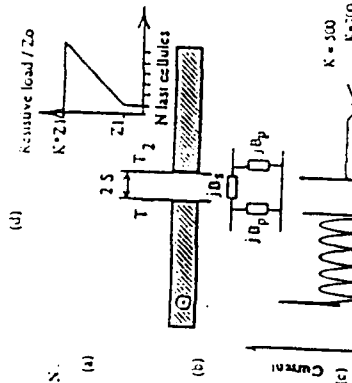


Fig. 2 (a) Top view, (b) equivalent circuit, (c) current distribution (d) A normalized return loss modeling of the N-lattice cells

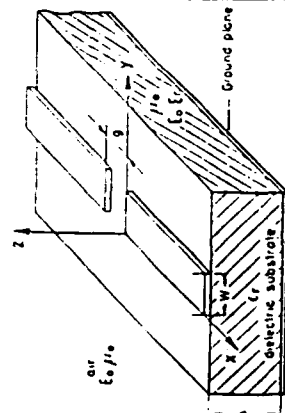
ANALYSIS

$$\vec{e}_x \times (\vec{E}(\vec{r}_1) + \vec{E}^s(\vec{r}_1)) = \vec{e}_z \times [Z_S J_S(\vec{r}_1)] \quad (1)$$

$$\vec{E}^d(\vec{r}_1) = -j\omega A(\vec{r}_1) - \nabla V(\vec{r}_1) \quad (2)$$

$$\vec{A} = \int_{S'} \vec{G}_A \cdot \vec{J}_S dS' \quad (3); \quad V = \int_{S'} G_V \rho_s dS' \quad (4)$$

$$\vec{e}_x \times \left[ \int_{S'} (j\omega \vec{A}_{xy}) \cdot \vec{J}_S + V G_{V,xy} \rho_s \right] dS' + Z_S J_S = \vec{e}_x \times \vec{E}_0^i \quad (5)$$



METALLIC LOSSES  
EXCITATION  
CNET

INTEGRAL EQUATIONS  $\rightarrow$  MATRIX EQUATION

- CONDUCTOR RECTANGULAR CELLS
- CURRENT  $\rightarrow$  OVERLAPPING

$$\vec{J}_S(\vec{r}) = \sum_{n=1}^{N-1} \vec{T}(\vec{r}) \cdot \vec{e}_x \quad (7)$$

$$(\vec{P}_S = \frac{-1}{j\omega}) \quad (\vec{r}_1, \vec{r}_2)$$

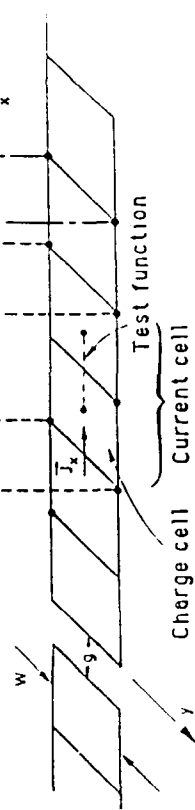


Fig. 22

SIMULATION OF A VIRTUAL MATCHED LOAD

The obtained new matrix equation, for the loaded structure, can be written in the form :

$$([Z] + [\Delta Z])[I] = [V]$$

where  $[Z]$  is the generalized impedance matrix of the unloaded structure,  $[I]$  is the unknown current vector,  $[V]$  is the feeding voltage vector and  $[\Delta Z]$  is a square diagonal impedance matrix

Fig 17 Variation of the resonator reflection coefficient as a function of frequency in the Ka band ( $L_0=4.45\text{mm}$ ).

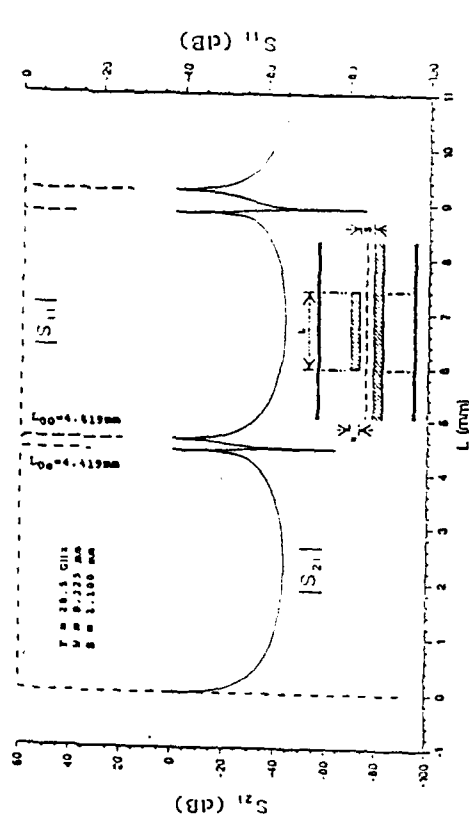
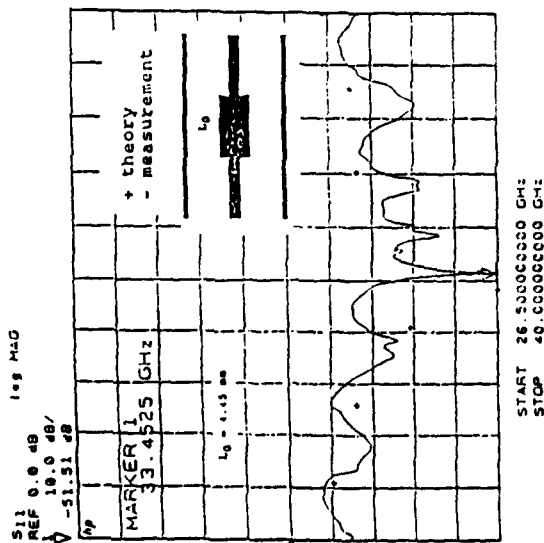


Fig 19 Variation of  $S_{11}$  (dB) and  $S_{21}$  (dB) as a function of the resonator length at  $F = 28.5 \text{ GHz}$ .

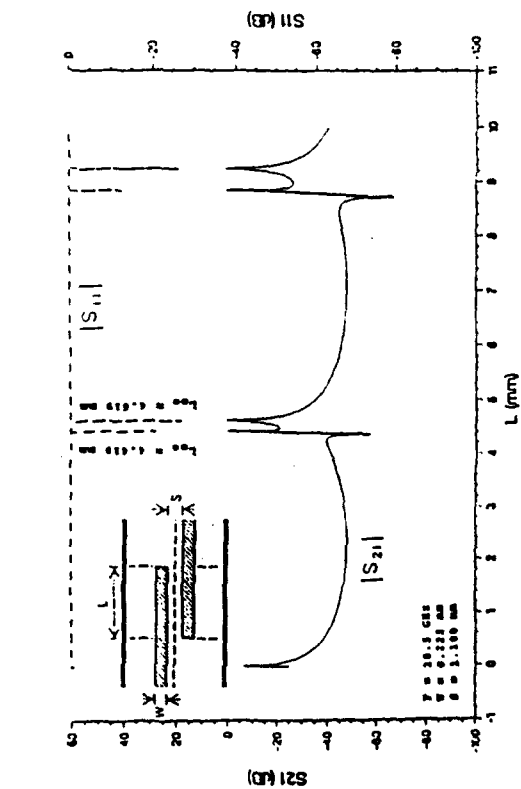


Fig 18 Variation of  $S_{11}$  (dB) and  $S_{21}$  (dB) as a function of the resonator length at  $F = 28.5 \text{ GHz}$ .

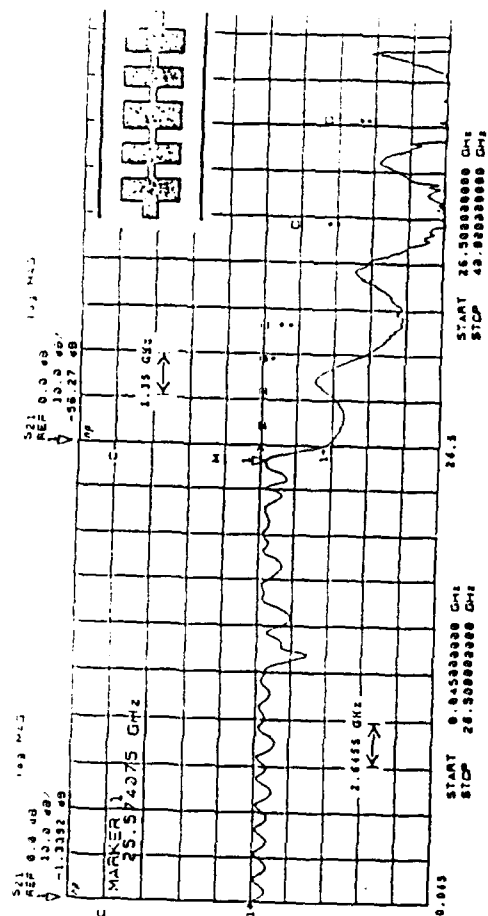


Fig 20 Theory, simulation and measure performances of the low-pass filter realized in SSL technology using RT/Duroid as substrate.

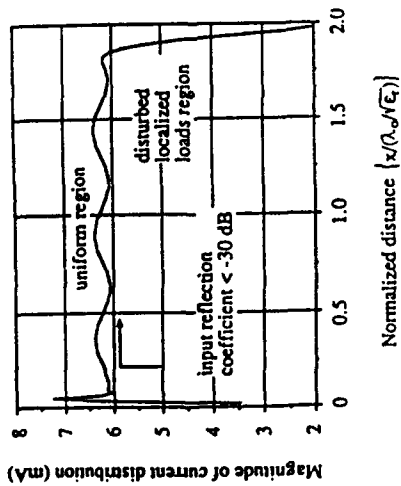


Fig 23 Current distribution along a loaded microstrip line

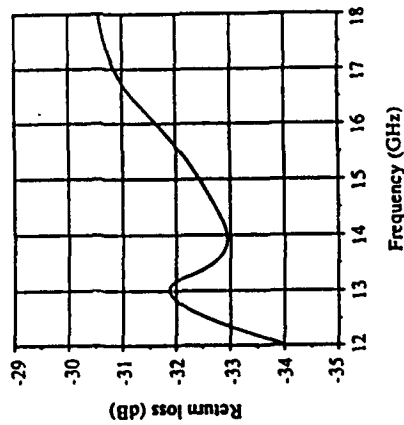


Fig 24 Variation of the return loss as a function of frequency in the Ku-band for a matched load on a microstrip line (a=80 mm, b=7.112 mm, c=3.56 mm, w=0.75 mm, h=254 μm, εr=2.22)

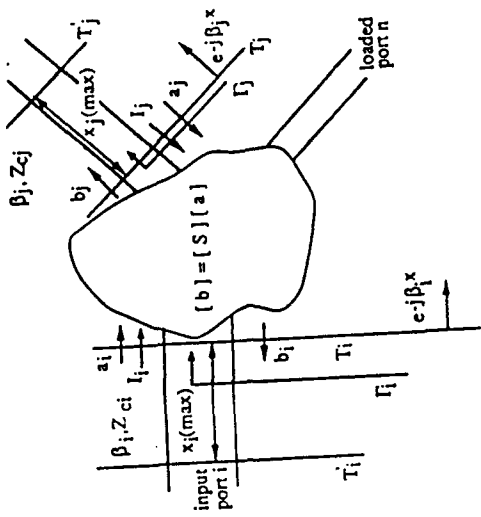


Fig 25: General N-port discontinuity configuration

$$|S_{ji}| = \frac{-\alpha_{ji}}{1 + \alpha_{ji}} + \frac{1}{2} \sqrt{\left( \frac{2\alpha_{ji}}{\alpha_{ji} + 1} \right)^2 - 4 \left( \frac{\alpha_{ji} - 1}{\alpha_{ji} + 1} \right)}$$

$$\alpha_{ji} = \sum_{j=1}^N k_{ji}$$

where

$$k_{ji} = \frac{|I_j(\max)|^2 Z_{cj}}{|I_j(\max)|^2 Z_{ci} (1 + |\Gamma_j|)^2}$$

$$\angle \Gamma_i - 2\beta_j x_j(\max) = (2m + 1)\pi, \angle \Gamma_j + 2\beta_i x_i(\max) = (2m + 1)\pi, m = 0, 1, \dots, \infty$$

$$S_{ji} = \frac{-I_j(\max) \sqrt{Z_{cj}} e^{j\beta_j x_j(\max)} (1 + |\Gamma_j|)}{I_i(\max) \sqrt{Z_{ci}} e^{j\beta_i x_i(\max)} (1 + |\Gamma_i|)}$$

$$\angle S_{ji} = \angle I_j(\max) - \angle I_i(\max) + \beta_j x_j(\max) + \beta_i x_i(\max)$$

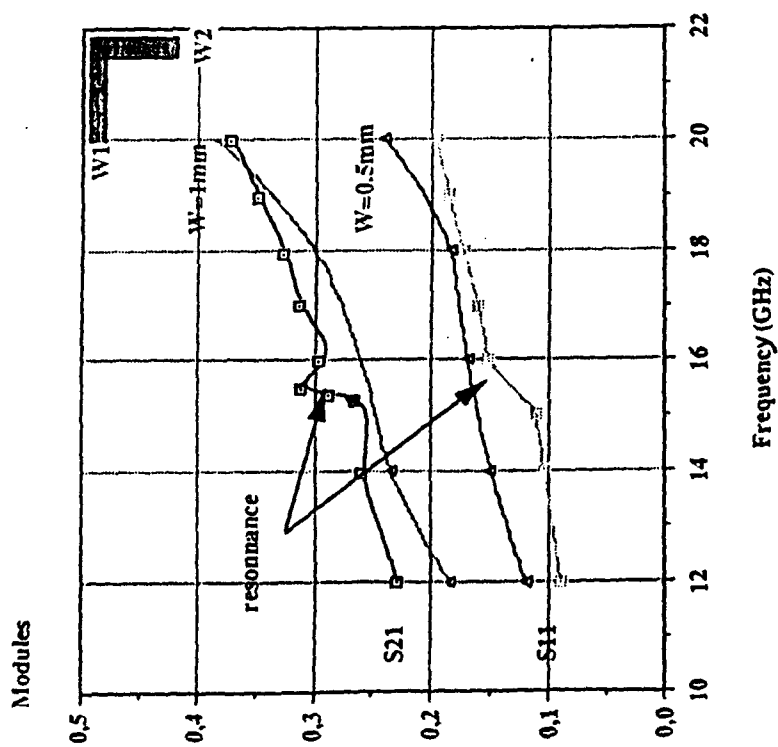


Fig.26 : Variation of S11 and S21 for a SSL bend discontinuity as a function of frequency ( $W1=W2=W$ ,  $a=b=12.7\text{mm}$ ,  $c=3.81\text{mm}$ ,  $\epsilon_r=9.8$ ,  $h=0.635\text{mm}$ ,  $\square$  our analysis,  $\Delta$  reference (28))

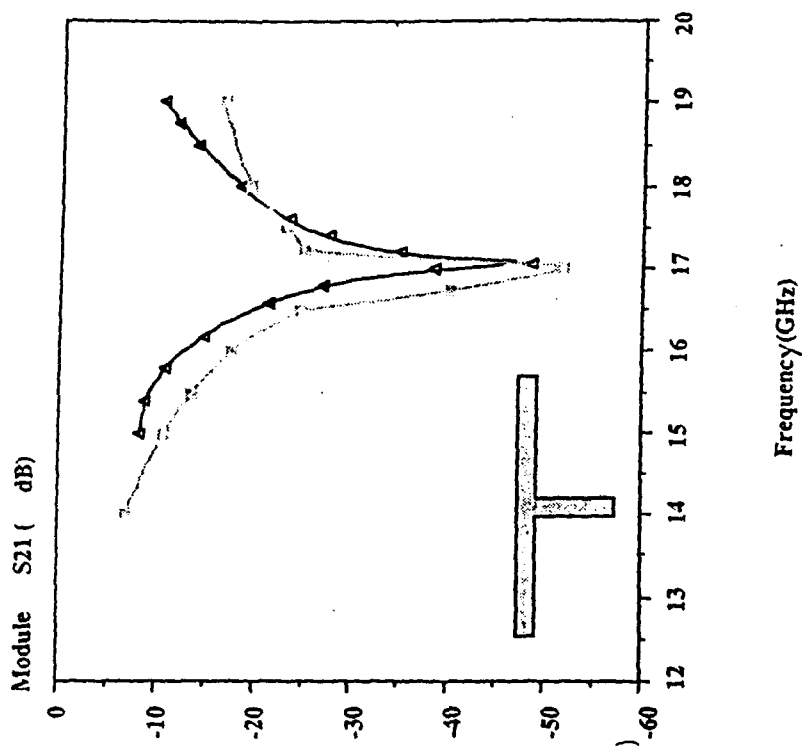


Fig. 27 : Variation of  $|S21|$  ( $\square$  Theory and  $\Delta$  measurements) for a SSL stub discontinuity.

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Module

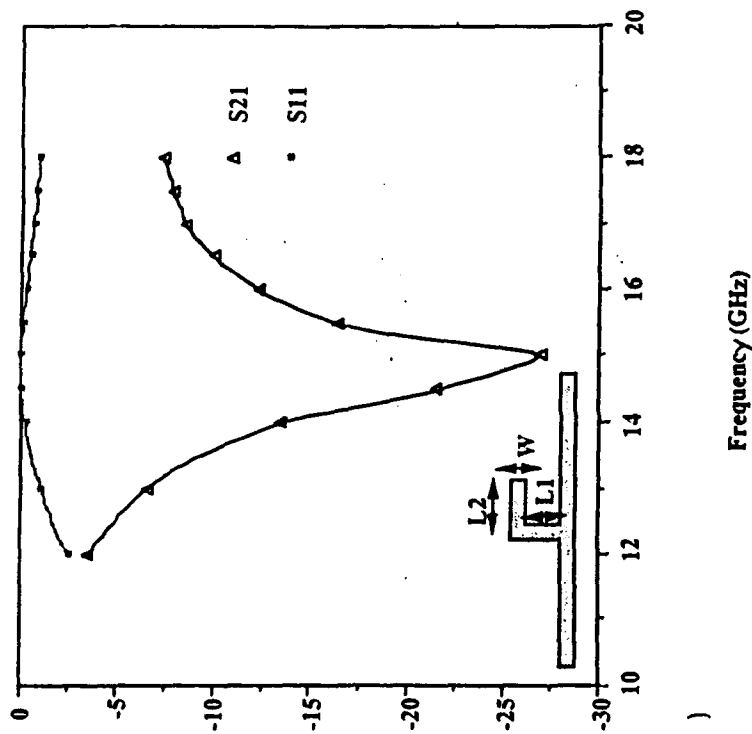


Fig. 28: Variation of S<sub>21</sub>, S<sub>11</sub> (dB) for a bent-stub discontinuity as a function of frequency (L<sub>1</sub>=2.358mm, L<sub>2</sub>=1.572mm, W=0.786mm).

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**LEAKAGE AND CROSSTALK EFFECTS IN MILLIMETER -  
WAVE INTEGRATED CIRCUITS**

**A. A. Oliner**

**Polytechnic University • USA**

## LEAKAGE AND CROSSTALK EFFECTS IN MILLIMETER-WAVE INTEGRATED CIRCUITS

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3rd International Workshop on Millimeter Waves  
Orvieto, Italy  
April 22-24, 1992

become leaky, then describe the physical properties of the leakage, and finally present several examples of such leakage. After that, we examine the effects of crosstalk due to such leakage and also to transmission-line discontinuities, along the lines indicated earlier. Finally we look at package effects, and examine how discontinuities and leakage may each influence package guided modes and participate in resonances of the package as a whole.

The stress in the presentation will be on the physical principles and the performance consequences.

## PRODUCTION

It is evident that crosstalk in microwave and millimeter-wave integrated circuits can seriously disturb and even ruin circuit performance. It is therefore important to understand how crosstalk can be produced and what circuit effects can result from it.

Unwanted coupling between two different portions of a circuit can occur when two lines, for example, are too closely spaced to each other, so that one is in the near field of the other. This type of crosstalk has been well understood for a long time, and it will not be discussed here. We will examine, instead, the crosstalk that results from the excitation of surface waves within the circuit. These surface waves can travel throughout the circuit, bounce around, and cause various effects, ranging from direct interaction to package resonances.

Surface waves can be excited by two principal mechanisms:

- (a) circuit discontinuities, and
- (b) power leakage from the transmission-line dominant mode.

The fact that surface waves are excited at transmission-line discontinuities is of course well known, but what effects these surface waves produce have been examined only infrequently. We will discuss here the conditions under which discontinuities can interact with other portions of an integrated circuit, and how discontinuities can excite package guided modes and package cavity resonances.

The fact that transmission-line dominant modes can leak became known only a few years ago, principally as a result of work by Shigesawa, Tsuji and Oliner. The leakage occurs in the form of a surface wave that propagates away at an angle from the axis of the transmission line. For most printed-circuit transmission lines, the dominant mode becomes leaky only at the higher frequencies, so that leakage effects become significant principally at millimeter wavelengths. The talk will discuss precisely when a dominant mode can become leaky, how the transmission line, when it leaks, can interact with other lines and with discontinuities, and how such leakage can excite package guided modes (whereas the dominant mode cannot interact with package guided modes when it is purely bound).

The present talk will begin with some necessary background material involving various constituent modes and their dispersion properties. The three printed-circuit lines that we consider are microstrip line, slot line and coplanar waveguide. We next discuss the way power can leak from the dominant mode on printed-circuit lines. In that connection, we first establish when a mode can

## OUTLINE OF TALK

### I. BACKGROUND AND REVIEW

#### A. Constituent Basic Modes on Printed-Circuit Lines

#### B. Dispersion Properties

- 1. Modes on printed-circuit lines
- 2. Surface waves

### II. POWER LEAKAGE FROM PRINTED-CIRCUIT LINES

#### A. When Can a Mode Become Leaky?

#### B. Physical Properties of the Leakage

#### C. Examples of Leakage

### III. SOURCES OF CROSSTALK

#### A. Waveguide Discontinuities

#### B. Leakage Effects

### IV. PACKAGE EFFECTS

#### A. Guided Modes and Cavity Modes

#### B. Coupling Between Leaky Modes and Guided Modes

#### C. Excitation by Discontinuities

## CONSTITUENT BASIC MODES

### 1. MODES ON PRINTED - CIRCUIT LINES

Microstrip Line  
Slot Line  
Coplanar Waveguide (CPW)

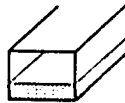
### 2. SURFACE WAVES ON THE SURROUNDING SUBSTRATE

(They are different when the line does or does not have a ground plane.)

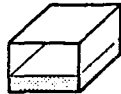


### 3. MODES OF THE PACKAGE

Guided Modes and Cavity Modes



waveguide



resonant cavity

## DISPERSION PROPERTIES

$$\text{wavenumber} = \beta - j\alpha$$

$\beta$  = phase constant

$\alpha$  = attenuation constant

$\alpha$  is due to material losses, and to leakage losses, where appropriate.

If only leakage is included, then  $\alpha$  is called the leakage constant.

Dispersion means the variation of wavenumber with frequency.

Customary to plot as  $\beta/k_0$  vs. frequency (or in normalized form, such as  $h/\lambda_0$ ).

$$k_0 = 2\pi/\lambda_0 = \text{free-space wavenumber}$$

$$\frac{\beta}{k_0} = \frac{\lambda_0}{\lambda_g} = \frac{c}{v_p}$$

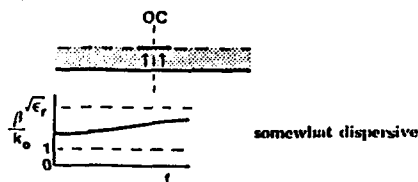
$c$  = velocity of light in free space

$v_p$  = phase velocity of the mode

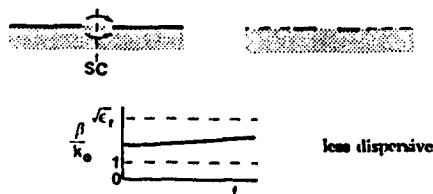
$$\epsilon_{eff} = \left(\frac{\beta}{k_0}\right)^2; \quad n_{eff} = \frac{\beta}{k_0}$$

## DISPERSION BEHAVIOR OF MODES ON PRINTED - CIRCUIT LINES

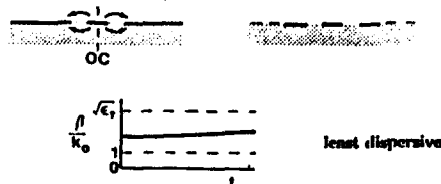
### 1. Microstrip Line



### 2. Slot Line

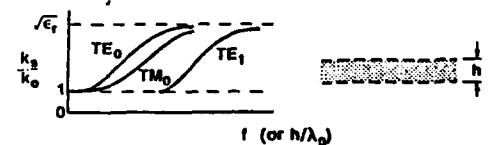


### 3. Coplanar Waveguide (CPW)



## DISPERSION BEHAVIOR OF SURFACE WAVES

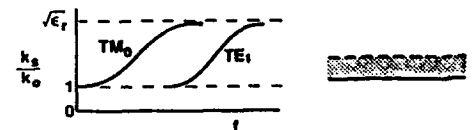
### 1. Ungrounded Layer



$k_z = \beta$  of the surface wave

Strongly dispersive at higher frequencies

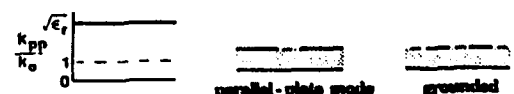
### 2. Grounded Layer



## EXAMPLES

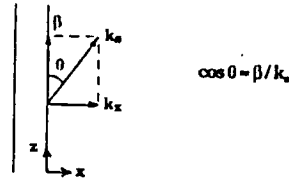


If conductor - backbit



## WHEN CAN A DOMINANT MODE BECOME LEAKY?

Suppose there is leakage:



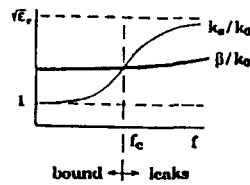
Therefore, for real power to leak away we must have

$$\beta < k_0$$

or

$$\beta / k_0 < k_0 / k_0$$

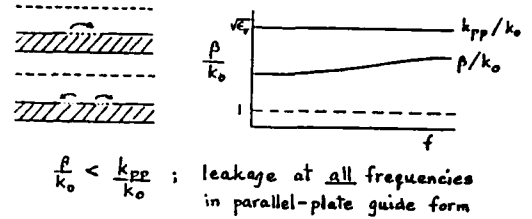
Example: CPW of finite width



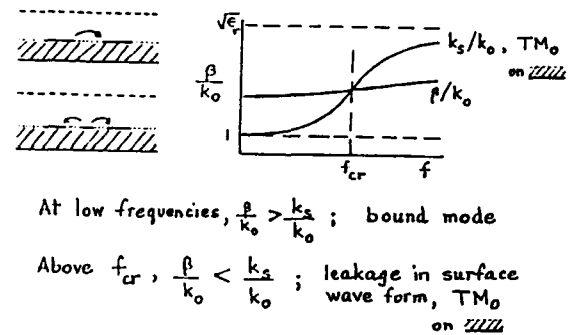
For  $f > f_c$ , the dominant mode on the CPW leaks.

## DISPERSION PLOTS FOR VARIOUS CASES

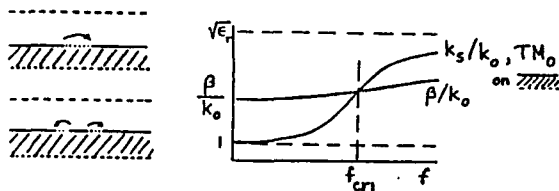
### 1. Conductor-backed slot line or CPW of infinite width



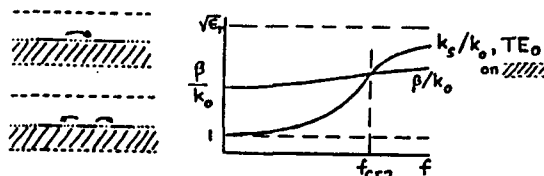
### 2. Conductor-backed slot line or CPW of finite width



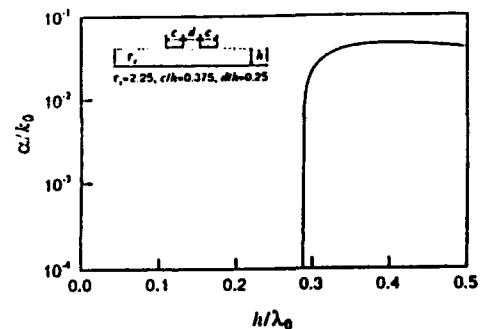
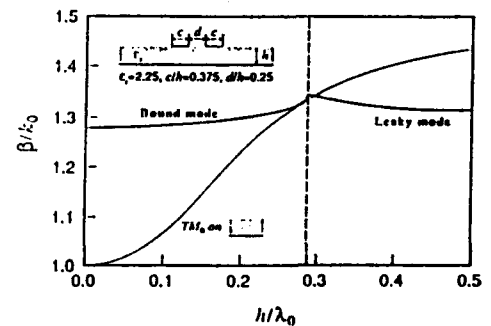
### 3. Conventional slot line or CPW of infinite width



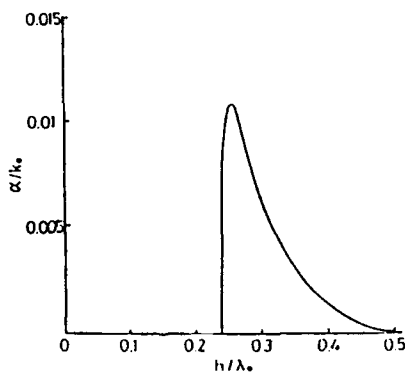
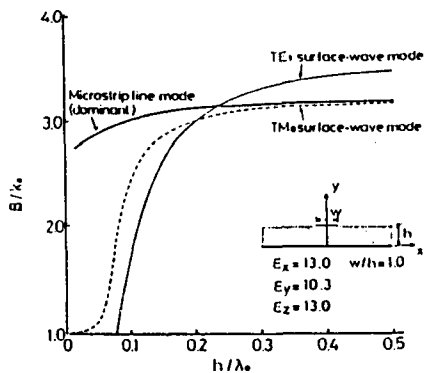
### 4. Conventional slot line or CPW of finite width



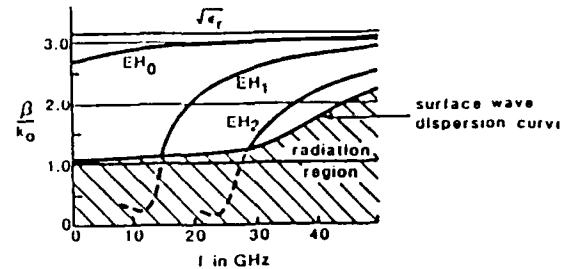
## NORMALIZED PHASE AND LEAKAGE CONSTANTS OF CONDUCTOR-BACKED SLOT LINE



# NORMALIZED PHASE AND LEAKAGE CONSTANTS FOR MICROSTRIP LINE ON AN EPSILAM 10 SUBSTRATE



## LEAKAGE FROM HIGHER MODES ON MICROSTRIP LINE



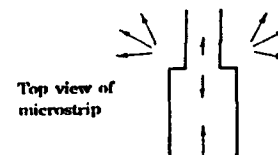
The solutions are leaky where the curves are shown dashed. In that region just above cutoff, the surface wave leaks over a wide range of angles, from endfire to almost broadside as the frequency decreases toward cutoff.

If there is no top cover, leakage also occurs into space waves when  $\beta/k_0 < 1$ .

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## WAVEGUIDE DISCONTINUITIES

What occurs physically?



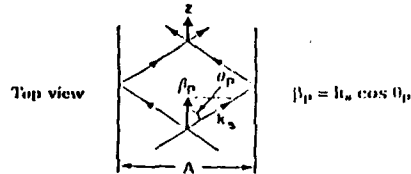
### SOURCES OF CROSSTALK

1. Waveguide Discontinuities
2. Leakage Effects

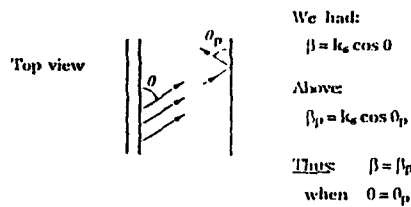
1. Higher modes on the line are excited, but decay away from it, and their power is essentially stored in the neighborhood of the discontinuity. They provide reactance to the equivalent network of the discontinuity.
2. The lowest surface wave (or waves) on the surrounding substrate is excited. It radiates power in surface-wave form over a wide range of angles, and it provides resistance to the equivalent network of the discontinuity.
3. If there are additional dominant modes, there will be some mode conversion into them. This effect contributes to resistance.

## COUPLING BETWEEN 'LEAKY' MODES AND PACKAGE GUIDED MODES

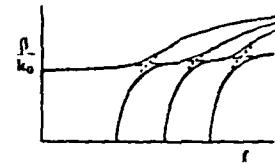
1. We can view the package guided mode in terms of a pair of surface waves at an angle.



2. The package guided mode can be excited by the "leaky" mode when the leakage angle  $\theta$  becomes the same as the package guided mode angle  $\theta_p$ . The two modes then couple to each other.

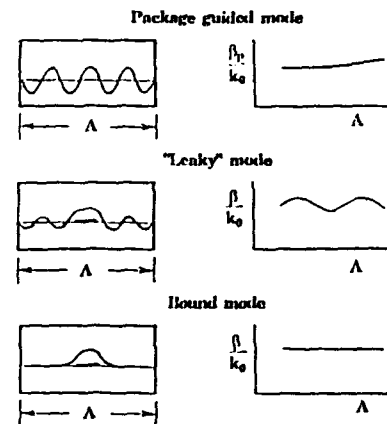


## COUPLING BETWEEN 'LEAKY' MODES AND PACKAGE GUIDED MODES

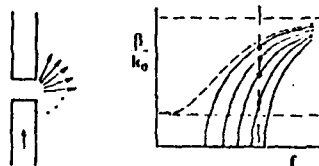


The dashed and dotted lines represent the original (unperturbed) transmission line mode and package guided modes, respectively. The solid lines represent the coupled mode result.

### FURTHER INFORMATION



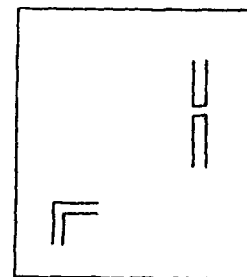
## EXCITATION OF PACKAGE GUIDED MODES BY DISCONTINUITIES



1. The discontinuity excites surface waves at all angles, so that their  $\beta$  values range from 0 to  $k_p$ .
2. Therefore, at a particular frequency, they can couple to all package guided modes with  $\beta$  values in that range, usually a large number of them.
3. In the above example, 4 package guided modes are excited.
4. By contrast, a transmission line mode, with a given  $\beta$  at frequency  $f$ , could at most excite (couple to) a single package guided mode, and perhaps none unless the frequency were shifted.



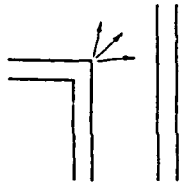
## EXCITATION OF PACKAGE CAVITY RESONANCES BY DISCONTINUITIES



These two discontinuities can interact via the surface wave supported by the substrate. The coupling is ordinarily quite small, but gets large at frequencies corresponding to package cavity resonances. For a large package, the resonances become closely spaced in frequency.

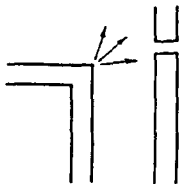
The cavity Q, and therefore the undesired coupling or crosstalk, can be reduced by introducing loss, but the resonances broaden and can form almost a continuum in frequency in large packages.

## CROSSTALK BETWEEN LINES AND DISCONTINUITIES



Is there  
crosstalk?  
NO

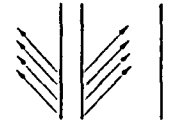
(The surface waves have wavenumbers that vary from  $k_0$  to 0 (or negative) in the component along the line length. They will couple only if the  $k_x$  component is  $= \beta$ . Here  $\beta > k_0$ )



Is there  
crosstalk?  
YES

## CROSSTALK DUE TO LEAKAGE

### 1. Two Identical Lines



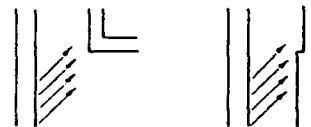
Do they couple?  
YES  
(Directional coupler  
action, at a distance.)

### 2. Two Different Lines



Do they couple?  
NO  
( $\alpha$  and  $\beta$  must be the  
same for coupling)

### 3. If There Is a Discontinuity on the Second Line



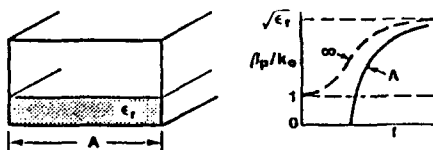
Crosstalk?  
YES

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## PACKAGE MODES

Two types: Guided modes  
Cavity modes

### EXAMPLE OF GUIDED MODES



If metal bounding walls

$$\beta_p^2 = \beta_s^2 - (n\pi/A)^2$$

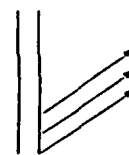
$$\beta_s = k_s$$

Thus, there is a simple relation between the package guided modes and surface waves on an unbounded substrate. There are many package guided modes for each possible surface wave. If  $A$  is large, they are closely spaced in wavenumber.

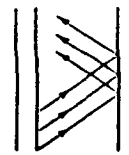
## WHAT HAPPENS TO LEAKAGE WHEN THE PACKAGE IS FINITE?

Infinite width

Finite width



Complex wavenumber  
 $\beta - j\alpha$



Real wavenumber  
 $\beta$

The "leakage" power is reflected from the boundaries, but it is still spread throughout the circuit, and still can cause crosstalk.

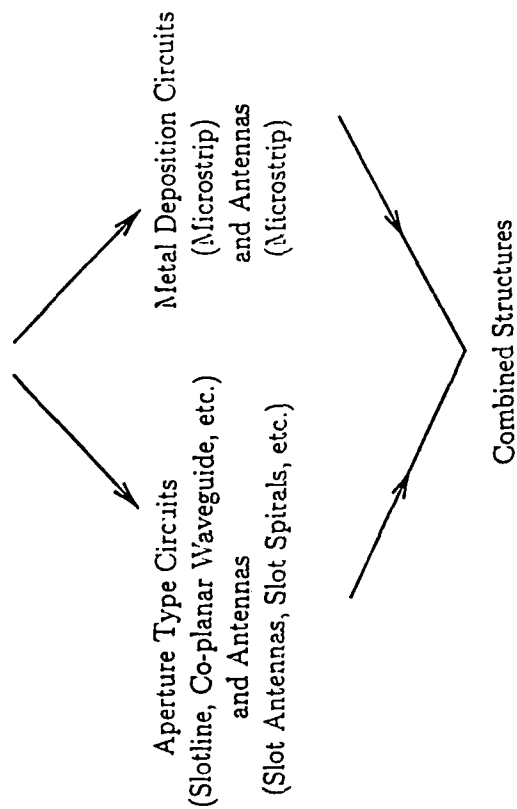
**3D MODELING OF MILLIMETER - WAVE CIRCUITS  
AND ANTENNAS**

**N. Alexopoulos**

**UCLA • USA**

## List of Some Advantages of this Philosophy

- a. Integral Equation Modeling
- b. Method of Moments Applies
- c. Green's Function can be derived
  - c.1 Multilayered Structures
  - c.2 Anisotropic, Gyrotropic Materials
  - c.3 Magnetic and Electric Sources



## Unified Theory for Printed Circuits, Antennas and Scatterers

N.G. Alexopoulos  
Electrical Engineering Department  
University of California  
Los Angeles  
CA 90024

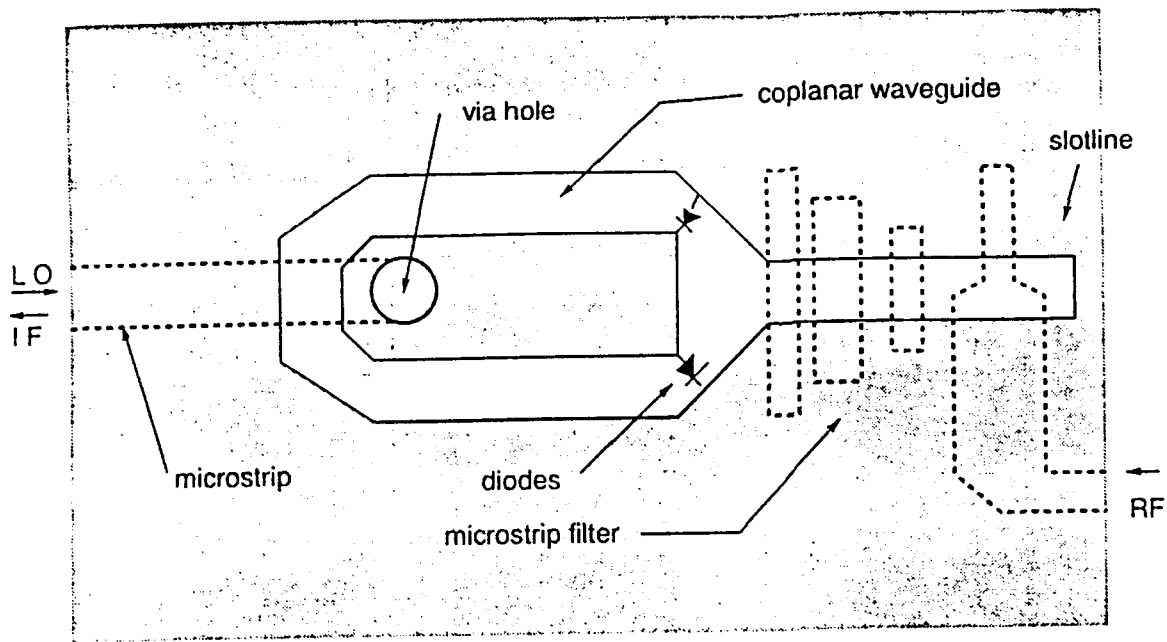
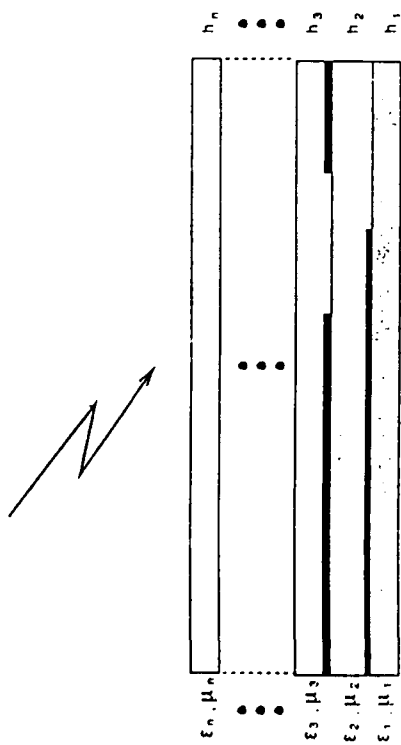
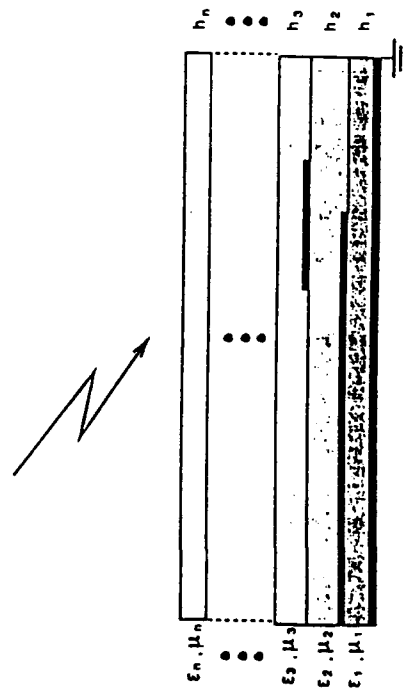


Fig. Balanced mixer

Microstrip Circuits - Aperture Antennas and Scatterers



Microstrip Circuits, Antennas and Scatterers



# Microstrip Linear Array of EMC Dipoles with a Corporate Feed

Tzyy-Sheng Horng \* and Nicolaos G. Alexopoulos  
Electrical Engineering Department  
University of California, Los Angeles  
Los Angeles, CA90024

June, 1991

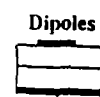


Fig. Linear array of microstrip dipoles

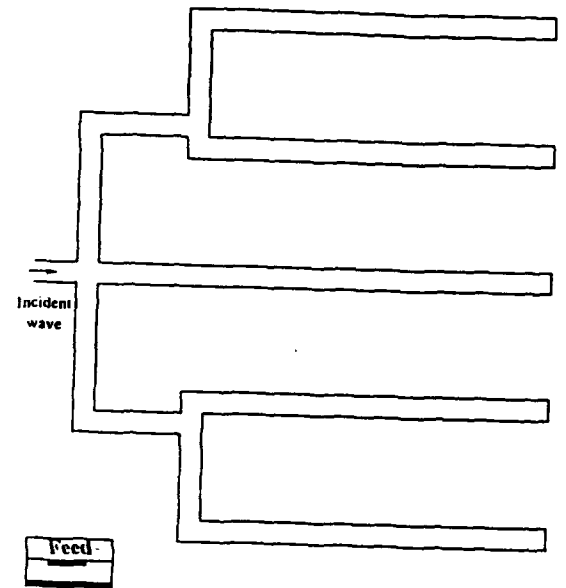


Fig Microstrip corporate feed

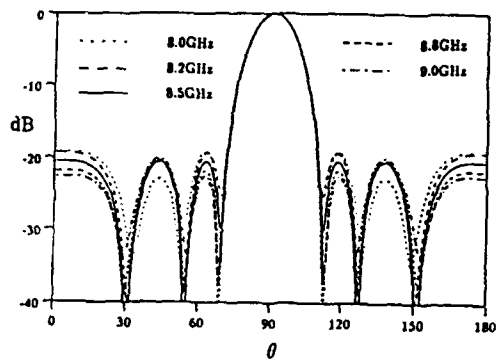


Fig. 7. Array pattern

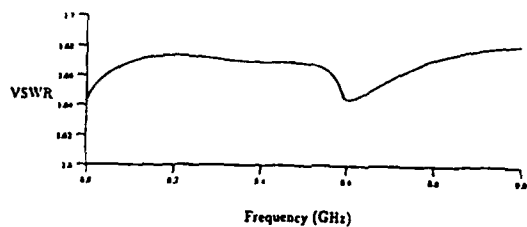


Fig. 8. VSWR versus Frequency

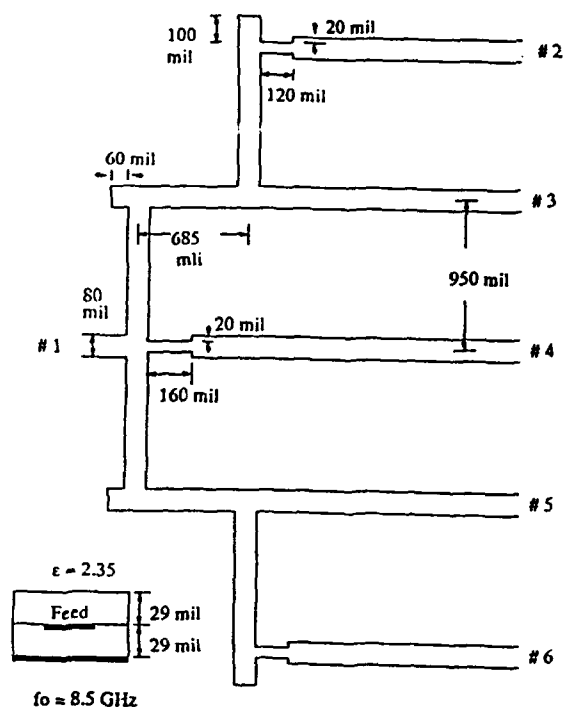
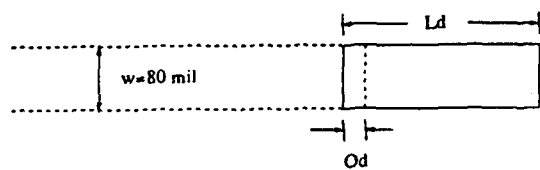


Fig. 6. An embedded microstrip corporate feed



$L_d = 442 \text{ mil}$ ,  $O_d = 15 \text{ mil}$

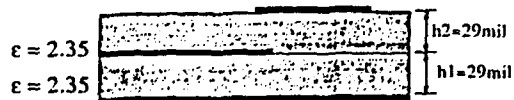


Fig. Electromagnetically coupled dipole fed by a microstripline

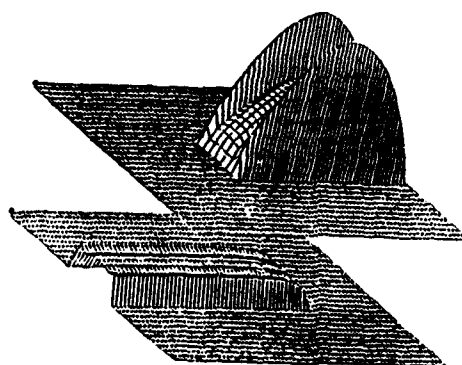
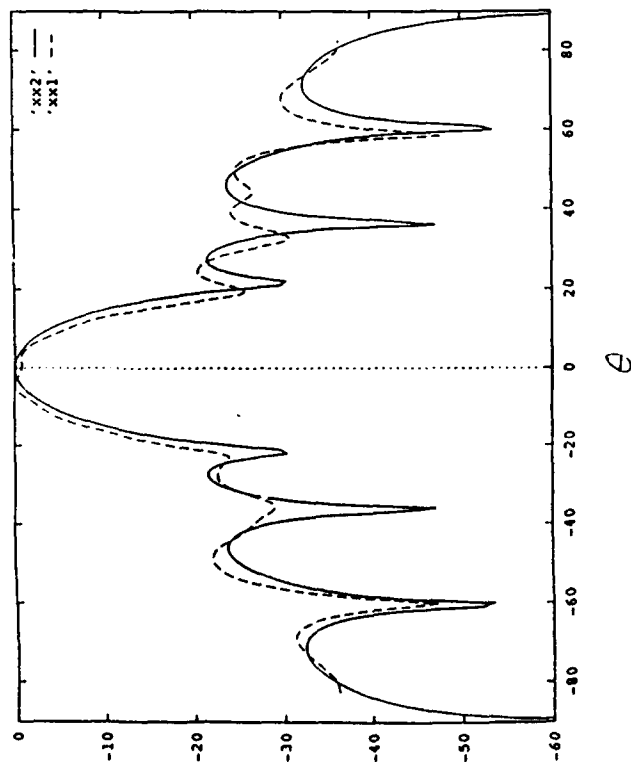
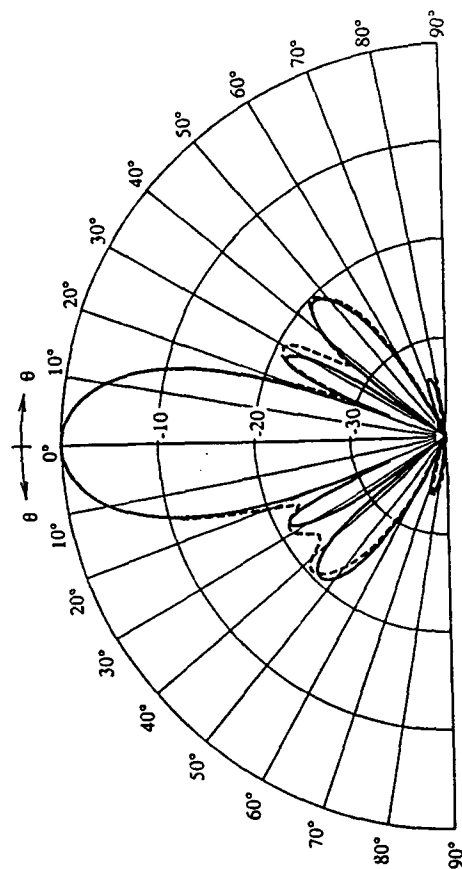
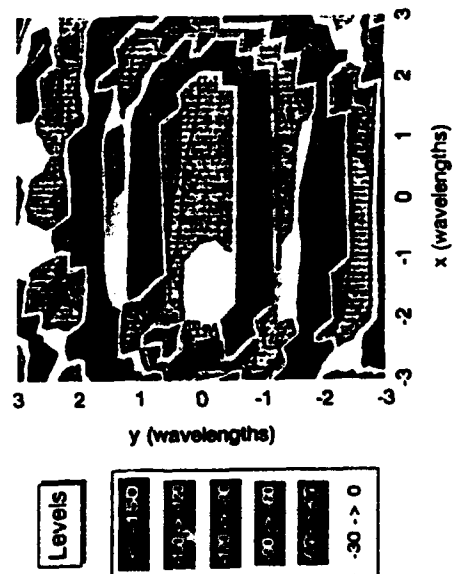


Fig. The magnitude of microstrip currents in an isolated EMC dipole



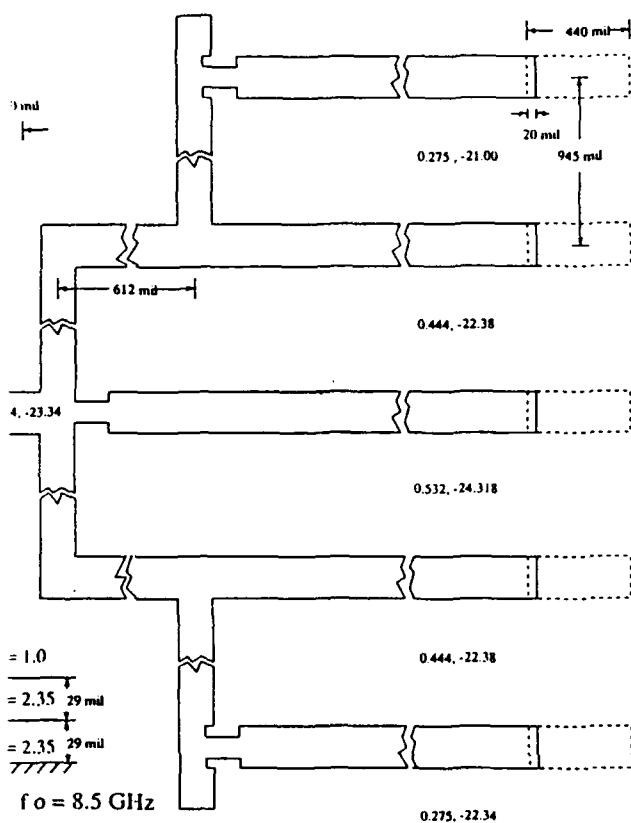
Phase Image of  
5 Element Dipole Array

file = jason1.dat  
freq = 8.5 GHz



H Plane Pattern for a Linear 5 Element Array Microstrip Dipoles

# Microstrip Linear Array of EMC Dipoles with a Corporate Feed



• Relative current distribution in the output ports

1/3.40° 2.42/1.22° 3.15/0° 2.42/1.22° 1/3.40°

• VSWR 1.67

• Directivity 5.77

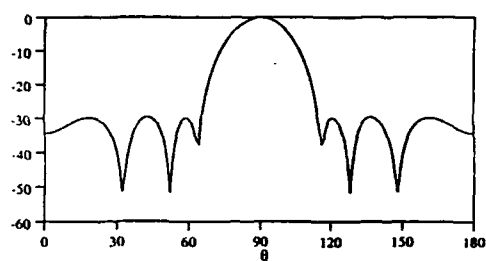
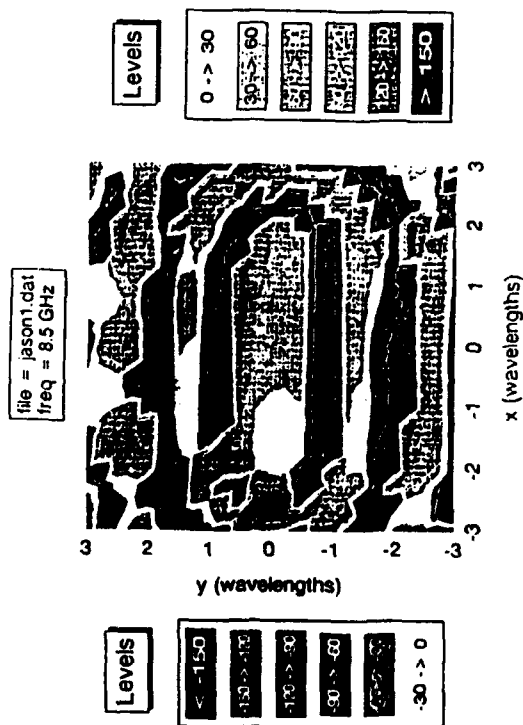
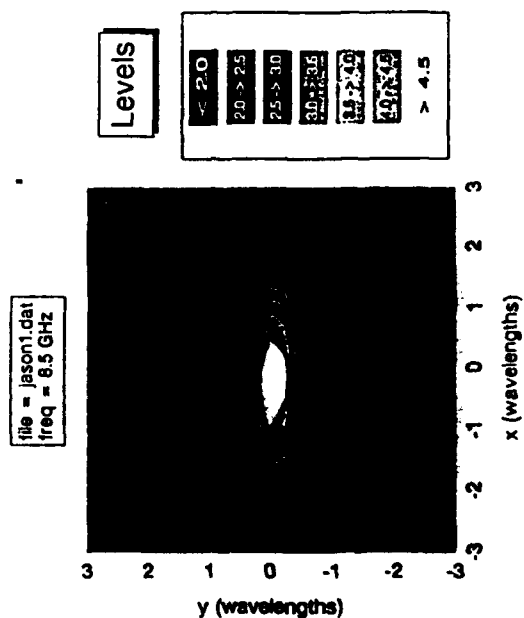


Fig. Array pattern

## Phase Image of 5 Element Dipole Array



## Magnitude Image of 5 Element Dipole Array



## • Computation Time

GOAL: To calculate the matrix elements  $Z_{ij}(m, n)$  between any two basis functions

where

$i, j$  identify type of basis function

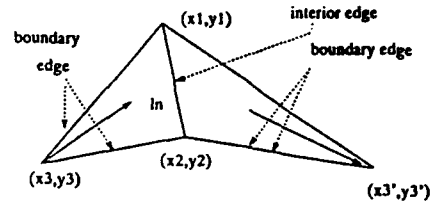
$$i, j = \begin{cases} 1, & \text{Rectangular subdomains} \\ 2, & \text{Triangular subdomains} \\ 3, & \text{Semi-infinite traveling wave modes} \end{cases}$$

$m, n$  are relative distance indices,  $m = [-50 : 50], n = [-50 : 50]$

Computation time on IBM-3090 for matrix elements

- $Z_{11}$ : 2100 sec    2 rect subdomains    setting up database
- $Z_{22}$ : 3900 sec    2 triang subdomains
- $Z_{13}$ : 8500 sec

## Triangular basis function



Ref: Rao, Wilton and Glisson, AP-30, pp.409-418, May 1982

For each interior edge  $l_n$ , the triangular basis function is defined as

$$\bar{F}_n(x, y) = \begin{cases} \frac{1}{2A_n}(x - x_3, y - y_3), & x, y \text{ in } T_n^+ \\ \frac{1}{2A_n}(x_3' - x, y_3' - y), & x, y \text{ in } T_n^- \\ 0, & \text{otherwise} \end{cases}$$

## • Method of Moments

Basis Functions

$$\bar{J}(x, y) = \sum_n C_n^e \bar{J}_n(x, y)$$

$$\bar{M}(x, y) = \sum_n C_n^m \bar{M}_n(x, y)$$

Integral Equations

$$\langle \bar{E}(x, y) \cdot \bar{J}_n(x, y) \rangle = 0, \quad n = 1, 2, \dots, N, \quad x, y \text{ on microstrips}$$

$$\langle \Delta \bar{H}(x, y) \cdot \bar{M}_n(x, y) \rangle = 0, \quad n = 1, 2, \dots, N, \quad x, y \text{ on slots}$$

Matrix Form

$$[Z]_{N \times N} [I]_N = [V]_N$$

max 5000 unknowns  
~ 600 Mbytes memory  
~ 40 min/mk2  
IBM 3090

Matrix Element

$$Z_{ij}(m, n) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \bar{G}_{ij}(k_x, k_y) \bar{J}_i^m(k_x, k_y) \bar{J}_j^n(k_x, k_y) dk_x dk_y$$

IMS package

## • Expansion Functions

1. Rectangular subdomain modes

$$\bar{J}_n, \bar{M}_n = \begin{cases} PWS(x)P(y)\hat{x}, & x\text{-directed currents} \\ PWS(y)P(x)\hat{y}, & y\text{-directed currents} \end{cases}$$

2. Triangular subdomain modes

$$\bar{J}_n, \bar{M}_n = \begin{cases} \frac{1}{2A_n}(x - x_3, y - y_3), & x, y \text{ in } T_n^+ \\ \frac{1}{2A_n}(x_3' - x, y_3' - y), & x, y \text{ in } T_n^- \\ 0, & \text{otherwise} \end{cases}$$

3. Semi-infinite traveling wave modes (entire domain modes)

$$\bar{J}_n, \bar{M}_n = \begin{cases} U(x - x_r)e^{\pm j\beta(x - z_r)}P(y)\hat{x}, & x\text{-directed currents} \\ U(y - y_r)e^{\pm j\beta(y - w)}P(x)\hat{y}, & y\text{-directed currents} \end{cases}$$

## • Testing Functions

Subdomain modes only

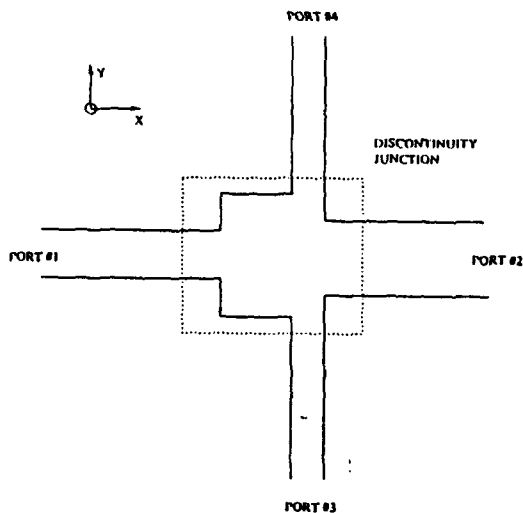


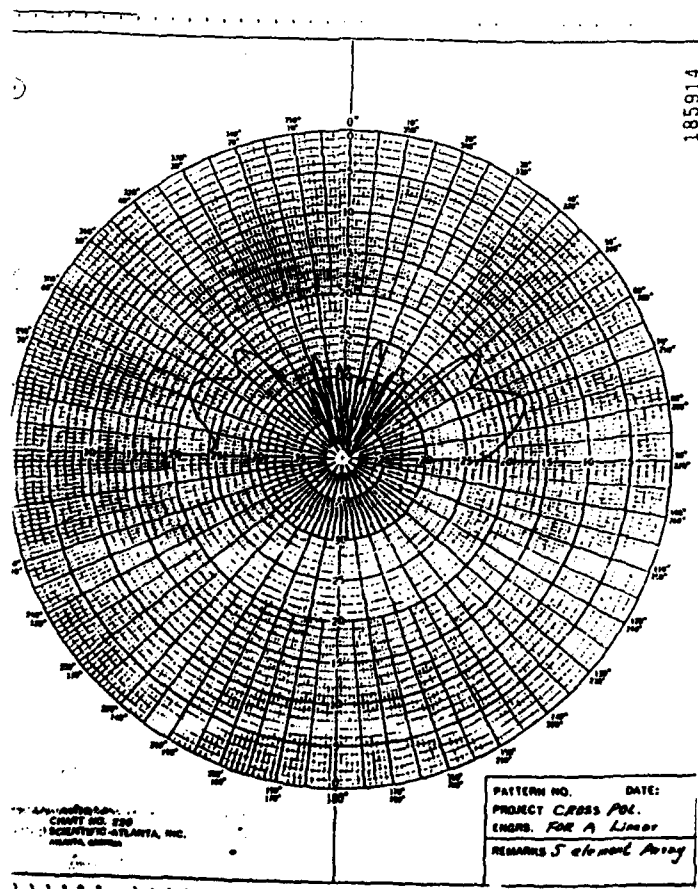
Fig. 1 A generic structure of a four port microstrip discontinuity

# • Spectral-Domain Dyadic Green's Function

$$\begin{bmatrix} \vec{E}(x, y) \\ \Delta \vec{H}(x, y) \end{bmatrix} = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \begin{bmatrix} \vec{G}_{ee}(k_x, k_y) & \vec{G}_{em}(k_x, k_y) \\ \vec{G}_{me}(k_x, k_y) & \vec{G}_{mm}(k_x, k_y) \end{bmatrix} \begin{bmatrix} \vec{J}(k_x, k_y) \\ \vec{M}(k_x, k_y) \end{bmatrix} e^{-jk_x x} e^{-jk_y y} dk_x dk_y$$

where

$$\vec{F}(k_x, k_y) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} F(x, y) e^{jk_x x} e^{jk_y y} dx dy$$



DISCONTINUITIES

JUNCTIONS

AND

TRANSITIONS

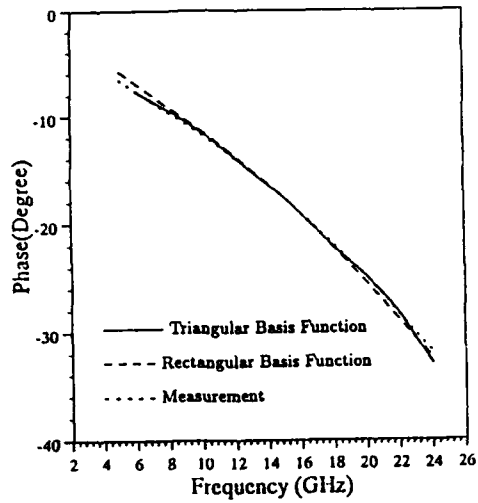


Fig. Phase of S parameters of an open-end ( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $h=25\text{mil}$ )

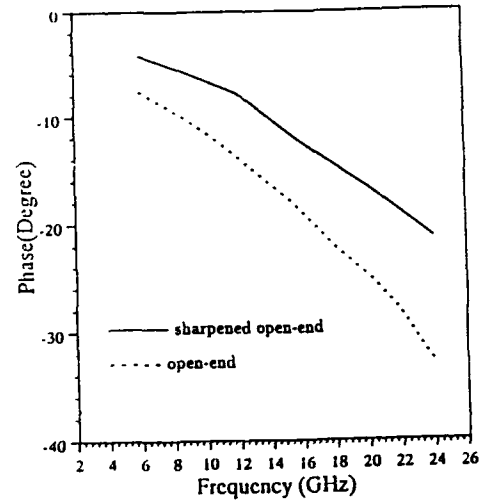


Fig. Phase of S parameters of a sharpened open-end and an open-end ( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $h=25\text{mil}$ )

#### Basic Features Using Triangular Subdomains

1. Transmission lines have finite length.
2. Ideal current source is used for excitation.
3. The grid of subdomains is translated in the x,y plane to facilitate the computation of reaction elements.
4. Subdomains which are too distant from each other to be included in the grid have reactions which are set to zero.

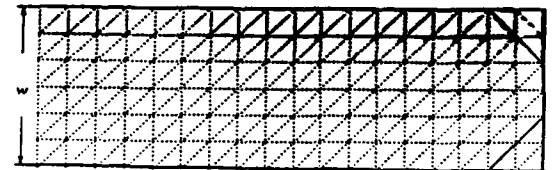


Fig. The geometry of an open-end and a sharpened open-end

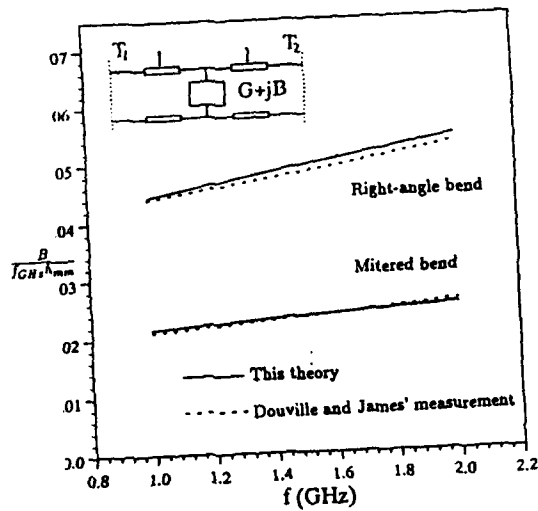


Fig. Frequency dependence of normalized susceptance for right-angle and mitered bend discontinuities ( $\epsilon_r=10.8$ ,  $w=4.572\text{mm}$ ,  $h=5.08\text{mm}$ )

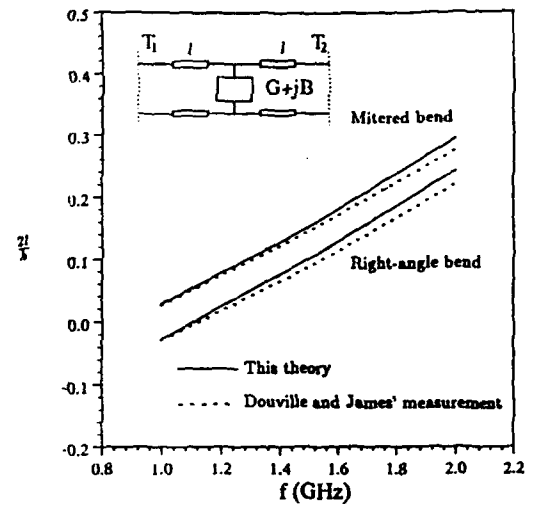


Fig. Frequency dependence of normalized electric length for right-angle bend and mitered bend discontinuities ( $\epsilon_r=10.8$ ,  $w=4.572\text{mm}$ ,  $h=5.08\text{mm}$ )

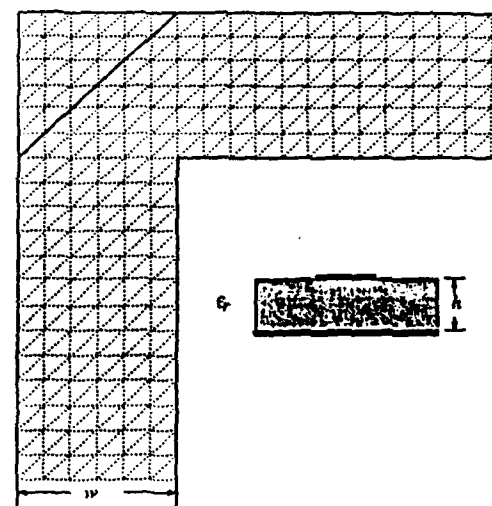
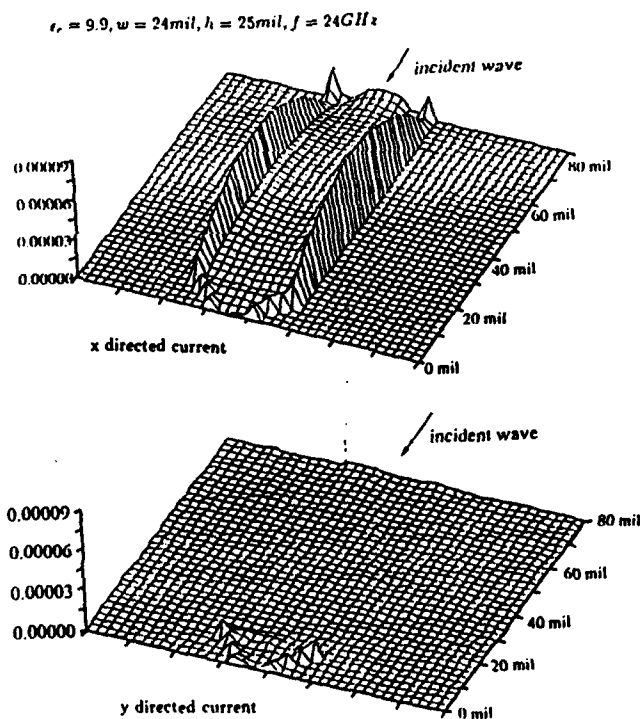


Fig. The geometry of right-angle bend and mitered bend discontinuities

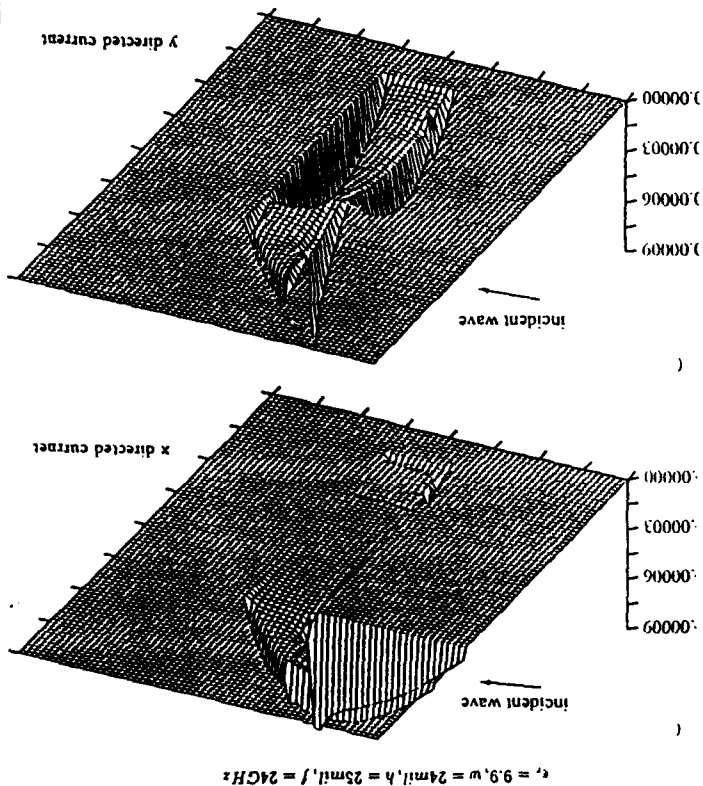
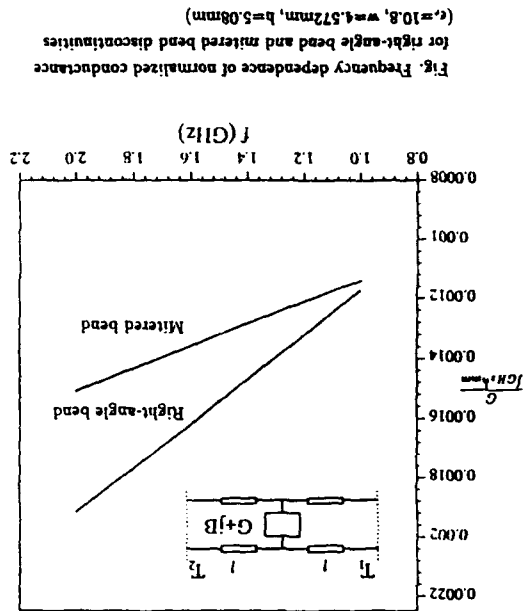
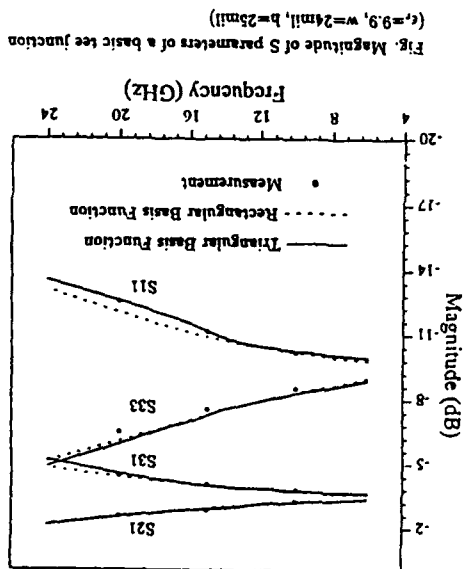
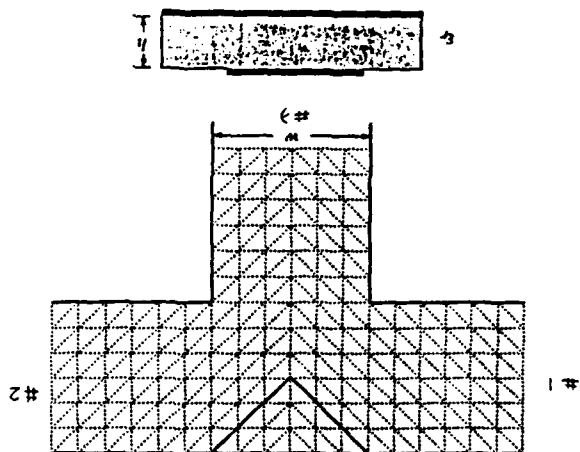


Fig. 2. The geometry of tee and mitered tee junctions



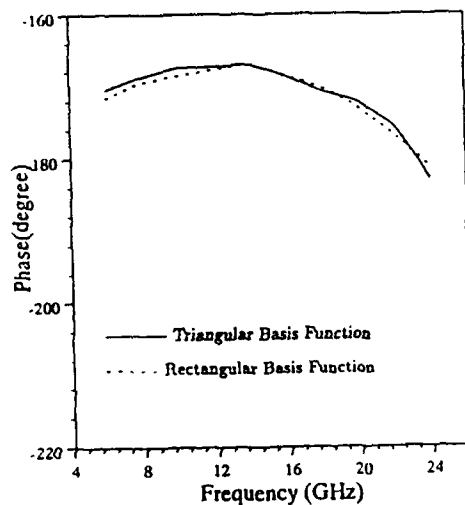


Fig. Phase of  $S_{33}$  of a basic tee junction  
( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $h=25\text{mil}$ )

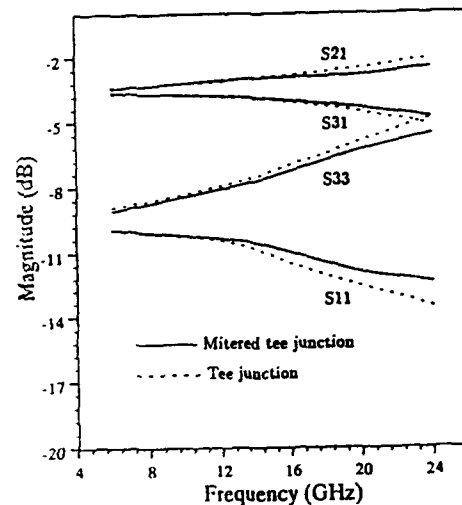


Fig. Magnitude of S parameters of tee and mitered tee junctions  
( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $h=25\text{mil}$ )

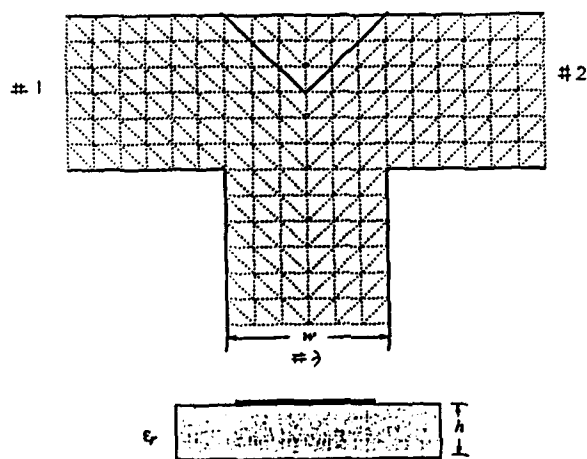


Fig. The geometry of tee and mitered tee junctions

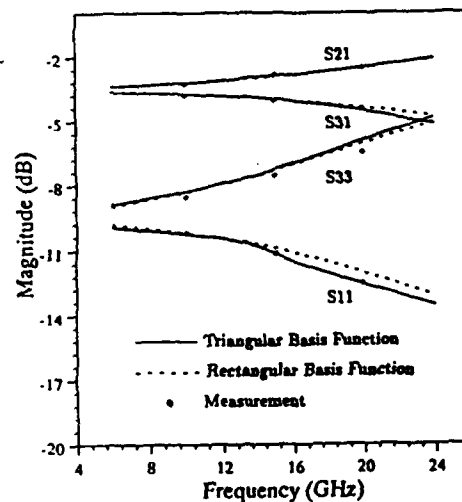


Fig. Magnitude of S parameters of a basic tee junction  
( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $h=25\text{mil}$ )

| Frequency<br>(GHz) | Total<br>Power | Reflected<br>Power | Transmitted<br>Power | Radiation<br>Loss | Surface-wave<br>Loss |
|--------------------|----------------|--------------------|----------------------|-------------------|----------------------|
| 20                 | 1.001          | 0.134              | 0.822                | 0.022             | 0.023                |
| 21                 | 0.997          | 0.153              | 0.787                | 0.027             | 0.030                |
| 22                 | 1.001          | 0.172              | 0.755                | 0.033             | 0.041                |
| 23                 | 1.005          | 0.193              | 0.719                | 0.039             | 0.054                |
| 24                 | 1.007          | 0.208              | 0.678                | 0.048             | 0.073                |

Table 1. Power conservation check for right-angle bend discontinuity.  
(Parameters are the same as those in Fig. 4.)

$$C_1 + Z_0 = \frac{h}{25} \left\{ -0.3154 + 1.1251 \tanh \left( 1.3576 \frac{W}{h} \right) \right\} (p F + \Omega) \quad (1)$$

$$C_2 + Z_0 = \frac{h}{25} \left\{ 0.9102 + 6.8272 \tanh \left( 0.0109 \frac{W}{h} \right) \right\} (p F + \Omega) \quad (2)$$

$$L / Z_0 = \frac{h}{25} \left\{ 0.0103 + 0.0083 \tanh \left( 0.5665 \frac{W}{h} \right) \right\} (n B / \Omega) \quad (3)$$

$$R / Z_0 = 1.1040 - 1.2385 \tanh \left( 0.1036 \frac{W}{h} \right) \quad (4)$$

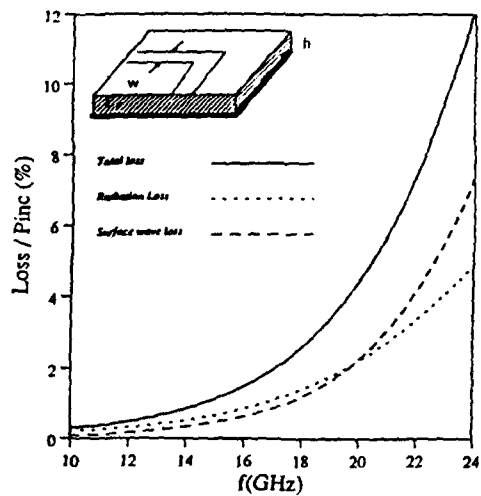


Fig. 4. Power losses versus frequency for right-angle bend discontinuity.  
( $\epsilon_r = 10.2$ ,  $w = 24\text{mil}$ ,  $h = 25\text{mil}$ )

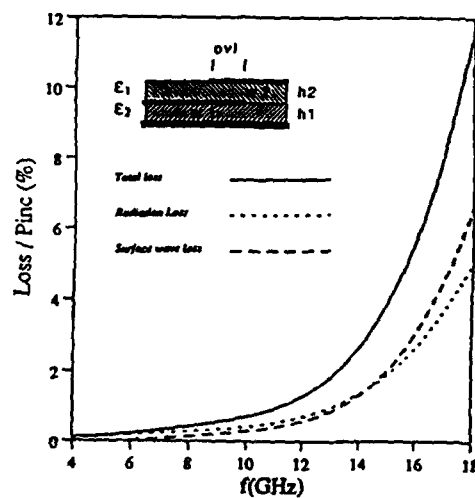


Fig. 5. Power losses versus frequency for overlay EMC lines.  
( $\epsilon_r = 2.2$ ,  $\epsilon_s = 10.2$ ,  $h1 = 25\text{mil}$ ,  $h2 = 25\text{mil}$ ,  
 $w1 = 42\text{mil}$ ,  $w2 = 76\text{mil}$ ,  $ovl = 83\text{mil}$ )

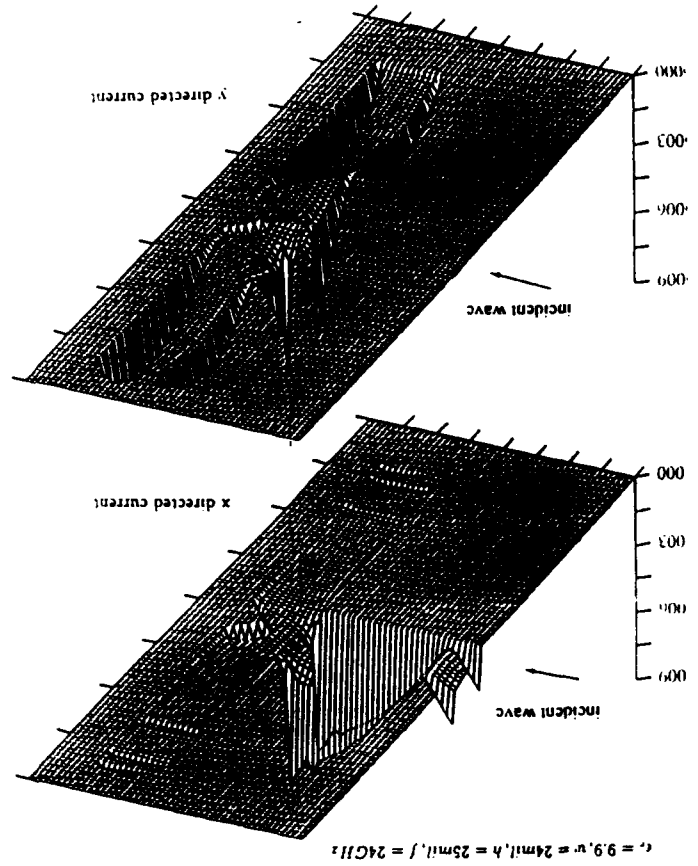


Fig. 1. Integration plane for calculating radiation loss.

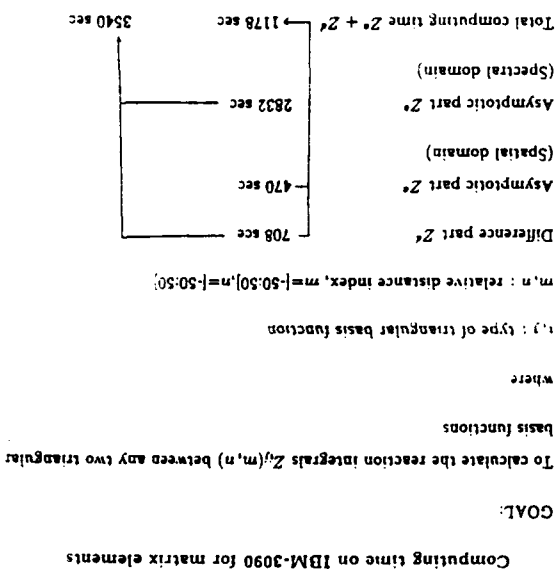
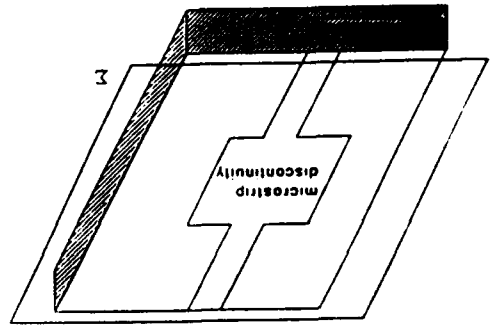
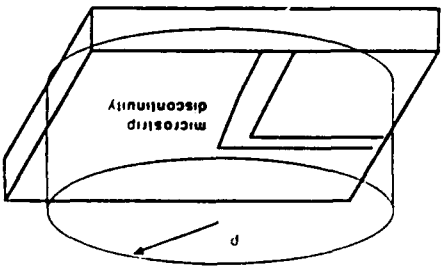


Fig. 2. Integration cylinder for calculating surface-wave loss.



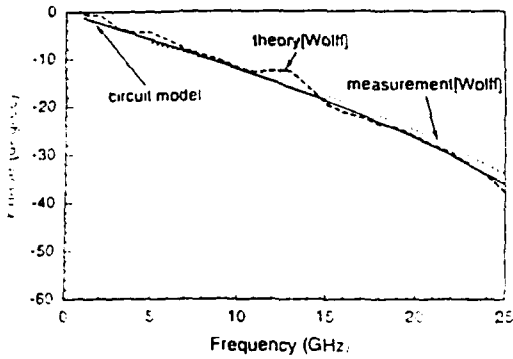
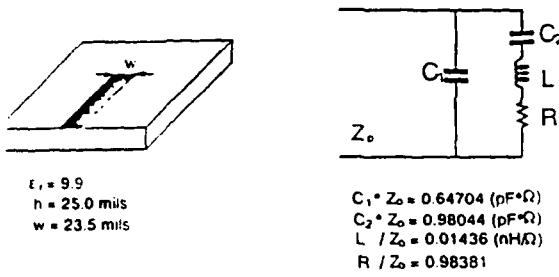
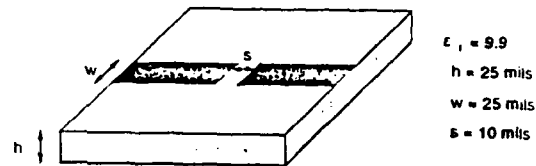


Fig. 3 Comparison of the Phase of the Open-End Discontinuity between the Proposed Circuit Model and the Measured Data

#### • Geometry of Microstrip Gap Discontinuity :



#### • Proposed Circuit Model :

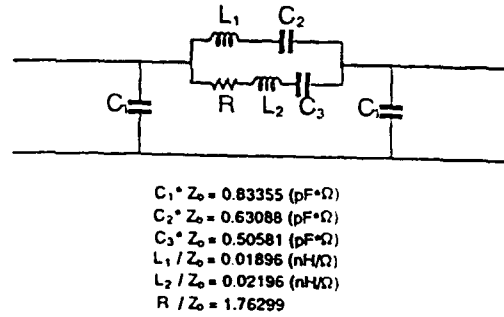
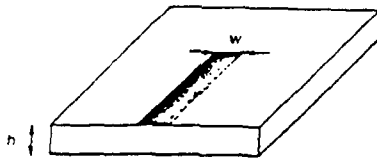


Fig. 4 Physical Dimension and the Equivalent Circuit Model of a Microstrip Gap Discontinuity

#### Microstrip Open-End Geometry :



#### Proposed Circuit Model :

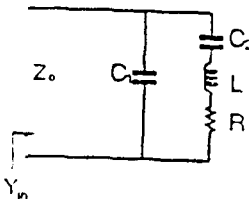


Fig. 5 Physical Dimension and the Equivalent Circuit Model of a Microstrip Open-End Discontinuity.

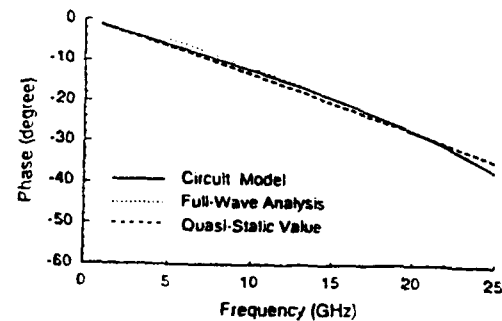
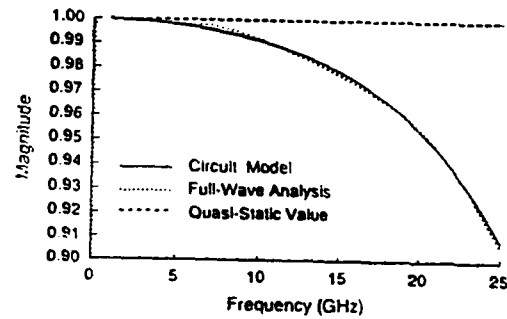


Fig. 2 Reflection Coefficient of a Microstrip Open-End ( $\epsilon_r = 9.9$ ,  $h = 25 \text{ mils}$ ,  $W = 25 \text{ mils}$ )

# Slot with Radial Open Ends

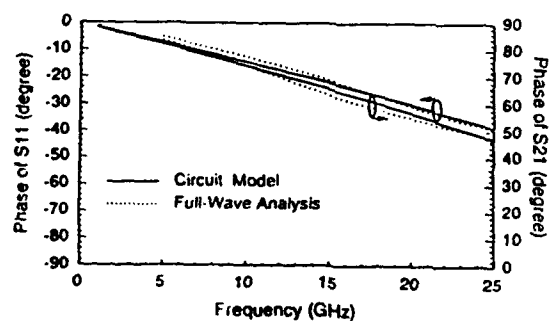
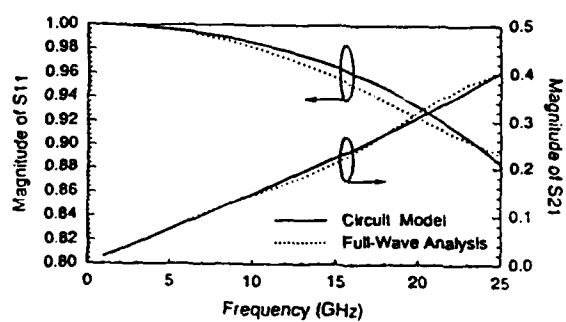
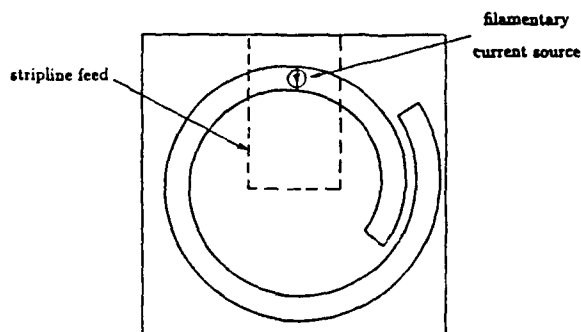
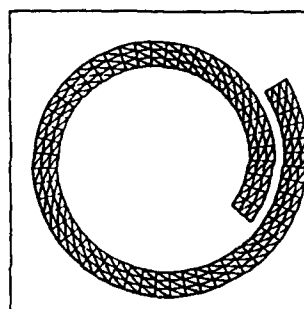
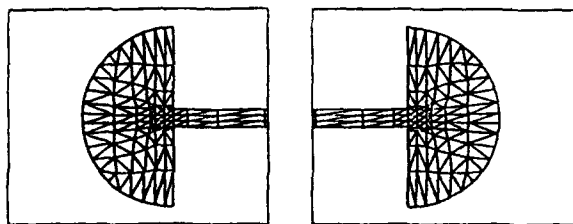
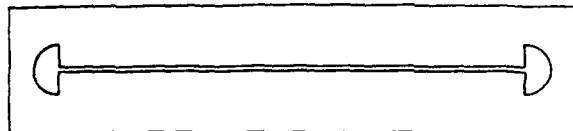
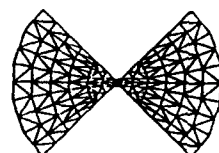


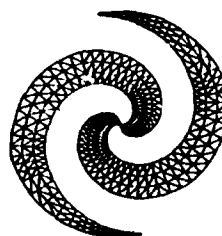
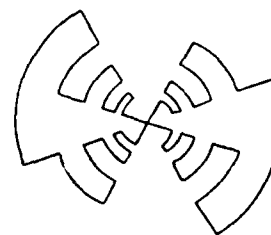
Fig. 5 S Parameters of a Microstrip Gap Discontinuity ( $\epsilon_r = 9.9$ ,  $h = 25$  mils,  $W = 25$  mils,  $a = 10$  mils)

## Potential Applications: Microwave and millimeterwave aperture-type antennas whose geometries include



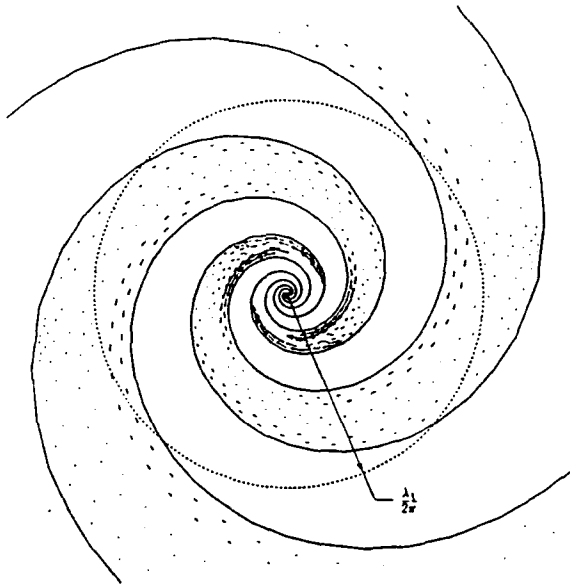
Bowtie

Log-periodic Bowtie



Archimedian Spiral

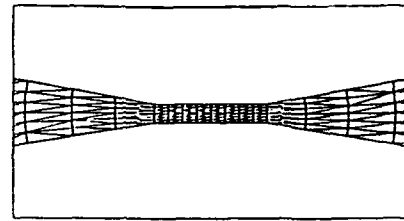
### Distribution of Magnetic Current



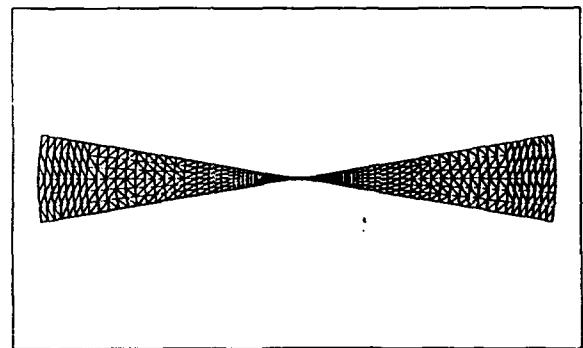
*not plotted  
the current  
is in the spiral*

$$\epsilon_r = 1.0, \quad b/R = 0.2, \quad R/\lambda_1 = 0.847$$

$$R_c = R_m = \dots$$



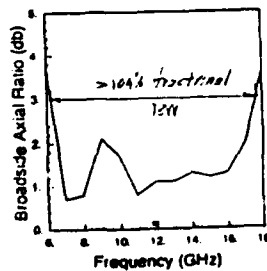
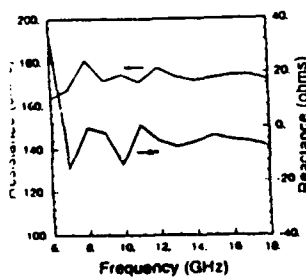
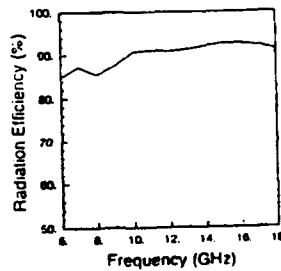
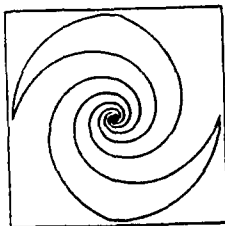
10° Bowtie Slot Antenna  
(898 edges, 2000 mil radius, 40 mil gap)



### Equiangular Spiral Slot Antenna

$$(\epsilon_r = 1.0, \quad b = 300 \text{ mils}, \quad R = 1000 \text{ mils})$$

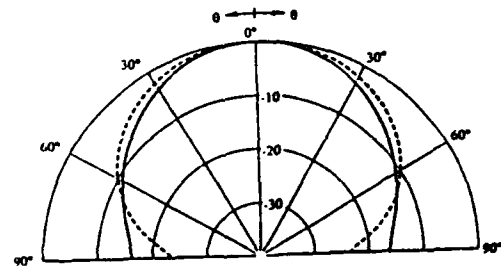
$$(\rho = e^{a\theta + j\phi}, \quad a = 0.3063/\text{rad}, \quad n \in \{0, 1, 2, 3\}, \quad 5 < \rho < 1000 \text{ mils})$$



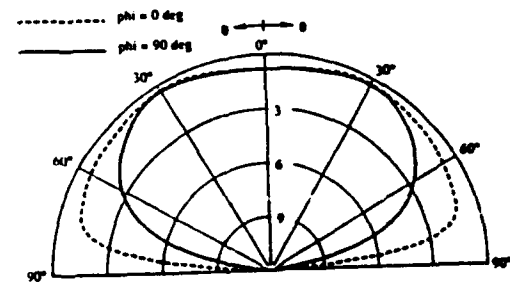
### Equiangular Spiral Slot Antenna

$$(\epsilon_r = 1.0, \quad b = 300 \text{ mils}, \quad R = 1000 \text{ mils})$$

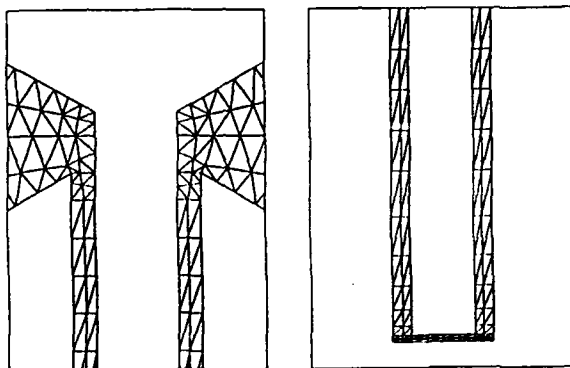
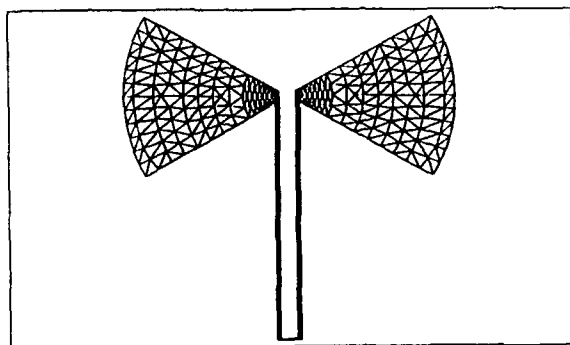
$$(\rho = e^{a\theta + j\phi}, \quad a = 0.3063/\text{rad}, \quad n \in \{0, 1, 2, 3\}, \quad 5 < \rho < 1000 \text{ mils})$$



Normalized Power Pattern (dB) at 8.0 GHz



Axial Ratio Pattern (dB) at 8.0 GHz

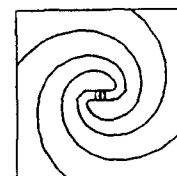


30° Bowtie Slot Antenna with 100Ω CPW Feed  
( 898 edges, 2000 mil radius)

### Equiangular Spiral Slot Antenna

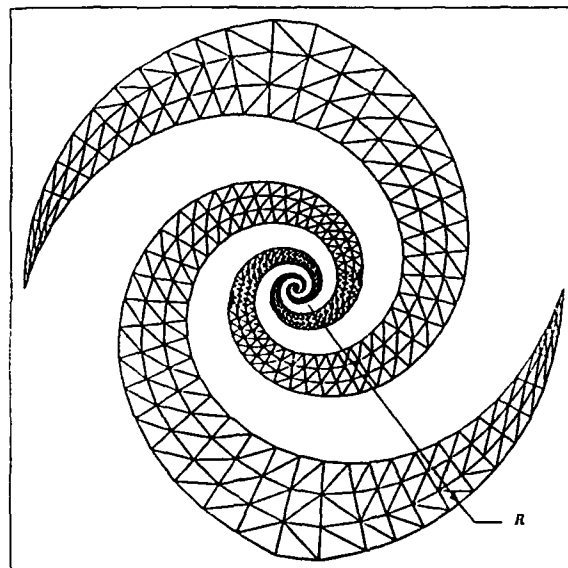
$$\rho = \exp \left[ 0.3063 \left( \phi - n \frac{\pi}{2} \right) \right]$$

$$n = 0, 1, 2, 3 \quad ; \quad 0.005R < \rho < R$$



- .020R

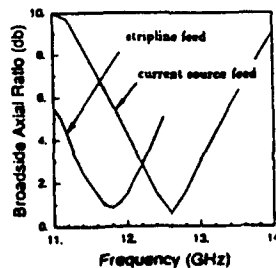
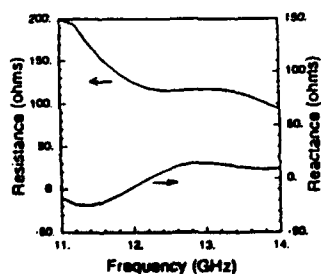
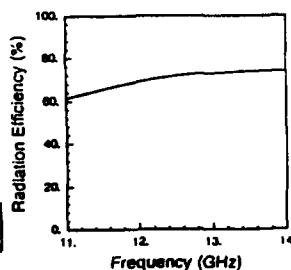
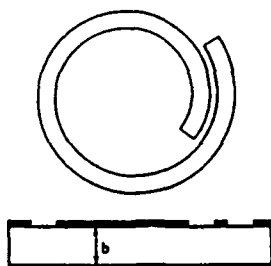
.020R



### Monofilar Archimedean Spiral Slot Antenna

$$(\epsilon_r = 2.2, b = 500 \text{ mils}, w = 40 \text{ mils}, \theta_{st} = 5.81\pi, \theta_{end} = 8.188\pi)$$

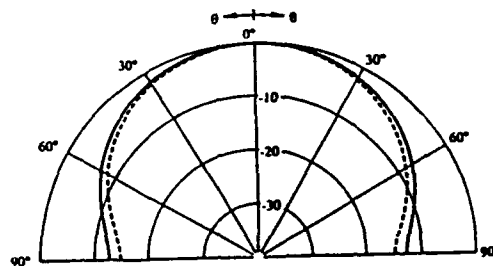
$$(\rho = a\theta + \Delta, a = 9.055 \text{ mils/rad}, \Delta = 7.087 \text{ mils})$$



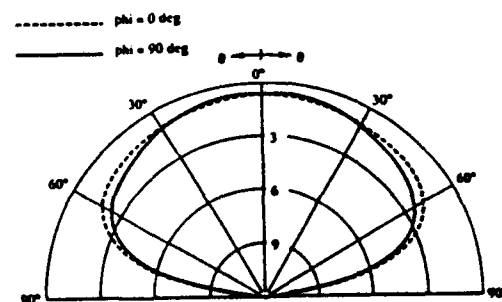
### Monofilar Archimedean Spiral Slot Antenna

$$(\epsilon_r = 2.2, b = 500 \text{ mils}, w = 40 \text{ mils}, \theta_{st} = 5.81\pi, \theta_{end} = 8.188\pi)$$

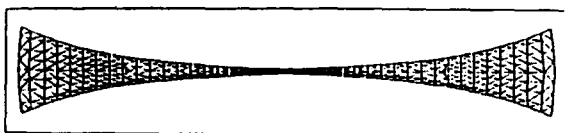
$$(\rho = a\theta + \Delta, a = 9.055 \text{ mils/rad}, \Delta = 7.087 \text{ mils})$$



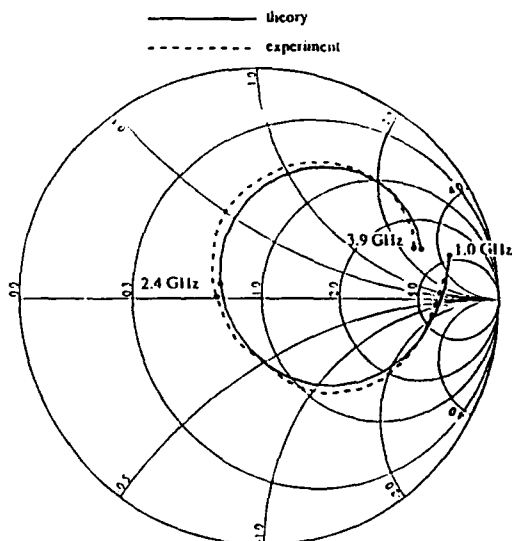
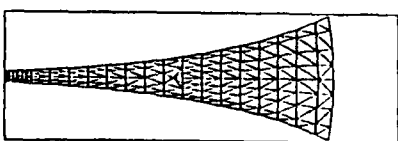
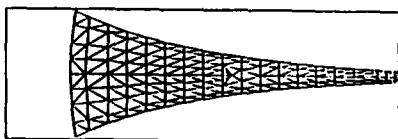
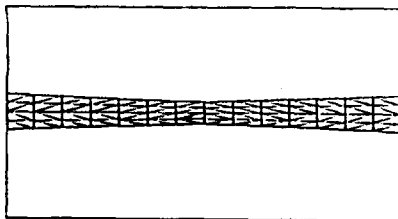
Normalized Power Pattern (dB) at 12.6 GHz



Axial Ratio Pattern (dB) at 12.6 GHz

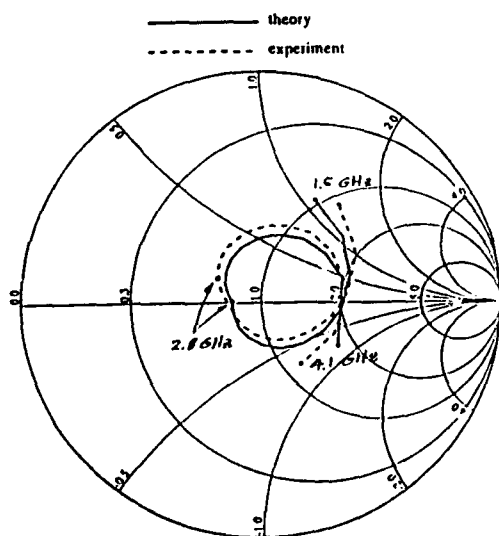


**Exponentially-Tapered Slot Antenna**  
( 736 edges, 2000 mil radius,  $w = 40 \exp(z/700) \text{ mils}$  )

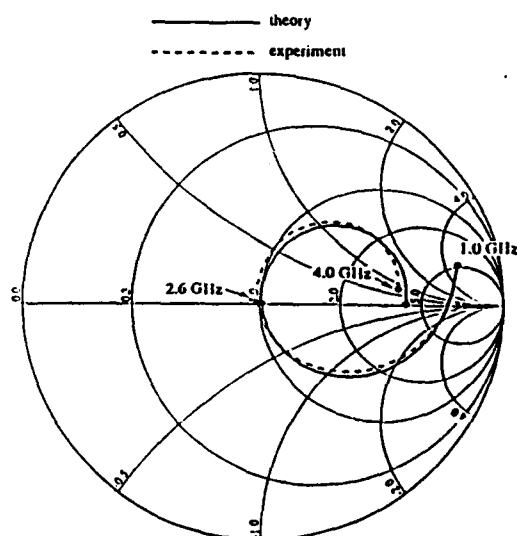


**Reflection Coefficient for the Exponentially-Tapered Slot**

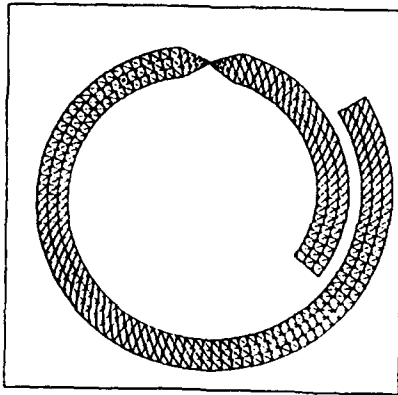
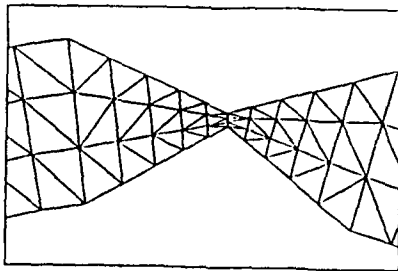
$Z_0 = 50\Omega$



**Reflection Coefficient, 30 Degree Bowtie w/ Feed**  
 $Z_0 = 100\Omega$



**Reflection Coefficient for the 10 Degree Bowtie**  
 $Z_0 = 50\Omega$

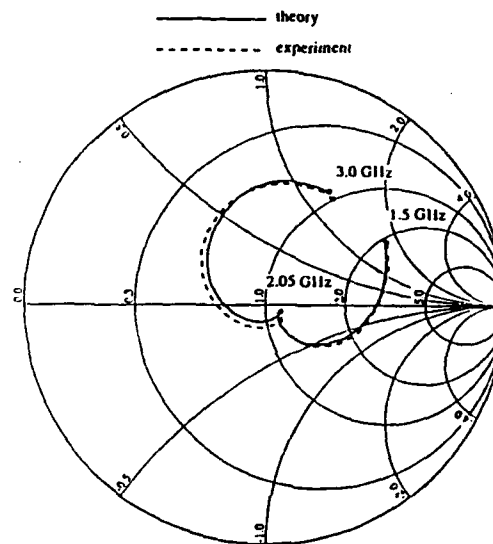


Geometry of the Monofilar Archimedean Spiral  
( 989 edges,  $w=254$  mils,  $r_{min} = 1004$  mils,  $r_{max} = 1434$  mils)

### Monofilar Archimedean Spiral Slot Antenna

$$(\epsilon_r = 1.08, b = 1000 \text{ mils}, w = 254 \text{ mils}, \theta_{sl} = 5.81\pi, \theta_{rad} = 8.188\pi)$$

$$(\rho = a\theta + \Delta, a = 57.5 \text{ mils/rad}, \Delta = 45 \text{ mils})$$

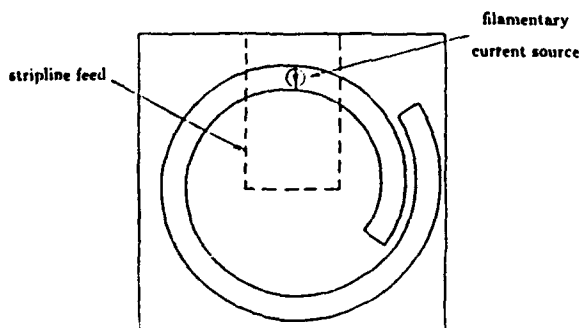
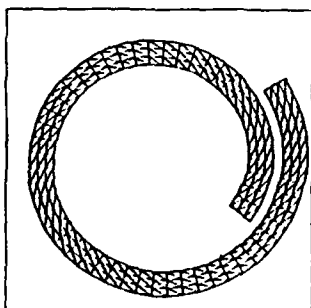


Reflection Coefficient for the Monofilar Spiral Slot Antenna

$$Z_0 = 100 \text{ ohms}$$

### Monofilar Archimedean Spiral Slot Antenna

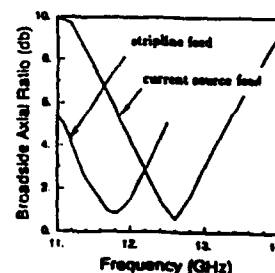
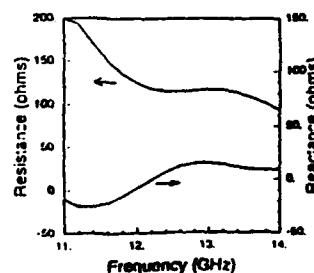
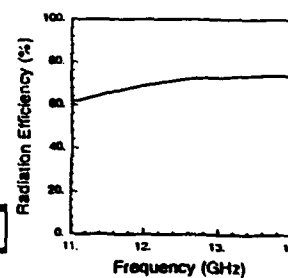
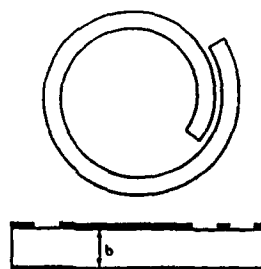
(Ref: Nakano, 1991 IEEE AP Symposium)



### Monofilar Archimedean Spiral Slot Antenna

$$(\epsilon_r = 2.2, b = 500 \text{ mils}, w = 40 \text{ mils}, \theta_{sl} = 5.81\pi, \theta_{rad} = 8.188\pi)$$

$$(\rho = a\theta + \Delta, a = 9.055 \text{ mils/rad}, \Delta = 7.087 \text{ mils})$$



- In each elementary waveguide the e.m. field can be considered as generated by equivalent current and charge sheets placed at the magnetic walls

$$J_s(x,y) = \pm u_x \times H(x,y) \quad \rho_s(x,y) = -\nabla \cdot J_s(x,y) / j\omega$$

$$\text{where } \nabla = u_x \frac{\partial}{\partial x} + u_y \frac{\partial}{\partial y} - j\beta u_z$$

- The electric field can be expressed as:

$$E(x,y) = -\nabla \int_{\text{magnetic walls}} F(x,y;x',y') \rho_s(x',y') dy' - j\omega \int_{\text{magnetic walls}} G(x,y;x',y') \omega \beta J_s(x',y') dy'$$

where the integrals represent the scalar and the vector potentials in the Coulomb gauge, and  $F$  and  $G$  are the pertinent Green's functions. They are the solution of the problem:

$$\nabla \cdot \epsilon \nabla F = -\delta(x-x') \delta(y-y')$$

$$\nabla \times \mu^{-1} \nabla \times G = \omega^2 \epsilon G = 1 \delta(x-x') \delta(y-y') - \epsilon \nabla \nabla F$$

with appropriate boundary conditions.

- Functions  $F$  and  $G$  can be determined analytically in each region.
- Enforcing the continuity of  $u_x \times E$  at the interfaces between the elementary waveguides a system of integral equations is obtained.
- The problem is put into algebraic form using the method of the moments, with base functions suitable to represent both the components of the unknown currents ( $J_s$ ) and the related charges ( $\rho_s$ ) and taking into account the singularities at the metallic and the dielectric edges.

## Standard integral equation approach

- For a given  $\beta$ ,  $\omega$  is found by solving a non-linear eigenvalue problem
- The solution of this problem requires the search of the zeros of a  $\omega$ -dependent matrix, which must be recalculated at each step in the search
- The search of the zeros is a time consuming procedure especially when:
  - the propagating frequencies of many modes are required
  - no estimate of the expected values is available
  - quasi-degenerate modes must be resolved
- The troubles of the integral equation approach derive from the fact that  $G$  is a transcendental function of  $\omega$

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### Microwave Group

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P. Arcioni  
M. Bressan  
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L. Perreggini

#### Present activities on numerical modelling of passive structures :

- Quasi planar waveguides and transmission lines
- Quasi planar 3 D circuits
- Cavities resonators

#### Development of non conventional approaches aiming at

- flexibility
- accuracy
- efficiency

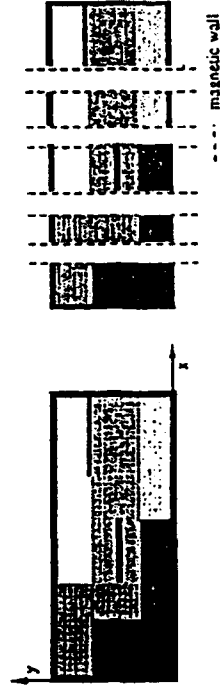
The following presentation aims at giving an idea of the approach under investigation.

### A new method for numerical modelling of general quasi-planar waveguides

M. Bressan, G. Conciauro, P. Gamba

(to be presented at U.R.S.I. Symposium on Electromagnetic Theory, Sydney, 17-20 August 1992)

- Most of the quasi-planar millimeter wave and optical waveguides can be segmented into side-coupled uniformly stratified rectangular waveguides.



- For any propagating mode, the e.m. field in the structure can be expressed as:

$$E = E(x,y) e^{j(\omega t - \beta z)} \quad H = H(x,y) e^{j(\omega t - \beta z)}$$

- The algorithm permits one to calculate the dispersion ( $\omega - \beta$ ) diagram and the modal fields.

## **PARTECIPANTS' CONTRIBUTIONS**

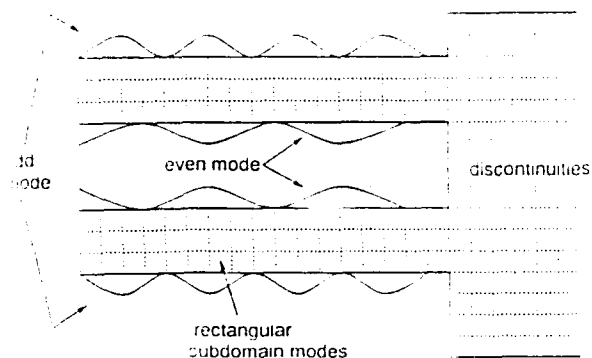


Fig. 2. Layout of expansion modes for magnetic currents on slots

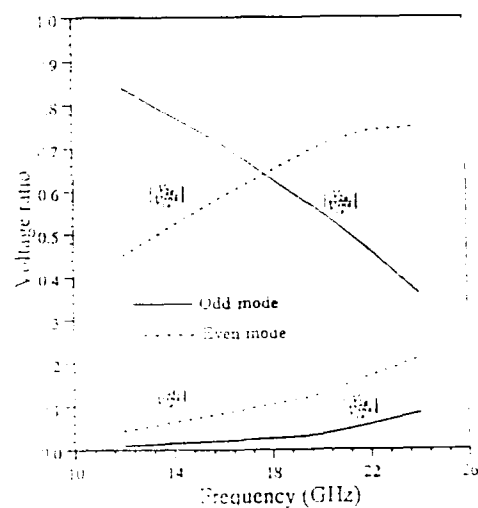
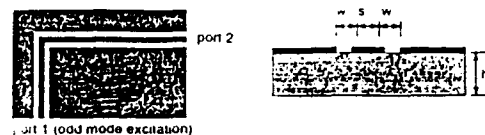


Fig. 3. The slot voltages of even modes and odd modes in a CPW bent-angle bend ( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $s=24\text{mil}$ ,  $h=25\text{mil}$ )

### Computational Performance of the Arbitrary-Geometry Modeling Software on the IBM 3090

| antenna<br>elements | no. of basis<br>functions | execution time<br>per frequency<br>(seconds) | storage<br>required<br>(megabytes) |
|---------------------|---------------------------|----------------------------------------------|------------------------------------|
| Slot Dipole         | 200 - 500                 | 50 - 300                                     | .05 - 4                            |
| Bowtie Slot         | 400 - 800                 | 250 - 500                                    | 2.5 - 11                           |
| Monofilar Spiral    | 500 - 800                 | 300 - 500                                    | 4 - 11                             |
| Archimedian Spiral  | 700 - 2000                | 450 - 2000                                   | 8 - 64                             |
| Equiangular Spiral  | 1000 - 3000               | 800 - 3500                                   | 16 - 150                           |

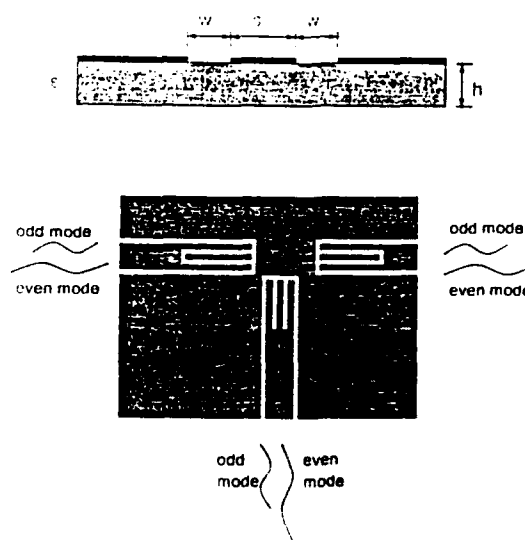


Fig. 1. Asymmetric coplanar waveguide discontinuities

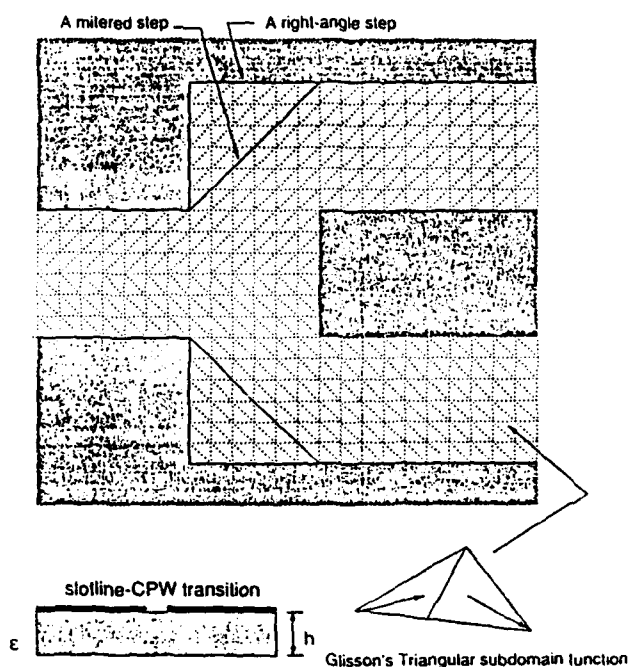


Fig. 1. Slot-to-CPW transition

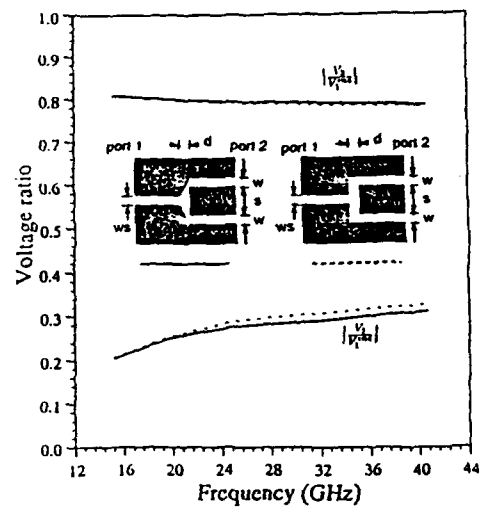


Fig. 2. The slot voltages in a slotline-CPW junction ( $\epsilon_r=9.9$ ,  $ws=0.24\text{mm}$ ,  $w=0.24\text{mm}$ ,  $s=0.24\text{mm}$ ,  $h=0.25\text{mm}$ ,  $d=0.24\text{mm}$ )

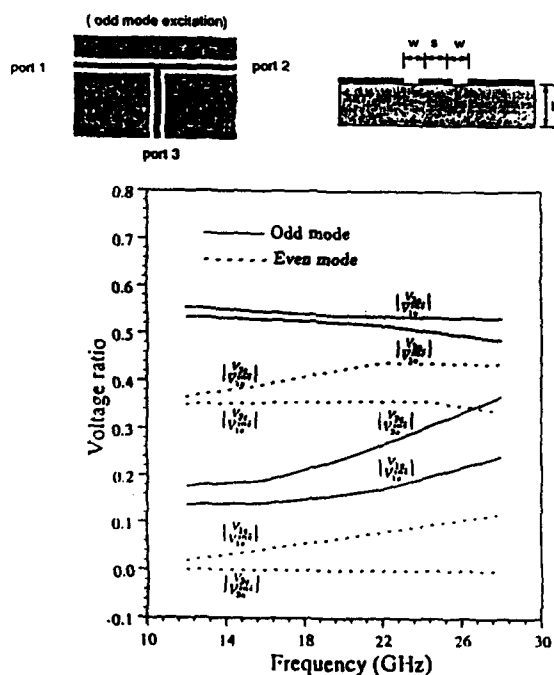


Fig. 4. The slot voltages of even modes and odd modes in a CPW tee junction ( $\epsilon_r=9.9$ ,  $w=24\text{mil}$ ,  $s=24\text{mil}$ ,  $h=25\text{mil}$ )

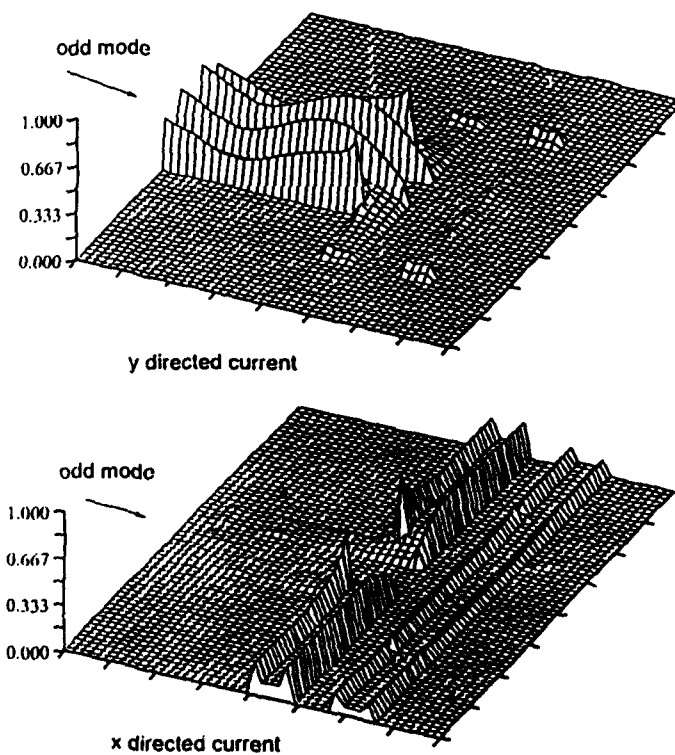
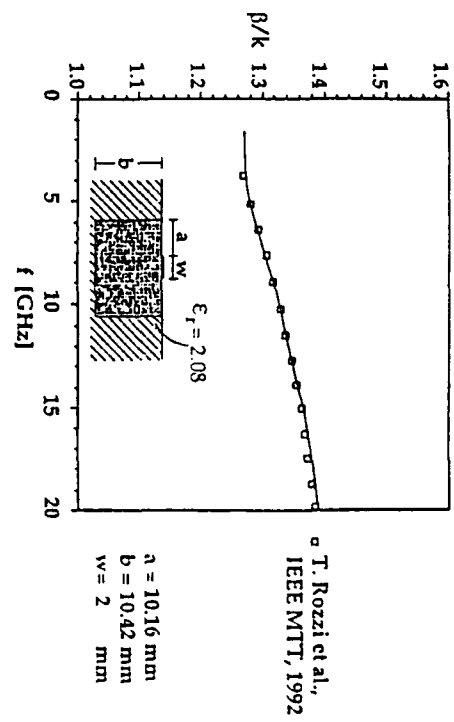


Fig. 5. Magnitude of equivalent magnetic current in a tee junction

A MILLIMETER WAVE SIX-PORT REFLECTOMETER  
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# Abstract

This paper presents a six- port reflectometer in the W band using Schottky diodes as power detectors. A linearization procedure and AC detection technique are used to improve the measurement accuracy. Moreover, a calibration procedure requiring two standards is described.



# Hybrid integral equation/mode matching approach

- $\underline{G}$  can be represented as:

$$\underline{G}(x,y,x',y',\omega,\beta) = \underline{G}(x,y,x',y',0,\beta) - \omega^2 \sum_m \frac{\underline{g}_m(x,y,\beta) \underline{g}_m(x',y',\beta)}{[\omega^2 - \omega_m^2(\beta)] \omega_m^2(\beta)}$$

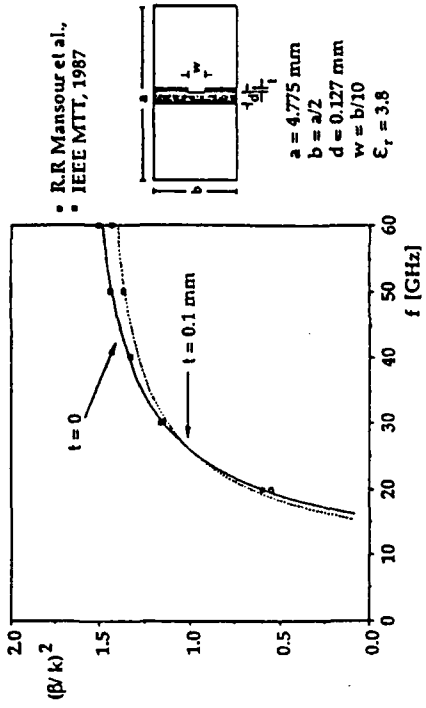
where  $\omega_m$  and  $\underline{g}_m$  represent the electric fields and the cutoff frequencies of the modes of the elementary waveguides. They are the eigenvalues and the eigenvectors of the problem:

$$\nabla \times \mu^{-1} \nabla \times \underline{g} - \omega^2 \epsilon \underline{g} = 0$$

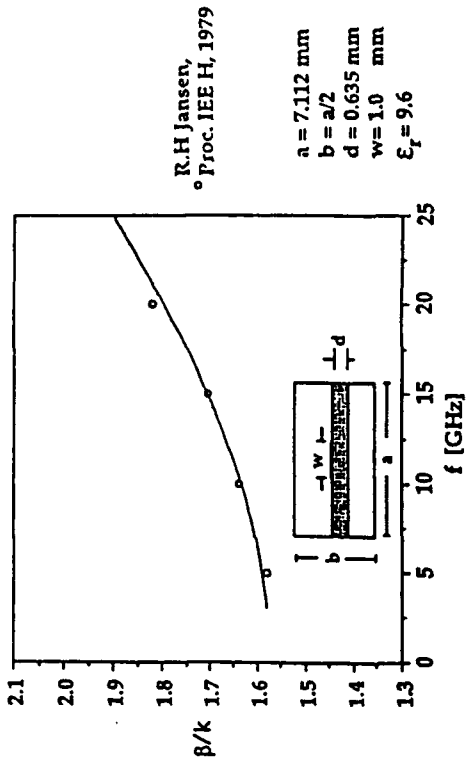
with suitable boundary conditions and normalization.

- Due to its fast convergence (like  $\omega_m^{-4}$ ) the series can be truncated without relative convergence problems.
- After truncation  $\underline{G}$  is represented by a rational function of  $\omega$ .
- Introducing a set of auxiliary variables related to the mode amplitudes in the elementary waveguides, the problem arising from the method of the moments can be converted into a *linear* eigenvalue problem, involving *frequency independent matrices*.
- Since the variables involved represent both the fields at the interfaces and the amplitudes of some modes, the described technique can be viewed as an hybrid between the boundary integral equation method and the mode matching technique.

## Unilateral Finline



## Suspended Strip Line



# **FREQUENCY DEPENDENT ANALYSIS OF PACKAGING PROBLEMS USING THE THREE-DIMENSIONAL INTEGRAL EQUATIONS TECHNIQUE**

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## **ABSTRACT**

The main problems associated with packaging technique for microwave and millimeter integrated or monolithic circuits are the shielding effects, discontinuity and chip connections characterization. In this paper, a dynamic analysis, based on the three-dimensional integral equations technique associated with the method of moments is presented for the characterization of shielded discontinuity problems. It uses the three-dimensional Green's function for a general current distribution inside a shielded box. The theory of loaded scatterers is also used to take into account the effect of existing localized and distributed active devices or passive loads. Using this last theory, a virtual matched load is simulated, numerically, by introducing localized absorbing loads at circuit sections corresponding to output ports. Consequently, a scattering parameters extraction technique, based on the use of these matched terminations, is developed, and hence the [S] parameters of the studied discontinuity are determined from the knowledge of only current or electric field maxima which are often accurately determined under good matching conditions. The shielding effects on a wide range of planar and three-dimensional discontinuities are given. The behaviour of a typical chip connection as a strip air bridge is also studied and the frequency dependent characteristics of these discontinuities are presented.

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## ABSTRACT

This paper presents an oscillator design using a new concept of a FET model, which allows to characterize the single side source grounded configuration of a FET, and therefore to predict the oscillator frequency in the mm-wave range very precisely. The freerunning oscillator was fabricated on a GaAs chip. At 36 GHz an output power of 14 dBm, and an external quality factor of greater than 100 was achieved.

## SUMMARY

For an accurate oscillator design in the mm-wave range the modeling of the active device is essential for a successful design. Especially oscillators are often layouted in a single side grounded source (SSG) configuration, that means only one source pad is grounded, while the second one is not connected. Depending on the gatewidth and the dimensions of the FET, distributed effects for low noise MESFETs have to be regarded above approximately 50 GHz [1, 2]. This assumption is only valid if both source pads are grounded. If the FET is used in a single side grounded source configuration, the FET should be regarded as distributed even above 15 GHz. To our knowledge the effect of SSG has not been regarded yet by any commercial CAD tool.

Especially for oscillator design the FET is often used in a SSG configuration [3] to reach the required instability, otherwise both source connections will have symmetrical networks which will increase the area on a chip. The difference in the measured S-parameters between single side source grounded (SSG) and double side source grounded (DSG) configuration is shown in Fig. 1. Above 15 GHz the S-parameters  $S_{12}$  and  $S_{11}$  show great differences. From the layout orientation of the 4-finger MESFET TEG4-50-25 (Fig. 2), a modular distributed model was deduced, Fig. 3. A 4-finger MESFET with a 0.25  $\mu$ m gate length and a total gatewidth of 200  $\mu$ m from TEG\* is used. The geometrical dimensions of the chip and source configuration are directly implemented as microstrip elements. The source pad which is not grounded in the SSG configuration is described by a tapered open line. The distributed model provides 3-port S-parameters for SSG configuration. Fig. 4 indicates the excellent agreement of the modeled and measured S-parameters for SSG configuration. The FET model is also able to describe the reverse channel operation.

S-parameters of the MESFET were performed using an inhouse developed testfixture with IRL calibration standards for microstrip measurements up to 40 GHz.

The basic concept of the oscillator is a reflection type oscillator with the output at the drain side [4, 5]. In Fig. 5 the layout of the 28 GHz oscillator is shown. The FET is positioned at the edge of the chip and only one sourcepad is extended with a microstrip transmission line to provide the required instability.

The GaAs oscillator has a chip size of 3x3 mm and is fabricated on a 150  $\mu$ m GaAs substrate. The performance of the 36 GHz freerunning Oscillator with  $U_{GS}=5V$  and  $I_{DS}=30mA$  has a power output  $P_{out}=14$  dBm, an external quality factor  $Q_{ext}=100$ .

\* Telefunken Electronic, Heilbronn, Germany

shown in Fig. 6 and Fig. 7, respectively.

Using a common FET model for this oscillator design, the simulated and measured resonance frequency had an offset of about  $\pm 3$  GHz. With the new distributed FET model the predicted resonance frequency of the freerunning oscillator was within a range of  $\pm 200$  MHz.

## CONCLUSION

This paper shows that the FET has to be regarded as distributed above 15 GHz if the FET is used in a SSG configuration. The deduced model describes very accurately the SSG configuration and uses only lumped elements and microstrip lines which makes it practicable for online simulation. This leads to a higher accuracy in oscillator design. The predicted resonance frequency of the freerunning oscillator lies within a range of  $\pm 200$  MHz from the design frequency. Oscillators designed with common FET models had their resonance frequency within a range of  $\pm 3$  GHz. For nonlinear modeling the intrinsic FET will be changed to a large signal model to provide also the possibility of nonlinear circuit design in SSG-configurations.

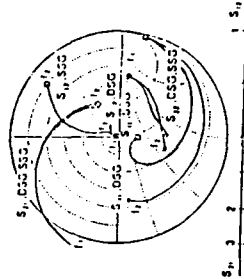


Fig. 1: Measured S-parameter in double side grounded (DSG) and single side grounded (SSG) configuration of a MESFET TEG4-50-25;  $f_1 = 1$  GHz,  $f_2 = 40$  GHz.

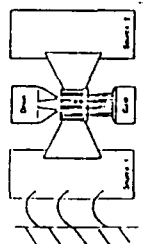


Fig. 2: Layout of the 4-finger MESFET in single side grounded configuration (SSG).

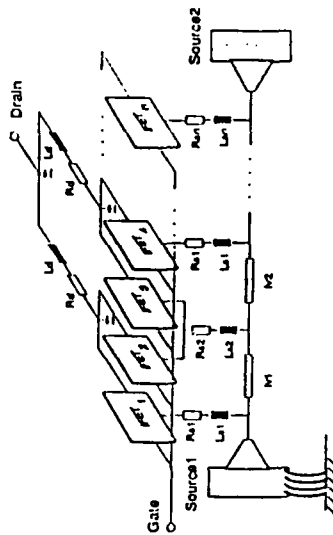


Fig. 3: Modular FET model with distributed source configuration for T-Gate FET in SSG configuration.

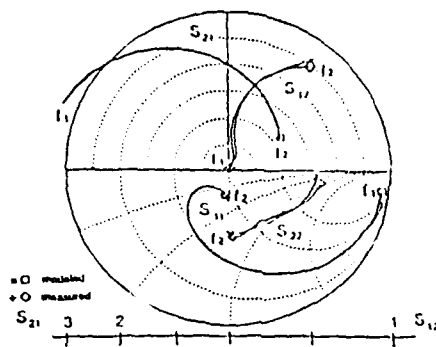


Fig. 4: Measured and modeled S-Parameter in single side grounded configuration (SSG).

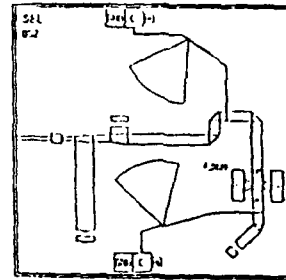


Fig. 5: Layout of the Oscillator.

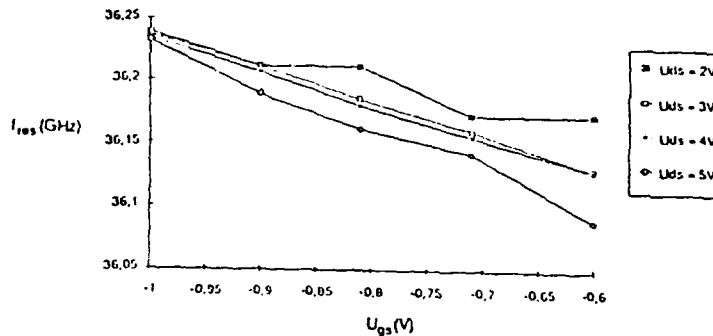


Fig. 6: Resonance frequency vs gate- and drain voltage.

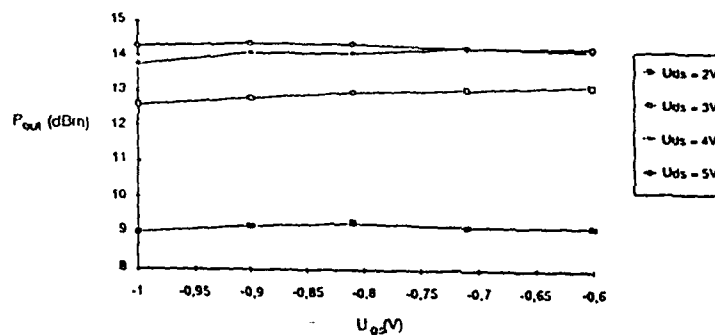


Fig. 7: Output power vs gate- and drain voltage.

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