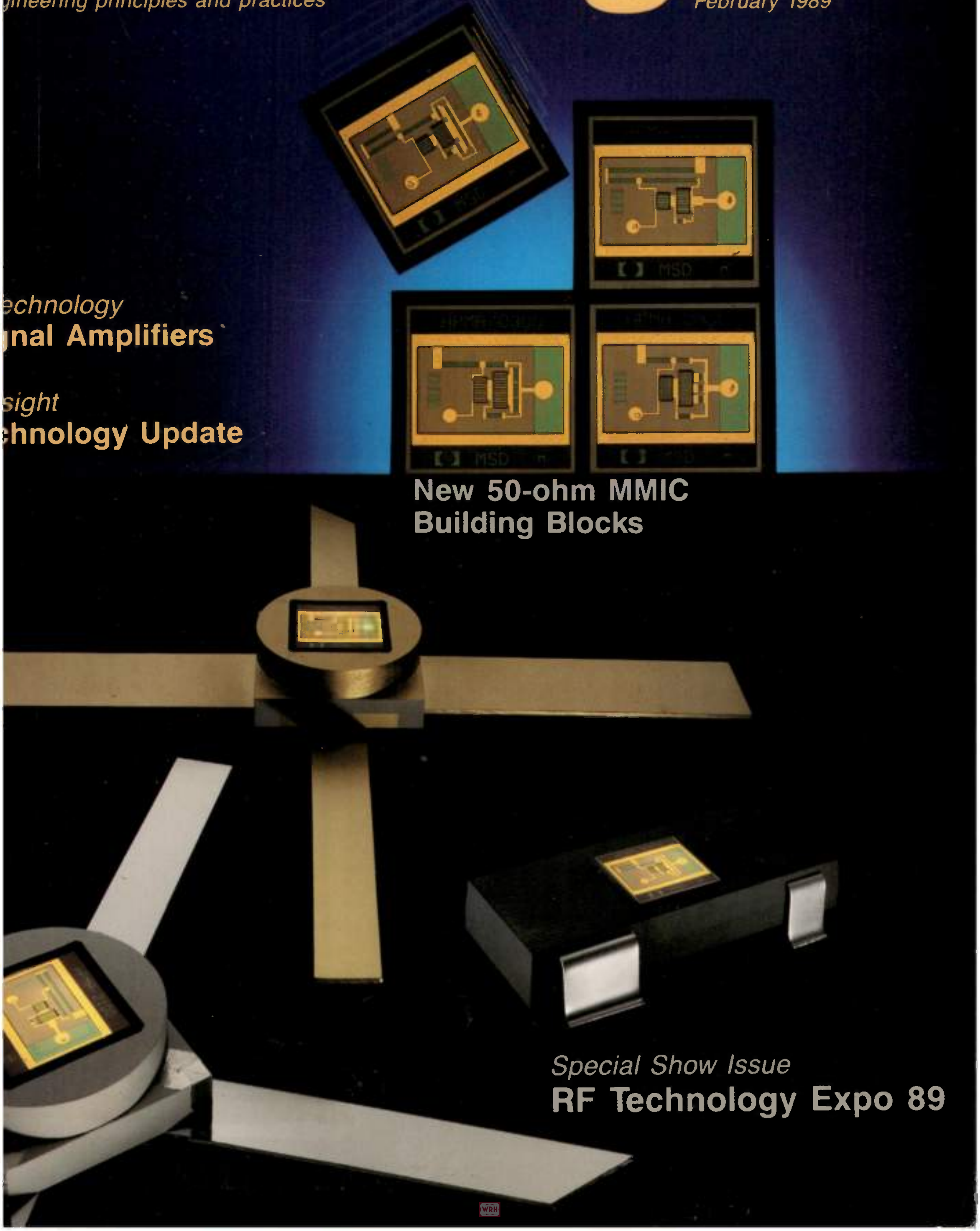
# rf design

*gineering principles and practices*

*February 1989*


*echnology*
**gnal Amplifiers**

*sight*
**chnology Update**



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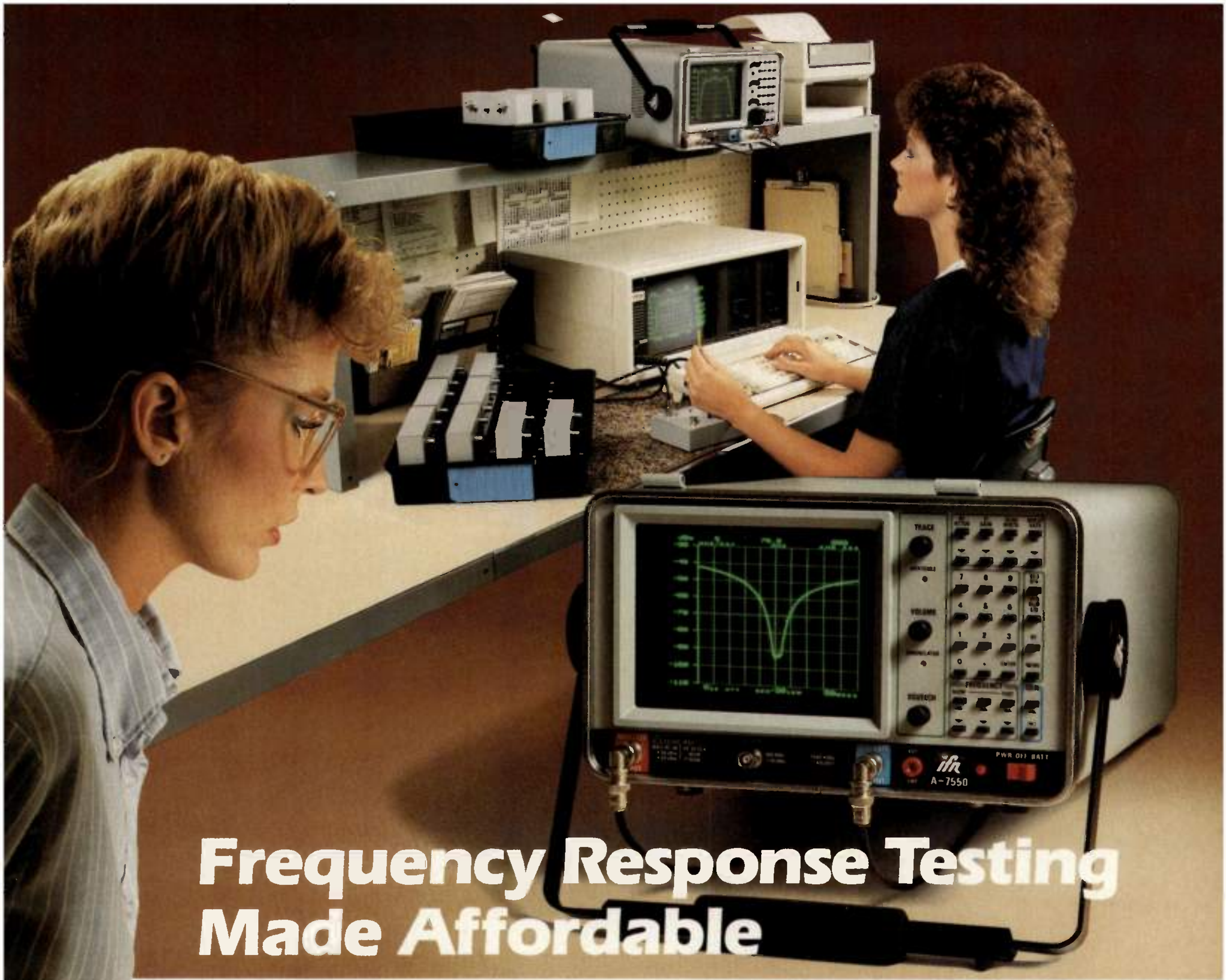


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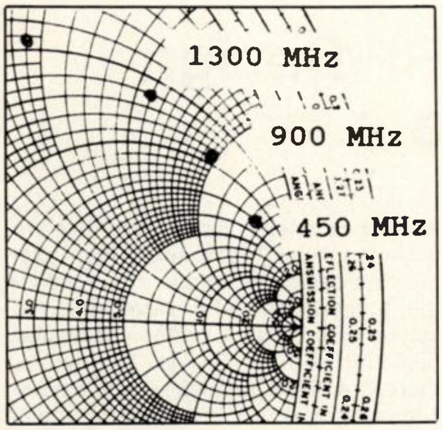


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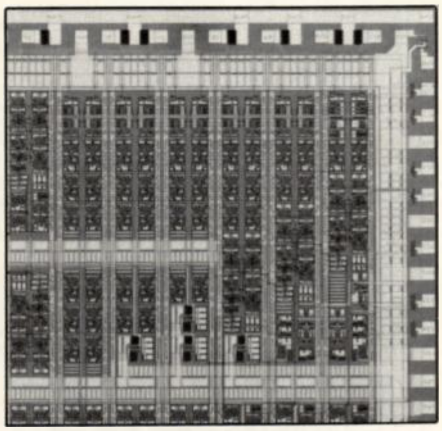

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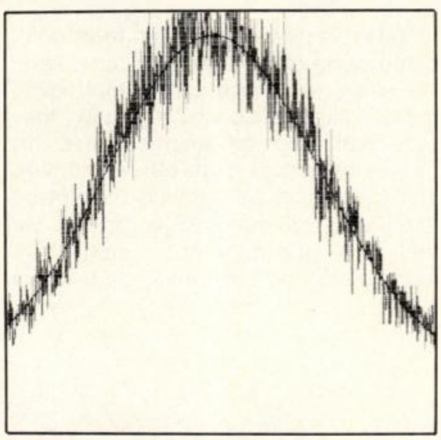

Page 113 — Noisy Signal Tracking



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*rf editorial*

# Introducing the RF Design Software Service

*By Gary A. Breed*
*Editor*

Not too long ago I bought a new car, and as the transaction proceeded, all the necessary forms for registration, bill of sale, tax reporting, etc., were filled out by computer. My insurance agent was even in their data base. In banking, insurance, retail stores, real estate, and so many other businesses we see outside our own offices, the computer has become the preferred method of record-keeping and information transfer.

It has happened more slowly in our industry, but the computer has now virtually replaced the calculator as the handiest tool for RF engineers. Many of us take pride in the fact that our expertise, RF engineering, has been more difficult to computerize than other areas. However, RF engineers will always find the best way to get their jobs done, and the time has come to embrace the computer 100 percent.

Evidence of the rapidly growing role of computers in RF engineering is the number of articles we receive that are accompanied by programs which execute the design technique described by the author. We even have received computer programs written to support past articles in *RF Design*. This month, there are three articles which involve computer-aided RF design techniques — for matching network, PLL, and MOSFET amplifier design.

**A New Service**

To expand our role in the exchange of engineering ideas, we have created the *RF Design Software Service*. For a small charge, our readers can receive programs published in *RF Design*, on either 5 1/4 or 3 1/2 inch diskettes. Initially, the service will provide programs in MS-DOS format, with additional formats to be considered according to demand. Instructions for ordering will be included with each article.

We want to make the service as useful as possible. Possible ideas include support of past articles, and perhaps exchanging programs that may not be published in *RF Design*. Let us know what you think.

Valuable new engineering tools don't come along often. Although computers, like slide rules and calculators, can't replace experience and ingenuity, they sure can help an engineer use his human resources more efficiently! Now that computers have found their place in the daily routine of RF engineers, we feel it is our duty to support them (computers *and* engineers) as best we can.

Gary A Breed

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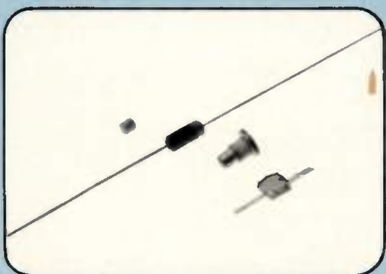

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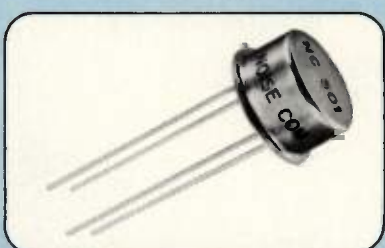

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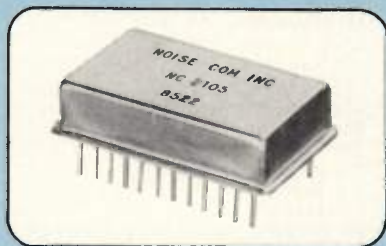


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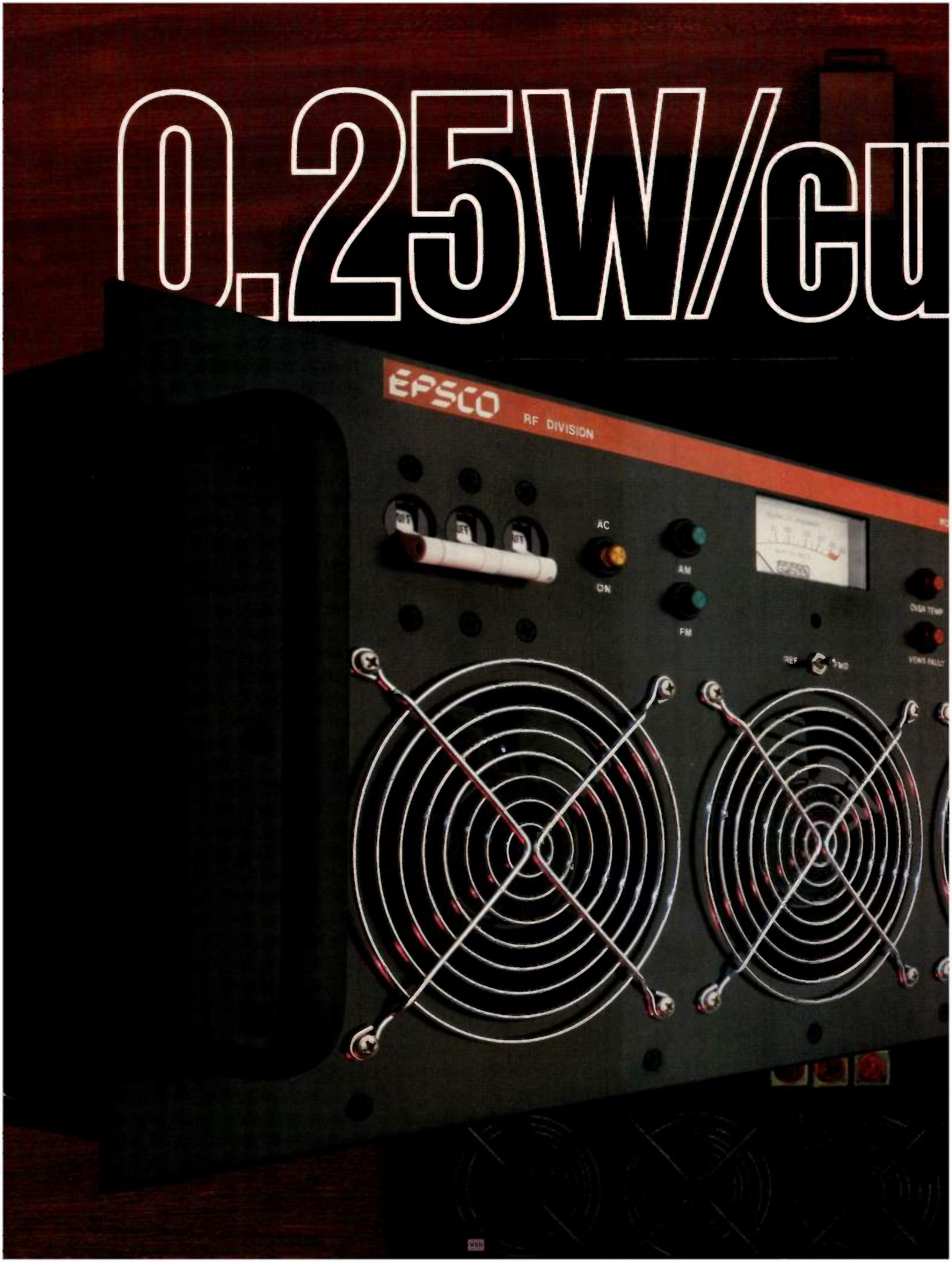
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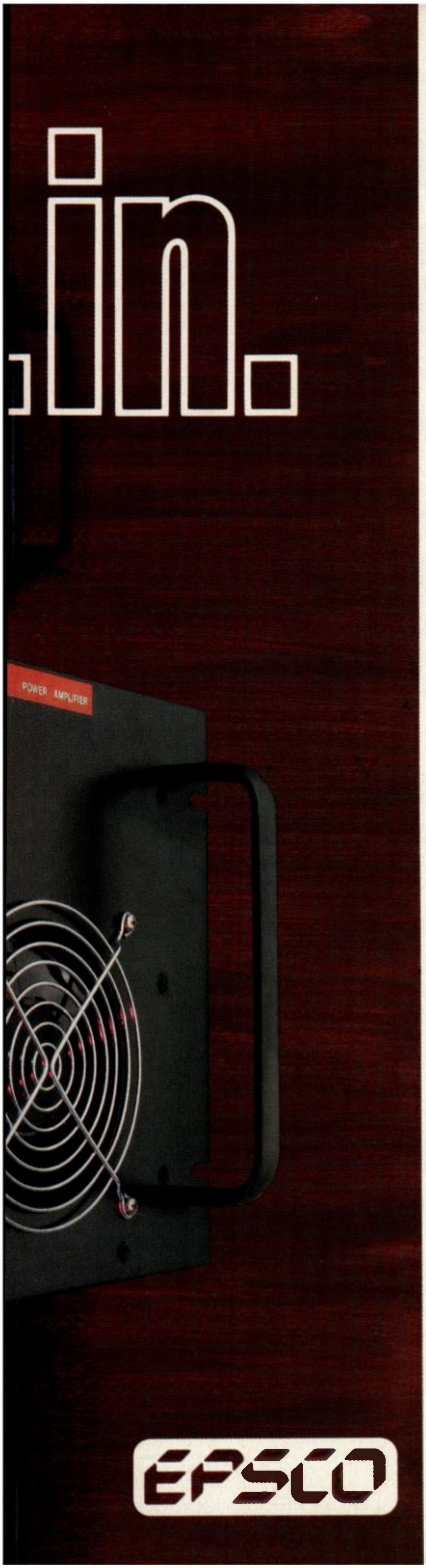


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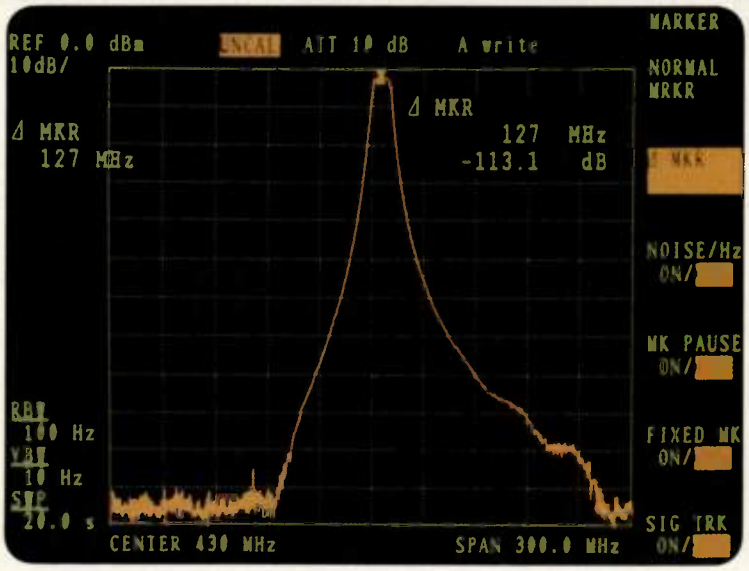


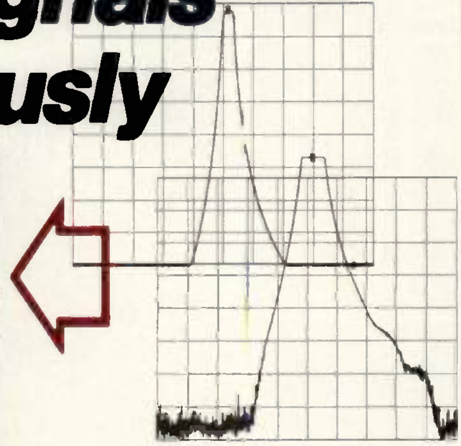

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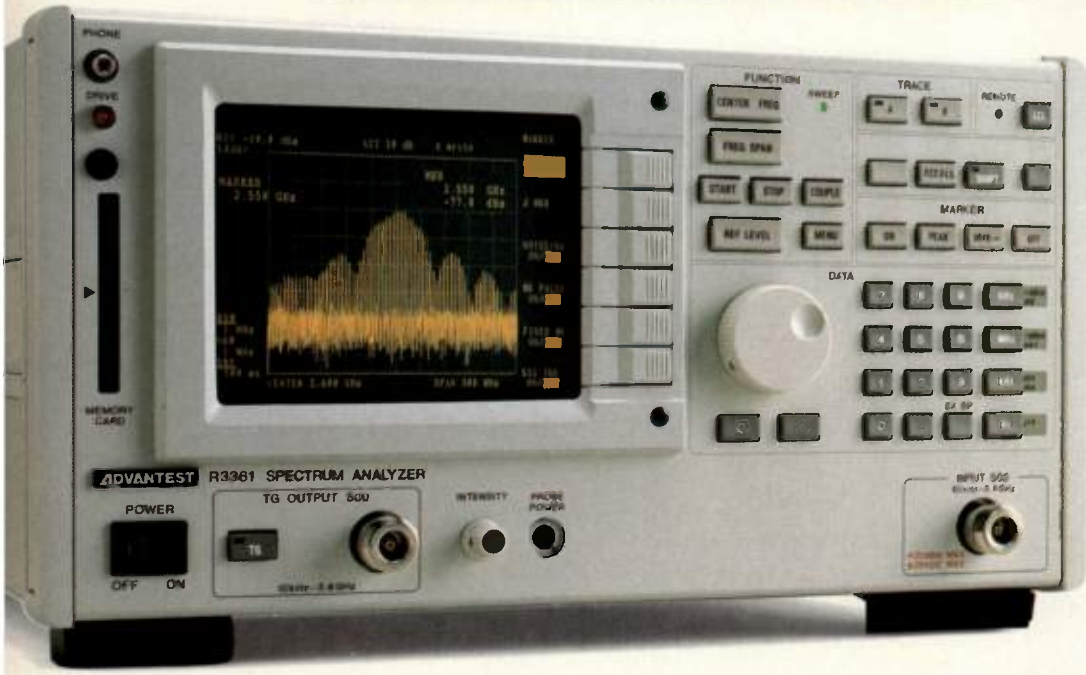


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# *rf letters*

Letters should be addressed to: Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.

## EMI Measurements — The Discussion Continues

Editor:

I am writing in reference to the article "Reflections on EMC Measurements" (October 1988, *RF Design*). The inventory of available spectrum analyzers capable of meeting the measurement requirements of existing regulations has increased in the past year. These spectrum analyzers, including the HP 8568B, have the required functional and data processing capacity.

The problem is not in the hardware, but in the measurement methodology and the available software, including the HP 85864A EMI software, which is still locked into an inefficient Q-P scan method. My paper "A Theory to Optimize the Detection and Measurement of EMI Signals" develops a mathematical basis to ensure detection and a process to ensure that all measurements are made under optimum receiver conditions. The process is a sequence of measurements, each performing separate functions and using data passed from the previous measurement to ensure optimum conditions. This procedure ensures a high probability of detection and accurate measurements. In addition, the sorting and culling process greatly reduces the actual number of measurements required. This method has been incorporated in a program "EMI Measurement Program Version 2.0" [described in "EMI Signal Measurement Automation," January 1989, *RF Design*] which demonstrates the validity of this approach.

In the past, EMC measurements were limited by the technology of the equipment available. Today this is no longer the case, and it is time that we as engineers re-examine our measurement methods in view of the basic physics of the problem and apply the mathematics and sciences necessary to solve the EMC measurement problem.

Roger Southwick
EMC Consulting
Tucson, Arizona

### Author's Reply

I can see how the program described by Mr. Southwick could speed up the measurement process in a closed site (e.g., a screen room or anechoic chamber), but it remains unclear to me how such a system would cope in an open site, where most of the signals being detected are of no concern. If a measurement were taken of the ambient and subtracted in some way from the measurement with the ambient plus the interfering device, there may still be some doubt as to the validity of the measurements. My primary concern is the inability of this modern equipment to be used easily to verify whether a signal is an ambient or not. Physics and mathematics are wonderful tools to use in our search for truth, but if I can't verify the results by some simpler, more basic observation technique, then I would feel uneasy with results generated by such a computer-controlled system, especially in an open site.

Tom Minnis
David Systems
Sunnyvale, California

### Alternative to the Double Lange

Editor:

I read Derek Fitzgerald's "Designing With the Double Lange Coupler" (October 1988, *RF Design*) with considerable interest, but feel obliged to point out that there is an alternative way of achieving in-phase or 180-degree or arbitrary phase hybrids which is simpler (it uses only one Lange coupler), performs better (it has better VSWR, broader bandwidth and lower loss), and is well known. I refer to the arrangement whereby one of the output arms of a single Lange coupler has the appropriate length of 50-ohm transmission line connected to it to achieve the desired phase split.

Figure 1 illustrates the arrangement for a 180-degree hybrid and Figure 2 shows the computed results for this arrangement in comparison with those for the double Lange design using Touchstone™. As can be seen, the 180-degree phase split is maintained over a broader bandwidth and the terminal VSWRs are better over a wider bandwidth as well. The latter effect is the result of ensuring that all ports of the hybrid (internal as well as external) are terminated in matched loads whilst the double Lange has two internal ports terminated in an open circuit. The terminal VSWRs of the double Lange can be improved by replacing the open circuits with 50-ohm loads. Finally, the arrangement of Figure 1 will have lower loss since the loss of a 50-ohm transmission line is less than that of a Lange coupler.

J.L.B. Walker
Thorn EMI Electronics Ltd.
Radar Division
Middlesex, England

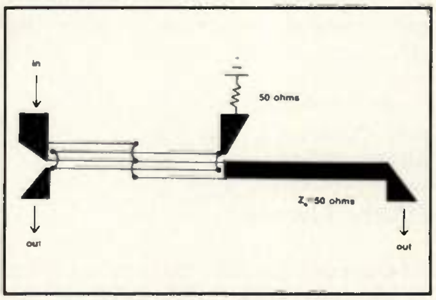



**Figure 1. 180-degree hybrid.**

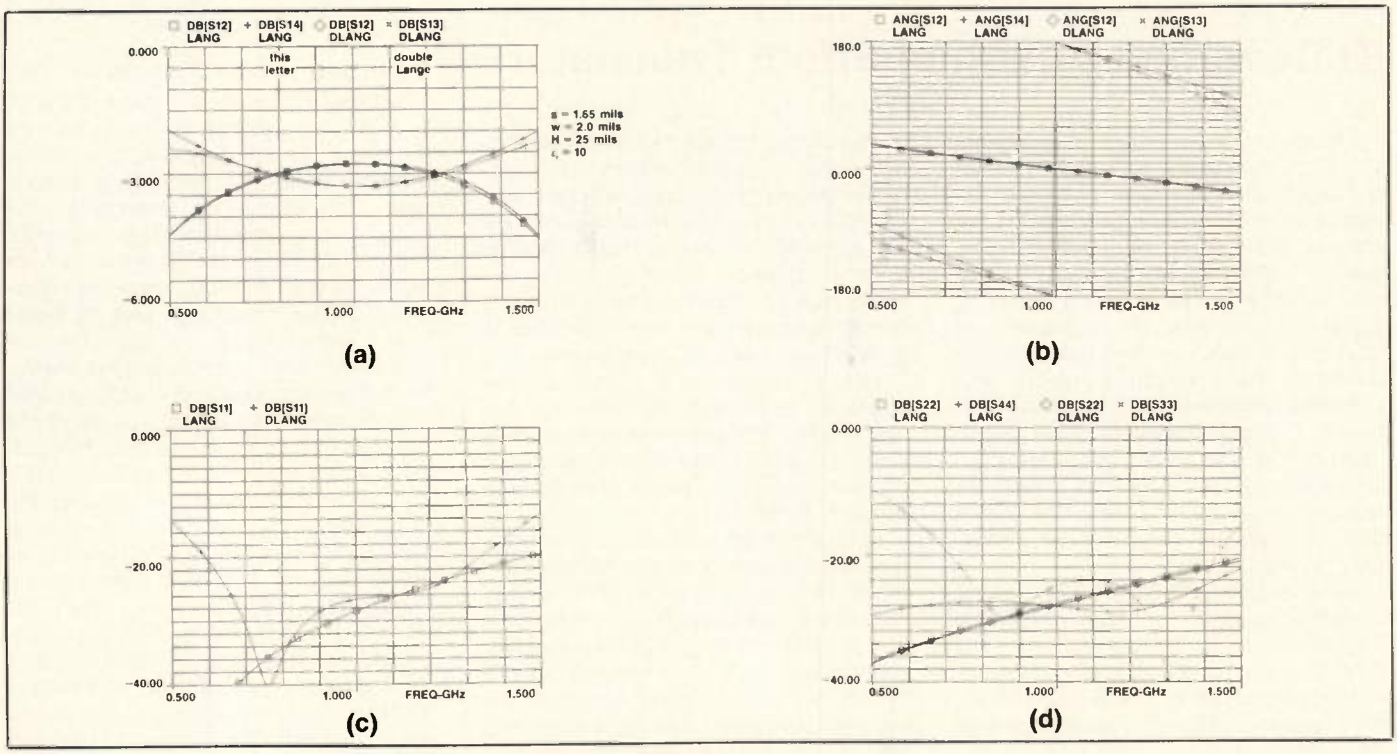



**Figure 2. Comparison with double Lange coupler design.**



INFO/CARD 26

# TI Develops Quantum Effect Transistor

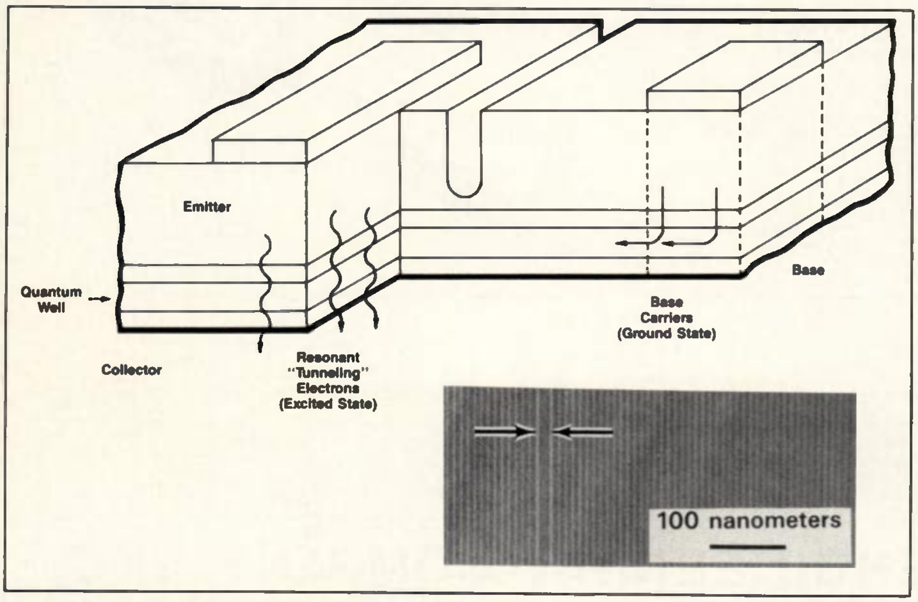
Researchers at Texas Instruments (TI) have fabricated a quantum effect transistor, characterized by critical dimensions 100 times smaller and transit speeds more than 1000 times faster than conventional transistors. The device, called a bipolar resonant tunneling transistor, operates on quantum mechanical effects — principles which dominate the behavior of matter and energy at dimensions of 0.02 micron and below. This new transistor is the first to contact and control a "quantum well" base directly. The quantum well is an ultrathin layer of the device which allows only electrons with certain discrete energies to pass.

Devices integrating quantum effect components such as tunneling diodes into conventional transistors are not new. In such devices, a quantum structure is embedded in one of the transistor's terminals, and the transistor exhibits certain quantum-related characteristics. However, the size and function of these transistors are essentially no different from conventional bipolar ones. In contrast, all essential components of the new TI device are confined to quantum dimensions, and its operation is based fundamentally on quantum effects. According to TI chief technical officer Dr. George Heilmeier, this development signals a significant step towards "a potential next generation of solid-state electronic devices."

The bipolar device has active regions measuring 10 to 20 nanometers wide. At these dimensions, quantum mechanical effects, in which electrons behave more like waves than particles, are the dominating factor. Electrons occupy discrete, non-overlapping energy levels or bands, and resonate when confined to a region the size of their wavelength. These properties are critical to the operation of the quantum effect device, which offers the potential for extremely precise and efficient switching at speeds thousands of times faster than modern semiconductor devices.

In quantum effect devices, the different discrete energy levels characteristic of the various materials in the base and the emitter and collector act as barriers to current flow. Current flows only when voltage applied to the transistor base is modulated so that these energy levels become precisely matched. Electrons can then resonate, enabling them to "tunnel" across the base, and thus provide current flow from emitter to collector.

Possible applications for devices incorporating quantum effect transistors include single-chip supercomputers and realtime image understanding systems, according to TI. With the device still in the laboratory development stage at this time, the transistor's practical applications are estimated to be about ten years in the future.




*TI's true resonant tunneling transistor. Inset is a photomicrograph of the transistor, showing the individually fabricated light regions measuring approximately 10 nm wide.*

**Volt and Ohm Standards Will Change in 1990**—Effective January 1, 1990, the world's industrial nations will share, for the first time, a common practical basis for measuring voltage and resistance. The change is the outcome of a recent meeting in Sevres, France of national representatives from the world's weights and measures community. Representatives met to adopt new "conventional values" for the Josephson and von Klitzing constants, fundamental physical constants required to determine operational values of the volt and the ohm. Before the changes, there was a difference of approximately 1.2 parts-per-million (ppm) between the U.S. voltage standard and that of most European countries purely because of differences in the way the national standards were maintained. With the advent of modern, high-precision voltmeters, such differences have become increasingly significant to U.S. firms seeking to export high-precision equipment. For the United States, the new values mean that the National Institute of Standards and Technology (NIST) will adjust the U.S. voltage standard by about 9.3 ppm, and the U.S. resistance standard by about 1.7 ppm. Precision electrical measuring instruments will have to be adjusted or recalibrated to maintain consistency with the new national standards.

**DoD Confirms Plans for Funding HDTV**—The Department of Defense (DoD) has confirmed its plans to fund development work on High-Definition Television (HDTV) technology as part of its 1990 budget, and has announced a request for research proposals. An estimated $30 million will be spent on the development of high-definition display and display processor technology. It is anticipated DoD's support of HDTV research and development will have a significant impact for commercial producers in the HDTV market.

**Phoenix Monolithics to Acquire MSC**—Phoenix Monolithics of Telford, Pa., has agreed to acquire Microwave Semiconductor Corp. (MSC) from Siemens AG. The agreement includes Phoenix Monolithics' acquisition of the MSC silicon and GaAs product lines, consisting of RF/microwave devices, amplifiers, assemblies and subsystem components. A "gallium arsenide leadership program" in place at MSC will be discontinued by Phoenix Monolithics, with the resultant layoff of nearly 60 employees. The sale is expected to be completed in early 1989.

**Call for Antenna Measurement Techniques Papers**—The 11th Annual Meeting and Symposium of the Antenna Measurement Techniques Association will take place October 9-13, 1989 in Monterey, Calif. Authors are invited to submit abstracts of proposed papers for consideration. Suggested topics include, but are not limited to: advanced antenna measurement techniques, instrumentation and systems; practical aspects of measurement equipment modifications, including hardware and software; systems and equipment interfacing; theory and application of antenna measurement techniques; range design, automation, modification and evaluation; nearfield techniques and their application; radar cross-section measurements; millimeter-wave antenna measurements; anechoic chamber and absorber material design and evaluation; phased array testing; and compact range design and evaluation. Four copies of a 200-word abstract should be submitted by May 5, 1989 to: Dr. Doren Hess, Scientific-Atlanta, Mailstop 28 I, P.O. Box 105027, Atlanta, GA 30348.

**Motorola Sells Clock Oscillator Business**—Motorola has announced the sale of its OEM crystal clock oscillator business to the newly formed Champion Technologies Inc. The new company is headed by William Deutschmann, former sales and marketing manager for Motorola's Components Division. Motorola will continue to manufacture crystals and non-clock crystal products for internal use only. Champion will maintain intact the same OEM clock oscillator products that were offered by Motorola, including TTL, NMOS, CMOS and ECL compatible clock oscillators, voltage-controlled crystal oscillators (VCXOs), and NiCd battery for memory backup applications. Champion plans to bring out its own line of temperature-compensated crystal oscillators (TCXOs) in the first quarter of 1989.

**Rockwell International to Acquire AIL**—Rockwell International has confirmed its decision to acquire Eaton Corp.'s subsidiary AIL. The sale is expected to be finalized in early 1989. AIL, based in Deer Park, N.Y., has been troubled by technical problems and financial losses on its B-1B AN/ALQ-161 defensive avionics system. Rockwell, prime contractor for the B-1B bomber program, hopes that the combining of Rockwell and AIL technical and management resources will help bring the B-1B to its full potential.

**Sokal Named IEEE Fellow**—Nathan O. Sokal, president and founder of Design Automation Inc., has been elected a fellow of the IEEE. The official citation reads: "For his contribution to the technology of high-efficiency power conversion and RF power amplification." Mr. Sokal, inventor of the class-E RF power amplifier, heads a Lexington, Mass., consulting firm working in a wide variety of RF-related areas, including amplifiers, digital signal processing, process control, analog circuit design and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custom design and design review. He received his M.S. in electrical engineering from the Massachusetts Institute of Technology in 1950.

**Free Course Calendar Available From ICT**—A free pullout calendar listing the 1988/1989 dates and locations for their advanced technology and management training courses is being offered by Integrated Computer Systems. The calendar lists 45 courses covering management and business skills, and eight principal technology areas: Software and Systems Engineering; Programming Languages and Operating Systems; Expert Systems and DB Systems; Microprocessors; Digital Processing and Computer Systems; Datacomm and Computer Networks; Local Area Networks; and Communication Systems. A free copy of the calendar can be obtained by contacting: John Valenti, Integrated Computer Systems, 5800 Hannum Avenue, P.O. Box 3614, Culver City, CA 90231-3614. Tel: (800) 421-8166; (213) 417-8888



**ARFTG Call for Papers**—The Automated RF Techniques Group (ARFTG) will hold its 33rd Annual Conference June 15-16, 1989 in Long Beach, Calif., in conjunction with the 1989 International Microwave Symposium. Papers are being solicited which will report on automated test systems, software tools, test techniques and test hardware which have been effective in increasing productivity and product quality and have decreased product manufacturing times. Papers on other topics related to computer-aided RF measurement or design are also welcome. Those interested in submitting a paper should contact: Mark Roos, EIP Microwave Inc., 2731 N. First Street, San Jose, CA 95134. Tel: (408) 433-5900

**Microsemi Acquires TI Military Silicon Discrete Product Line**—Microsemi Corporation (MSC) of Santa Ana, Calif., and Texas Instruments (TI) have signed an agreement for the transfer of TI's military discrete power and small-signal product lines, including radiation-hardened technologies, from TI's Military Semiconductor Division to MSC's Power Technology Components Division. Under terms of the agreement, MSC will receive the rights to utilize TI's process technology to manufacture TI's complete line of military discrete power and small-signal devices. TI will supply process and product specifications, masks and testing instructions, and technical training.

**HP and AT&T in CPU Support Agreement**—Hewlett-Packard Co. (HP) and AT&T Microelectronics have announced an agreement under which they will jointly support and develop development tools for the AT&T WER DSP32C digital signal processor (DSP). According to the agreement, HP will design, market and sell a new real-time DSP emulator, which emulates AT&T's DSP32C microchip. AT&T will provide integrated software-development tools which will be compatible with HP's new emulator, and personal or desk-top computers. In addition, the companies will share technical information to assist in the design of development tools.

# rf calendar

**February 14-16, 1989**
**RF Technology Expo 89**
Santa Clara Convention Center, Santa Clara, CA
Information: Linda Fortunato, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600; (800) 525-9154

**March 1-3, 1989**
**EMC Japan '89**
Sunshine City Convention Center, Tokyo, Japan
Information: Hirotaka Suzuki, Japanese Management Association, 3-1-22, Shiba-Koen, Minato-ku, Tokyo 105, Japan. Tel: (03) 434-1377; Fax: (03) 434-1836; Telex: J25870

**March 6-9, 1989**
**Advanced Materials Conference**
Embassy Suites Hotel, Denver, CO
Information: Dr. Jerome Morse, Advanced Materials Institute, Colorado School of Mines, Golden, CO 80401. Tel: (303) 273-3852

**March 6-9, 1989**
**National Electronic Packaging and Production Conference (NEPCON) West '89**
Anaheim Convention Center, Anaheim, CA
Information: NEPCON WEST '89, P.O. Box 7207, North Suburban, IL 60199-7207. Tel: (312) 299-9311

**March 21-23, 1989**
**3rd European Frequency and Time Forum**
Le Kursaal, Place Granvelle, Besancon, France
Information: A. Remond, 41bis, avenue de l'Observatoire, 25044 Besancon Cedex, France. Tel: 81.80.22.66

**April 4-7, 1989**
**6th International Conference on Antennas and Propagation**
Coventry, England
Information: ICAP 89 Secretariat, Conference Services, IEE, Savoy Place, London WC2R OBL, England.

**April 11-13, 1989**
**1989 IEEE VLSI Test Workshop**
Bally's Park Place Casino Hotel, Atlantic City, NJ
Information: Wesley E. Radcliffe, IBM, East Fishkill, Dept. 277, Bldg. 321-5E1, Hopewell, NY 12533. Tel: (201) 323-2560

**April 11-13, 1989**
**Electro/89**
Jacob K. Javits Convention Center, New York, NY
Information: Electro/89, 8110 Airport Boulevard, Los Angeles, CA 90045. Tel: (800) 421-6816; (800) 262-4208 (in California)

**April 26-28, 1989**
**Aerospace and Defense 89**
Santa Clara Convention Center, Santa Clara, CA
Information: Chuck Jungi, AEA, 5201 Great American Parkway, Santa Clara, CA 95054. Tel: (408) 987-4202

**April 29-May 2, 1989**
**National Association of Broadcasters Annual Convention**
Las Vegas Convention Center, Las Vegas, NV
Information: NAB, Conventions and Meetings, 1771 N Street, N.W., Washington, DC 20036. Tel: (202) 429-5300

# The Gallium Arsenide Trend

## Lower prices, higher levels of integration and more commercial participation will be the highlights for 1989

*By Mark Gomez*
*Technical Editor*

After a slow start that suprised many analysts, gallium arsenide (GaAs) technology now appears to be moving quickly in the RF industry. "As far as we are concerned, there is strong indication that things are going to move very rapidly," said Chuni Ghosh, president of Tachonics Corporation. "We closed out 1988 with a $5.5 million backlog as opposed to $1 million in 1987." He added that at the beginning of 1988, the orders for Tachonics were for 10- to 20-piece quantities and that this number has risen to several hundreds and thousands at the present time. One reason for this is the growing awareness, on the design engineer's part, of the reliability of GaAs MMICs. "Theoretically, there is no reason why MMICs should not be more reliable than hybrids," said Ghosh.

The number of players in the GaAs MMIC market is on the downward trend. "There is evidence of a slight shrinkage of companies that are in and serving the GaAs MMIC marketplace," said Mike Gagnon, director of sales and marketing at Anadigics, "and this is based on the number of companies that have reduced their efforts or are not in operation any longer."

Gagnon sees contracts with nominal quantities for 1989. He notes that as companies introduce more standard catalog products, there is continuance towards a larger number of contracts for nominal quantities of GaAs devices.

For 1989, GaAs products with better performance will surface in the RF industry. "Consumers will see amplifiers with better gain flatness and lower power consumption, higher accuracy attenuators and broader families of absorptive and reflective switches," said Gagnon. As far as new products go, Ghosh sees new limiting amplifiers, VCOs, single chip transceivers and 6-bit digital attenuators being introduced.

The interest in GaAs MMIC technology is at its highest point to date. "There is interesting potential business and new designs going on," noted Louis Pengue, product marketing manager at TriQuint Semiconductor. He noted that the commercial sector is presenting more opportunity for high volume in the short time frame. This viewpoint is also shared by other companies. "There is more drive in the commercial marketplace for doing a variety of different things than has been seen in the military area," said Mike Malbon, vice president of GaAs technology at Avantek. "In the commercial area there is a lot more comfort in using GaAs," he added.

From a marketing stand point, GaAs will have to compete dollarwise with silicon while offering more performance. According to Pengue, you have to be able to sell GaAs as a technology before you can sell your products. "For GaAs to be competitive, you must offer cost parity with greater performance. For example, you can mix signal processing or other digital functions on the same die as the microwave stages," he noted. "A customer will not pay twice the price for twice the performance." As Ghosh sees it, customers are influenced by the greater density that MMICs offer.

The RF and microwave market has always been seen as a somewhat custom market. In the past, Malbon has seen a lot of requests for modifications to standard products. Examples of such requests include higher dynamic range, lower noise figure and more power.

The price on GaAs MMICs is on a downward trend. Doug Lockie, cofounder and manager of special products at Pacific Monolithics, attributes the anticipated price drop to market pressure and lower costs of raw materials. "Three years ago, a 3-inch wafer was priced at approximately $2000. This figure has dropped to a current price of about $100," he noted. According to Paul Schurr, technical marketing manager at Texas Instruments, pricing will fall somewhat over the next twelve months, and considerably over the next two years. "The major issue facing this industry is getting the volume up to move down the learning curve."

The custom market is being served by the companies that provide foundry services. "This market will grow by about $5 million next year," remarked Schurr. The theory that the foundry business will see growth is shared by various other companies as well. "After being in the foundry service market for 18 months, we have over 30 clients," said Gagnon of Anadigics. "This is an excellent alternative for the customer who has internal design skills and wants to design a chip for which he may use only 100 or 500 over a lifetime, but the chip is not available as a catalog product and is not a large enough quantity to get anyone excited to design it as a catalog unit," he added. "Customers like having that blank piece of canvas," said Pengue of TriQuint. "The foundry service market is growing," said Joe Barrera, vice president and general manager of Harris Microwave Semiconductor, "and a lot of it is still exploratory."

GaAs MMICs, like almost any new technology, seem to be taking time to catch on. "Significant volume turn-on is not going to happen until the early to mid-nineties," added Barrera. "This is primarily true in the analog market." He noted that the earlier predictions for GaAs were way off and that it will take another five to seven years to make a real difference.

The gallium arsenice industry has seen several different phases. The future trend seems to be towards higher levels of integration, lower prices and a growing marketplace with more commercial activity, where vendors anticipate a faster turnaround than that possible with the military industry.

The GaAs industry has been slow taking off and this has caused several companies such as Gain Electronics and Microwave Semiconductor Corporation to leave the industry. The remaining players seem to have a solid business established, are key players in the military program with funding, and/or are currently pursuing other niches where GaAs can be used.

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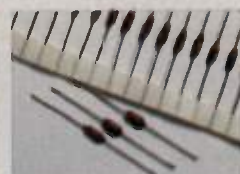

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# rf courses

## The George Washington University

**Electromagnetic Pulse and Its Effects on Systems**
February 27-March 1, 1989, Washington, DC
**Introduction to Receivers**
March 6-7, 1989, London, England
**Electromagnetic Interference and Control**
March 6-10, 1989, Washington, DC
**Modern Receiver Design**
March 8-10, 1989, London, England
**Grounding, Bonding, Shielding and Transient Protection**
March 13-16, 1989, Orlando, FL
**Modern Radar System Analysis**
March 13-17, 1989, Orlando, FL
**Pulsed-Power Technology**
March 27-29, 1989, Washington, DC
**Microwave Radio Systems**
March 30-31, 1989, Washington, DC
**Microwave High-Power Tubes and Transmitters**
April 10-14, 1989, Washington, DC
**Modern Radar System Analysis**
April 10-14, 1989, London, England
**Modern Communications and Signal Processing**
April 17-21, 1989, Washington, DC

Information: Misael Rodriguez, Continuing Engineering Education, George Washington University, Washington, DC 20052. Tel: (800) 424-9773; (202) 994-6106

## Georgia Tech Education Extension

**Antenna Engineering**
February 28-March 3, 1989, Atlanta, GA
**Electronic Support Measures**
March 21-23, 1989, Atlanta, GA
**Radar Cross Section Reduction**
March 21-23, 1989, Atlanta, GA

Information: Education Extension, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel: (404) 894-2547

## Southeastern Center for Electrical Engineering Education (SCEEE)

**Antennas: Principles, Design and Measurements**
March 15-18, 1989, St. Cloud, FL

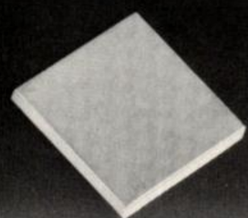
Information: Kelly DuVuyst, SCEEE, 1101 Massachusetts Avenue, St. Cloud, FL 32769. Tel: (305) 892-6146

## UCLA Extension

**Power Electronic Circuits**
March 6-10, 1989, Los Angeles, CA
**Introduction to Automatic Testing and ATE Engineering**
March 13-16, 1989, Los Angeles, CA
**Modern Microwave Techniques**
March 13-16, 1989, Los Angeles, CA

Information: UCLA Extension, P.O. Box 24901, Los Angeles, CA 90024-0901. Tel: (213) 825-1047





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2. **Engineering:** Engineering is the application of technology to solve a problem or meet a design goal. Entrants should clearly identify how their circuit was created in response to such a need. Judges will evaluate performance, practicality, reproducibility and economy.
3. **Documentation:** Communicating ideas to others is the business of *RF Design* and a necessary part of good engineering. Each entry will be judged on its description, analysis and graphical material. Each circuit should have a complete list of components, explanation of functions, and a summary of performance and test data.

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# *rf courses* Continued

## Amador Corporation

**Electromagnetic Emissions: Testing and Design**
March 28, 1989, Minneapolis, MN
**ESD Immunity for Electronic Equipment**
March 29, 1989, Minneapolis, MN

Information: Diane Swenson, Amador Corporation, Wild Mountain Road, Taylors Falls, MN 55084-0270. Tel: (612) 465-3911

## Compliance Engineering

**EMI**
April 25, 1989, Boston, MA
**Safety**
April 26, 1989, Boston, MA
**ESD**
April 27, 1989, Boston, MA
**Telecom**
April 28, 1989, Boston, MA

Information: Compliance Engineering, 271 Great Road, Acton, MA 01720. Tel: (508) 264-4208.

## Integrated Computer Systems

**Fiber Optic Communication Systems**
February 21-24, 1989, San Francisco, CA
March 14-17, 1989, Washington, DC
March 28-31, 1989, San Diego, CA
**C Advanced Programming Techniques and Data Structures**
February 28-March 3, 1989, Washington, DC
April 18-21, 1989, Washington, DC
April 25-28, 1989, Boston, MA
**Digital Signal Processing: Techniques and Applications**
March 7-10, 1989, San Diego, CA
March 14-17, 1989, Boston, MA
April 4-7, 1989, Washington, DC
**C Programming Hands-On Workshop**
March 14-17, 1989, Washington, DC
April 4-7, 1989, Albuquerque, NM
April 11-14, 1989, Washington, DC

Information: John Valenti, Integrated Computer Systems, 5800 Hannum Avenue, P.O. Box 3614, Culver City, CA 90231-3614. Tel: (800) 421-8166; (213) 417-8888

## Interference Control Technologies Inc.

**TEMPEST Design and Measurement**
February 21-24, 1989, Washington, DC
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Information: Penny Caran, Registrar, Interference Control Technologies Inc., State Route 625, P.O. Box D, Gainsville, VA 22056. Tel: (703) 347-0030

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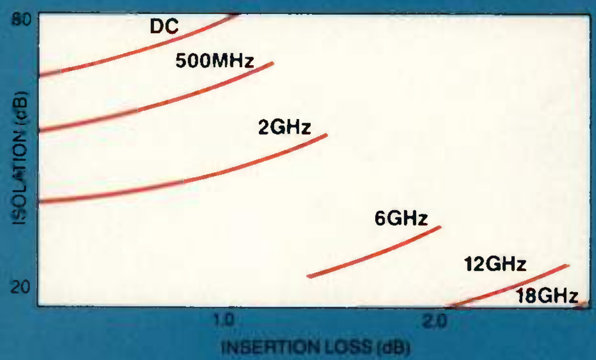


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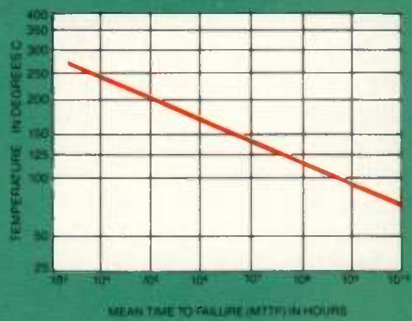


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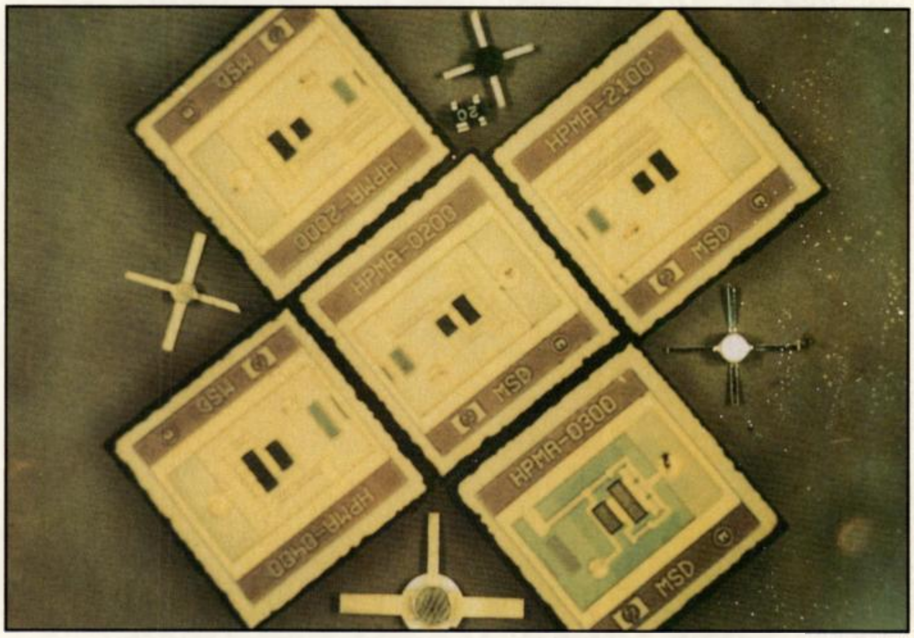
The initial HPMA products consist of five devices, offered in packaged and chip form. The packages include: 1. Bare chips for hybrid applications, 2. The SOT-143 package for SMT manufacturing (three devices are currently available), 3. The hermetic Micro-Plus package, and 4. Two types of Micro-Plastic packages, the "85" with straight microstrip leads, and the "86," with bent leads for surface-mounting. The Micro-Plastic package will be available later this spring. All other packages are currently available.

Gain, bandwidth and power output capability of the family varies, depending on the packaging. The highest gain device is the HPMA-21xx series, with 20 dB gain at 1 GHz. The top power handler is the HPMA-04xx group, with typical 1 dB compression output of 12.5 dBm. A high-performance version of this device on a Beryllium Oxide substrate (HPMA-0420) allows a 10-volt supply to be used for 16.0 dBm output at 1 GHz, with a 4 GHz -3 dB bandwidth. Performance of the chip and Micro-Plastic

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|---|---|---|---|---|---|---|---|
| (Chip Devices) | | | | | | | |
| HPMA-0200 | 7 | 5 | 25 | 12.0 | 6.5 | 4.5 | DC-2700 |
| HPMA-0300 | 7 | 5 | 35 | 12.0 | 6.0 | 10.0 | DC-2400 |
| HPMA-0400 | 7 | 5.25 | 50 | 8.3 | 6.5 | 12.5 | DC-3800 |
| HPMA-2000 | 7 | 5 | 32 | 18.0 | 4.5 | 8.5 | DC-1500 |
| HPMA-2100 | 7 | 5 | 29 | 20.0 | 4.0 | 7.5 | DC-800 |
| (SOT-143 Package) | | | | | | | |
| HPMA-0211 | 7 | 5 | 25 | 12.0 | 6.5 | 4.5 | DC-2600 |
| HPMA-0311 | 5 | 3.8 | 22 | 12.0 | 5.0 | 5.5 | DC-1900 |
| HPMA-2011 | 7 | 5 | 32 | 18.0 | 4.5 | 8.5 | DC-1500 |
| (Plastic Package — xx85 straight-leads, xx86 bent leads) | | | | | | | |
| HPMA-0285/86 | 7 | 5 | 25 | 12.0 | 6.5 | 4.5 | DC-2200 |
| HPMA-0385/86 | 7 | 5 | 35 | 12.0 | 6.0 | 10.0 | DC-2600 |
| HPMA-0485/86 | 7 | 5.3 | 50 | 8.0 | 7.0 | 12.5 | DC-2800 |
| HPMA-2085/86 | 7 | 5 | 32 | 18.0 | 4.5 | 8.5 | DC-1500 |
| HPMA-2185/86 | 7 | 5 | 29 | 20.0 | 4.0 | 7.4 | DC-800 |

**Table 1. Specifications for HPMA amplifiers in three of the available packages.**

devices is summarized in Table 1. Typical performance curves (gain and 1 dB compression power output) are shown in Figure 2. The curves clearly illustrate the gain performance of the HPMA-21xx family, and the power output characteristics of the HPMA-04xx devices.

The advantages of 50-ohm matched amplifiers are well known. Design time, board space, and manufacturing costs can all be reduced by using these simple devices. With the introduction of a new line, Hewlett-Packard gives RF engineers a greater choice of available components, plus ongoing, experienced applications support.

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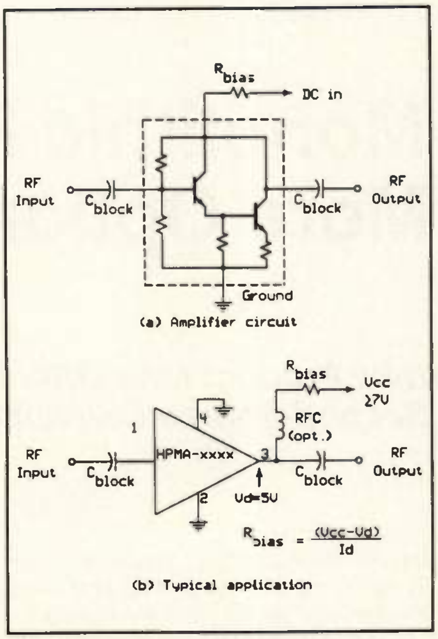


Figure 1. Circuit diagram and external connections for HPMA series amplifiers.

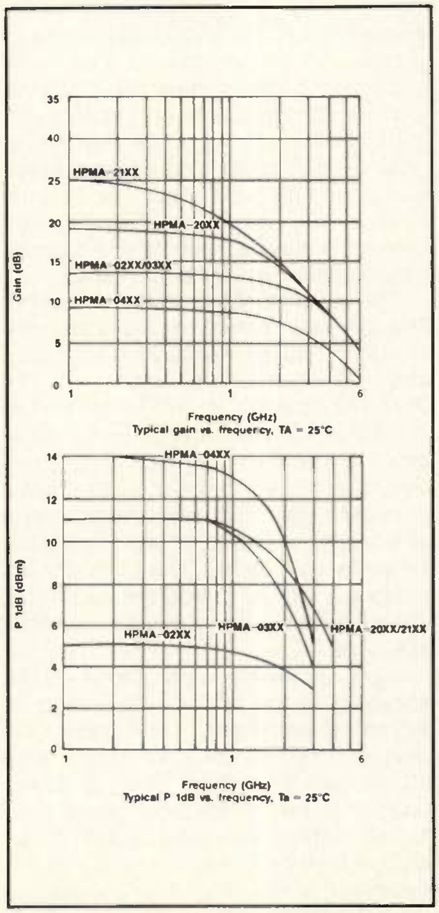


Figure 2. Typical performance curves for the HDMA amplifier series.

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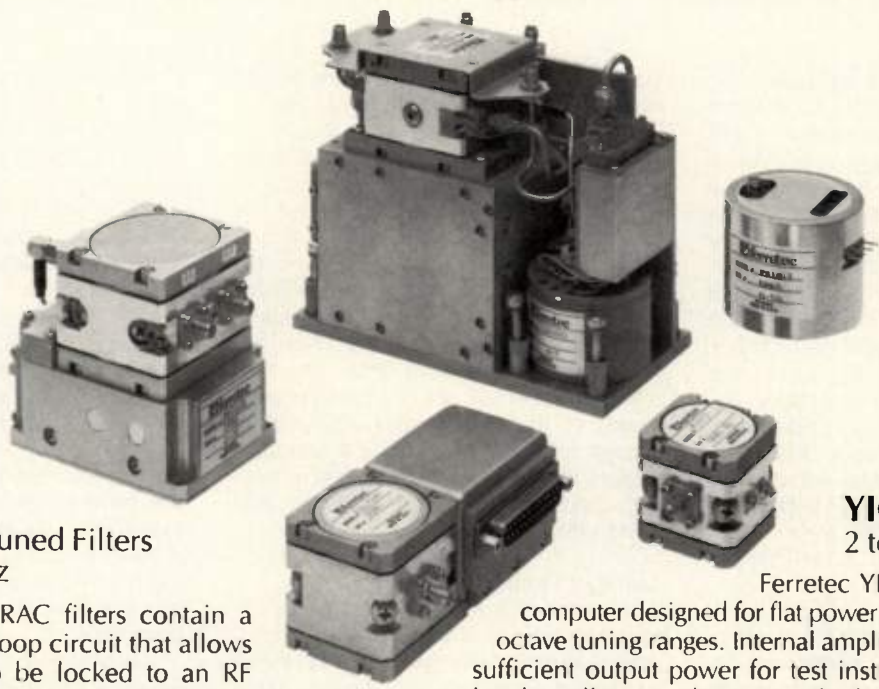

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## *rf featured technology*

# Low-Noise VHF and L-Band GaAs FET Amplifiers

*By Al Ward*
*Avantek Inc.*

*GaAs FET devices are typically used in low-noise amplifiers in the microwave region, where silicon transistors can't provide the required gain and noise performance. There are, however, many applications in the frequency range below 2000 MHz where the low noise figures and high gain of GaAs FETs can improve receiver sensitivity. This article describes a series of three low-noise amplifiers that use identical circuit topology. The only differences are in the proper choice of three inductors depending on the frequency of operation. The designs are centered at 450 MHz, 900 MHz and 1300 MHz, but can be scaled for any frequency within the region of 400 to 1600 MHz. Each amplifier has a usable bandwidth of about 30 to 40 percent.*

Using a high-gain, high-frequency GaAs FET at VHF poses special problems. Of greatest concern is the problem of designing the amplifier for unconditional stability. Typically, GaAs FETs have greater gain as frequency is decreased, e.g., 25 dB maximum stable gain at 500 MHz. A second problem is that matching the typical microwave GaAs FET at lower frequencies for minimum noise figure does not necessarily produce minimum input VSWR.

Achieving the lowest possible noise figure requires matching the device to $\Gamma_{opt}$ (the source match required for minimum noise figure). At higher microwave frequencies this will generally produce a reasonable input VSWR, since $\Gamma_{opt}$ and the complex conjugate of the device input reflection coefficient $S_{11}$ are usually close on the Smith Chart. At lower frequencies, special consideration needs to be given to the input circuit design and to the tradeoffs required to ensure low noise figure while still achieving moderate gain, low VSWR and unconditional stability.

### Design Technique

The Avantek ATF-10135, supplied in the commercial 0.085 inch "micro-X" metal/ceramic package, is used in these examples. Examination of the data sheet reveals that the device is capable of 0.4 dB noise figure at frequencies below 2 GHz with an associated gain of greater than 15 dB. The noise parameters and S-parameters of this transistor are summarized in Table 1.

Achieving the associated gain of which the device is capable is difficult since the device is not inherently stable. It is not enough that the amplifier be stable at the operating frequency — it must be stable at all frequencies. Any out-of-band oscillation will make the amplifier unusable.

The simplest technique to ensure broad-band stability is to resistively load the drain. Resistive loading produces a constant impedance on the device over a wide frequency range. In the case of the ATF-10135, a 47 ohm carbon resistor is used to load the output of the device, with the series inductance from the resistor leads also used to better match the device to 50 ohms. This produces acceptable gain while ensuring a good output match and retaining stability over as wide a bandwidth as possible.

Obtaining the lowest possible noise figure from the device requires that the input matching network convert the nominal 50 ohm source impedance to $\Gamma_{opt}$. This produces a deliberate impedance mismatch that, while minimizing amplifier noise figure, produces a high input VSWR. The ideal situation is

ATF-10135

Typical Scattering Parameters, Common Source, Vds=2V, Ids=20mA

| Frequency GHz | S11 Mag | S11 Ang | S21 dB | S21 Mag | S21 Ang | S12 dB | S12 Mag | S12 Ang | S22 Mag | S22 Ang |
|---|---|---|---|---|---|---|---|---|---|---|
| 0.5 | .98 | -18 | 14.5 | 5.32 | 163 | -34.0 | .020 | 78 | .35 | -9 |
| 1.0 | .93 | -33 | 14.3 | 5.19 | 147 | -28.4 | .038 | 67 | .36 | -19 |
| 2.0 | .79 | -66 | 13.3 | 4.64 | 113 | -22.6 | .074 | 59 | .30 | -31 |

Noise Parameters, Vds=2V, Ids=20mA

| Frequency * GHz | NF dB | Gamma Opt Mag | Gamma Opt Ang | Rn/50 |
|---|---|---|---|---|
| 0.45 | 0.35 | .93 | 12 | .80 |
| 0.90 | 0.36 | .87 | 21 | .70 |
| 1.30 | 0.38 | .81 | 31 | .62 |
| 2.00 | 0.40 | .70 | 47 | .46 |

* Noise parameters at frequencies below 2 GHz are extrapolated from measured data at higher frequencies.

**Table 1. Scattering and noise parameters for the Avantek ATF-10135 transistor, common source, $V_{DS}$= 2V, $I_{DS}$= 20mA.**

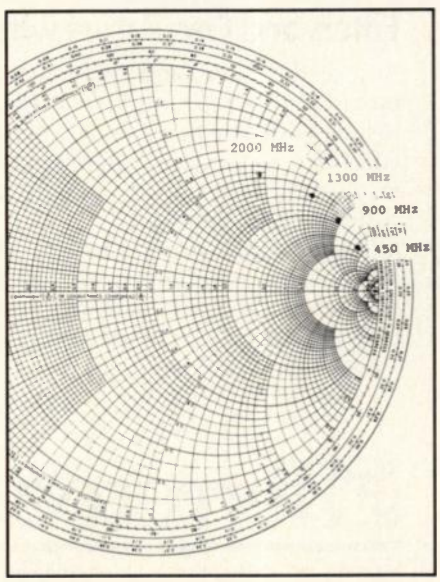


**Figure 1. ATF-10135 $\Gamma_{opt}$ vs. frequency, $V_{DS}$ = 2V and $I_{DS}$= 20mA.**

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| Lead Length | NF | Gain | $\|S_{11}\|^2$ | $\|S_{22}\|^2$ | k | k (@ 9 GHz) |
|---|---|---|---|---|---|---|
| 0 inch | 0.46 dB | 20.1 dB | -5.4 dB | -8.0 dB | .75 | 2.79 |
| 0.1 inch | 0.48 dB | 17.2 dB | -14.3 dB | -16.0 dB | 1.30 | 1.94 |
| 0.2 inch | 0.52 dB | 14.9 dB | -10.6 dB | -18.4 dB | 1.55 | 0.92 |

**Table 2. Performance vs. source lead length at 900 MHz.**

| Frequency | | NF | Gain | $\|S_{11}\|^2$ | $\|S_{22}\|^2$ | k |
|---|---|---|---|---|---|---|
| 450 MHz | (Measured): | 0.45 dB | 21.3 dB | -8.0 dB | -7.6 dB | -- |
| | (Simulated): | 0.54 dB | 20.6 dB | -9.9 | -7.0 | 1.161 |
| 900 MHz | (Measured): | 0.40 dB | 16.5 dB | -10.7 dB | -14.4 dB | -- |
| | (Simulated): | 0.48 dB | 17.2 dB | -14.3 dB | -16.0 dB | 1.30 |
| 1300 MHz | (Measured): | 0.50 dB | 14.5 dB | -8.5 dB | -17.8 dB | -- |
| | (Simulated): | 0.45 dB | 15.7 dB | -13.0 dB | -19.3 dB | 1.20 |

**Table 3. Measured performance vs. computer simulation.**

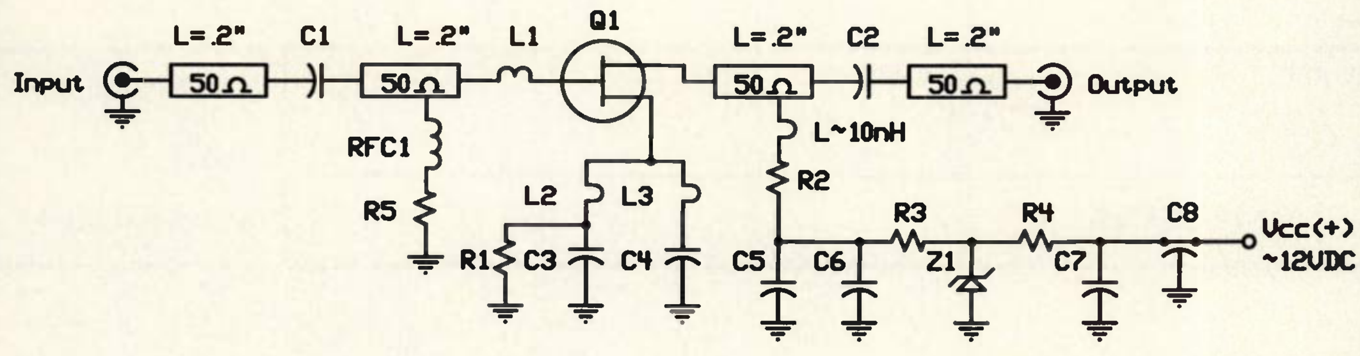
where $\Gamma_{opt}$ is the complex conjugate of $S_{11}$ (i.e., $S_{11}^*$). For this condition, minimum noise figure is achieved when the device is matched for minimum VSWR. This situation occurs predominantly above 2 GHz and tends to diverge at lower frequencies, where $S_{11}$ approaches 1.

High input VSWR has varying significance, depending on the application. Most noteworthy is the increased uncertainty of the noise figure measurement due to reflections between the noise source and amplifier input. Similarly, when the amplifier is connected to a receive antenna, high input VSWR creates added uncertainty in overall system performance. The effect is difficult to analyze unless an isolator is placed at the input to the amplifier. The use of an isolator, however, adds excessive loss and, at VHF frequencies, the size of the isolator is often prohibitively large. In the case of systems using pulse position modulation (PPM), reflections due to VSWR manifest themselves as displaced pulses, which create direction-finding errors.

To examine the alternatives, constant noise figure and constant gain circles can be constructed to assess the impact of trading increased noise figure for a decrease in input VSWR and a corresponding increase in amplifier gain. In most instances, the result is a much higher noise figure than really desired.

An option is to use source feedback. This subject has already been covered by several authors (References 1-3). Source feedback, in the form of source inductance, can improve input VSWR with minimal noise figure degradation. The drawback of utilizing source inductance is a gain reduction of up to several decibels. However, GaAs FET devices often have more gain than desired at low frequencies, so the penalty is not severe.

The effect of source inductance on amplifier input match is best studied with the help of a computer simulation. The computer was used to analyze $S_{11}$ of the amplifier with the proposed output matching network. $S_{11}$ was measured looking directly into the gate of the device with the source inductance added between the source and ground. With the Avantek ATF-10135 at 500 MHz, adding the




$I_D$ = 20 mA
$V_D$ = 1 V @ 450 and 900 MHz, 1.4 V @ 1300 MHz

Components common to all versions:

| | |
|---|---|
| Q1: | Avantek ATF-10135 low-noise GaAs FET |
| C1, C2: | 100 pF, 0.050-inch square chip capacitor |
| C3, C4: | 1000 pF, 0.050-inch square chip capacitor |
| C5: | 1000 pF, 0.110-inch or 0.050-inch square chip capacitor |
| C6: | 0.1 μF, 0.110-inch square chip capacitor |
| C7: | 1 μF electrolytic capacitor |
| C8: | 1000 pF feedthrough capacitor |
| L2, L3: | Source inductance: lead length on Q1 per table |
| R1: | 18-33Ω, 1/4-W carbon resistor--adjust for $I_D$ ≈ 20 mA |
| R2: | 47Ω, 1/4-W carbon resistor with 0.15" leads, tap 0.1" from Q1 |
| R3: | 68-120Ω, 1/4-W carbon resistor--adjust for $V_{DS}$ |
| R4: | 220Ω, 1/4-W carbon resistor |
| R5: | 100Ω, 1/4-W carbon resistor |
| Z1: | 1N752, 5.6-V Zener diode |

Components changed with frequency of operation:

| Frequency | 450 MHz | 900 MHz | 1300 MHz |
|---|---|---|---|
| RFC1: | 0.47 μH | 0.33 μH | 0.33 μH (min. molded RFC) |
| L1: | 8T #28, 1/8" I.D. close-wound | 3T #28, 1/8" I.D. ←Spaced 1 wire diameter→ | 3T #28, 0.050" I.D. ←Spaced 1 wire diameter→ |
| L2, L3 | 0.125" | 0.10" | .065" |

50Ω microstriplines are .080" wide on RT/Duroid 5880, .031"-thick

**Figure 2. Schematic of the GaAs FET amplifier circuit. The only change made to modify the operating frequency range from 400 to 1500 MHz is changing the values of RFC1 and L1.**

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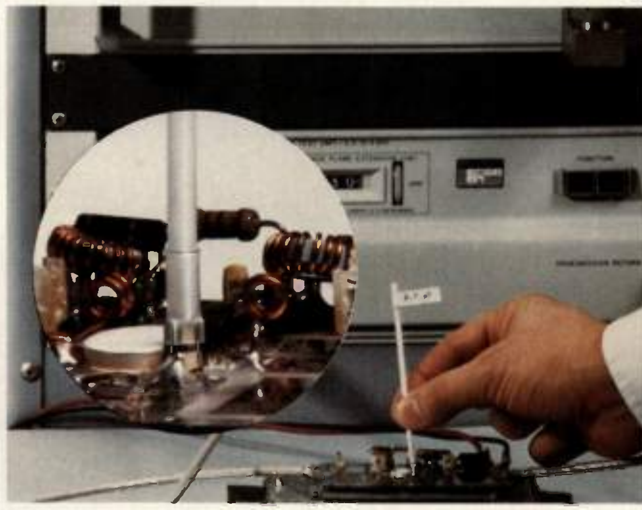

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| CAPACITY VALUE (pf) | TOLERANCE |
|---|---|
| 4.7<br>5.6 | C |
| 6.8<br>8.2<br>10<br>47<br>68<br>100 | J |
| 470<br>1000 | K |

TOLERANCE CODE: B = ±0.1 pf; C = ±0.25 pf; J = ±5%; K = ±10%

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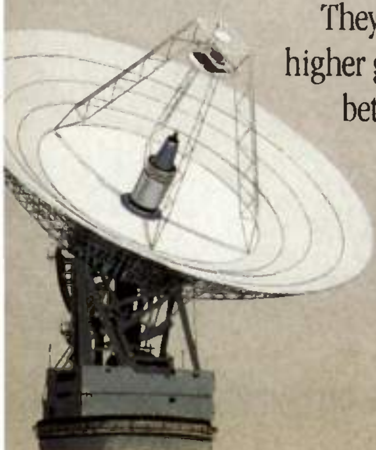

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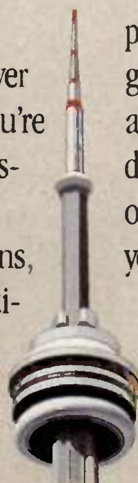

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equivalent source inductance of two 0.10 inch leads causes the value of $S_{11}$ to decrease from 0.987 to 0.960. Angle remains relatively constant at about -16 degrees. Comparing $S_{11}$ to $\Gamma_{opt}$ at 500 MHz now shows them to be nearly identical. The result is that minimum noise figure and minimum VSWR will coincide more closely with one another when matching the device for minimum noise figure. Plotting $\Gamma_{opt}$ for the ATF-10135 device from 450 MHz through 2 GHz in Figure 1 shows that $\Gamma_{opt}$ lies very near the $R/Z_o=1$ curve. This implies that a series inductance will provide the necessary match to attain minimum noise figure.

The simplest way to incorporate source inductance is to use the device source leads. Using device leads as inductors produces approximately 1.3 nH per 0.100 inch of source lead, or 0.65 nH for two source leads in parallel. With the help of Touchstone™, the effect of the lead inductance can be analyzed by simulating the inductance as a high-impedance transmission line. The TUNE mode was invaluable for determining the optimum lead length for a given performance. Table 2 shows the effect of lead length on gain, noise figure, stability, and input and output VSWR at 900 MHz. It is clear that lead lengths of 0.1 inch or less have a mincr effect on noise figure while improving input match substantially. Gain does suffer, but this is not a major concern.

An added benefit of using source inductance is enhanced stability as evidenced by the Rollett stability factor, K. Excessive source inductance can have an adverse effect on stability at the higher frequencies. In the case of the 900 MHz amplifier, zero length source leads create potential instability at low frequencies while longer source lead length creates a potential instability at high frequencies; a 9 GHz, 0.100 inch source lead length is an optimum choice based on all parameters. The optimum source lead length varies with frequency of operation. In the case of the 450 MHz model, 0.125 inch lead length provides the best overall performance with K>1 at frequencies of 450 MHz and higher. According to Touchstone™, low frequency stability can be enhanced with 0.200 inch lead length at a penalty of 2.5

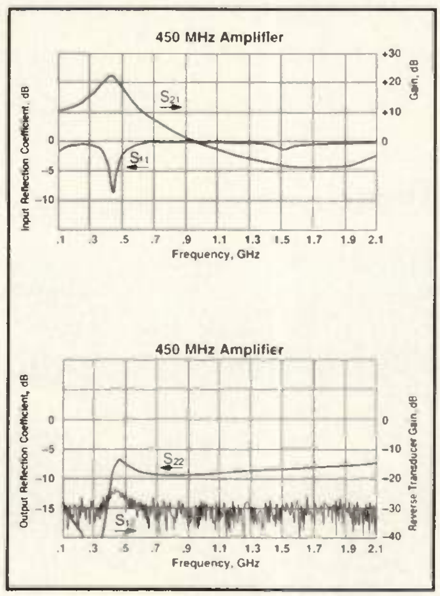


**Figure 3. Swept performance of the amplifier circuit with 450 MHz component values.**

dB of inband gain. For the 1300 MHz model, 0.065 inch source lead length provides optimum performance with unconditional stability up to 11 GHz. Decreasing source lead length improves stability at 12 GHz while making K<1 at 400 MHz. For applications below 450 MHz, greater source inductance will no doubt be required to retain K>1 and to obtain a reasonable input VSWR.

The amplifier circuit actually built is shown in Figure 2. For simplicity, the FET is self-biased. The loss associated with the bypassed source resistor is no greater than 0.1 dB at these frequencies. Zener diode regulation worked well. Although there is interaction, the source resistor, R1, primarily sets the drain current while R3 sets the drain voltage. Improved regulation over temperature is possible with any of the popular active bias networks discussed in References 4 and 5. The active bias network sets both the drain voltage and drain current regardless of device variations.

### Measurements on Amplifiers

The performance of all three amplifiers is comparable to that predicted by the computer simulation. Table 3 summarizes the gain, noise figure and VSWR parameters. The actual noise figure is within 0.1 dB, and the gain within 1.2 dB of the prediction. The VSWR performance is not as good as predicted by the simulation, but still very acceptable. Stability is very good with no problems noted when cascading stages.

The swept gain plots (included in Figures 3-5) show the wide bandwidth of these amplifiers. Low noise figure is also retained over the bandwidths. The 450 MHz amplifier has less than 0.5 dB noise figure between 400 and 500 MHz while the 900 MHz amplifier has less than 0.6 dB noise figure between 800 and 1000 MHz. Similarly, the 1300 MHz amplifier has less than 0.65 noise figure from 1200 MHz to 1500 MHz.

At frequencies above 2 GHz, the ATF-10135 is rated for minimum noise figure when operated at $V_{DS}$ of 2 V, and $I_{DS}$ of 20 mA. At frequencies below 2 GHz, it was empirically determined that

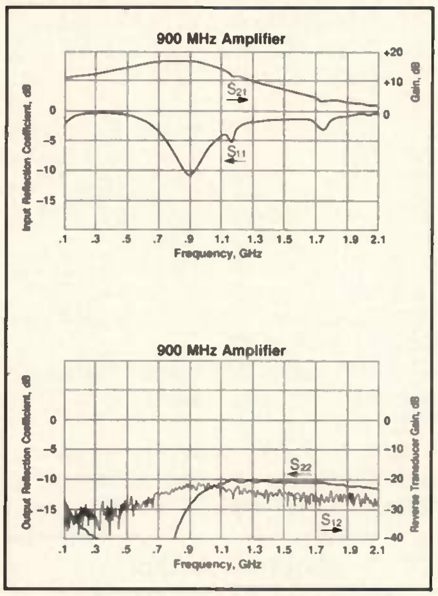


**Figure 4. Swept performance of the amplifier circuit with 900 MHz component values. (See Appendix 1 for the computer modeling of the circuit at 900 MHz.)**

an additional 0.1 dB reduction in noise figure is possible if the device is re-biased. At 1300 MHz, the optimum $V_{DS}$ is 1.4 V, while at 450 and 900 MHz, 1 V gave the lowest noise figure.

The amplifiers were built on RT/Duroid™ 5880 dielectric material of 0.031 inch thickness. The 50 ohm microstriplines are 0.080 inch in width. Epoxy-glass dielectric material should also work if the microstripline widths are properly scaled.

The most critical factor in construction is assuring proper ground returns for the bypass capacitors. In this prototyping work, grounds for the bypass capacitors are obtained by using 0.1-inch-wide "z" wires to connect the top groundplane to the bottom groundplane. Plated through-holes (vias) are typically used in high-volume production.

For enclosures, standard Hammond or Bud diecast aluminum boxes are appropriate; the microstripline board fits nicely into the lid of a Hammond 1590A diecast aluminum box. Flange mount SMA- type connectors are suggested to ensure mechanical rigidity. The connector mounting hardware is used to provide a good mechanical and RF connection between the connector, the box and the groundplane side of the microstripline board.

### Amplifier Tuning

The inherent broad bandwidth of these amplifiers drastically reduces the time required to get them into operation. Setting up each amplifier is simple.

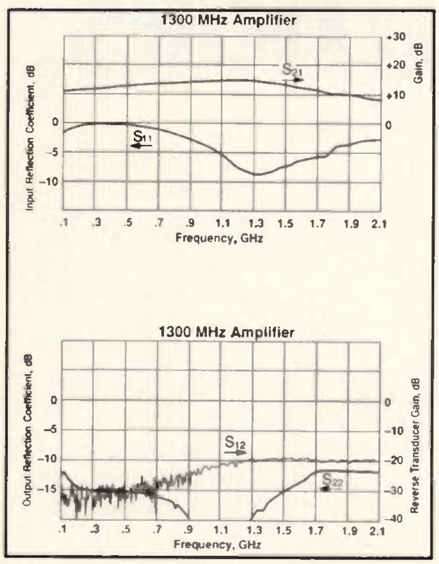


**Figure 5. Swept performance of the amplifier circuit with 1300 MHz component values.**

Once the device is set up for the proper DC operating parameters for the frequency of interest, noise figure and gain performance should be comparable to that shown in Table 3. If necessary, adjust the turns spacing on the input inductor for the desired input VSWR. This will automatically coincide with minimum noise figure and maximum gain. As shown in the foregoing data, the noise figure varies very little over a wide bandwidth, so it might be advantageous to tune for minimum input VSWR as opposed to noise figure. Without the source inductance, the input VSWR will be considerably higher.

The simple series L/R matching network in the output circuit forces a good broadband low VSWR output match. Due to the finite amount of reverse isolation of the device, the output match is affected by the input match and vice versa. Therefore the frequency of best output VSWR is somewhat dependent on where the input network is optimum.

### Using the Design at Other Frequencies

The basic amplifier design can be adapted for any frequency in the 400 to 1600 MHz range. Merely scaling the input inductor for the desired frequency will allow operation on a different frequency. The graph shown in Figure 6 gives some idea of the relationship of L vs. frequency — source feedback should be adjusted accordingly. The ATF-10135 has been used successfully in circuits operating at as low as 150 MHz with similar results.

### Conclusion

The results show that high-frequency GaAs FETs can be used very successfully in the 400 to 1600 MHz frequency range. This same technique can be used down to 150 MHz and up to 1.7 GHz with similar results. Conventional microstripline matching techniques will still offer the best performance above 1.76 GHz.

The single-element match in the input network provides very good performance in this frequency range and offers the greatest bandwidth. There is no doubt that noise figure and input VSWR performance can be further enhanced by the use of a two-element matching network. A shunt capacitor can be used on the device side of the input inductor, but this may necessitate additional tuning for very little improvement in performance. Present noise figure performance is already within 0.1 dB of that specified in the data sheet.

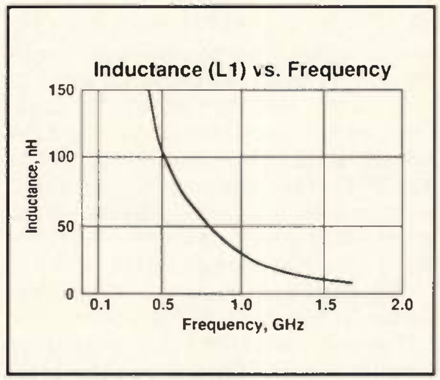


**Figure 6. Inductance of L1 vs. frequency in the amplifier circuit.**

Improved VSWR performance at the expense of increased noise figure can be achieved by a further increase in source inductance. Overall amplifier performance is best analyzed with the help of the computer.

When using any GaAs FET in the VHF region, it becomes even more difficult to obtain broadband stability due to the high gain available from the device. For this reason, a broadband resistive load was chosen for the output network as opposed to the typical L/C tank circuit. Some gain is sacrificed for the added benefit of increased stability.

[Appendix 1 follows on p. 48]

### References

1. L. Besser, "Stability Considerations of Low Noise Transistor Amplifiers With Simultaneous Noise and Power Match," *Low Noise Microwave Transistors and Amplifiers*, H. Fukui, Editor, IEEE Press, 1981, pp. 272-274.
2. G.D. Vendelin, "Feedback Effects on the Noise Performance of GaAs MESFETs," *Low Noise Microwave Transistors and Amplifiers*, H. Fukui, Editor, IEEE Press, 1981, pp. 294-296.
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4. G. Gonzalez, *Microwave Transistor Amplifiers*, Prentice-Hall 1984, pp. 131-133.
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### About the Author

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INFO/CARD 41

```
Touchstone (TM) - Configuration( 100 1500 100 16437 2364 1000 1 3294 :
              900AMP.CKT     Thu Dec 29 15:19:38 1988
DIM
        FREQ  GHZ
        IND   NH
        CAP   PF
        LNG   IN
VAR
        LL=0.1
        CC=1000

CKT
        MSUB   ER=2.1  H=.031  T=.001  RHO=1  RGH=0
        MLIN  1   2   W=.080  L=.2
        SLC   2   3   L=.5  C=100
        MLIN  3   4   W=.080  L=.2
        IND   4   5   L\40
        IND   4   6   L=330
        RES   6   0   R=100
        DEF2P 1   5   NAIN

        S2PA  5   8   9   A:\S_DATA\10135N.S2P
        DEF3P 5   8   9   NA2P

        MLIN  9   10   W=.020  L^LL
        MLIN  9   11   W=.020  L^LL
        SLC   11  12   L=.5  C^CC
        RIBBON 12  0   W=.1  L=.031  RHO=1
        SLC   10  13   L=.5  C^CC
        RIBBON 13  0   W=.1  L=.031  RHO=1
        RES   10   0   R=47
        DEF1P 9   NASER

        MLIN  8   14  W=.080  L=.1
        IND   14  15  L=10
        RES   15  16  R=47
        SLC   16  0   L=.5   C=1000
        SLC   16  0   L=.5   C=10000
        MLIN  14  17  W=.080  L=.1
        SLC   17  18  L=.5   C=100
        MLIN  18  19  W=.080  L=.2

        DEF2P 8   19  NAOUT

        NAIN  1   5
        NA2P  5   8   9
        NASER 9
        NAOUT 8   19
        DEF2P 1   19  AMP

FREQ
       !STEP  .900
        SWEEP 0.100  12.0   .2
       !STEP   9.0

OUT
        AMP  DB[S11]
        AMP  DB[S22]
        AMP  DB[S21]
        AMP  DB[S12]
        AMP  DB[NF]
        AMP  K

OPT
        AMP  DB[NF]<.2
       !AMP  DB[S21]>25
```

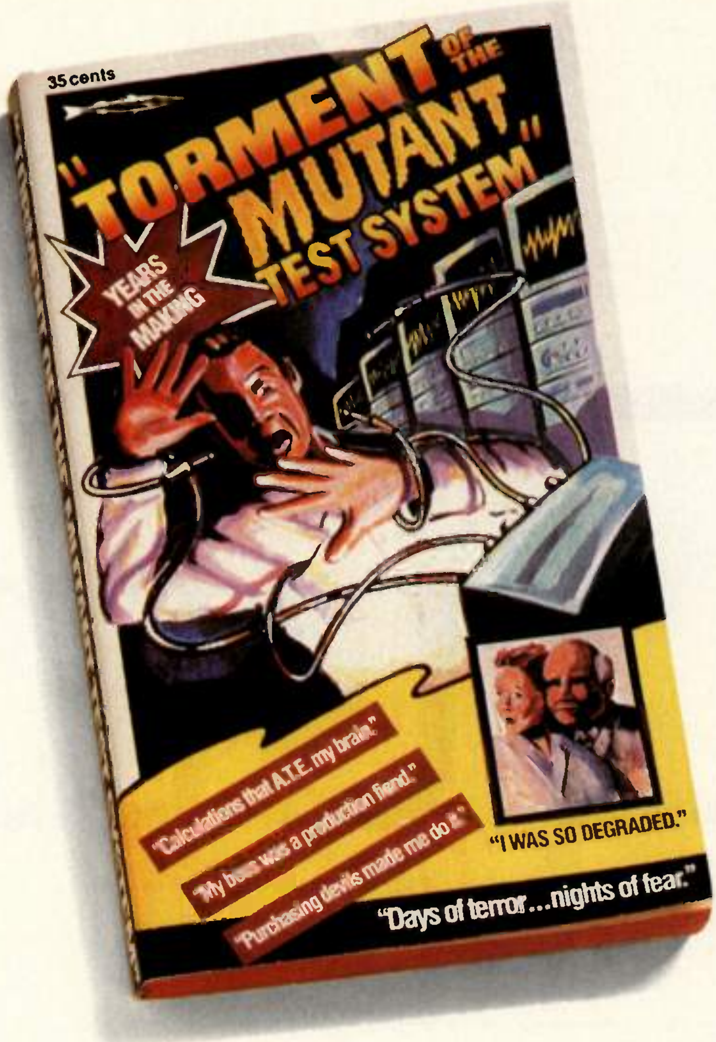
Touchstone (TM) - Configuration( 100 1500 100 16437 2364 1000 1 3294 )
900AMP.OUT Thu Dec 29 15:47:35 1988 LL = 0.0 inch.

| FREQ-GHZ | DB[S11] AMP | DB[S22] AMP | DB[S21] AMP | DB[S12] AMP | DB[NF] AMP | K AMP |
|---|---|---|---|---|---|---|
| 0.10000 | -1.524 | -11.737 | 9.624 | -49.925 | 3.093 | 14.219 |
| 0.30000 | -0.172 | -14.979 | 11.608 | -40.835 | 2.243 | 0.667 |
| 0.50000 | -0.366 | -15.860 | 14.535 | -33.837 | 1.769 | 0.571 |
| 0.70000 | -2.267 | -14.581 | 19.363 | -26.114 | 0.906 | 0.681 |
| 0.90000 | -5.390 | -8.046 | 20.141 | -23.061 | 0.458 | 0.755 |
| 1.10000 | -1.162 | -8.603 | 13.754 | -27.446 | 0.994 | 0.834 |
| 1.30000 | -0.501 | -8.961 | 9.020 | -30.305 | 2.500 | 0.901 |
| 1.50000 | -0.287 | -8.887 | 5.603 | -32.024 | 4.413 | 0.941 |
| 1.70000 | -0.189 | -8.665 | 2.926 | -33.147 | 6.302 | 0.965 |
| 1.90000 | -0.135 | -8.397 | 0.721 | -33.919 | 7.999 | 0.980 |
| 2.10000 | -0.105 | -8.131 | -1.093 | -34.450 | 9.893 | 0.998 |
| 2.30000 | -0.087 | -7.897 | -2.617 | -34.832 | 11.983 | 1.021 |
| 2.50000 | -0.074 | -7.707 | -3.958 | -35.116 | 13.864 | 1.044 |
| 2.70000 | -0.066 | -7.565 | -5.147 | -35.323 | 15.567 | 1.069 |
| 2.90000 | -0.059 | -7.477 | -6.203 | -35.467 | 17.119 | 1.095 |
| 3.10000 | -0.055 | -7.438 | -7.127 | -35.608 | 18.540 | 1.123 |
| 3.30000 | -0.051 | -7.450 | -7.939 | -35.749 | 19.841 | 1.154 |
| 3.50000 | -0.049 | -7.525 | -8.668 | -35.846 | 21.025 | 1.190 |
| 3.70000 | -0.048 | -7.669 | -9.324 | -35.905 | 22.082 | 1.231 |
| 3.90000 | -0.047 | -7.888 | -9.916 | -35.935 | 22.995 | 1.278 |
| 4.10000 | -0.047 | -8.185 | -10.486 | -35.974 | 23.700 | 1.334 |
| 4.30000 | -0.047 | -8.568 | -11.030 | -36.023 | 24.271 | 1.400 |
| 4.50000 | -0.048 | -9.054 | -11.521 | -36.051 | 24.769 | 1.473 |
| 4.70000 | -0.048 | -9.665 | -11.969 | -36.068 | 25.182 | 1.554 |
| 4.90000 | -0.049 | -10.427 | -12.384 | -36.082 | 25.493 | 1.642 |
| 5.10000 | -0.050 | -11.039 | -12.795 | -36.099 | 25.680 | 1.733 |
| 5.30000 | -0.050 | -11.411 | -13.204 | -36.109 | 25.721 | 1.822 |
| 5.50000 | -0.051 | -11.888 | -13.595 | -36.112 | 25.597 | 1.911 |
| 5.70000 | -0.051 | -12.522 | -13.973 | -36.120 | 25.286 | 1.999 |
| 5.90000 | -0.051 | -13.358 | -14.344 | -36.141 | 24.770 | 2.088 |
| 6.10000 | -0.051 | -14.065 | -14.735 | -36.281 | 25.014 | 2.189 |
| 6.30000 | -0.050 | -13.942 | -15.155 | -36.505 | 25.978 | 2.303 |
| 6.50000 | -0.049 | -13.465 | -15.585 | -36.694 | 26.698 | 2.405 |
| 6.70000 | -0.048 | -13.038 | -16.022 | -36.848 | 27.219 | 2.497 |
| 6.90000 | -0.047 | -12.889 | -16.465 | -36.968 | 27.582 | 2.578 |
| 7.10000 | -0.045 | -12.468 | -16.950 | -37.152 | 27.826 | 2.666 |
| 7.30000 | -0.043 | -11.527 | -17.471 | -37.406 | 27.980 | 2.757 |
| 7.50000 | -0.041 | -10.671 | -17.981 | -37.633 | 28.058 | 2.831 |
| 7.70000 | -0.039 | -9.917 | -18.473 | -37.828 | 28.076 | 2.886 |
| 7.90000 | -0.037 | -9.264 | -18.941 | -37.983 | 28.043 | 2.918 |
| 8.10000 | -0.035 | -8.574 | -19.381 | -38.158 | 28.247 | 2.935 |
| 8.30000 | -0.033 | -7.817 | -19.801 | -38.375 | 28.690 | 2.939 |
| 8.50000 | -0.032 | -7.117 | -20.198 | -38.571 | 29.092 | 2.923 |
| 8.70000 | -0.030 | -6.473 | -20.564 | -38.740 | 29.459 | 2.887 |
| 8.90000 | -0.028 | -5.884 | -20.896 | -38.875 | 29.792 | 2.829 |
| 9.10000 | -0.027 | -5.480 | -21.208 | -39.036 | 30.092 | 2.792 |
| 9.30000 | -0.026 | -5.251 | -21.505 | -39.223 | 30.359 | 2.780 |
| 9.50000 | -0.025 | -5.059 | -21.762 | -39.366 | 30.597 | 2.760 |
| 9.70000 | -0.024 | -4.904 | -21.973 | -39.457 | 30.806 | 2.729 |
| 9.90000 | -0.023 | -4.787 | -22.131 | -39.490 | 30.988 | 2.690 |
| 10.1000 | -0.022 | -4.781 | -22.305 | -39.521 | 31.146 | 2.682 |
| 10.3000 | -0.022 | -4.889 | -22.499 | -39.546 | 31.281 | 2.707 |
| 10.5000 | -0.022 | -5.051 | -22.633 | -39.496 | 31.389 | 2.724 |
| 10.7000 | -0.021 | -5.278 | -22.702 | -39.363 | 31.471 | 2.731 |
| 10.9000 | -0.021 | -5.592 | -22.698 | -39.141 | 31.527 | 2.725 |
| 11.1000 | -0.022 | -5.895 | -22.636 | -38.936 | 31.559 | 2.715 |
| 11.3000 | -0.022 | -6.179 | -22.526 | -38.765 | 31.566 | 2.704 |
| 11.5000 | -0.022 | -6.588 | -22.346 | -38.514 | 31.545 | 2.685 |
| 11.7000 | -0.023 | -7.172 | -22.090 | -38.175 | 31.495 | 2.656 |
| 11.9000 | -0.025 | -8.018 | -21.755 | -37.746 | 31.416 | 2.615 |
| 12.0000 | -0.025 | -8.585 | -21.556 | -37.496 | 31.365 | 2.590 |

Touchstone (TM) - Configuration( 100 1500 100 16437 2364 1000 1 3294 )
900AMP.OUT Thu Dec 29 15:40:06 1988 LL = 0.1 inch

| FREQ-GHZ | DB[S11] AMP | DB[S22] AMP | DB[S21] AMP | DB[S12] AMP | DB[NF] AMP | K AMP |
|---|---|---|---|---|---|---|
| 0.10000 | -1.545 | -11.796 | 9.619 | -50.033 | 3.093 | 14.596 |
| 0.30000 | -0.321 | -14.549 | 11.407 | -40.783 | 2.243 | 1.148 |
| 0.50000 | -0.919 | -14.361 | 13.751 | -33.875 | 1.773 | 1.151 |
| 0.70000 | -4.230 | -14.979 | 16.816 | -27.388 | 0.912 | 1.262 |
| 0.90000 | -14.264 | -15.983 | 17.206 | -24.221 | 0.482 | 1.297 |
| 1.10000 | -2.997 | -12.778 | 13.040 | -25.844 | 1.044 | 1.305 |
| 1.30000 | -1.244 | -11.060 | 8.888 | -27.571 | 2.563 | 1.287 |
| 1.50000 | -0.681 | -9.960 | 5.626 | -28.677 | 4.470 | 1.252 |
| 1.70000 | -0.431 | -9.130 | 2.993 | -29.383 | 6.340 | 1.215 |
| 1.90000 | -0.297 | -8.462 | 0.796 | -29.848 | 8.013 | 1.178 |
| 2.10000 | -0.222 | -7.933 | -1.029 | -30.187 | 9.883 | 1.156 |
| 2.30000 | -0.176 | -7.521 | -2.569 | -30.458 | 11.948 | 1.148 |
| 2.50000 | -0.145 | -7.192 | -3.924 | -30.664 | 13.803 | 1.142 |
| 2.70000 | -0.122 | -6.939 | -5.122 | -30.819 | 15.480 | 1.140 |
| 2.90000 | -0.106 | -6.755 | -6.183 | -30.935 | 17.006 | 1.141 |
| 3.10000 | -0.094 | -6.631 | -7.108 | -31.039 | 18.404 | 1.142 |
| 3.30000 | -0.085 | -6.562 | -7.917 | -31.135 | 19.685 | 1.144 |
| 3.50000 | -0.078 | -6.556 | -8.641 | -31.203 | 20.849 | 1.148 |
| 3.70000 | -0.073 | -6.612 | -9.290 | -31.250 | 21.890 | 1.156 |
| 3.90000 | -0.069 | -6.734 | -9.874 | -31.280 | 22.796 | 1.167 |
| 4.10000 | -0.066 | -6.926 | -10.434 | -31.328 | 23.482 | 1.185 |
| 4.30000 | -0.064 | -7.188 | -10.966 | -31.391 | 24.041 | 1.211 |
| 4.50000 | -0.062 | -7.530 | -11.443 | -31.450 | 24.531 | 1.241 |
| 4.70000 | -0.062 | -7.965 | -11.877 | -31.511 | 24.941 | 1.275 |
| 4.90000 | -0.061 | -8.508 | -12.277 | -31.583 | 25.254 | 1.313 |
| 5.10000 | -0.061 | -8.948 | -12.678 | -31.659 | 25.443 | 1.356 |
| 5.30000 | -0.060 | -9.213 | -13.080 | -31.713 | 25.482 | 1.399 |
| 5.50000 | -0.060 | -9.540 | -13.464 | -31.754 | 25.358 | 1.441 |
| 5.70000 | -0.059 | -9.972 | -13.835 | -31.795 | 25.053 | 1.480 |
| 5.90000 | -0.058 | -10.549 | -14.198 | -31.847 | 24.548 | 1.517 |
| 6.10000 | -0.057 | -11.226 | -14.579 | -32.041 | 24.815 | 1.565 |
| 6.30000 | -0.055 | -11.550 | -14.992 | -32.335 | 25.802 | 1.627 |
| 6.50000 | -0.054 | -11.539 | -15.418 | -32.572 | 26.538 | 1.681 |
| 6.70000 | -0.052 | -11.397 | -15.856 | -32.733 | 27.070 | 1.725 |
| 6.90000 | -0.050 | -11.323 | -16.303 | -32.809 | 27.437 | 1.754 |
| 7.10000 | -0.047 | -11.157 | -16.793 | -32.973 | 27.694 | 1.790 |
| 7.30000 | -0.045 | -10.667 | -17.320 | -33.277 | 27.873 | 1.840 |
| 7.50000 | -0.042 | -10.119 | -17.842 | -33.564 | 27.976 | 1.882 |
| 7.70000 | -0.040 | -9.563 | -18.349 | -33.828 | 28.020 | 1.916 |
| 7.90000 | -0.037 | -9.033 | -18.837 | -34.065 | 28.014 | 1.940 |
| 8.10000 | -0.035 | -8.476 | -19.298 | -34.330 | 28.253 | 1.954 |
| 8.30000 | -0.032 | -7.880 | -19.739 | -34.652 | 28.741 | 1.964 |
| 8.50000 | -0.030 | -7.323 | -20.155 | -34.984 | 29.198 | 1.966 |
| 8.70000 | -0.028 | -6.813 | -20.541 | -35.320 | 29.626 | 1.961 |
| 8.90000 | -0.026 | -6.358 | -20.886 | -35.656 | 30.030 | 1.948 |
| 9.10000 | -0.024 | -6.097 | -21.211 | -35.949 | 30.405 | 1.934 |
| 9.30000 | -0.022 | -6.019 | -21.526 | -36.175 | 30.748 | 1.919 |
| 9.50000 | -0.021 | -5.991 | -21.804 | -36.358 | 31.069 | 1.897 |
| 9.70000 | -0.019 | -6.021 | -22.040 | -36.494 | 31.370 | 1.869 |
| 9.90000 | -0.018 | -6.117 | -22.232 | -36.577 | 31.650 | 1.834 |
| 10.1000 | -0.017 | -6.402 | -22.470 | -36.581 | 31.930 | 1.807 |
| 10.3000 | -0.016 | -6.857 | -22.784 | -36.502 | 32.210 | 1.791 |
| 10.5000 | -0.015 | -7.329 | -23.094 | -36.379 | 32.473 | 1.775 |
| 10.7000 | -0.015 | -7.737 | -23.415 | -36.235 | 32.721 | 1.761 |
| 10.9000 | -0.015 | -7.938 | -23.766 | -36.101 | 32.955 | 1.753 |
| 11.1000 | -0.015 | -7.862 | -24.143 | -35.997 | 33.189 | 1.738 |
| 11.3000 | -0.015 | -7.509 | -24.566 | -35.917 | 33.423 | 1.719 |
| 11.5000 | -0.016 | -6.824 | -25.080 | -35.897 | 33.643 | 1.714 |
| 11.7000 | -0.017 | -5.926 | -25.702 | -35.970 | 33.847 | 1.727 |
| 11.9000 | -0.018 | -4.969 | -26.440 | -36.157 | 34.035 | 1.762 |
| 12.0000 | -0.019 | -4.507 | -26.851 | -36.297 | 34.122 | 1.790 |

**Appendix 1.**
This is the Touchstone™ run for simulating the amplifier operating at 900 MHz. The first printout shows the configuration. The first printout of amplifier performance is for source lead length of approximately 0 (as short as possible), the second for lead length of 0.1 inch (the optimum value).

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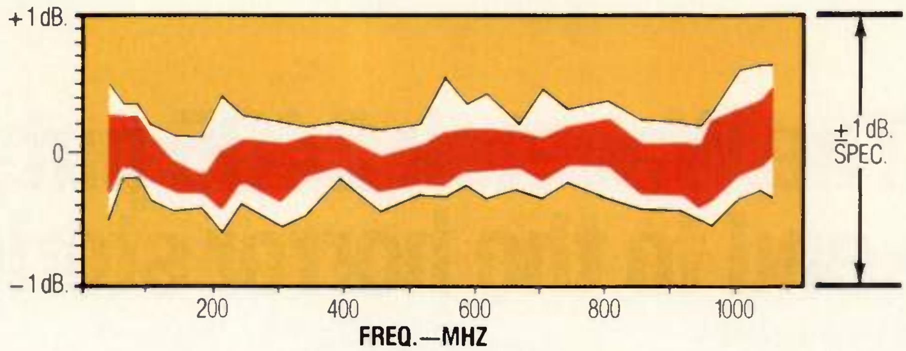


**6060B typical level accuracy vs. frequency at -127 dBm.**

Sample: 38 units. Solid line: worst case. Shaded: Typical (75%).

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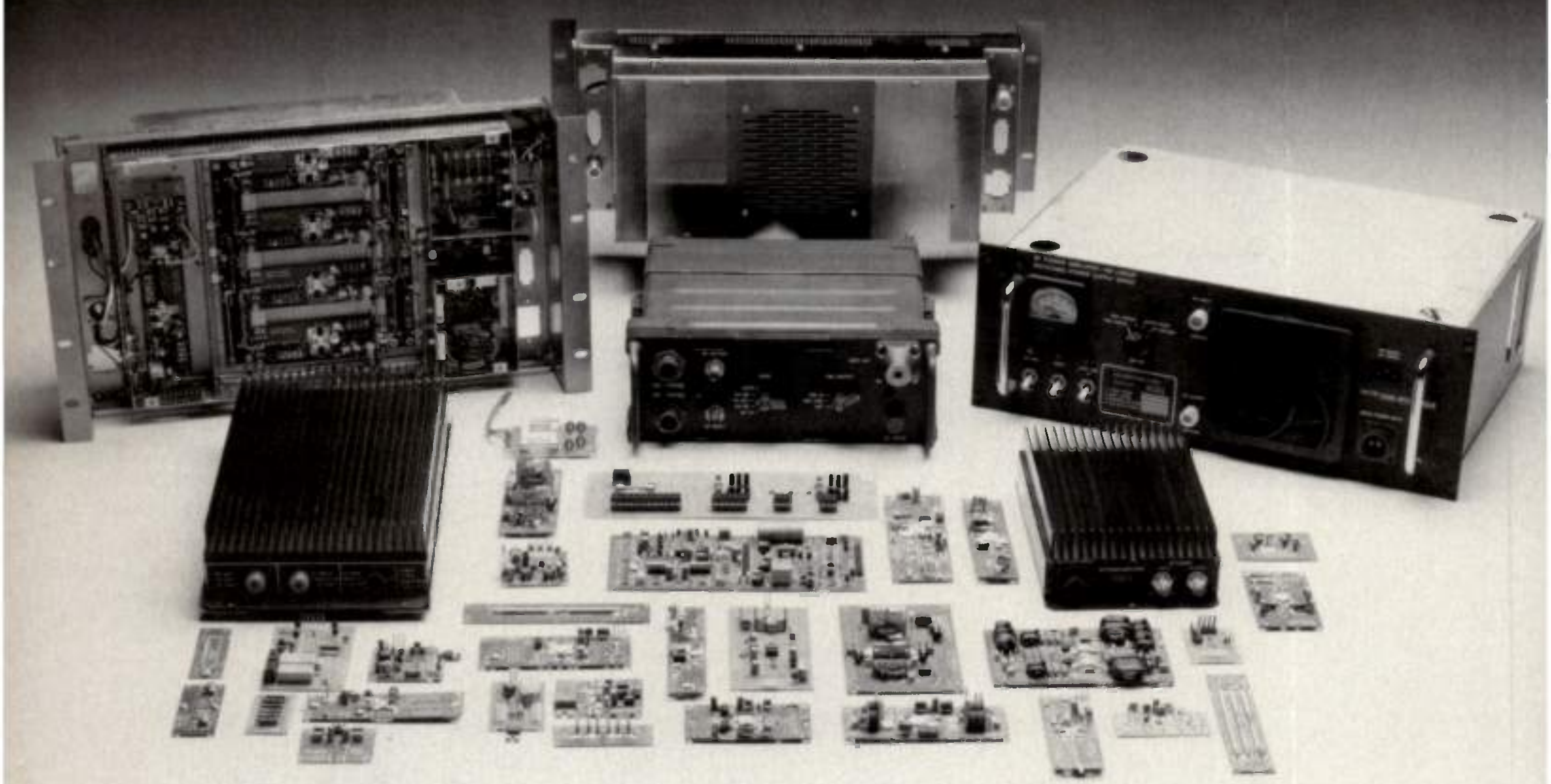

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*By Victor Koren*
*Tadiran, Israel*

*In this article, the author describes new negative feedback amplifier circuits that offer equal or better performance than previously possible, while using a simpler design. Also included is a review of currently used negative feedback amplifier circuits.*

Broadband linear RF amplifier performance is measured by these main characteristics: bandwidth, low noise figure, linearity, strong-signal handling, input/output impedance match, reverse isolation and stability of parameters. In the past, most designs used negative feedback to achieve desired performance. The following section discusses four patented negative feedback amplifiers. This highlights the major steps in broadband linear negative feedback amplifier design.

### Negative Feedback Amplifiers

*Amplifier A (1)* — This amplifier (Figure 1) uses resistive parallel and series

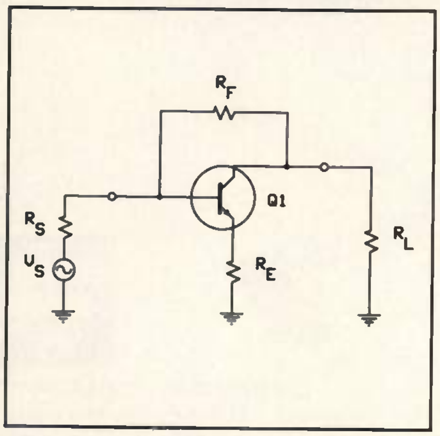


Figure 1. Amplifier A.

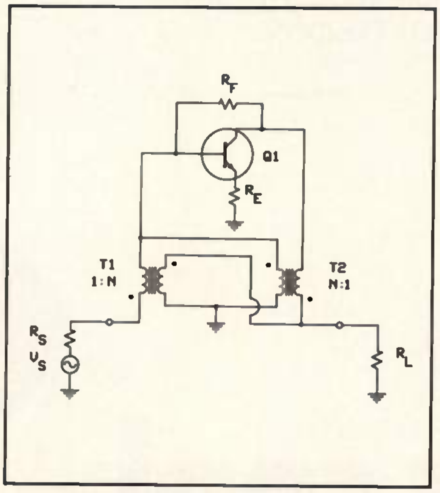


Figure 2. Amplifier B.

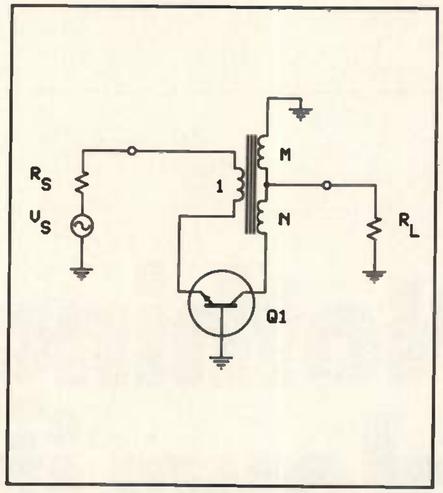


Figure 3. Amplifier C.

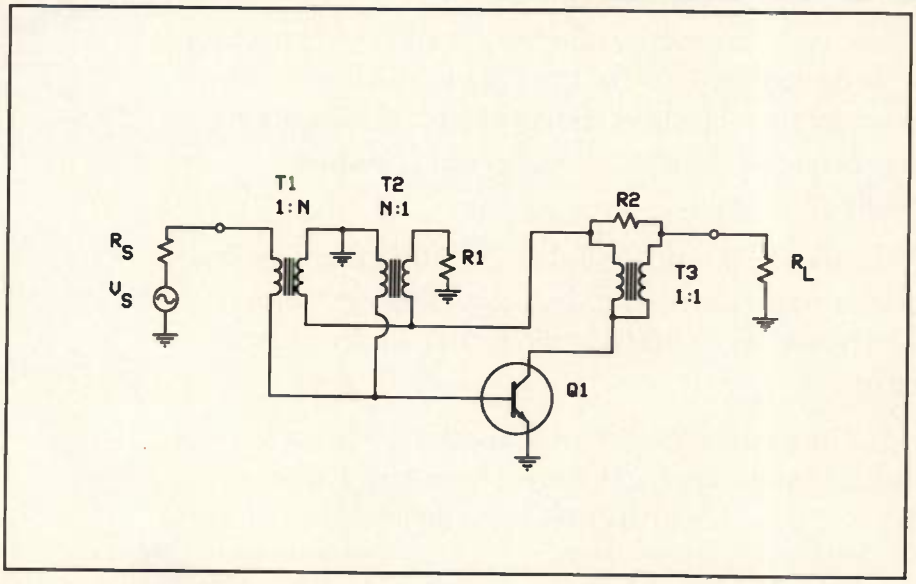


Figure 4. Amplifier D.

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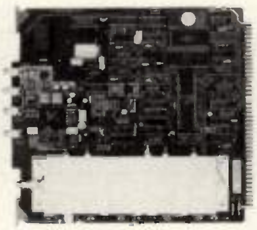

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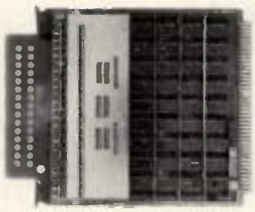

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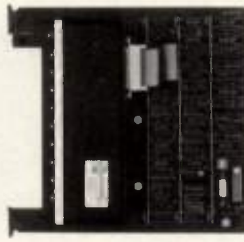

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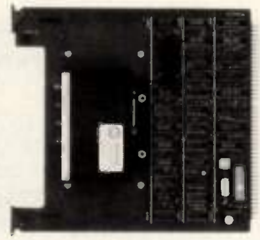

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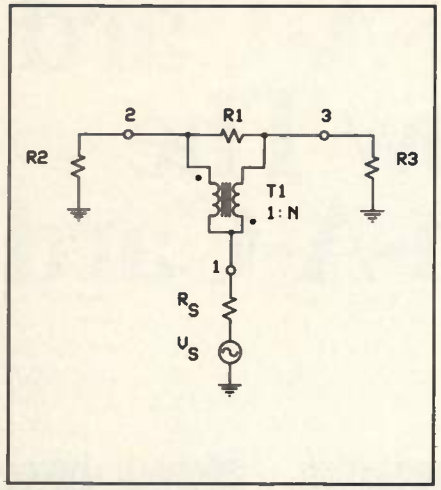



Figure 5. Non-symmetrical power combiner-splitter.

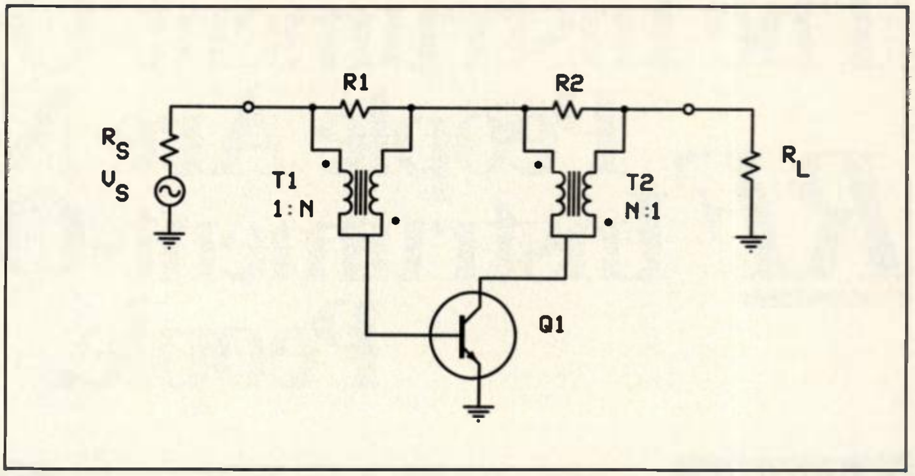



Figure 6. The new design.

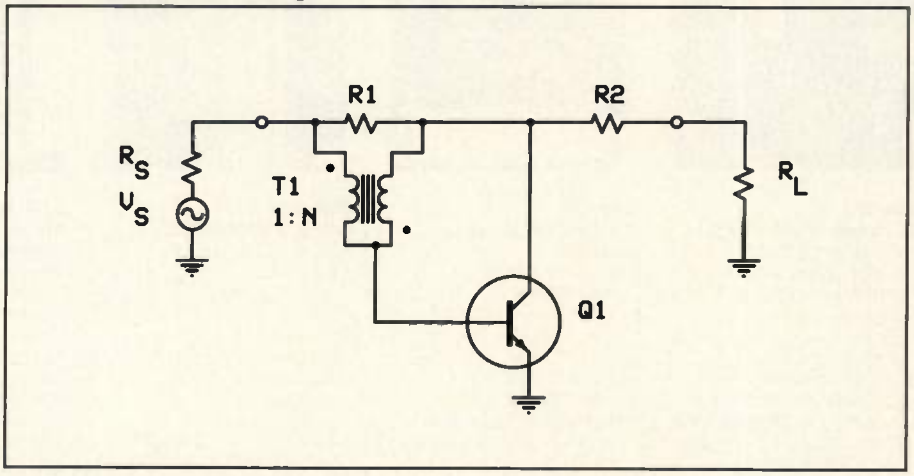



Figure 7. The alternative design.

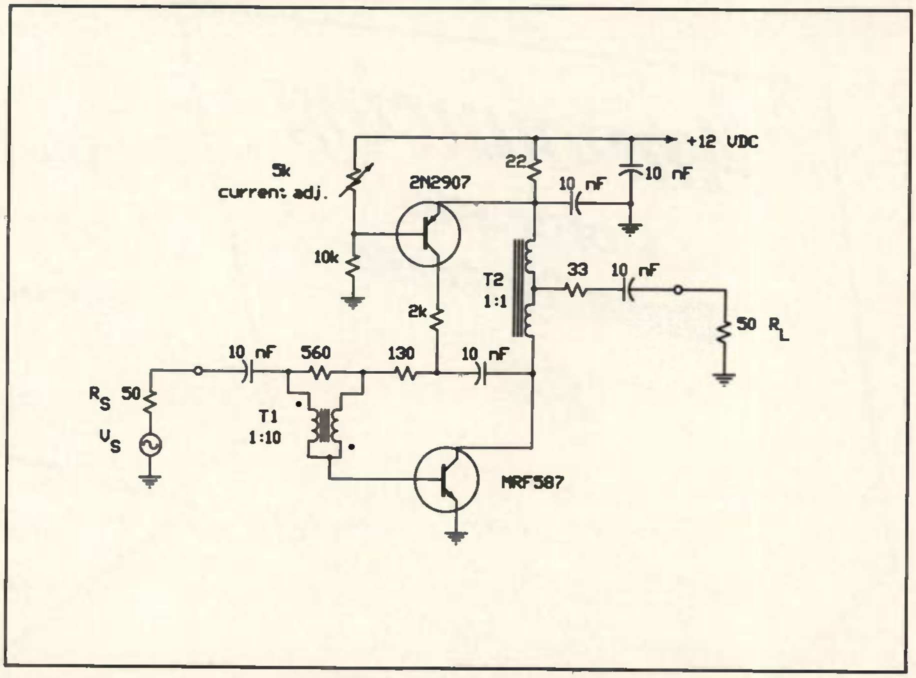



Figure 8. A prototype negative feedback amplifier.

feedback to stabilize gain and impedance matching. The disadvantage with this method is degradation of noise figure and strong signal handling capability due to the resistive loss, and low reverse isolation due to the direct connection from output to input.

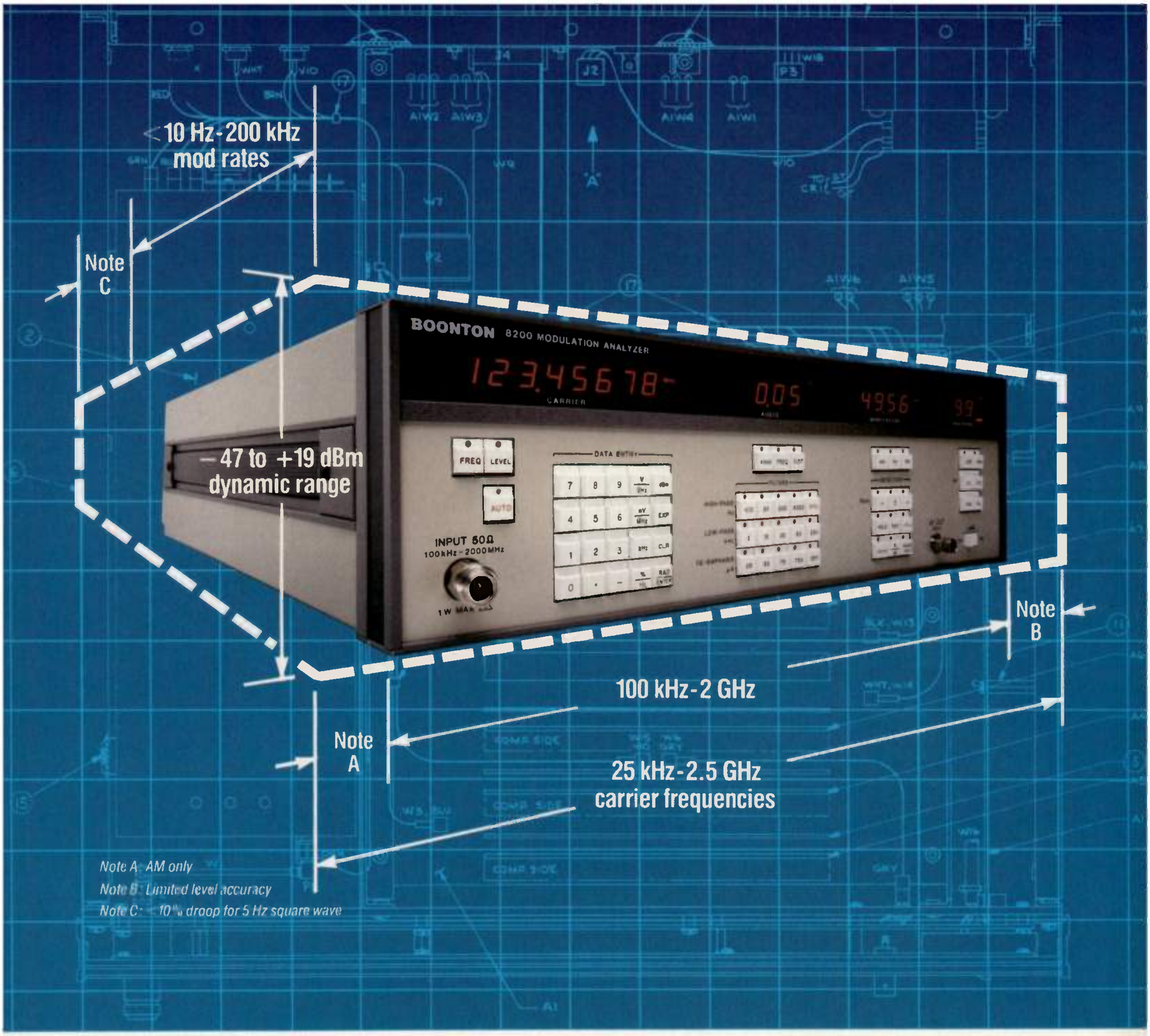
*Amplifier B (2)* — The Lossless Feedback amplifier, shown in Figure 2, uses a resistive feedback as in amplifier A. It has a much weaker feedback, and the main feedback is performed with a directional coupler that feeds back part of the output power. Most implementations do not use the resistive feedback, gaining better noise figure and strong signal handling capability, but with degradation of reverse isolation.

*Amplifier C (3)* — This amplifier is a common base amplifier that uses a transformer as the negative feedback element (Figure 3). The transformer defines the gain and the relationship between input and output impedances. It has good noise figure and good strong signal handling capability, but has low reverse isolation.

*Amplifier D (4)* — Until now, amplifiers A, B and C had poor reverse isolation (slightly higher than -G dB). Amplifier D, illustrated in Figure 4, solved the problem of reverse isolation by using a 3 dB power splitter at the output, isolating the load from the feedback path. The feedback is performed with a directional coupler at the input of the amplifier. The use of the power splitter at the output causes a 3 dB power loss, degrading the strong signal handling capability by the same amount.

## The Proposed Design

This design is based on a non-

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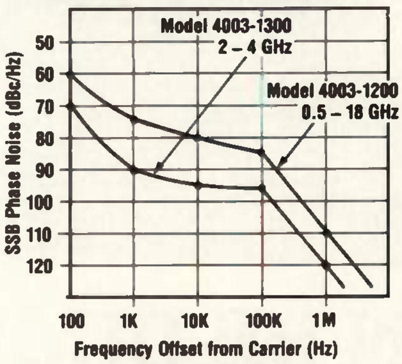


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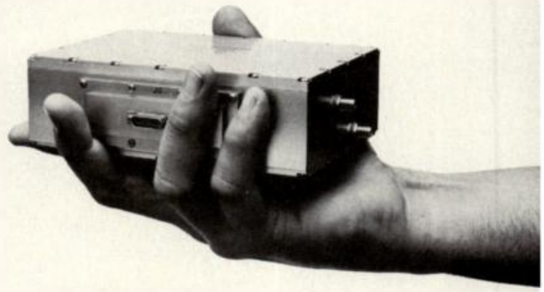

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symmetrical power combiner-splitter (Figure 5) which uses the following relationships:

$R_2 = R_s(1 + N)/N$

$R_3 = (1+N)R_s$

$R_1 = R_2 + R_3$

All the ports are matched and ports 2 and 3 are isolated. The input power will be devided between $R_2$ and $R_3$ according to the inverse of their ratio:

$P(R_2)/P(R_3) = R_3/R_2$

As a combiner, the power loss ratio from port 2 to 1 is N/(N+1), and from port 3 to 1 is 1/(N+1). This non-symmetrical power combiner-splitter can be used in the design of a new amplifier (Figure 6), as a splitter at the output to take a sample of the output power, and as a combiner at the input to combine the power sample with the input signal.

The feedback through the combiner and splitter defines the gain and the input/output impedance. The reverse isolation is high because of the inherent isolation between the ports of the combiner and splitter.

For higher N, the input and output signal loss in the combiner and splitter is lower. For example, if N=4, then the loss is about 1 dB; if N=8 then the loss is about 0.5 dB.

The only point of caution is the effect of the phase shift of the signal passing through the power splitter and combiner. This phase shift, added to the phase shift of the transistor, defines the phase margin at the point that the loop gain is equal to 1. Good construction of the transformers and choice of transistor can assure a stable amplifier. A second amplifier, shown in Figure 7, was designed. It uses only one non-symmetrical power combiner with a 3 dB power loss at the output as in amplifier D.

The feedback consists of parallel feedback at the output by connecting the port of the power combiner directly to the collector. This causes the output impedance at the collector to drop to a very low value. So, a resistor in series to the output defines the output impedance: $Z_{out} = R_2$, with a 3 dB power loss.

To achieve signal cancellation at the base of the transistor (virtual ground), the negative feedback forces the input impedance to be $Z_{in} = R_1/(1+N)$. Noticing the virtual ground, the voltage ratio between the collector and input is N, but the load voltage is half the collector voltage, so the voltage gain is N/2.

The summary of the design parameters for the amplifier (Figure 7) are:

$Z_{in} = R_1/(N+1)$

$Z_{out} = R_2$

$Av = -N/2$

From these equations, the designer can calculate $R_1$, $R_2$ and N, after establishing the desired values for $Z_{in}$, $Z_{out}$ and Av.

In a practical circuit, a small resistor (50-300 ohm) is connected between the collector and the feedback path, to suppress parasitic oscillations. This resistor will change only Av, making it a little larger.

The input loss in the power combiner, according to the power combiner equations, is N/(N+1), or in dB: 10 log (N/N+1). This loss adds to the noise figure of the transistor, defining the noise figure of the amplifier.

The frequency response is limited by two main factors: frequency response of the amplifier with feedback disconnected and frequency response of the transformer in the power combiner. Higher collector impedance, and higher turns ratio (N) make a higher gain amplifier but with lower bandwidth.

In this circuit (Figure 8), the collector voltage is approximately 10 times the input voltage, because N=10. This voltage is stepped down by $T_2$ and connected to the 50 ohm load through a 33 ohm resistor (only 0.301 of the collector voltage reaches the load). Hence, the voltage gain is 10(0.301)=3.01 or 9.6 dB. In the practical circuit, the gain is 10.8 to 11 dB because an additional 130 ohm resistor was connected in series with the feedback path to suppress parasitic oscillation at UHF.

This article has illustrated two circuits that can achieve or surpass the performance of conventional negative feedback amplifier designs, but with simpler configurations.


**References**

1. U.S. Patent 3,493,882
2. U.S. Patent 3,624,536
3. U.S. Patent 3,891,934
4. U.S. Patent 4,042,887



**About the Author**

Victor Koren is a project engineer at Tadiran, P.O. Box 267, Holon 58102, Israel. He holds a BScEE from Tel Aviv University.

# CAD Optimizes the Gain of Dual Gate MOSFET VHF Amplifiers

*By Amy Purushotham and S.V.K. Shastry*
*ISRO Satellite Centre, India*

*A software method that optimizes the gain of dual gate MOSFET VHF amplifiers and reduces design time is described in this article. The software needs only center frequency, bandwidth, source and load impedance data and eliminates the cumbersome calculations that are involved in the conventional design technique using Smith charts. This article is designed not only to discuss the advantages of using CAD, but also to provide information on the design of dual gate MOSFET RF amplifiers.*

Using a polynomial curve-fit routine, the y-parameters versus frequency curves of the device are stored in memory in the form of polynomial coefficients. After ensuring the stability of the circuit, the input and output matching network circuit parameters are determined. The network chosen has the advantage that the amplifier bandwidth is adjustable using a single capacitive element. The theoretically determined gain curve is found to fit well with the experimentally measured one.

RF amplifiers have been designed using the y-parameter technique for some time. Of major importance is the design of suitable networks to match the input and output impedances of the device to the source and load impedances respectively, over the operating bandwidth. A graphical approach (1) using the Smith chart is commonly employed for this task, but tedious calculations are required before the chart can be used.

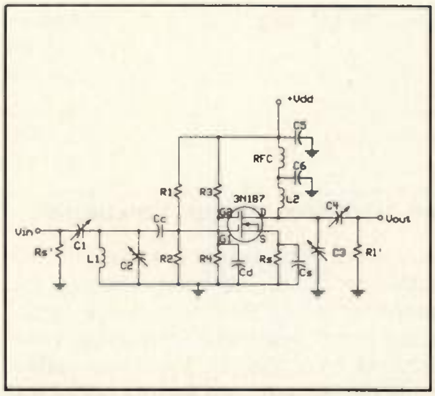


**Figure 1. Dual gate MOSFET RF amplifier (200 MHz).**

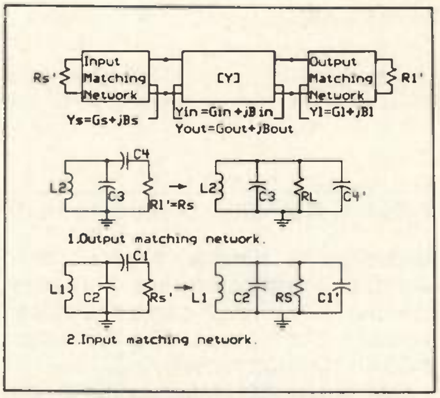


**Figure 2. Network for optimum load transformation.**

The FORTRAN 77 program described here uses a curve-fit routine to compute the y-parameters of the device at the desired frequencies. The experimentation to validate the CAD is done using a dual gate MOSFET. This device is particularly useful in VHF amplification and in mixer service (2).

The design of a dual gate MOSFET RF amplifier is, in theory, simpler than that of a bipolar amplifier because of the very low internal feedback associated with the dual gate device. A second gate is available for either AGC or local oscillator injection. Highly stable RF and conversion gains are easily obtained with inexpensive handmade coils, without the need for neutralization.

The RF small-signal performance of a transistor can be completely charac-

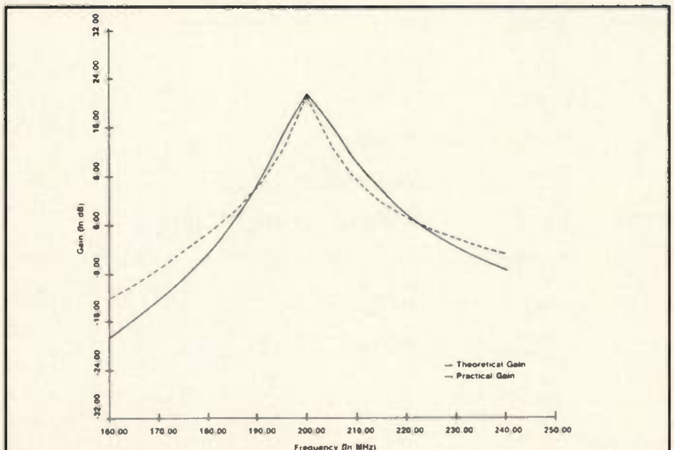


**Figure 3. Frequency vs. gain curve.**

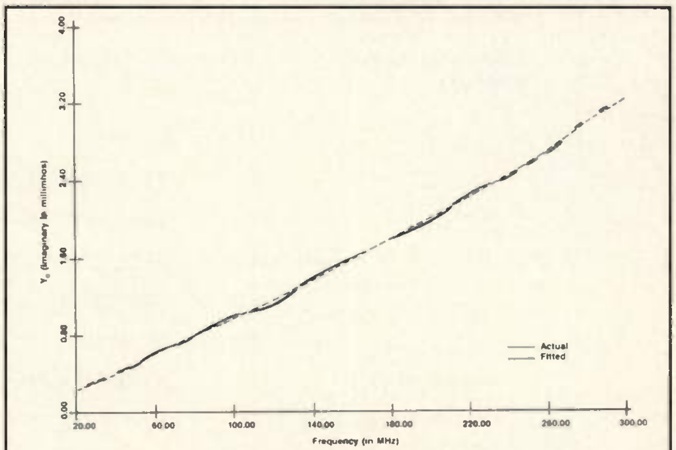


**Figure 4. Curve-fit example II.**

| Particulars | Equations | Remarks |
|---|---|---|
| Linvill stability factor (C) | $y = y_r y_f$<br>$C = \lvert y \rvert / 2g_i g_o - \mathrm{Re}\,(y)$ | C<1: Device is unconditionally stable<br>C>1: Device is potentially unstable |
| Stern's stability factor (K) | $K = 2(g_i + G_s)\,(g_o + G_l) / \lvert y \rvert + \mathrm{Re}\,(y)$ | K>1: Device is unconditionally stable<br>K<1: Device is potentially unstable |
| Source and load admittances for simultaneous conjugate match | $G_s = [2g_i g_o - \mathrm{Re}\,(y)]^2 - \lvert y \rvert\,^2/2g_o$<br>$B_s = jb_i + \mathrm{Im}\,(y)\,/2g_o$<br>$G_l = G_s g_o/g_i$<br>$B_l = jb_o + \mathrm{Im}\,(y)/2g_i$ | If simultaneous conjugate match is not possible, carry out a mismatch design. |
| Mismatch design to overcome potential instability | $R = \sqrt{K\,[\,\lvert y \rvert + \mathrm{Re}\,(y)]\,/2g_i g_o} - 1$<br>$G_s = Rg_o$<br>$G_l = Rg_i$<br>$B_l = -\,jb_o + [\mathrm{Im}\,(y)/2g_i]$ | Assume K = 3 or 4 for determining the mismatch ratio "R".<br>$G_s$ = Source Conductance<br>$B_s$ = Source Susceptance<br>$G_l$ = Load Conductance<br>$B_l$ = Load Susceptance |

**Table 1. Stability, conjugate match and mismatch design equations.**

terized by its admittance parameters. Based on these parameters, equations can be written for finding a suitable transistor and for completing the design once the transistor is selected.

One of the first requirements in any amplifier design is to choose the device which is best suited for the task. Two of the most important considerations in choosing a device for use in amplifier design are its stability and its maximum available gain (MAG). Stability, as it is used here, is a measure of the device's tendency toward oscillation. MAG is a figure of merit for the transistor which indicates the maximum theoretical power gain that can be expected from the device when it is conjugately matched to its source and load impedance.

Two factors are used to determine the potential stability of transistors in RF amplifiers. One factor is known as the Linvill factor (C), and the other as the Stern factor (K). Both factors are calculated from equations requiring y-parameter information. The main difference between the two factors is that the Linvill factor assumes the device is not connected to load, while the Stern factor includes the effects of source and load admittances.

If C<1 and K>1, the device is unconditionally stable. In practical design, it is recommended that a K-factor of 3 or 4 be used to provide a margin of safety. If C>1 and K<1, the device is potentially unstable.

There are two basic solutions to the problem of unstable RF amplifiers. First, the amplifier can be neutralized. This permits the amplifier to be matched perfectly to the source and load impedances. However, neutralization requires extra components and creates a problem when the frequency is changed. The other solution is to introduce some mismatch into either the source or load tuning networks. This method, sometimes known as the Stern's solution, requires no extra components but does produce a reduction in gain.

## Design Procedure (4)

A typical design procedure uses the following steps:

1. Characterize the amplifier in terms of its center frequency, maximum available gain, bandwidth, and source and load impedances.

2. Select a suitable device and decide the operating point, i.e., the bias to produce a given drain current, gain, NF, etc. Once the operating point has been selected, the biasing network is designed.

| Particulars | Output Network | Input Network |
|---|---|---|
| Matching network design | $C4 = 1/[2(\pi)(F)(XC4)]$ | $C1 = 1/[2(\pi)(F)(XC1)]$ |
| | $XC4 = R_l'\sqrt{R_l/R_l' - 1}$ | $XC1 = R_s'\sqrt{R_s/R_s' - 1}$ |
| | $C_{out} = B_{out}/2(\pi)(F)$ | $C_{in} = B_{in}/2(\pi)(F)$ |
| | $C_t = 1/[2(\pi)(BW)(1/G_{out} + G_l)]$ | $C_t = 1/[2(\pi)(BW)(1/G_{out} + G_l)]$ |
| | $C4' = 1/[2(\pi)(F)(XC4)(1 + R_l'/XC4)^2]$ | $C1' = 1/[2(\pi)(F)(XC1)(1 + R_s'/XC1)^2]$ |
| | $C3 = C_t - C_{out} - C4'$ | $C2 = C_t - C_{out} - C1'$ |
| | $L2 = 1/[4(\pi^2)(F^2)C_t]$ | $L1 = 1/[4(\pi^2)(F^2)C_t]$ |
| | $R_l' = 50$ ohms | $R_s' = 50$ ohms |
| Power Gain ($G_p$) | $G_p = [\lvert y_f \rvert^2 G_l]/[\lvert Y_i y_o \rvert^2 \mathrm{Re}(Y_{in})]$ | |
| Transducer Gain ($G_T$) | $G_t = \lvert\,(y_i + Y_s)\,(y_o + Y_l) - y_f y_r \rvert^2$ | |
| | $G_T = 4G_s G_l \lvert y_f \rvert^2/G_t$ | |

**Table 2. Equations for matching network and gain calculation.**

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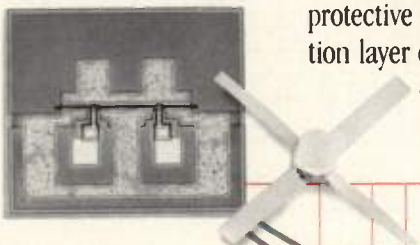

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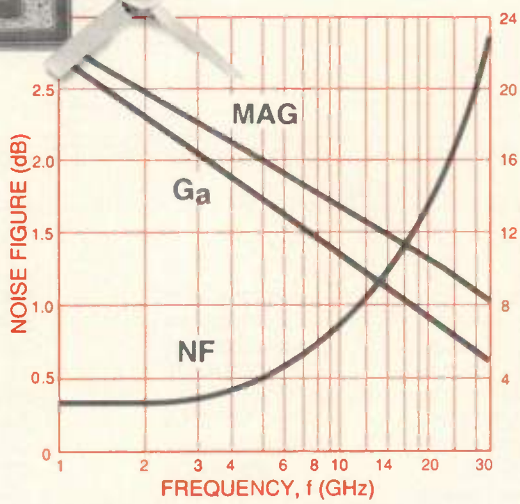


NE32083 typical Noise Figure, Gain and Available Gain

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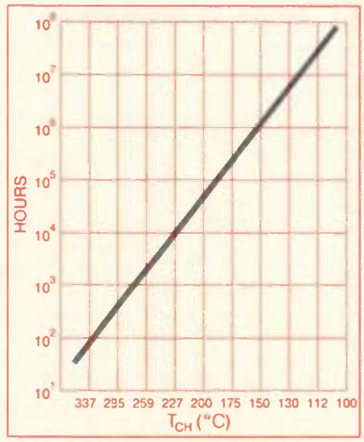


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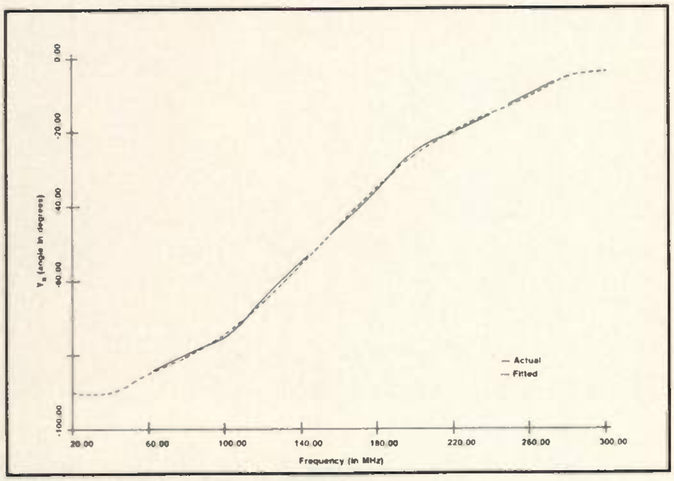


**Figure 5. Curve-fit example I.**

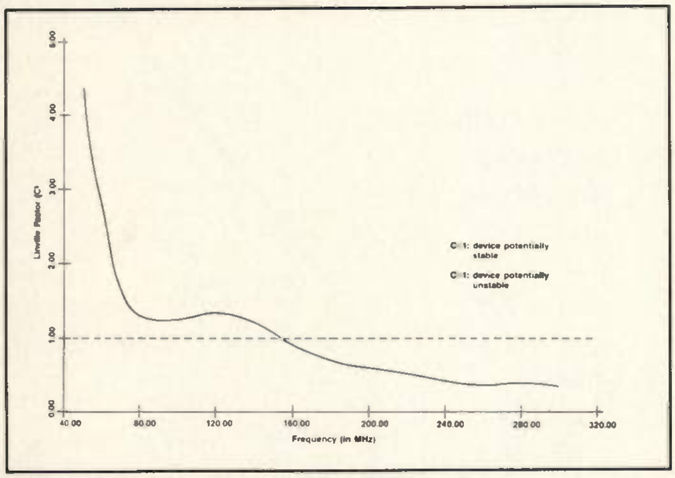


**Figure 6. Frequency vs. Linville factor.**

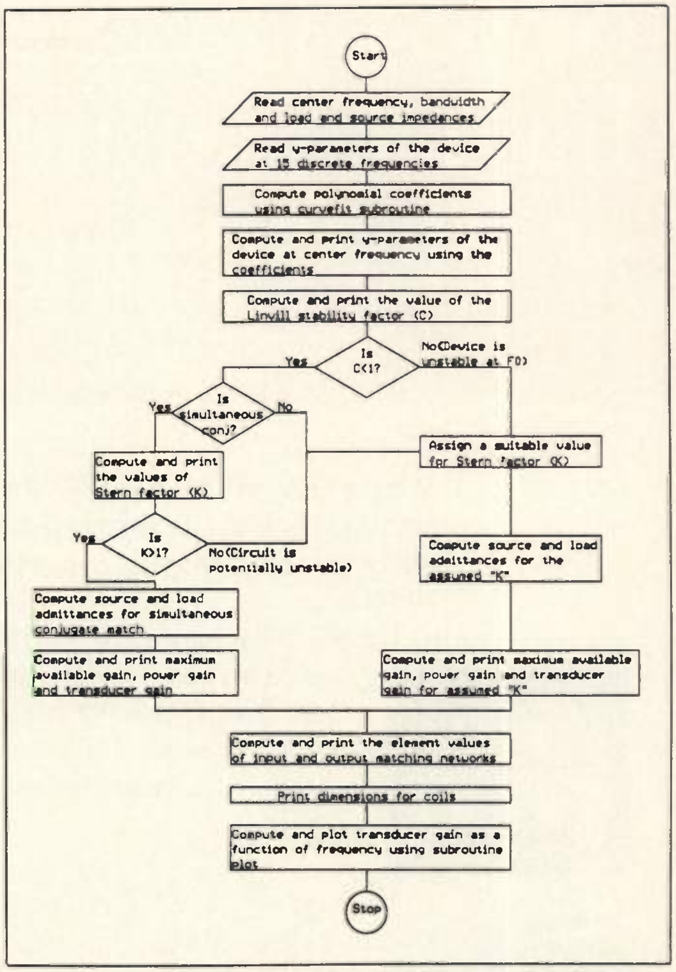


**Figure 7. CAD flowchart for dual gate MOSFET RF amplifiers.**

| | $Y_R$-MAG | $Y_R$-ANG | $Y_F$-MAG | $Y_F$-ANG |
|---|---|---|---|---|
| K 0 | .26630189-001 | -.49622651+002 | .11908608+002 | .33864470–000 |
| K 1 | -.35941459-002 | -.44512158+001 | .18118453-001 | .45065370–000 |
| K 2 | .21139669-003 | .17969174+000 | -.55947220-003 | -.26060640-001 |
| K 3 | -.57045806-005 | -.36659164-002 | .78543478-005 | .45021363-003 |
| K 4 | .83683442-007 | .43988771-004 | -.73176790-007 | -.39929301-005 |
| K 5 | -.71924611-009 | -.32527465-006 | .58716173-009 | .18441323-007 |
| K 6 | .37264139-011 | .15089259-008 | -.38032808-011 | -.33806799-010 |
| K 7 | -.11453677-013 | -.42943657-011 | .15838199-013 | -.51190030-013 |
| K 8 | .19234819-016 | .68641502-014 | -.35270853-016 | .31739501-015 |
| K 9 | -.13595044-019 | -.47192646-017 | .31619594-019 | -.37887754-018 |
| | $Y_I$-REAL | $Y_I$-IMAG | $Y_O$-REAL | $Y_O$-IMAG |
| K 0 | .94166907+000 | -.12362629+001 | .12234468-001 | .57793106+000 |
| K 1 | -.10812794+000 | .16872135+000 | .71404563-003 | -.48803179-001 |
| K 2 | .48931375-002 | -.47936081-002 | .56209898-004 | .22874115-002 |
| K 3 | -.11066784-003 | .77779498-004 | -.14985186-005 | -.45482221-004 |
| K 4 | .14552626-005 | -.59196917-006 | .21775734-007 | .52299931-006 |
| K 5 | -.11718854-007 | .14094339-008 | -.18237733-009 | -.37073858-008 |
| K 6 | .58518926-010 | .84283349-011 | .90423247-012 | .16552324-010 |
| K 7 | -.17641496-012 | -.67289400-013 | -.26218070-014 | -.45560885-013 |
| K 8 | .29369786-015 | .17379415-015 | .41087776-017 | .70853198-016 |
| K 9 | -.20712943-018 | -.16088380-018 | -.26883653-020 | -.47708040-019 |

**Table 3. Polynomial coefficients from curve-fit subroutine.**

3. Create a y-parameter data file for a minimum of 10 discrete frequencies in the frequency band of interest.

4. Express the y-parameters in terms of a polynomial of suitable degree in f (frequency) and obtain the polynomial coefficient to achieve a least square curve-fit (3). Using these coefficients, the y-parameters at any desired frequency may be determined. The y-parameter versus frequency plots for the device are stored in the computer for subsequent use.

5. Check for stability of the device. This involves computing the y-parameters at the center frequency and substituting them in the Linvill or Stern stability equations (see Table 1). If the device is potentially unstable, the stability may be ensured by mismatching input/output tuning circuits, thereby sacrificing some gain. If the device is stable, the gain may be optimized employing tuning circuits by simultaneously conjugate-matching the actual load and source admittances.

6. Obtain the gain versus frequency curve and component values for the construction of a stable amplifier.

Assume that the RF amplifier to be designed requires a center frequency of 200 MHz and bandwidth of 10 MHz. Further, let the source and load impedances be 50 ohms. The circuit that is proposed is depicted in Figure 1.

An N-channel dual gate MOSFET (3N187) was chosen as an active device to construct the RF amplifier. The operating point characteristics of the MOSFET (as taken from the data sheets) are as follows:

a) $V_{dd}$ = 15 volts c) $V_{g2s}$ = 4.0 volts
b) $V_{g1s}$ = −0.5 volts d) $I_d$ = 10mA
e) $R_s$ = 270 ohms (This has been chosen empirically to give the most suitable self-bias for dual gate MOSFET.)

The biasing network may be designed in the usual manner to achieve the above conditions.

From the data sheets, the y-parameter values are sampled at N discrete frequencies. Subroutine "FIT" is used to fit polynomials of several different degrees to the given set of N data points ($y_i$, $F_i$). This subroutine determines the coefficients of polynomials of degrees 1,2,3...(N-1). In other words, up to N-1 different sets of polynomial coefficients can be found for the same set of N data points.

In the example considered, a 10th degree polynomial is fitted to 15 data points. The polynomial equation is of the form:

$$Y'_i = K_0 + K_1 f_1 + K_2 f_2^2 + \ldots + K_m f_i^m \quad (1)$$

The least square fit algorithm is used to arrive at the values of the constants $K_o$ through $K_m$. This method involves finding a minimum value of S, where:

$$S = \sum_{i=1}^{N} (Y'_i - y_i)^2 = \sum_{i=1}^{N} (K_0 + K_1 f_1 + \ldots K_m f_i^m - y_i)^2 \quad (2)$$

The coefficients $K_o$ through $K_m$ may be found by solving a system of linear equations obtained by setting the first partial derivative (with respect to $K_o \ldots K_m$) of S to zero. These coefficients are stored in another data file and the original file containing y-parameter values can be discarded. Thus the y-parameters for any frequency of interest

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can be extracted using Equation 1 and the new data file. The stability calculations for the device can now be performed by finding the y-parameters at 200 MHz.

Optimum power gain is obtained from a transistor when the input admittance $Y_i$ and the output admittance $Y_o$ are conjugately matched to source admittance $Y_s$ and load admittance $Y_l$, respectively. Expressions for $Y_s$ and $Y_l$ for simultaneous conjugate match for maximum power transfer (from source to load) are given in Table 1.

### Designing with Potentially Unstable Transistors

If the transistor is unstable, there are several options available that will enable the use of the transistor in a stable configuration:

1. Select a new bias point for the transistor.
2. Neutralize the transistor.
3. Selectively mismatch the input and output impedances of the transistors to reduce the gain of the stage.

For reasons stated earlier, only option 3 is described here. This method, sometimes called the Stern solution, makes $G_s$ and $G_l$ sufficiently large so as to force K to become greater than 1. Hence, the amplifier remains stable for those terminations. The gain of the amplifier must be less than that which would be possible with a simultaneous conjugate match. The procedure for a design using unstable devices is as follows:

1. Assign a value of K that will assure a stable amplifier (K = 3 or 4).
2. This value of K is substituted in the equation for mismatch ratio R and $G_s$, $G_l$ and $B_l$ are computed (see Table 1).
3. The transistors input admittance ($Y_{in}$) for the load chosen in step 2 is computed using the formula,

$$Y_{in} = y_i - ((y_f y_r)/y_o + Y_l)) \qquad (3)$$

4. Once $Y_{in}$ is known, set $B_s$ equal to the negative of the imaginary part of $Y_{in}$.
5. The gain of the stage is calculated using equations given in Table 2.

### Matching Network Design

The y-parameters of the device at the frequency of interest are first calculated, as well as the source and load admittances for the required mismatch and gain at center frequency.

The 50 ohm load impedance must be transformed to the optimum load for the MOSFET. This transformation can be performed by the network shown in Figure 2. The equations for computing the values of inductive and capacitive elements of the matching networks are given in Table 2. An advantage of the matching network is that the bandwidth of the amplifier can be adjusted using the capacitive elements C2 and C3.

The inductance used in the circuit is an aircore device. The number of turns for the unit is calculated by the subroutine "INDUCT" when the length of the coil, internal diameter and the wire diameter are specified.

Input matching network calculations are performed in a similar manner. In any practical RF amplifier, it is important that the circuit be well-bypassed to ground at the signal frequency, since only a small impedance to ground may cause instability or loss of gain. The

bypass capacitor should be such that the reactance is about 1 or 2 ohms at the operating frequency. These conditions are met in the design by a proper choice of capacitors C5, C6 and $C_d$.

### Gain Versus Frequency Curve

Two types of gains have been defined for any RF amplifier — power gain ($G_p$) and transducer gain ($G_t$). Expressions for these gains are given in Table 2.

When the input of the amplifier is conjugately matched to the generator, $G_p$ becomes equal to $G_t$. Hence, at the frequency of interest when the circuit is conjugately matched at the output and input, the transducer gain will be maximum, whereas at other frequencies below and above center frequency there is a decrease in gain. This fact is best illustrated in Figure 3. The gains at frequencies other than the center frequency are calculated by the subroutine "PLOT." This graph illustrates the measure of amplifier selectivity.

### Computer-Aided Design Example

A data file consisting of 15 discrete frequencies along with corresponding y-parameter values is to be given as the input to the FORTRAN 77 program. Once the data file is fed in, the program computes the coefficients of the 10th degree polynomial that fits the data. The polynomial coefficients for computing both real and imaginary parts of the y-parameters are given in Table 3.

With the help of these coefficients, the y-parameters at any given frequency in the range of 20-300 MHz may be computed. Graphs of actual and curve-fit for some of the parameter values are shown in Figures 4 and 5. These graphs clearly illustrate the closeness of the actual and fitted values.

The y-parameters at the center frequency are substituted in the Linvill equation to determine the stability of the device. The calculated value of the stability factor is 0.560. The result clearly shows that the device is unconditionally stable at the center frequency (200 MHz). Figure 6 depicts the Linvill factor versus frequency curve for 3N187.

The source and load admittances for simultaneous conjugate match at 200 MHz are determined using equations given in Table 1. The Stern factor (K) for this condition is 4.175. The computer values of gains are:
Transducer gain ($G_t$) = 21.5 dB
Power gain ($G_p$) = 21.5 dB
Equality in gains clearly illustrates the case of simultaneous conjugate match.

### Matching Network Design (4)

Networks shown in Figure 2 are used to match the input and output impedances of the transistor to 50 ohm source and load impedances. The computed values of the components of the networks for matching at the center frequency are as follows:

a) Output network components:
C3 = 1.7 pF (0.6 to 4.5 pF); C4 = 3.8 pF (0.6 to 4.5 pF); L1 = 0.087 $\mu$H; Coil diameter = 10 mm; Coil length = 2 mm; Wire diameter = 1 mm; Number of turns = 2.

b) Input network components:
C1 = 2.9 pF (0.6 to 4.5 pF); C2 = 11.5 pF (1.4 to 15 pF); L1 = 0.03 $\mu$H; Coil diameter = 10 mm; Coil length = 3 mm; Wire diameter = 1 mm; No. of turns = 2.

For devices other than 3N1887, it is

necessary to create a new y-parameter data file. For use with bipolars, in addition to creating a new y-parameter data file, the bias network design has to be slightly modified. Figure 7 gives the flowchart for the CAD of dual gate MOSFET RF amplifiers.

## Experimental Results

A 3N187 dual gate MOSFET was chosen for the experimentation, and a set of y-parameters (for the desired operating point) was taken from datasheets. After entering this data, along with center frequency and bandwidth information, the program produced the information needed to build the circuit.

The measured gain of the amplifier at 200 MHz is 20 dB, which compares favorably with the design figure of 21.5 dB. The difference may be attributed to the insertion loss introduced by the matching network and device-to-device variations in y-parameters. The theoretical and measured (using the HP 8566B spectrum analyzer) response curves of the amplifier are given in Figure 3. It is seen that the gain curve predicted by the program closely matches the experimentally measured one.

## Acknowledgements

The authors wish to thank Shri. V.R. Katti, division head, and Shri. K.N.S. Shamanna, section head, for their encouragement and help. Special thanks are due to Mr. T.V.S. Rambabu for his help in developing the software.

Further information on the software may be obtained directly from the authors.

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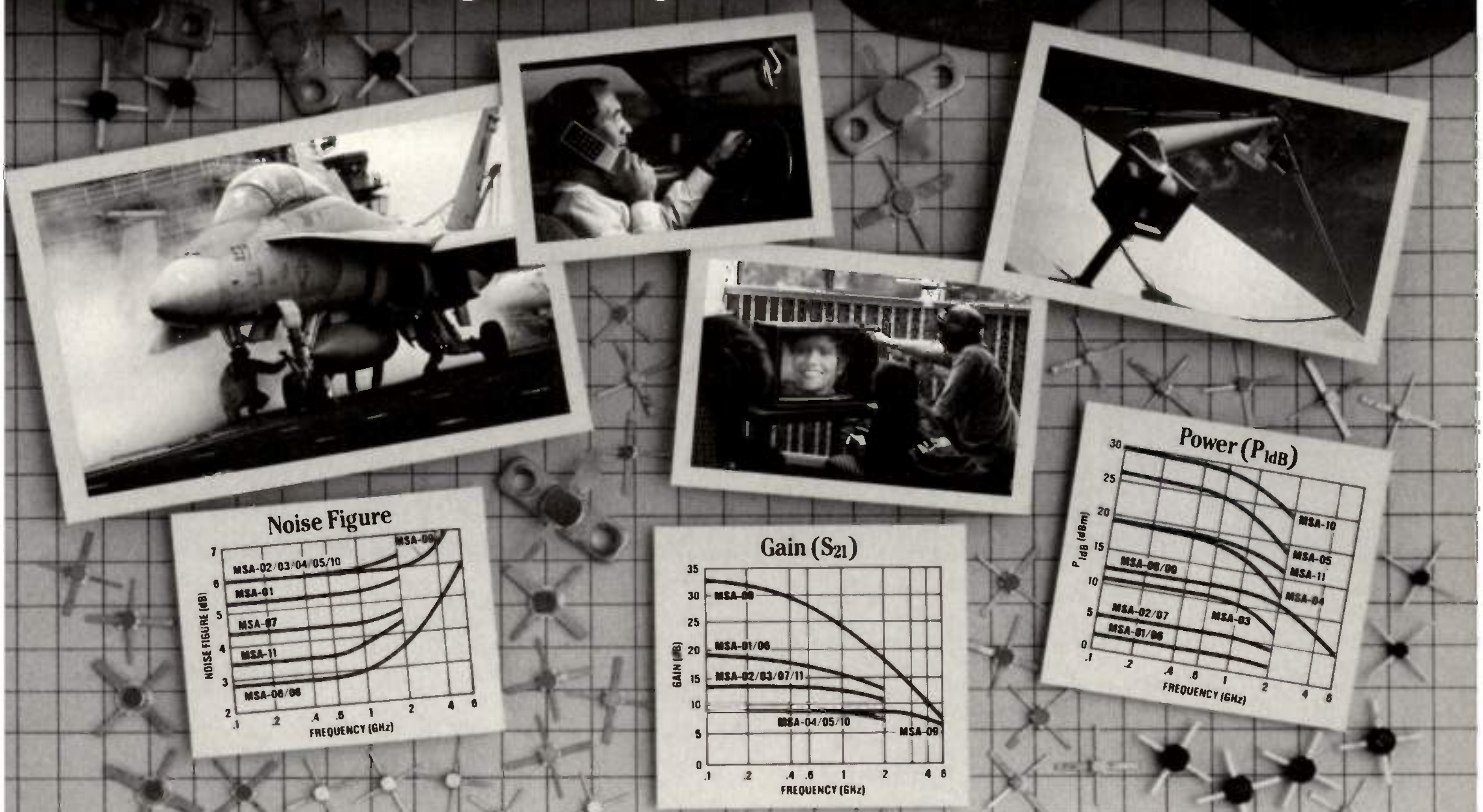


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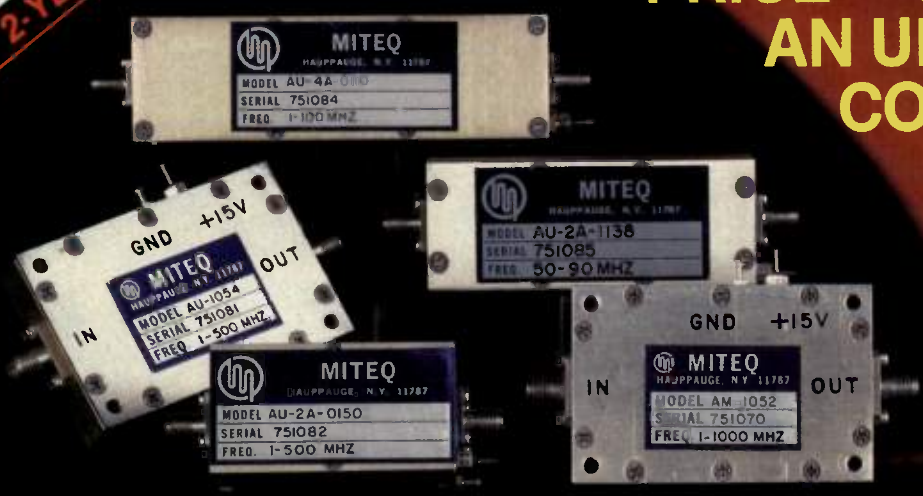


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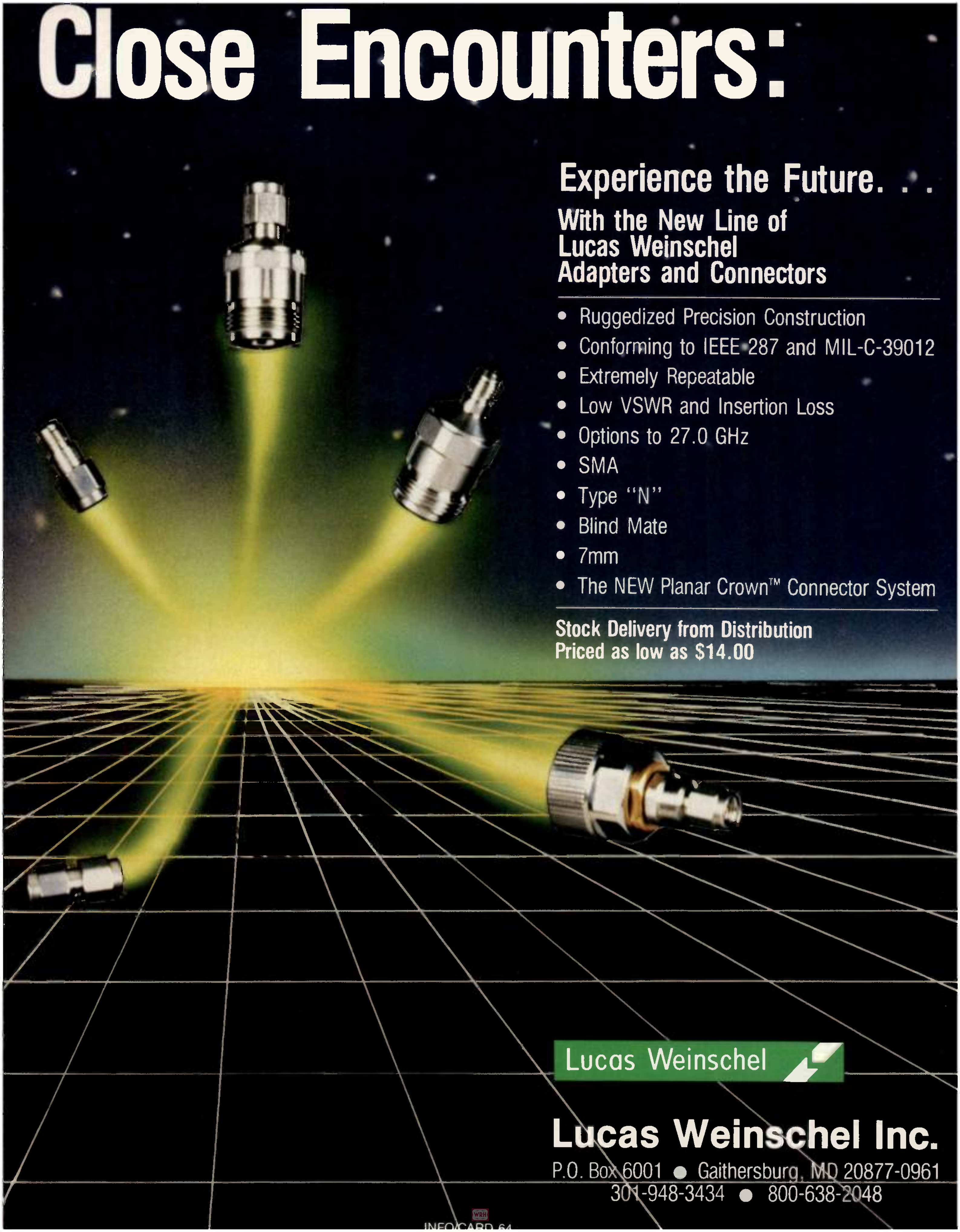
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*Gerald Hall, Editor*
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This book differs from previous amateur publications in several ways. First, the outright size of the book (over 700 pages in 8 1/2 x 11 in. format) indicates the scope of coverage. Also, many of the classic antenna types are accompanied by radiation patterns modeled using MININEC, a program developed by the Naval Ocean Systems Center for analysis of antennas constructed from wire or tubing. Antenna systems for space communications is another area where new information is available to the amateur, with both land-based and spacecraft antenna configurations described.

The book includes current information on loop, yagi, vertical, log periodic, quad, and long-wire antennas, plus substantial information on arrays of antenna elements. Chapters on safety, radiowave propagation, transmission lines, matching and coupling methods, measurements and calculations provide the necessary support for the antenna descriptions.

The only notable error is in the chapter on log periodic antennas. The book uses data from the early log periodic work by Carrell, which was later found to provide inaccurate directive gain computations. Fortunately, this error can easily be corrected by subtracting 1.5 dB from all of the gain figures given in that chapter.

This book should prove valuable to an engineer who needs to know about the construction of practical antennas. When more information is needed, there are extensive references given at the end of each chapter, citing resources in both amateur and engineering literature.

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*By Andrew Bateman*
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*Andrew F. Inglis, Editor-in-Chief*
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*Sprague Semiconductor Group*

*This article describes a noise blanker for AM-band entertainment radios which is capable of virtually eliminating impulse noise caused by automotive ignitions, SCR power controls, meteorological disturbances, and other sources. The system is simple, requiring none of the often-used complex audio-frequency phase shift networks. This allows it to be almost entirely realized in a monolithic integrated circuit. The system (Figure 1) takes advantage of differences in bandwidth and timing between the RF and IF selectivity.*

Noise has been a problem almost from the beginning of over-the-air communication. The first reference to a method of noise reduction at the receiver seems to be in an 1899 British patent by none other than Lord Rayleigh. Although it was a two-diode arrangement, which today would more properly be called a "noise limiter," it suggests the pervasive nature of impulse noise.

To describe the function of noise blanking it is first necessary to examine the nature of impulse noise and its effect on AM receivers. The common spark plug noise is a true impulse with a low repetition rate. Spikes from power supply diodes and SCR line controls are also fast rise time, low repetition rate impulse noise sources. A fast rise time square wave is a good model for this type of noise. Different types of noise, generated by power lines, brush motors, and fluorescent lamps, can be much more complicated, producing numerous closely spaced spikes that repeat at the line rate.

Noise blankers can be divided into two basic categories:

1. *Audio blanking* detects the incoming noise pulse in the RF or IF, then blanks the audio. This has the disadvantage that nothing is done to protect the receiver RF and IF from overload and the noise pulse is stretched to a value equivalent to at least the period corresponding to the entire selectivity of the receiver (see Figure 2). This produces a long and variable-length audio pulse.

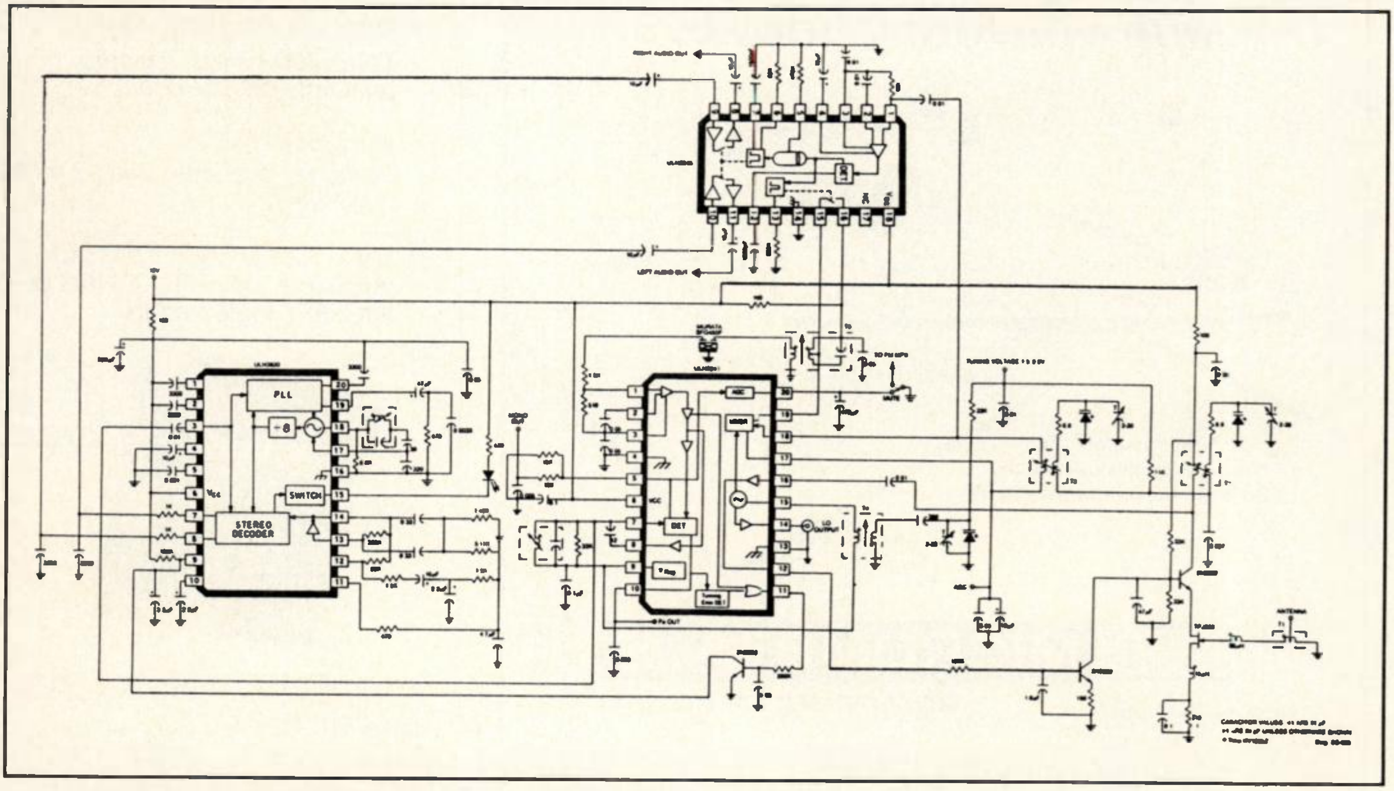


**Figure 1. ETR AM stereo receiver with noise blanking.**

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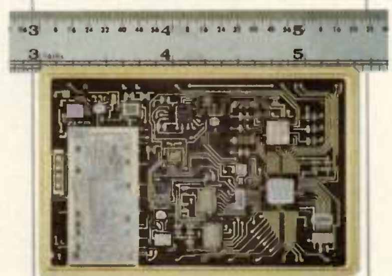

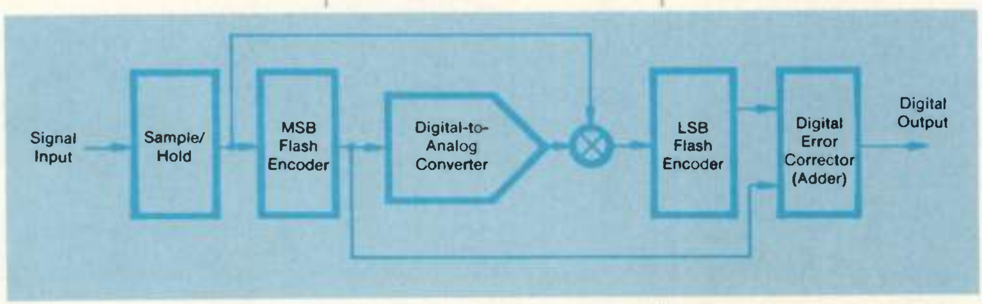


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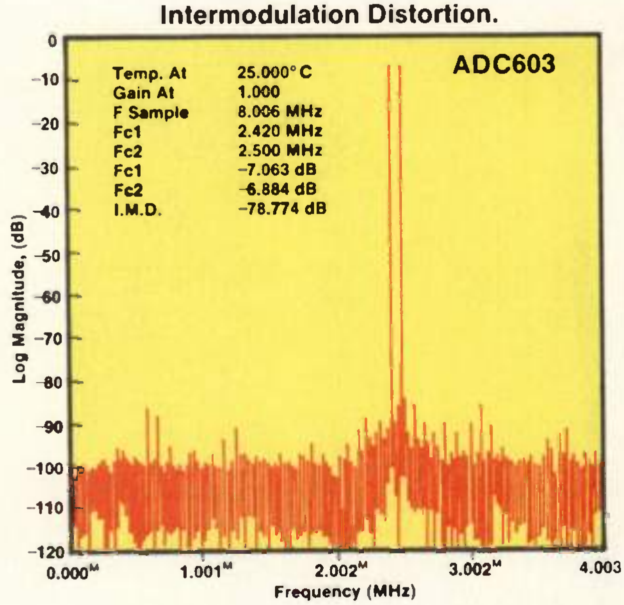


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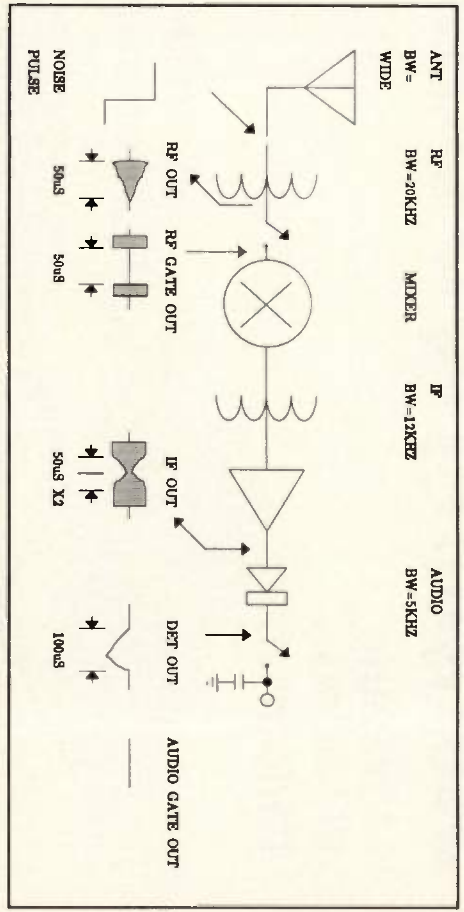


Figure 2. Noise pulse progressing through a typical AM receiver.

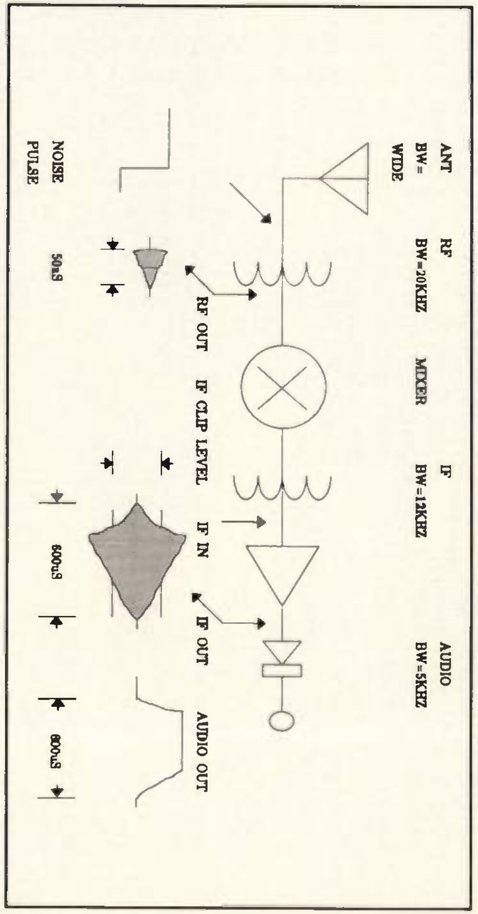


Figure 3. Sprague AM noise blanker.

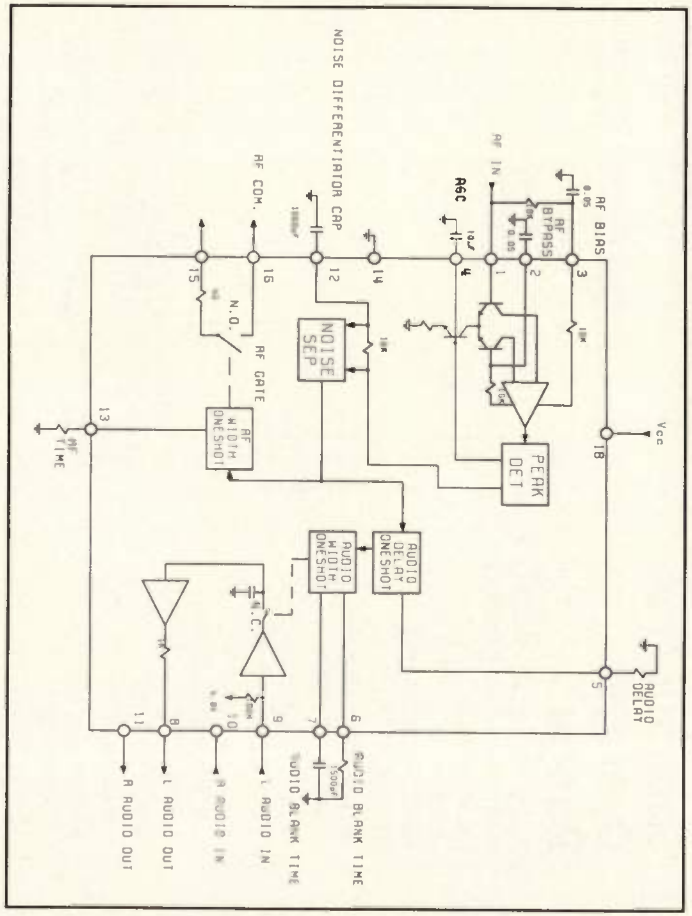

Figure 4. ULN3845A block diagram.

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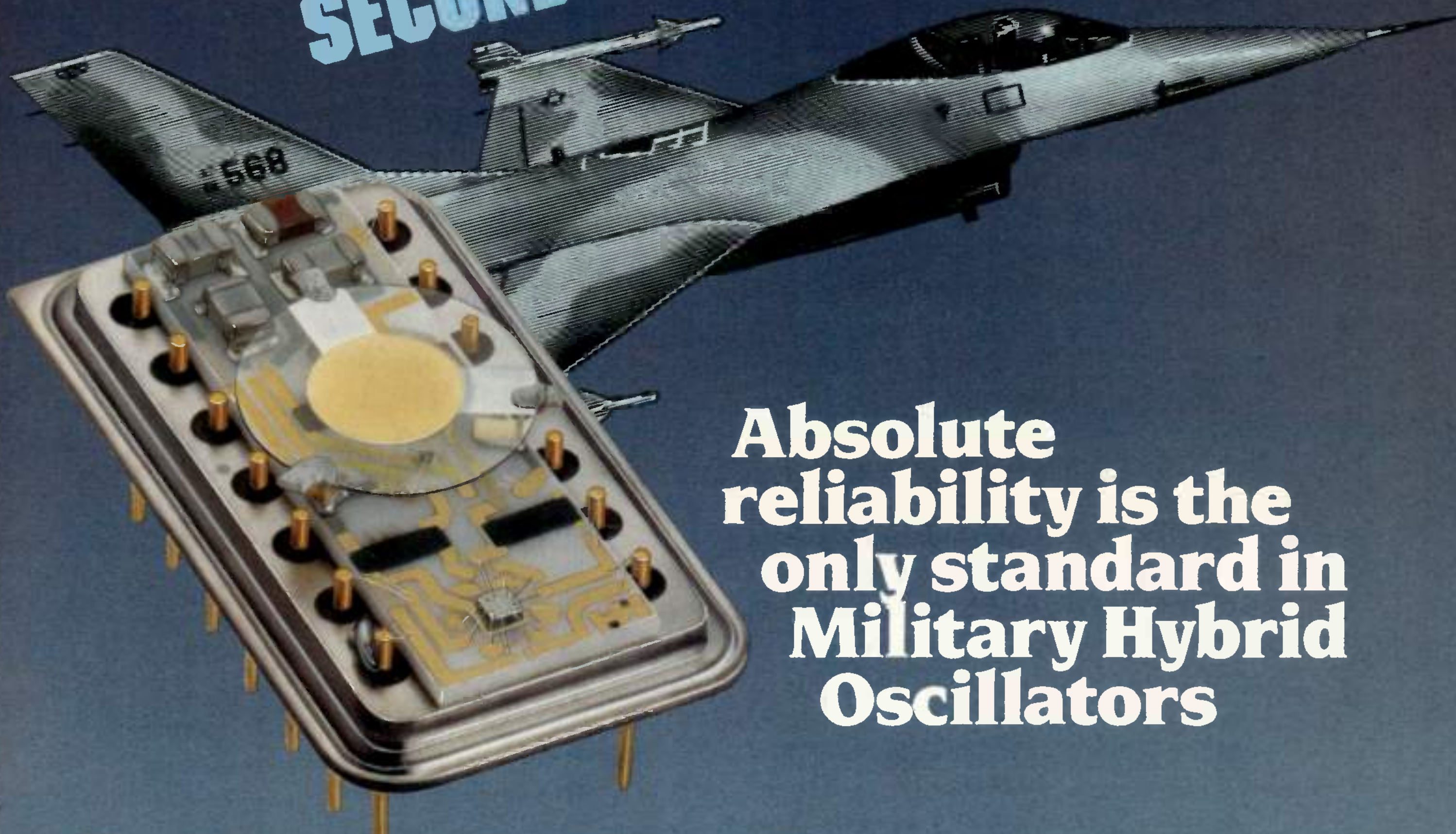

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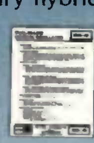

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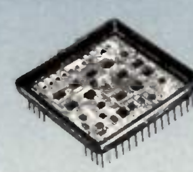

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Since the pulse is stretched to a length determined by the IF bandwidth, the pulse contains audio components that are less than the desired maximum audio response, which is also determined by the IF passband. At higher noise pulse levels, the audio pulse is even longer. Under very-high pulse levels, or rapid pulse repetition rates, the receiver AGC may also be activated to the point that the desired signal is heavily attenuated. The slow overall response time of this type of blanker makes it poorly suited for eliminating fast repetition rate noise pulses.

Audio blanking does have the advantages of simple timing requirements and good noise elimination under unmodulated carrier conditions. However, the long blank period, typically 600 $\mu s$, produces audible damage to the audio content.

*2.* A more advanced system known as *Lamb noise blanking* (1) puts the switch in the RF or IF, ahead of most of the gain and selectivity. This overcomes much of the stretching and AGC blocking described earlier. Variations of this system are widely used in communications and navigation gear today.

### A New Insight

The system described here can be viewed as an improvement of the Lamb system to achieve the performance requirements of an entertainment receiver. The Lamb system can blank the incoming noise pulse at a point in the receiver RF that has wide enough bandwidth to permit a narrow blanking pulse which is effectively faster than the audio response and the IF bandwidth. This does, however, leave a hole in the carrier (and modulation) of the desired signal. This hole can only approach 100 percent negative modulation (zero carrier). It is faster than the response time of the IF selectivity, and therefore is filled by the finite decay and rise time of the IF selectivity. The duration of the hole after the IF selectivity is the fall time commencing at the loss of carrier (at the time of opening of the RF switch), plus the rise time starting at the restoration of carrier (upon closure of the RF switch). Since both fall and rise time are essentially the same, the hole in the carrier is stretched, theoretically to twice the RF blank time, and is reduced in amplitude.

This system (Figure 3) recognizes the synergistic relationship between RF blanking, the brief noise created by the hole, and the role of audio gating. The holes introduced into the carrier are an annoying crackle which would be unacceptable in entertainment receivers. Due to the short period of the hole, a sample-and-hold gate can be employed in series with the audio (operating at slightly more than twice the RF gate period) to remove the crackle created by the holes. The audio gate timing is determined by the RF bandwidth, not the IF bandwidth that would be the case in audio-only blanking. The short nature of this gate period, and its sample-and-hold implementation, place all of the sound created by the audio blanking beyond the highest frequency that can be reproduced by the IF and can be filtered out at the audio frequency without a loss of frequency response, as determined by the IF bandwidth.

## A Practical System

A monolithic IC has been developed, the ULN3845, incorporating the noise pulse detector, RF gate, and dual audio gate (for AM stereo applications) needed to implement this noise blanking system. A block diagram of the ULN3845 is shown in Figure 4. Note that the input stage is the open base of a differential amplifier. The overall sensitivity of the RF amplifier is about 5 $\mu$V at 1 MHz, at the antenna.

If the blanker is connected to the RF stage output, a noise signal of 40 $\mu$V will trigger the blanker. The noise differentiating capacitor from the threshold circuit to ground is connected to one side of a differential amplifier which has both inputs supplied from the detector through resistors. The capacitor is selected so the audio signals do not cause triggering, but impulses do.

The RF gate is controlled by the RF one-shot, and its blanking time is determined by a resistor from the control pin to ground (500k ohms gives a blanking time of about 50 $\mu$s). The ON resistance of the series FET is about 50 ohms. The circuit is set up internally so that when an impulse occurs, the shunt FET turns on after a delay of about 0.5 $\mu$s. The internal circuit also includes capacitors to the input and output of the gates so that switching transients are cancelled and do not appear at the output. These features ensure transient-free switching even when the RF gate is connected to the low-level input stages of a receiver.

As mentioned above, blanking in the RF section of the receiver removes most of the interference, but a small amount still remains due to the hole punched in the carrier. This residual noise is theoretically only twice the RF blanking pulse width and much smaller than that which the impulse would normally produce in a receiver without blanking. The audio delay, audio one-shot, and audio gates are included to eliminate this residual signal.

The same trigger signal to the RF one-shot also goes to the audio delay one-shot. The delay is set with an external resistor and the amount required depends on the IF filtering characteristics of the particular receiver design. After the audio delay time, the audio one-shot is triggered. Its output pulse is controlled by an external resistor and capacitor, with 1 nF and 120k ohms giving a blanking time of about 150 $\mu$s.

The ULN3845 includes two sample-and-hold audio gates for stereo applications. The FET gates are also compensated so switching transients are not fed through. A MOSFET buffer with a capacitor to ground is connected to the output of the gates to form a sample-and-hold circuit. There is also an internal buffer amplifier feeding the output, providing a low output impedance, on the order of 1000 ohms.

A typical application using the ULN3845 in a car radio AM tuner is shown in Figure 1. Note that the RF input and RF gate are common. Although there is a 0.5 μs delay from the start of the impulse to the start of blanking, this is small compared to the impulse response time of the receiver; it takes almost 10 μs for the RF burst to reach 70 percent amplitude at the mixer input. The RF input could also be connected to the collector of the RF amplifier, but the bandwidth is much wider there, and false triggering may occur from other strong AM signals. Another possibility is to connect the RF input to a small sense antenna, since the device is useful to approximately 30 MHz. The sensitivity would be much lower and the bandwidth much greater for this arrangement, but impulse noise is usually much stronger that the received signal, and will usually trigger the blanker.

### Conclusion

Although this noise blanker was originally conceived for entertainment equipment, the principle of operation is applicable to any receiver which has an RF or IF bandwidth that is substantially wider than the audio bandwidth. rf

### References


1. James J. Lamb, "A Noise Silencing IF for Superhet Receivers," *QST*, February 1936.



**About the Author**

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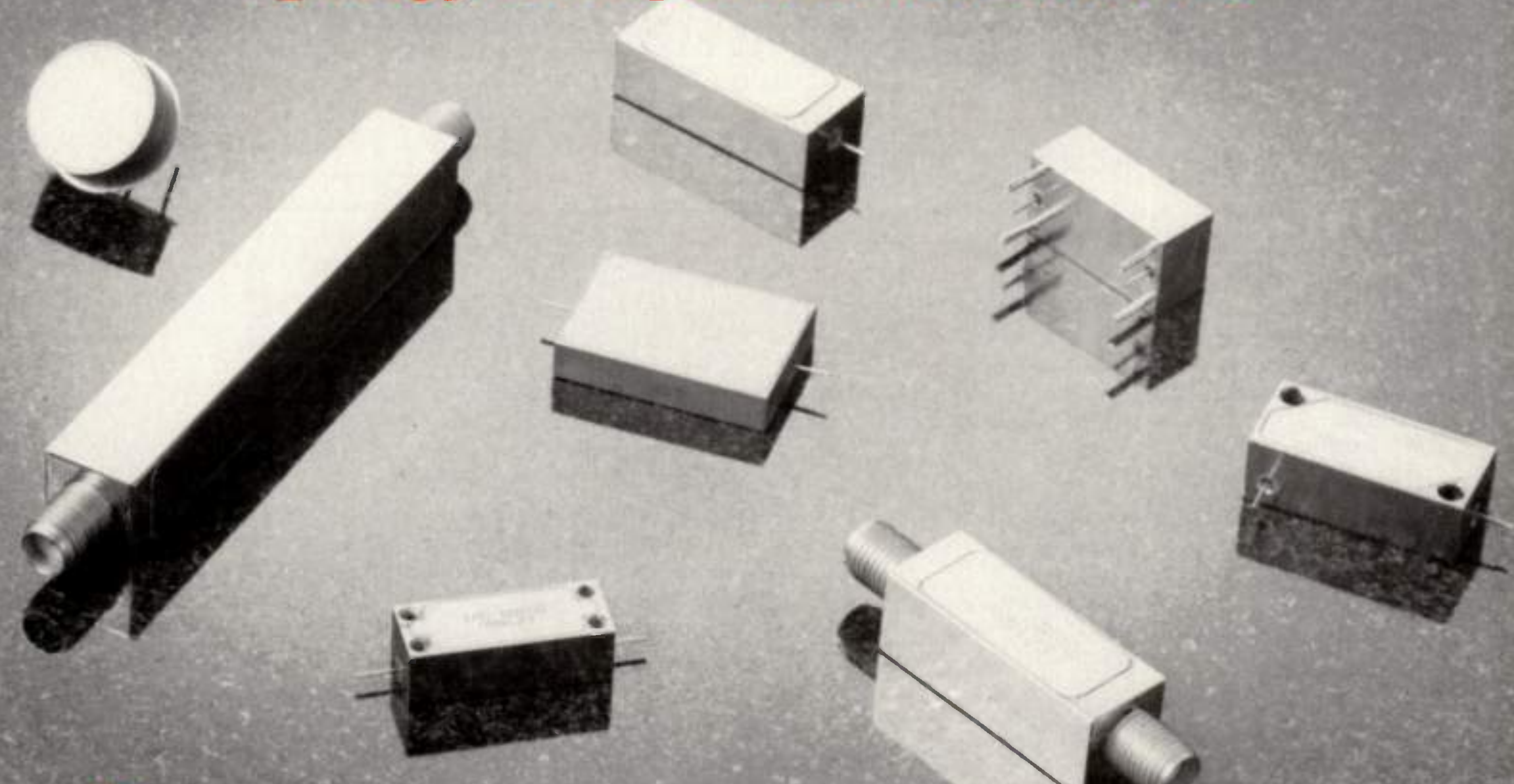

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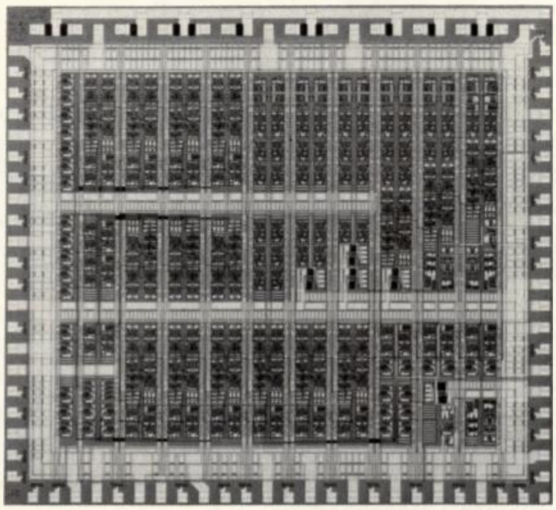
# New Products Featured at RF Technology Expo 89

## 32-Bit DDS Accumulator from GigaBit



The 10G102 32-bit phase accumulator generates digital sine or cosine functions when used with an external 14GX048 sine or cosine lookup table ROM. It can be used in digital signal processing or in conjunction with a D/A converter to provide high resolution frequency synthesis with virtually instantaneous frequency switching.

The 10G102 maintains a record of phase, which is accurate to 32 bits of resolution. The 12 MSBs of the accumulator represent the current phase of the synthesized sine or cosine function. Specifications include 1 GHz typical clock frequency, under 20 ns frequency switching time and a 12-bit parallel output.

Also from GigaBit is a 512 X 8-bit mask programmable ROM with 1 GByte/s access rate, 1 ns typical address access time, 500 ps typical OE access time and ECL compatible I/O levels. Model 14GX048 features open source follower outputs which permit wire-OR connection in bus organized systems. **GigaBit Logic, Inc., Newbury Park, CA. Please circle INFO/CARD #209.**

### Miniature RF Power Amplifier
### Alliance Technologies

The AT225-50 is a miniature RF power MOSFET amplifier module. The frequency range of the device is 225 to 400 MHz and output power is 50 watts. It is available with SMA or 50 ohm pin terminations, feed-through capacitor gain control and DC input pins. Price is $595 each when purchased in 250-piece quantity. **Alliance Technologies, Inc., Redmond, WA. INFO/CARD #208.**

### VXI Chassis for ATE
### Racal-Dana

Racal-Dana introduces a C-size VXIbus chassis designed to meet the requirements of both military and commercial automatic test systems. Featured is an eight-output power supply with voltages at +5, +12, -12, +26, -24, -5.2, -2, and +5 V. A 12-layer backplane that uses DIP switches to route the daisy-chained signals through the backplane, rather than requiring jumpers to be utilized across any empty slots in the chassis, is included.

Also being introduced is the Series 1260 switching system which is specified from DC to 26.5 GHz. Modules are available to handle power, signal and coaxial switching requirements. A control module is required to convert the first module in a chain to a message-based device. This module then utilizes the VXIbus local bus to control the other switching cards in a master-slave arrangement. **Racal-Dana Instruments, Inc., Irvine, CA. INFO/CARD #207.**

### MMIC Converters
### Pacific Monolithics

The PM-CVXXXX Series of MMIC converters comes with an integrated RF amplifier, mixer, and IF and LO amplifiers. Frequency range is 0.01 to 3 GHz with conversion frequencies from 20 MHz to 2 GHz. The devices are offered in surface-mount packages—0.180 in. square and 0.270 in. square. **Pacific Monolithics, Sunnyvale, CA. INFO/CARD #206.**

### SPDT GaAs Switch
### Custom Microwave Components

This non-reflective single-pole double-throw GaAs switch with integral TTL compatible driver features a frequency range of 50 to 250 MHz with insertion loss under 1.2 dB. Isolation is greater than 55 dB and switching speed is below 35 ns. In 100-piece quantity, price is $270. **Custom Microwave Components, Inc., Fremont, CA. INFO/CARD #205.**

### Cellular Chip Set
### Signetics

Signetics introduces a cellular chip set which includes an FM IF prescalar, frequency synthesizer, audio processor, data processor and microcontroller in six integrated circuits. **Signetics, Sunnyvale, CA. INFO/CARD #125.**

### Horn Antennas
### Amplifier Research

Models AT4000 and AT4001 cover the 200 to 1000 MHz and 400 to 1000 MHz frequency range, respectively. At 1000 MHz, the antennas exhibit a gain of 15 to 18 dB and field strength reaches 1000 V/m at an input power of 100 watts. **Amplifier Research, Souderton, PA. INFO/CARD #204.**

### RF Analysis and Optimization Software
### Ingsoft

RF Designer™ is a program for RF analysis and optimization. It is written for the Macintosh® and offers full-feature schematic capture, network analyzer interface and full integration with desktop publishing. **Ingsoft, Willowdale, Ontario, Canada. INFO/CARD #203.**

### Miniaturized Phase Comparators
### Merrimac

Merrimac introduces a line of 0 to 360 degree phase comparators with a frequency range of 10 to 500 MHz. The PCP-3S is also available at 30, 60, 70 and 160 MHz. At the specified center frequency, the devices have an output amplitude balance of ± 5 mV and phase error not exceeding ± 5 degrees.

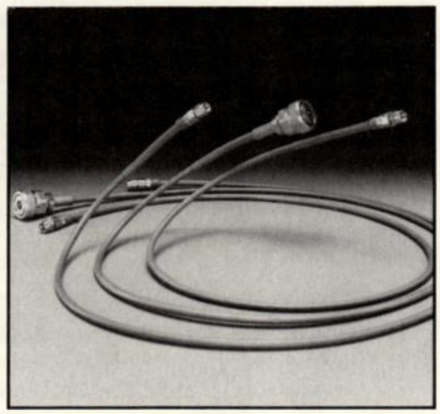
The WDP-2R Series of octave bandwidth discriminators has a typical linearity of 2 percent over one octave. Frequency range is 160 MHz to 1 GHz and the device is packaged in a 20 mm square package. Price commences at $425. **Merrimac Industries, Inc., West Caldwell, NJ. INFO/CARD #202.**

### Log Amplifiers
### Radar Technology

These miniature log amplifiers feature a frequency range of 10 to 160 MHz and dynamic range of -80 to 0 dBm. Typical sensitivity is 30 mV/dB and log accuracy is ± 1 dB. The RTL-6 measures 3.5 in. X 1.5 in. X 0.5 in. **Radar Technology, Inc., Haverhill, MA. INFO/CARD #201.**

### Reformable Cable Assemblies
### Connecting Devices

Models HF 501 and 502 feature 0.141 and 0.085 in. cable diameters respectively, terminated in 5319SF and 5285-2SF SMA male connectors. With VSWR of 1.25:1 out to 18 GHz, insertion loss of 0.9 dB/ft., the cable is ideal for calibration and test as well as operational cable assemblies. **Connecting Devices, Inc., Long Beach, CA. Please circle INFO/CARD #200.**



### MESFET MMIC Switch
### Doty Scientific

Doty introduces an SPDT switch with 4 ns rise and fall times from DC to 2 GHz. Insertion loss is 1 dB at 25 mW. **Doty Scientific, Inc., Columbia, SC. INFO/CARD #199.**

### Chip Attenuators
### Mini-Systems

Mini-Systems introduces a line of thin-film chip attenuators. Attenuation is available from 1 to 24 dB. **Mini-Systems, Inc., North Attleboro, MA. INFO/CARD #198.**

### SMT Trimmer Capacitors
### Voltronics

The trimmers are O-ring sealed, multi-turn devices with glass, air or sapphire dielectrics. The glass units have a maximum capacitance of 60 pF, Q of 500 and are usable up to 200 MHz. Units that use the air dielectric have a maximum capacitance of 14 pF, Q of 3000 and are usable up to 1.5 GHz. The specifications for the sapphire units include maximum capacitance of 8 pF and Q of 5000 with maximum frequency of 5 GHz. **Voltronics Corp., East Hanover, NJ. INFO/CARD #197.**

### Schottky Ring Quads
### M-Pulse Microwave

These devices are optimized for operation in the 1 to 40 GHz range. Five barrier heights are available and screening to MIL-S-19500 is available. **M-Pulse Microwave, San Jose, CA. Please circle INFO/CARD #196.**

### Mechanical Phase Shifter
### Sage Laboratories

Model 6705K is equipped with SMA compatible K connectors. Frequency range is DC to 26.5 GHz and the unit measures 3.83. in. X 1.50 in. X 0.38 in. plus K connectors and adjustment shaft. **Sage Laboratories, Inc., Natick, MA. INFO/CARD #195.**

### Miniature Bandpass Filters
### TTE, Inc.

The T Series covers the 50 MHz to 1 GHz and any number of poles from 2 to 6 may be specified. The filters measure from 0.5 in. X 0.8 in. X 0.38 in. to 0.5 in. X 1.5 in. X 0.38 in. **TTE, Inc., Los Angeles, CA. INFO/CARD #194.**

### Bipolar Oscillator Transistor
### California Eastern Laboratories

NE64700 features a frequency range of 2 to 26 GHz, gain bandwidth product of 11 GHz at 2 GHz and typical $h_{FE}$ of 100. Output capacitance is 0.09 pF. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #193.**

### Receiver Software
### Webb Laboratories

Webb Labs introduces Receiver Advantage, a PC-based system simulation package. It combines CommView with a spectrum analyzer emulator. The software predicts overall performance for user-entered receiving systems with pre- and post-detection signal-to-noise ratios, bit error probability, AGC characteristics and compression displayed as functions of receiver input power. **Webb Laboratories, Hartland, WI. Please circle INFO/CARD #192.**

### Connector Adapter Kit
### RF Industries

RFA-4021 contains type N, UHF and BNC adapters. The devices are built with machined brass bodies with either silver or nickel plating. **RF Industries, San Diego, CA. INFO/CARD #191.**

### Tuning Stick Kits
### American Technical Ceramics

ATC introduces a tuning stick kit that includes values from 0.1 pF to 1000 pF. It contains radial wire leaded capacitors labeled with the specific values. The kit is priced at $79.95. **American Technical Ceramics, Huntington Station, NY. INFO/CARD #190.**

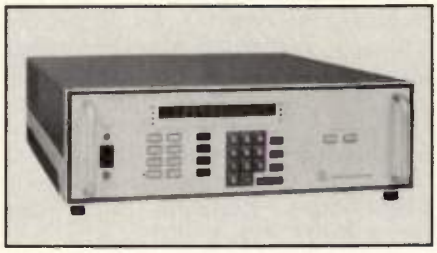
### SAW Filter Module
### Phonon Corp.

Centered at 160 MHz with a 20 MHz bandwidth, the FB160-20 filter provides greater than 70 dB rejection and a 1.2 to 1 shape factor. Maximum input power is +10 dBm and the unit is offered in a unity gain module with 75 dB dynamic range. **Phonon Corp., Simsbury, CT. INFO/CARD #189**

# rf expo products Continued

## Frequency Synthesizer
## Comstron Corp.

Comstron's FS-2000 combines fast switching, low phase noise and low spurious. It achieves direct analog synthesis over an octave range. **Comstron Corp., Melville, NY. INFO/CARD #188.**



## Synthesized Signal Generator
## Ramsey Electronics

RSG-10 is a synthesized signal generator with a 1 GHz frequency range and a calibrated output level from -127 to +7 dBm. It features 10 memory locations, membrane keyboard and a price of $2500. **Ramsey Electronics Inc., Penfield, NY. INFO/CARD #187.**

## Right Angle Cable Assemblies
## Penstock

Penstock introduces swept right angle cable assemblies that are specified up to 26.5 GHz. When purchased in 100-piece quantities, the assemblies cost $20 each.

Also being introduced are custom-built instrumentation grade cable assemblies and custom-built semi-rigid and flex cable assemblies. **Penstock, Inc., Sunnyvale, CA. INFO/CARD#186.**

## Noise Parameter Test Set
## Cascade Microtech

The NPT18 noise parameter test set simultaneously measures both S-parameters and noise parameters of active linear two-port devices with frequency coverage of 2 to 18 GHz. Measurement time is less than 10 seconds per frequency point.

The RTP Series replaceable-tip microwave probe offers performance through 26.5 GHz. The connection repeatability is better than -40 dB. Specifications include 10 dB return loss and 2 dB insertion loss. **Cascade Microtech, Beaverton, OR. INFO/CARD #185.**

## Diplexers
## FSY Microwave

FSY introduces a line of diplexers with crossover frequencies covering the 1 to 12.4 GHz range and passbands from DC to 20 GHz. Passband VSWR is 1.5:1 max and selectivity is greater than 60 dB at ± 15 percent. **FSY Microwave Inc., Rockville, MD. INFO/CARD #184.**

## Microwave Cable Assembly
## Huber + Suhner

Sucoflex 204 is a flexible microwave cable assembly for applications in temperatures up to +70 degrees C. Frequency range is DC to 26.5 GHz and attenuation is 1.95 dB at 18 GHz for 1 m inclusive of connector pair. SMA, TNC, N and PC3.5 connectors are available for the assembly. **Huber + Suhner, Inc., Woburn, MA. Please circle INFO/CARD #183.**


## Thick Film Switches
## JFW Industries

JFW will be displaying a line of RF switches including thick film, high power and relay type. The line includes internally terminated switches. **JFW Industries, Inc., Indianapolis, IN. Please circle INFO/CARD #182.**

## Voltage Controlled SAW Oscillators
## RF Monolithics

RFM introduces a line of voltage controlled SAW oscillators that cover up to 2 GHz. Typical SSB phase noise for a 500 MHz VCSO is -115 dBc at 1 kHz offset. The packaged device measures 1.27 in. X 0.77 in. X 0.020 in. **RF Monolithics, Inc., Dallas, TX. Please circle INFO/CARD #181.**

## 24-Pin DCXOs
## Vectron

The Series CO-285W is available for frequencies up to 500 MHz and the CO-286W is designed for frequencies from 500 MHz to 1.2 GHz. Both are available with output levels up to +13 dBm into 50 ohms. Also being introduced is the Model CO-233GEQ ECL oscillator which uses GaAs to provide fast switching for frequencies up to 600 MHz. **Vectron Laboratories, Inc., Norwalk, CT. INFO/CARD #180.**

### Translation Software
### Step Electronics

The ASM 500 AutoCAD-to-Gerber translation software is optimized for use in RF, microwave and hybrid circuits. This PC-based package is designed to reduce mask preparation time. **Step Electronics, Inc., Campbell, CA. Please circle INFO/CARD #179.**

### PC Board Shields
### Instrument Specialties

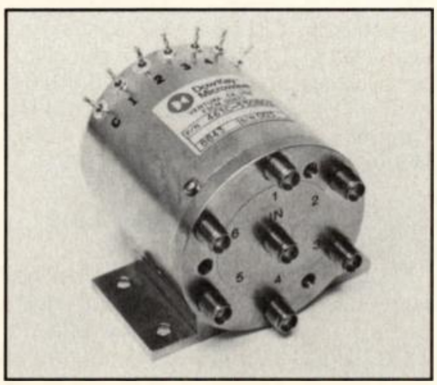
The PC board shield permits shielding of PC board components without the effort of forming punches and dies to create specially shaped boxes. It controls EMI emissions, susceptibility and cross-talk and can be used on any through-pin, multi-layer board designed to accept it. **Instrument Specialties Co., Inc., Delaware Gap, PA. Please circle INFO/CARD #178.**

### MOSFET Amplifier
### Kalmus Engineering

Model 702LC covers 18 octaves of bandwidth from 3 kHz to 1000 MHz at a power output level of 1.5 watts. The device has a 33 dB gain and is priced at $1,295. **Kalmus Engineering International, Ltd., Woodinville, WA. Please circle INFO/CARD #177.**

### Coaxial Switch
### Dow-Key

The 461 Series multi-position switches are available from SP3T to SP6T. Isolation is 60 dB minimum at 18 GHz and VSWR is 1.5:1 max with insertion loss of less than 0.5 dB. **Dow-Key Microwave Corp., Ventura, CA. Please circle INFO/CARD #176.**



### Current Feedback Amplifier
### Harris Semiconductor

Model HA-5004 is a current feedback video/wideband amplifier with 100 MHz bandwidth unity gain. This value drops to 65 MHz at a gain of 10. It is suited for applications such as a video gain block, high-speed peak detector, fiber optic transmitter, zero insertion loss transmission line driver, current-to-voltage converter and radar systems. **Harris Corp., Semiconductor Sector, Melbourne, FL. INFO/CARD #175.**

### SPDT Switch
### KDI/triangle Electronics

KDI introduces an SPDT switch that uses high-speed PIN diodes to achieve 10 ns switching speeds. VSWR is less than 1.5:1 and insertion loss is typically

1.5 dB. **KDI/triangle Electronics, Whippany, NJ. INFO/CARD #174.**

### Digital Storage Oscilloscope
### John Fluke Mfg.

The Philips PM 3308 digital storage oscilloscope from Fluke features a 100 MHz bandwidth with a maximum 40 Ms/s sampling rate on one channel and 8 kbyte acquisition memory. It displays up to four channels simultaneously plus computed traces based on addition, subtraction, multiplication, division, integration and differentiation. GPIB and RS 232 interfaces are standard. **John Fluke Mfg. Co., Inc., Everett, WA. INFO/CARD #173.**

### Waveform Analyzer
### Tektronix

Model 2510 is a multi-channel analyzer that features a card modular expansion capability for up to eight acquisition channels. The channel cards have a 50 kHz bandwidth and independent real-time sampling on each channel for simultaneous waveform digitizing and storage. Waveform acquisition, storage analysis and data management are integrated with control through a spreadsheet-style user interface.

Also being introduced is the 2402 TekMate — a software and hardware product that offers waveform processing, storage and communication capabilities. It is compatible with the 2400 Series digital oscilloscopes and includes two 3.5 in., 720k floppy disks for storing waveform procedures and system software. This package is priced at $2,990. **Tektronix, Inc., Beaverton, OR. Please circle INFO/CARD #172.**

### Miniature SMA Connectors
### Texscan

Texscan unveils their FP-18 Series of miniature 18 GHz SMA connectors. These minipads meet MIL-A-3933 and are available in 1, 2, 3, 6, 10 and 20 dB values. VSWR at 18 GHz is 1.35:1 while average power input is 2 watts CW. **Texscan Corp., Indianapolis, IN. Please circle INFO/CARD #171.**

### Single Channel Power Meter
### Boonton

The Model 4220 has a 4 1/2 digit readout which displays measurements in dBm or watts. The frequency range is 100 kHz to 110 GHz, and depending on the sensor used, levels from -70 to +30 dBm can be measured. An analog meter is provided for nulling or peaking. Price is $1750 ($2250 with GPIB). **Boonton Electronics Corp., Randolph, NJ. INFO/CARD #170.**

### Variable Attenuators
### Alan Industries

Alan Industries introduces two 50-ohm variable attenuators. Model 50HV9 has a frequency range of DC to 18 GHz with an attenuation range of 0 to 9 dB in 1 dB steps and is priced at $625. Model 50HV60 has a frequency range of DC to 12.4 GHz with an attenuation range 0 to 60 dB in 10 dB steps and costs $525. **Alan Industries, Inc., Columbus, IN. INFO/CARD #169.**

### Small Step Frequency Synthesizers
### Communication Techniques

This synthesizer is available as a stand-alone VHF synthesizer or as part of a microwave synthesizer unit which uses frequency agile phase-locked signal sources to multiply the VHF signal to frequencies up to 23 GHz. The VHF unit is available in bands from 30 MHz to 500 MHz while the microwave synthesizer is available from 500 MHz to 23 GHz.

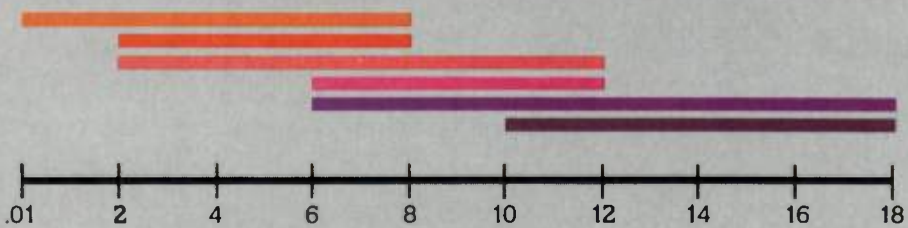
Also from CTI is a VHF synthesizer with a 90 to 154 MHz range. Specifications include phase noise of 113 dBc/Hz at 100 kHz and 132 dBc/Hz at 10 kHz. **Communication Techniques, Inc., Whippany, NJ. INFO/CARD #168.**

### Digitally Controlled Delay Line
### TriQuint Semiconductor

TQ6411 is a digitally controlled gallium arsenide-based delay line which provides a selectable pipeline delay for digital signals up to 1 GHz. It features

# It's a Giga-Sweeper


## The Microwave Sweeper with Auto-Lock




Frequency Ranges (GHz)

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Giga-tronics Series 610 Microwave Sweepers are all you expect from a sweeper, and a lot more.

- Sweep speeds to 600MHz/msec for fast, real-time testing
- Five (8 via the bus) frequency identifying markers
- F1/F2, marker to marker and $\Delta$CF sweep modes with direct BW readout
- Ten milliwatts of leveled output power and 119dB of attenuation
- Excellent CW phase noise characteristics
- Wide range AM and FM
- Save/recall of complete panel settings and IEEE bus control

Only the Series 610 has them all.

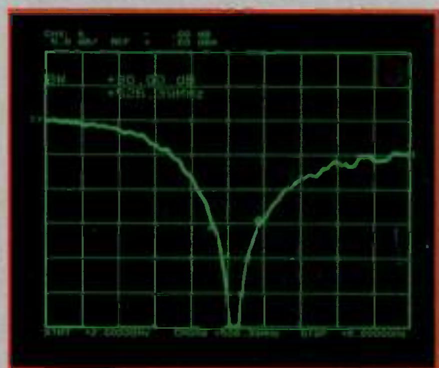
### Plus Auto-Lock

Auto-Lock means CW accuracy and stability. Accuracy is 1 part per million; a few kilohertz vs the old standard of tens of megahertz. Stability is equally good, over time, temperature and line or load variation. It also provides digital lock at the start of each sweep. Like adding a $6,000 lock-box or source synchronizer, but Giga-Sweeper has it built-in at no extra cost; the perfect match for your network analyzer.

### And More

More Giga-Sweeper extras; pulse modulation, spectral purity and step and lock mode. Pulse modulate a Giga-Sweeper with 100nsec pulses, you'll get on/off ratios >80dB and rise/fall times <25nsec. Harmonics (>40dB down) are 15 to 20dB better than your old sweeper. Spurs are >55dB down. CW SSB phase noise typically <-75dBc/Hz at 10kHz offset. Step and lock stops at precise frequencies, giving you time to record, or make adjustment, before stepping on.

A Giga-Sweeper is a sweeper, a lock-box and a signal generator for the price of the sweeper. Prices start at about $15,000 and don't go over $22,000.



For all your sweeper requirements, talk to the microwave signal generator people.

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# rf expo products *Continued*

500 ps delay-chain steps and is controlled by a six-digit word capable of producing 64 discrete steps for a total delay of 32 ns. The digital inputs and differential delayed signal outputs are ECL compatible. **TriQuint Semiconductor, Beaverton, OR. INFO/CARD #167.**

## Probe Sets
## Hewlett-Packard

HP introduces three probe sets which provide DC to 250 MHz coverage with 1 mohm and 8 pF input loading. The probes are accessories for the HP 8980A and 8981A vector-modulation analyzers. Maximum input voltage is ± 1 V or ± 10 V with the furnished 10:1 dividers.

HP will also show surface mount PIN and Schottky diodes. **Hewlett-Packard Co., Palo Alto, CA. INFO/CARD #166.**

## Optical Spectrum Analyzer
## Advantest

The Q8382 optical spectrum analyzer has a 60 dB dynamic range at 1 nm and 50 dB at 0.5 from the peak wavelength. The measurement range covers 0.6 to 1.75 μm with a resolution of 0.1 nm and measurement accuracy of 0.5 nm at 25 degrees C. **Advantest America, Inc., Lincolnshire, IL. INFO/CARD #165.**

## High Data Rate Amplifier
## Avantek

Avantek introduces a 0.1 to 4000 MHz amplifier combining 19 dB gain, ± 0.5 dB full-band gain flatness, low pulse overshoot and less than 1.8:1 input and output VSWR. The ACT-4032 is suitable for use in high data rate ( 1 GHz) fiber optic systems as well as in pulse amplification and instrumentation applications. **Avantek, Inc., Santa Clara, CA. Please circle INFO/CARD #164.**

# Design of Line Matching Networks

*By Peter Martin*
*University of Manchester*

*The use of a Smith chart to design transmission line matching networks is well known. However, the process can be time-consuming and perhaps error-prone, especially if the chart has to be renormalized to accommodate lines of differing characteristic impedance.*

The use of simple analytic expressions, which can be handled by a microcomputer or even a programmable calculator, offers an alternative method. Consider the case of two complex admittances, $Y_1$ and $Y_2$, connected by a transmission line of length $\phi$ (degrees) and characteristic admittance Y (Figure 1).

$$Y_1 = G_a + jB_a \tag{1}$$

$$Y_2 = G_b + jB_b \tag{2}$$

The conditions for this transmission line to match the two admittances are expressed as:

$$Y = \sqrt{\frac{(G_a^2 + B_a^2)\,G_b - (B_b^2 + G_b^2)\,G_a}{G_a - G_b}} \tag{3}$$

$$\phi = \tan^{-1}\left[\frac{Y\,(G_b - G_a)}{B_aG_b - B_bG_a}\right] \tag{4}$$

A common requirement is to design a matching network between a conductance (real) and an admittance (complex). This admittance can be represented by $Y_2$ and the conductance by $G_a$. The complex function, $jB_a$, can then represent either a capacitive or inductive element depending on its sign (+ and – respectively).

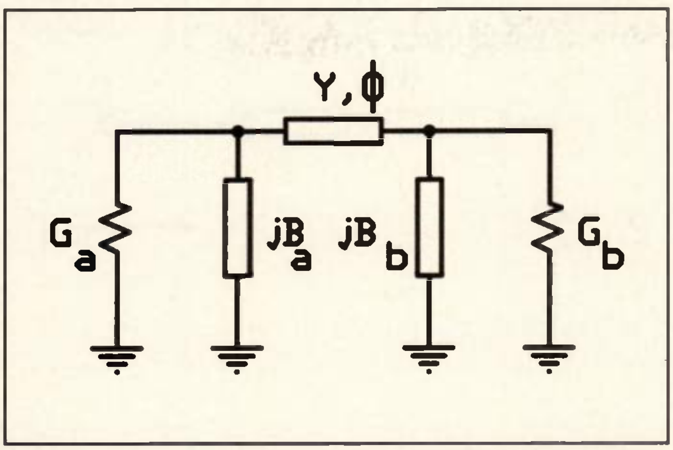


**Figure 1. Two complex admittances connected by a transmission line.**

The variables in equations 1 to 4 can be more easily handled by normalizing to the system conductance ($Z_0^{-1}$) and expressing the values as lower case. Putting $G_a = 1$ and rearranging equation 2 for $b_a$ gives:

$$b_a = \pm\sqrt{\frac{y^2(1 - g_b) + b_b^2 + g_b^2 - g_b}{g_b}} \tag{5}$$

The length of the associated transmission line has two solutions:

$$\phi_1 = \tan^{-1}\left[\frac{y\,(g_b - 1)}{|b_a|g_b - b_b}\right] \quad \text{for } b_a > 0 \tag{6}$$

$$\phi_2 = \tan^{-1}\left[\frac{y\,(1 - g_b)}{|b_a|g_b + b_b}\right] \quad \text{for } b_a < 0 \tag{7}$$

It is useful to calculate which values of $g_a$ can be directly matched to $y_2$ with $b_a = 0$. Solving equation 3 for $g_a$ gives:

$$g_a = \frac{(b_b^2 + g_b^2 + y^2) \pm \sqrt{(b_b^2 + g_b^2 + y^2) - 4y^2g_b^2}}{2g_b} \tag{8}$$

It is often more convenient to express $Y_2$ as a reflection coefficient, such that Γ represents Y*. Consider three matching networks that make the transformation coefficient as represented on a Smith chart normalized to $Z_0$. With $\Gamma = Y_2^*$, $b_b$ and $g_b$ are expressed as:

$$b_b = \frac{2r\sin\phi}{1 + 2r\cos\phi t r^2} \tag{9}$$

$$g_b = \frac{1 - r^2}{1 + 2r\cos\phi + r^2} \tag{10}$$

Figure 2a represents a matching network consisting of a short circuit stub and a transmission line whose impedances $Z_1$ and $Z_2$ are specified. The lengths $n_1$ and $n_2$ are given by:

$$n_1 = \cot^{-1}\left(\frac{Z_1\,|b_a|}{Z_0}\right) \tag{11}$$

$$n_2 = \tan^{-1}\left[\frac{y\,(1 - g_b)}{|b_a|\,g_b + b_b}\right] \tag{12}$$

$$\text{where } y = \frac{Z_0}{Z_2} \text{ and } |b_a| = \sqrt{\frac{y^2(1 - g_b) + b_b^2 + g_b^2 - g_b}{g_b}}$$

Figure 2b represents a matching network consisting of an open circuit stub and a transmission line whose impedances $Z_3$ and $Z_4$ are specified. The lengths $n_3$ and $n_4$ are given by:

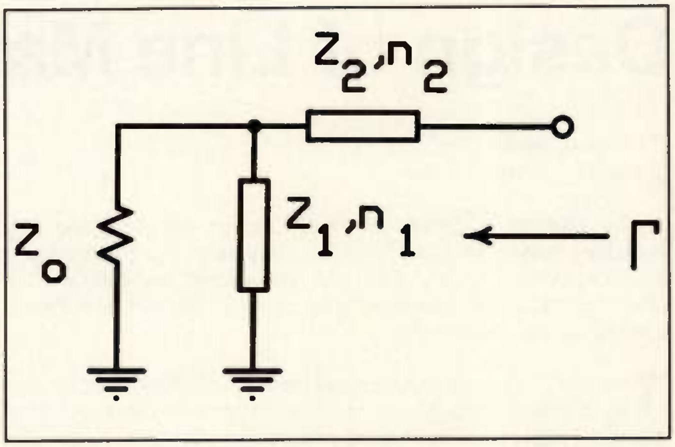


Figure 2a. Matching network consisting of a short circuit stub and transmission line.

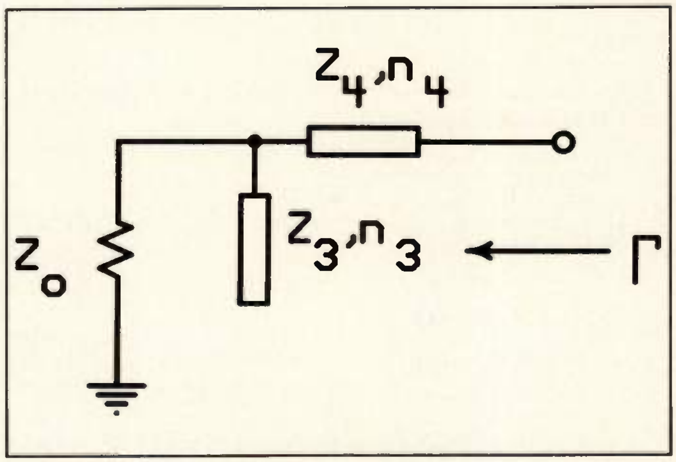


Figure 2b. Matching network consisting of an open circuit stub and transmission line.

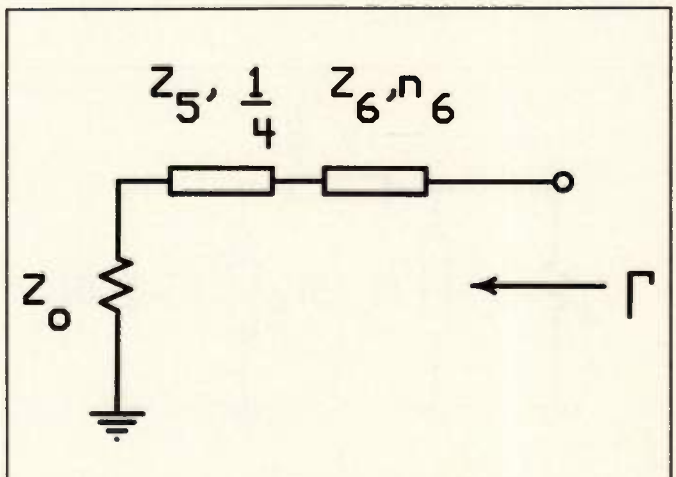


Figure 2c. Matching network consisting of a quarter wave transformer and transmission line.

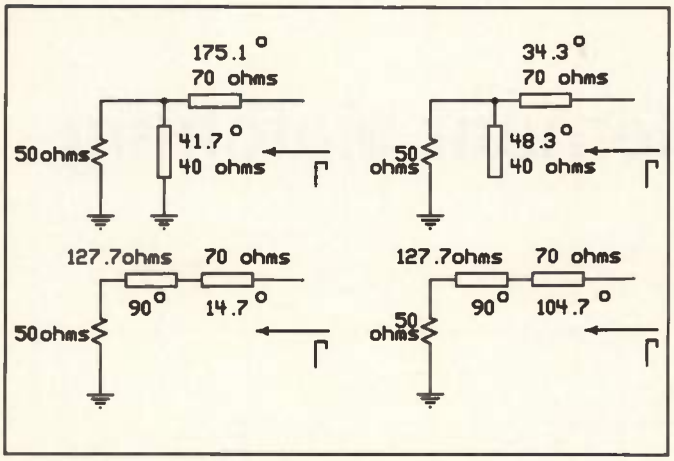


**Figure 3. Transformation examples.**

$$n_3 = \tan^{-1}\left(\frac{Z_3}{Z_0}|b_a|\right); \quad n_4 = \tan^{-1}\left[\frac{y\,(g_b - 1)}{|b_a|\,g_b - b_b}\right] \qquad (13,14)$$

$$\text{where } y = \frac{Z_0}{Z_4} \text{ and } |b_a| = \sqrt{\frac{y^2(1 - g_b) + b_b^2 + g_b^2 - g_b}{g_b}}$$

Figure 2c represents a matching network consisting of a quarter wave transformer of impedance $Z_5$ and transmission line of specified impedance $Z_6$ and length $n_6$. There are two associated solutions for the impedance of $Z_5$ and length $n_6$:

$$Z_{5A} = \frac{Z_0}{\sqrt{g_{a1}}} \qquad (15)$$

$$Z_{5B} = \frac{Z_0}{\sqrt{g_{a2}}} \qquad (16)$$

$$n_{6A} = \tan^{-1}\left[\frac{(g_{a1} - g_b)y}{b_b g_{a1}}\right]; \quad n_{6B} = \tan^{-1}\left[\frac{(g_{a2} - g_b)\,g}{b_b g_{a2}}\right] \qquad (17,18)$$

where $y = \frac{Z_0}{Z_6}$ and $g_{a1}$, $g_{a2}$ are as defined in Equation 7.

A BASIC program called MATCH has been written to calculate the above expressions. As an example, four possible solutions to effect the transformation of 50 ohms to Γ =0.56 at 138 degrees are shown in Figure 3.

### References


1. Kurokawa, "Power Waves and the Scattering Matrix," *IEEE Trans. on Microwave Theory and Techniques*, Vol. MTT-13, No. 2, March 1965.
2. *S Parameters, Circuit Analysis and Design*, HP Application Note 95, Hewlett-Packard, 1967.
3. R. Compton and D. Rutledge, *PUFF*, California Institute of Technology, 1987.


### About the Author


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# CAD For Lumped Element Matching Circuits

*By Stanley Novak*
*Polytechnic University, Brooklyn*

*The necessity of matching arbitrary impedances to resistive impedances of transmission lines, or to any other impedance, requires the use of matching circuits with ideal reactive elements. Such elements can be realized with inductances or capacitances or by transmission lines. Note that for simplicity (and to obtain relations which are easily solved), it is normal to ignore losses in matching elements. This is particularly common at high frequencies where matching can be employed effectively, since the resulting values of the elements are small.*

The simplest networks used in RF are shown in Figures 1a and 1b, and are called "L", "T" and "PI" networks. In reality, the T and PI networks are derived from the L network. This is shown here by comparing the equations for elements of matching networks for various cases.

Each network can be characterized by its quality factor "Q". It is important to realize how this factor is calculated for various networks. Confusion is usually generated by the use of parallel and series equivalent networks for the impedances which are to be matched.

In normal circumstances, arbitrary impedance in its series form Z=R+jX is used. Unfortunately, the same symbols are frequently used to describe parallel combinations of the resistive and reactive elements, which can be more appropriately defined by Y=G+jB. This more clearly represents the parallel transformation of Z, and should be used solely for describing parallel combinations.

When using equations for various matching networks, the designer must use both series and parallel forms. It is important to remember that when an element in a matching network is in parallel with the terminals of the matching network, the adjoining impedance to be matched must be converted to a

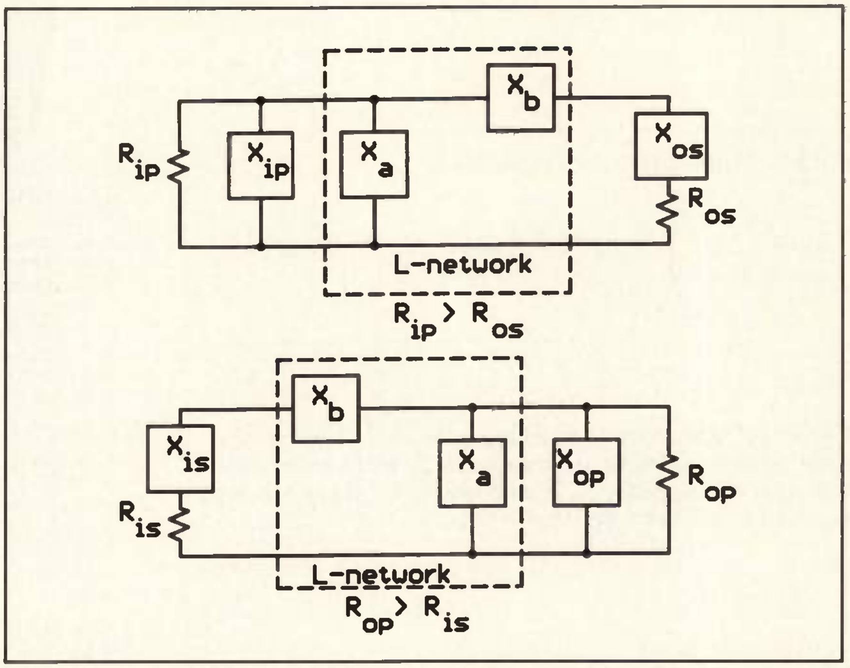


**Figure 1a. Configurations for two possible L networks.**

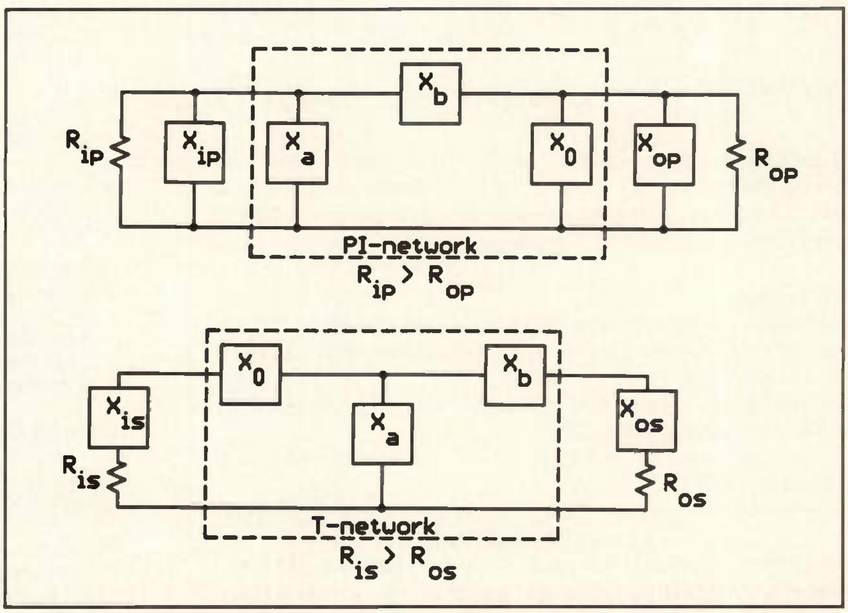


**Figure 1b. Configurations for PI and T networks.**

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**ST1016**

| RF | | DC* | |
|---|---|---|---|
| FREQ: | 400 MHz | gM: | .8 ℧ |
| PO: | 20 W | CISS: | 28 pF |
| GAIN: | 13 dB | CRSS: | 3.5 pF |
| VDS: | 28 V | COSS: | 19 pF |

**REPLACES:** ACR UMIL20FT; PHO UF2840G; POL F1016

**COMMENTS:** Usable in push-pull configurations to 2 GHz.

**TYPICAL APPLICATION:** 10 W, Class AB, 15 dB gain, 100-400 MHz broad band.

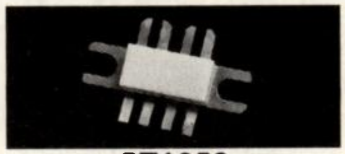
**ST1018**

| RF | | DC* | |
|---|---|---|---|
| FREQ: | 550 MHz | gM: | 2.4 ℧ |
| PO: | 80 W | CISS: | 84 pF |
| GAIN: | 10 dB | CRSS: | 10.5 pF |
| VDS: | 28 V | COSS: | 57 pF |

**REPLACES:** PHO 28100V; POL F1018

**COMMENTS:** High power, broad band, push pull.

**TYPICAL APPLICATION:** 100-500 MHz broad band power amplifiers.



**ST1053**

| RF | | DC* | |
|---|---|---|---|
| FREQ: | 1000 MHz | gM: | .8 ℧ |
| PO: | 50 W | CISS: | 28 pF |
| GAIN | 10 dB | CRSS: | 3.5 pF |
| VDS: | 28 V | COSS: | 19 pF |

**REPLACES:** ACR 0510-50; POL F1053

**TYPICAL APPLICATION:** Dual push pull, 500-1000 MHz

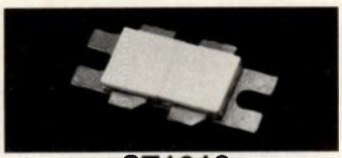
**ST1027**

| RF | | DC* | |
|---|---|---|---|
| FREQ: | 175 MHz | gM; | 4.8 ℧ |
| PO: | 250 W | CISS: | 168 pF |
| GAIN: | 10 dB | CRSS: | 21 pF |
| VDS: | 28 V | COSS: | 114 pF |

**REPLACES:** MRF 171G; PHO DU28200M; POL F1027; POL F3001

**COMMENTS:** Gemini package, very high power to 450 MHz.

**TYPICAL APPLICATION:** Broad band, 225-400 MHz, 200 W



**ST1012**

| RF | | DC* | |
|---|---|---|---|
| FREQ: | 550 MHz | gM: | 2.4 ℧ |
| PO: | 80 W | CISS: | 84 pF |
| GAIN: | 10 dB | CRSS: | 10.5 pF |
| VDS: | 28 V | COSS: | 57 pF |

**REPLACES:** PHO DU2880V; POL F1012

**COMMENTS:** Gemini package

**TYPICAL APPLICATION:** 100-500 MHz broad band

**ST1008**

| RF | | DC* | |
|---|---|---|---|
| FREQ: | 400 MHz | gM: | 1.6 ℧ |
| PO: | 40 W | CISS: | 56 pF |
| GAIN: | 13 dB | CRSS: | 7 pF |
| VDS: | 28 V | COSS: | 38 pF |

**REPLACES:** ACR UMIL40FT; PHO UF2840G; POL F1008

**TYPICAL APPLICATION:** 20-400 MHz, Class AB, broad band.

*** Per side**

***Other Devices Available:*** ST1001; ST1002; ST1003; ST1004; ST1005; ST1006; ST1007; ST1009; ST1013; ST1014; ST1015; ST10017; ST1020; ST1021; ST1022.

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| | $X_1$ | $X_2$ | $X_\varnothing$ | Q |
|---|---|---|---|---|
| PI | $Q(R_{op})$ | $\frac{-1}{[Q/R_{ip} + 1/X_{ip}]}$ | $-X_{op}$ | $\sqrt{[R_{ip}/R_{op}] - 1}$ |
| L | $Q(R_{os}) - X_{os}$ | $\frac{-1}{[Q/R_{op} + 1/X_{op}]}$ | | $\sqrt{[R_{ip}/R_{os}] - 1}$<br>$\sqrt{[R_{op}/R_{is}] - 1}$ |
| T | $Q(R_{os}) - X_{os}$ | $-R_{is}/Q$ | $-X_{is}$ | $\sqrt{[R_{is}/R_{os}] - 1}$ |

**Table 1. Comparison of circuit elements for the three network configurations.**



parallel form for all subsequent calculations. If the element in the matching network is in series with regard to the terminals, the adjoining impedance to be matched must be converted to a series form.

Now, using $R_p$ and $X_p$ for the parallel form (although it is more appropriate to use G and B), and $R_s$ and $X_s$ for the series form, plus the letters i for input and o for output, then orienting the network so that the higher resistance (series or parallel, depending on the first element in the matching network), is on the left side (i.e. "input" as shown in Figures 1a and 1b), the quality factor Q for the T network is:

$$Q = \sqrt{\frac{R_{is}}{R_{os}} - 1}$$

Note that in the T network, both impedances are in series form since the adjoining elements are in series with the terminals. For the PI network (impedances oriented as before),

$$Q = \sqrt{\frac{R_{ip}}{R_{op}} - 1}$$

The situation with the L network is more complicated. In many cases the designer could realize two values for Q, depending on the magnitude of resistances in parallel and series transformation. Assuming that the matching element in parallel with the terminals is on the left, the parallel transformation of the impedance to the matched must have a larger value than the series resistance of the second impedance (which is in series with the series element of the matching network). In this case,

$$Q = \sqrt{\frac{R_{ip}}{R_{os}} - 1}$$

Now, in the case where the parallel resistance of the second impedance is larger than the resistance of the series form of the first impedance, another matching network is obtained with the equation below:

$$Q = \sqrt{\frac{R_{op}}{R_{is}} - 1}$$

Obviously, in each case there are two solutions, since a square root can be either positive or negative.

### Similarity of the Three Basic Configurations

Since the T and PI networks are

variations of the basic L network, one could expect similarities between various equations used for calculating the elements of the networks. Arranging the elements of matching networks so that $X_a(X_1, (X_3))$ are series elements, $X_b(X_2, (X_4))$ are parallel elements and $X_\varnothing$ an additional element to complete the design for T or PI networks to be series or parallel respectively, the elements can be arranged in Table 1 as shown.

Note that the suffixes assigned to the real part of the input impedance $R_i$ or output impedance $R_o$ indicate whether the equivalent network for the impedance is parallel or series. For example, $R_{ip}$ indicates that a parallel equivalent network for input impedance must be used.

This approach makes the realization of a simple BASIC program easy. Considering that equations are repeated for more than one matching circuit, a flow graph, as shown in Figure 2, can be used to write a short program which will take care of the calculations for all three matching circuits.

To include the conditions for designing T and PI matching circuits with higher than minimum Q, called $Q_n$, substitute $X_\varnothing$ and $X_a$ for the T network or $X_\varnothing$ and $X_b$ for the PI network, as given in Table 2.

Higher values of Q will result in narrower bandwidth than that for minimum Q. This is sometimes undesirable. However, wider bandwidths than that obtained here can only be obtained by using multiple element networks.

Since Qs are square roots, alternate solutions using the opposite signs in the expressions can be obtained. This results in values for $X_3$ and $X_4$. These alternate solutions are, for the sake of convenience, called "highpass" in contrast to original "lowpass" solutions. Note that this is actually valid for certain cases only. The same convenience is used for extended equations where square roots are also present. Traditionally, the lowpass networks do have a higher number of poles than zeros and highpass have a higher number of zeros than poles.

Since equations for higher Q in the cases of PI and T circuits are valid for all Qs including minimum Q (a special case), these equations can be used for all cases, thus avoiding the necessity to branch the program more than needed. This keeps the delay in processing minimal.

The complete program given here takes care of all possible combinations.

| | $X_1$ | $X_2$ | $X_\varnothing$ |
|---|---|---|---|
| PI | $\dfrac{R_{ip}[Q_n + \sqrt{(R_{os}/R_{ip})(1+Q_n^2) - 1}]}{1+Q_n^2}$ | $\dfrac{-1}{[Q/R_{ip} + 1/X_{ip}]}$ | $\dfrac{1}{-[1/R_{op}\sqrt{(R_{op}/R_{ip})(1+Q_n^2) - 1} + 1/X_{op}]}$ |
| T | $QR_{os} - X_{os}$ | $\dfrac{R_{os}(1+Q_n^2)}{Q_n + \sqrt{[(R_{os}/R_{is})(1+Q_n^2) - 1]}}$ | $-R_{is}\sqrt{(R_{os}/R_{is})(1+Q_n^2) - 1} - X_{is}$ |

**Table 2. Extended values for $Q_n > Q$.**

The user can realize matching between impedances, impedance to resistance, or resistance to resistance, and the program provides all the available solutions for all cases. It also indicates if a solution is not available. Examples of solutions for various types of matching networks are given in Tables 3 to 7, with indication of data parameters entered at the program end.

### Computer Program For Calculation of Matching Elements

Since the program was designed as a subroutine for a program evaluating amplifiers (2,3), the numbering starts at 2610. The initial parameters are stored in a data line in the following order: frequency (in MHz), real and imaginary parts of the first impedance ($R_s$, $X_s$), and real and imaginary parts of the second impedance ($Z_\varnothing$, $l_l$). Impedance may be entered arbitrarily, as the program sorts the higher appropriate resistance accordingly as the equations require. This is done in lines 2680-2730. The program then proceeds to evaluate the matching elements for a selected type of matching network. Calculations are done for both types of networks, lowpass and highpass, from which the operator may choose the desired network. Calculated elements may be obtained by eliminating REM in lines 3160 and 3170.

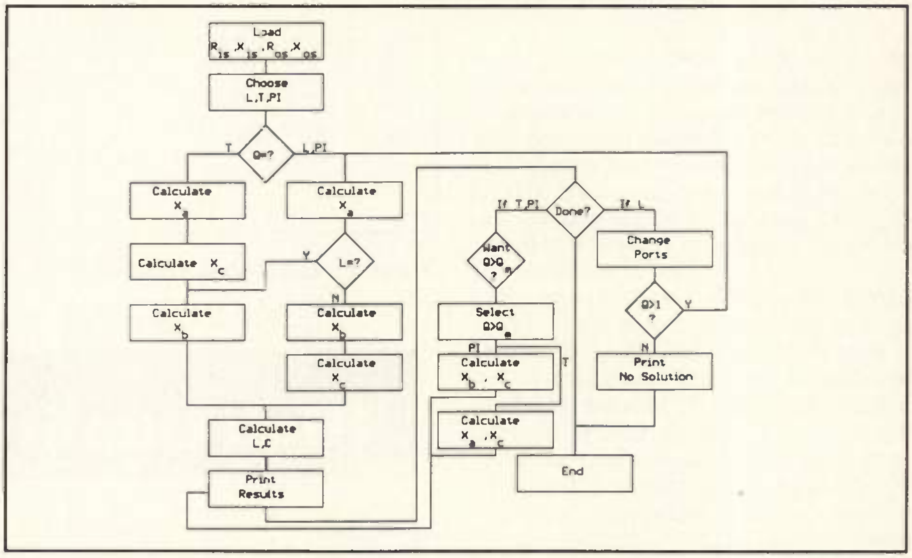


**Figure 2. Flow graph for lumped parameter matching for L, T and PI networks.**

In the case of the L network, it is sometimes possible to find another solution. In this case the operator is provided with the additional solution as well. In the case of a T or PI network

```
CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? L

LOW OR HIGH PASS MATCH?[L/H]? L
RH= 4.13 OHMS, RS= 2 OHMS    ,QUALITY Q= 4.897

LOW-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P           S

XP= 49.871 [PF],  XS=  6.04  [NH]

RH= 2 OHMS, RS= 4.13 OHMS    ,QUALITY Q= 1.96

LOW-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P           S

XP= 394.725 [PF],  XS= 34.784  [NH]


CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? L

LOW OR HIGH PASS MATCH?[L/H]
RH= 4.13 OHMS ,RS= 2 OHMS    ,QUALITY Q= 4.897

HIGH-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P           S

XP= 9.663999 [NH],  XS= 100.76 [PF]

RH= 2 OHMS, RS= 4.13 OHMS    ,QUALITY Q= 1.96

HIGH-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P           S

XP= 82.74 [PF],  XS= 9.014 [NH]
```

**Table 3. L network calculations.**

there is only one solution, but it is possible to find solutions for higher Q if desired. In all cases a simple graphic printout is provided indicating the placement of elements in relation to the resistive parts of the impedances to be matched at a particular frequency.

The whole process for evaluating various networks is best shown on the flow graph, which illustrates the procedure. It is shown that, as some equations are the same for different networks, the computer branches to equations suited for a particular selection of the network and ignores non-required equations. When this is done, the actual values of elements in picofarads and nanohenries are calculated and displayed.

In some cases only a limited number of solutions are available, particularly when one or both impedances are resistive only. Examples for such cases are given in Tables 6 and 7 with appropriate data entries, showing that only a limited number of solutions exist. Some border line cases may result in negative or zero Qs. For these cases the results should be judged with caution as the program does not have any error-trapping routines. *The program described in this article is available on disk. See page 127 for details.*

[Program listing and Tables 4 - 7 are on page 108.]

## References


1. A.B. Przedpelski, "Simplify Conjugate Bilateral Matching," *Electronics Design 5*, March 1, 1978, pp. 54-62.
2. S. Novak, "Computer Enhanced Amplifier Design," *RF Design*, Feb. 1987, pp. 97-109.
3. S. Novak, "CAD Amplifier Matching With Microstrip Lines," *RF Design*, June 1988, pp.43-53.


### About the Author

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```
2600 ' LUMPED ELEMENT MATCHING PROGRAM. S.NOVAK,MAY 88
2610 PRINT"PROGRAM FOR CALCULATION OF LUMPED MATCHING NETWORKS"
2620 E=1000:PI=3.14159
2630 'PARAMETERS;F=FREQUENCY(MHZ);RS,XS FIRST IMPEDANCE;Z0,II-SECOND IMPEDANCE
2640 READ F,RS,XS,Z0,II
2650 'CALC.L,T,PI MATCH ************************************
2660 PRINT:PRINT"CALCULATION FOR MATCH";LEFT$(F$,6);" AT";F/1000000!;"MHZ"
2670 INPUT"SELECT L, T OR PI NETWORK MATCH [L/T/P]";A$
2680 IF Z0>RS THEN RH=Z0
2690 IF Z0>RS THEN XH=II
2700 IF Z0<RS THEN RH=RS
2710 IF Z0<RS THEN XH=XS
2720 IF Z0<RS THEN XS=II
2730 IF Z0<RS THEN RS=Z0
2740 PRINT
2750 INPUT"LOW OR HIGH PASS MATCH?[L/H]";C$
2760 GOSUB 2900
2770 IF A$="T" OR A$="P"THEN 2830
2780 R2=RH:RH=RS:RS=R2:R1=XH:XH=XS:XS=R1
2790 PRINT
2800 GOSUB 2920
2810 GOTO 3650
2820 R2=RH:RH=RS:RS=R2:R1=XH:XH=XS:XS=R1:GOTO 2920
2830 INPUT"WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]";B$
2840 IF B$="N" THEN 3650
2850 INPUT"SELECT Q>QMIN";Q
2860 IF Q<QM THEN 2880
2870 GOSUB 3020
2880 GOTO 2830
2890 A$="L":PRINT"*** ONLY L-MATCH AVAILABLE! ***":GOTO 2920
2900 'CALC.LUMPED ELEMENTS **************************
2910 IF A$="T"AND XH=0 THEN 2890
2920 X=RH*RH+XH*XH:Y=RS*RS+XS*XS
2930 Q=X/(RH*RS)-1
2940 IF A$="T" THEN Q=RH/RS-1
2950 IF A$="P" THEN Q=(X/RH)/(Y/RS)-.999999
2960 IF A$="P"AND INT(Q)=-1 THEN 2820
2970 IF A$="P"AND XS=0 THEN 2890
2980 IF INT(Q)=-1 THEN RETURN
2990 Q=SQR(Q)
3000 QM=Q:'STORE MIN.Q
3010 PRINT"RH="INT(RH*E)/E;"OHMS";",RS="INT(RS*E)/E;"OHMS",
     ",QUALITY Q=";INT(Q*E)/E
3020 A=1+Q*Q
3030 IF A$="T" THEN 3110
3040 X2=-X/(Q*RH+XH)
3050 X4=-X/(-Q*RH+XH)
3060 IF A$="L" THEN 3140
3070 X1=(X/RH)*(Q+SQR((Y/RS)/(X/RH)*A-.999999))/A
3080 X3=-(X/RH)*(Q-SQR((Y/RS)/(X/RH)*A-.999999))/A
3090 X0=-1/(RS/Y*SQR((Y/RS)/(X/RH)*A-.999999)+XS/Y)
3100 GOTO 3160
3110 X2=-RS*A/(Q-SQR((RS/RH)*A-1))
3120 X4=RS*(1+Q*Q)/(Q+SQR((RS/RH)*A-1))
3130 X0=-RH*SQR((RS/RH)*A-1)-XH
3140 X1=Q*RS-XS
3150 X3=-Q*RS-XS
3160 'PRINT"XLP-SER","XLP-PAR","XHP-SER","XHP-PAR","X0"
3170 'PRINT INT(X1*E)/E,INT(X2*E)/E,INT(X3*E)/E,INT(X4*E)/E,INT(X0*E)/E
3180 'CALC.ELEM.VALUES ****************************
3190 PRINT
3200 IF C$="H" THEN D$="HIGH-PASS"
3210 IF C$="L" THEN D$="LOW-PASS"
3220 PRINT:D$;" SOLUTION FOR ";A$;" NETWORK"
3230 IF A$="L" THEN GOSUB 3490
3240 IF A$="T" THEN GOSUB 3540
3250 IF A$="P" THEN GOSUB 3590
3260 IF C$="L" THEN 3280
3270 X1=X3:X2=X4
3280 IF X0=>0 THEN 3320
3290 E$="[PF]"
3300 X0=-1/(2*PI*F*X0)*1E+12
3310 GOTO 3340
3320 E$="[NH]"
3330 X0=X0/(2*PI*F)*1E+09
3340 IF A$="T" OR A$="P"THEN PRINT"X0=";INT(X0*E)/E;E$;", ";
3350 IF X1=>0 THEN 3390
3360 F$="[PF]"
3370 X1=-1/(2*PI*F*X1)*1E+12
3380 GOTO 3410
3390 F$="[NH]"
3400 X1=X1/(2*PI*F)*1E+09
3410 IF X2=>0 THEN 3450
3420 G$="[PF]"
3430 X2=-1/(2*PI*F*X2)*1E+12
3440 GOTO 3470
3450 G$="[NH]"
3460 X2=X2/(2*PI*F)*1E+09
3470 PRINT"XP=";INT(X2*E)/E;G$;
     ", XS=";INT(X1*E)/E;F$
3480 RETURN
3490 PRINT
3500 PRINT"R ****X*** XS *** R"
3510 PRINT"H     P        S"
3520 PRINT
3530 RETURN
3540 PRINT
3550 PRINT"R **** X0 ****X*** XS *** R"
3560 PRINT"H             P        S"
3570 PRINT
3580 RETURN
3590 PRINT
3600 PRINT"R ****X*** XS *** X *** R"
3610 PRINT"H     P           O     S"
3620 PRINT
3630 RETURN
3635 DATA 1.7E9,45.241,-95.518,50,0
3640 DATA 1E8,2,6,4.13,-13.76
3650 DATA 1E8,1,15,50,0
3660 DATA 1E8,10,0,50,0
3670 DATA 1E8,1,0,50,15
3680 DATA 1E8,20,35,50,0
3690 END
```

This program is available on disk. See page 127 for details.

```
PROGRAM FOR CALCULATION OF LUMPED MATCHING NETWORKS

CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? T

LOW OR HIGH PASS MATCH?[L/H]? L
RH= 4.13 OHMS ,RS= 2 OHMS     ,QUALITY Q= 1.031

LOW-PASS SOLUTION FOR T NETWORK

R **** X0 ****X*** XS *** R
H             P        S

X0= 21.899 [NH], XP= 397.69 [PF], XS= 404.355 [PF]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? Y
SELECT Q>QMIN? 10

LOW-PASS SOLUTION FOR T NETWORK

R **** X0 ****X*** XS *** R
H             P        b

X0= 107.343 [PF], XP= 24.253 [PF],  XS= 22.281 [NH]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? N


CALUCLATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? T

LOW OR HIGH PASS MATCH?[L/H]? H
RH= 4.13 OHMS ,RS= 2 OHMS     ,QUALITY Q= 1.031

HIGH-PASS SOLUTION FOR T NETWORK

R **** X0 ****X*** XS *** R
H             P        S

X0= 21.899 [NH], XP= 6.369 [NH],  XS= 197.365 [PF]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? Y
SELECT Q>QMIN? 10

HIGH-PASS SOLUTION FOR T NETWORK

R **** X0 ****X*** XS *** R
H             P        S

X0= 107.343 [PF], XP+ 18.998 [NH], XS=61.213 [PF]
```

Table 4. T network calculations.

```
CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? L

LOW OR HIGH PASS MATCH?[L/H]? L
RH= 50 OHMS ,RS= 1 OHMS     ,QUALITY Q= 7

LOW-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P        S

XP= 222.817 [PF],  XS= 198.943 [PF]

RH= 1 OHMS ,RS= 50 OHMS     ,QUALITY Q= 1.876

LOW-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P        S

XP= 118.846 [PF], XS= 149.3 [NH]
OK


PROGRAM FOR CALCULATION OF LUMPED MATCHING NETWORKS

CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? L

LOW OR HIGH PASS MATCH?[L/H]?L
RH= 50 OHMS ,RS= 10 OHMS     ,QUALITY Q= 2

LOW-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P        S

XP+ 63.662 [PF],  XS= 31.831 [NH]
```

Table 6. L network calculations.

```
CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? P

LOW OR HIGH PASS MATCH?[L/H]? L
RH= 4.13 OHMS ,RS= 2 OHMS     ,QUALITY Q= 1.224

LOW-PASS SOLUTION FOR P NETWORK

R ****X*** XS *** X *** R
H     P           O     S

X0= 238.78 [PF], XP= 37.739 [NH],  XS= 38.987 [NH]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? Y
SELECT Q>QMIN? 10

LOW-PASS SOLUTION FOR P NETWORK

R ****X*** XS *** X *** R
H     P           O     S

X0= 738.367 [PF], XP= 212.366 [PF],  XS= 12.819 [NH]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? N
OK


CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? P

LOW OR HIGH PASS MATCH?[L/H]? H
RH= 4.13 OHMS ,RS= 2 OHMS     ,QUALITY Q= 1.224

HIGH-PASS SOLUTION FOR P NETWORK

R ****X*** XS *** X *** R
H     P           O     S

X0= 238.78 [PF], XP= 17.457 [NH],  XS= 65.034 [PF]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? Y
SELECT Q>QMIN? 10

HIGH-PASS SOLUTION FOR P NETWORK

R ****X*** XS *** X *** R
H     P           O     S

X0= 738.367 [PF], XP= 5.965 [NH],   XS= 864.342 [PF]
```

Table 5. PI network calculations.

```
CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? P

LOW OR HIGH PASS MATCH?[L/H]? L
*** ONLY L-MATCH AVAILABLE! ***
RH= 50 OHMS ,RS= 1 OHMS     ,QUALITY Q= 7.314

LOW-PASS SOLUTION FOR L NETWORK

R ****X*** XS *** R
H     P        S

XP= 222.36 [PF],  XS= 11.641 [NH]


CALCULATION FOR MATCH AT 100 MHZ
SELECT L, T OR PI NETWORK MATCH [L/T/P]? T

LOW OR HIGH PASS MATCH?[L/H]? L
RH= 50 OHMS ,RS= 1 OHMS     ,QUALITY Q= 7

LOW-PASS SOLUTION FOR T NETWORK

R **** X0 ****X*** XS *** R
H             P        S

X0= 106.103 [PF], XP= 222.817 [PF],  XS= 11.14 [NH]
WANT ANOTHER Q (HIGHER THEN QMIN)?[Y/N]? Y
SELECT Q>QMIN? 10

LOW-PASS SOLUTION FOR T NETWORK

R **** X0 ****X*** XS *** R
H             P        S

X0= 24.299 [PF], XP= 141.664 [PF],  XS= 15.915 [NH]
```

Table 7. L and T network results.

# PLL Implementation

By Andrzej B. Przedpelski
A.R.F. Products, Inc.

*Coming up with a phase-lock loop (PLL) theoretical design is only half of the battle. The desired circuit has to be implemented using available hardware. The choice of the hardware may have a pronounced effect on the overall design.*

In practice, the overall proposed PLL design should be checked against available hardware during its development to ensure that the final configuration is the best compromise of often conflicting interrelated factors.

## Overall Configuration

The overall hardware configuration is quite often dictated not only by performance characteristics, but also by cost, available space, and other factors.

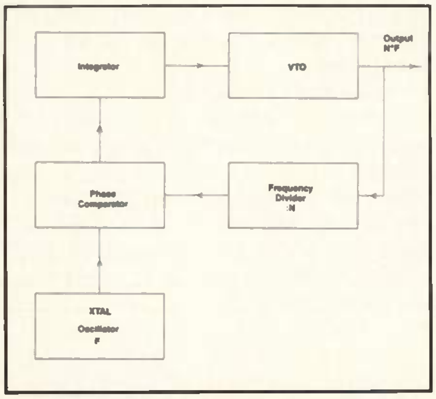


**Figure 1. Discrete PLL frequency multiplier.**

*Discrete IC Design*

When the PLL is used as a frequency multiplier (as when a low frequency crystal oscillator controls a higher, harmonically related output frequency), the best solution may be to use individual ICs for the frequency dividers and the phase detector (Figure 1). Since it is desirable to use the highest possible input frequency to the comparator (to reduce the number and amplitude of undesirable sidebands in the output), the crystal frequency is used as one of the inputs to the phase comparator and the divided output frequency as the other. Thus, this comparison frequency may be higher than the usual limit of frequency synthesizer ICs' built-in phase comparators. Using an external phase comparator, comparison frequencies up to about 1 GHz can be used (Giga-Bit Logic 16G044).

*Direct Digital Synthesizers*

There are cases when the usual PLL approach is not suitable for generation of multiple frequencies. One such case is the series of 20 tones, in the 7 kHz to 70 kHz range, used in flight termination systems. Even frequency spacing is not used, and it would be necessary to use very small steps to obtain the required frequency resolution. This would result not only in about 630 steps (of which only 20 would be used), but the purity of the output signal would be compromised by the many sidebands caused by the very low reference frequency, and the switching speed would be very slow. In this and similar cases, the direct digital synthesizer approach would be more desirable (the Plessey SP2001 IC, now in development, is an example).

*The Frequency Synthesizer IC*

This is the most common approach to multi-frequency (equal tuning step) frequency synthesizers and it will be further discussed. The usual IC used for this purpose has most of the following functions built-in to save space, cost and to provide proper interfaces between all the internal circuits, as shown in Figure 2. These functions include reference oscillator, programmable reference oscillator frequency divider, programmable input frequency divider, phase comparator, programming, and out-of-lock indicator. The use of this very desirable component does, however, put some restrictions on the overall design.

## Comparison Frequency

One of the first important decisions is the choice of the input frequency to the phase comparator. It is important, since it determines the tuning steps and has a pronounced effect on the switching speed, sideband content in the output, and output noise characteristics.

The highest possible comparison frequency should be normally used to permit maximum speed and minimum sidebands and noise. This is usually

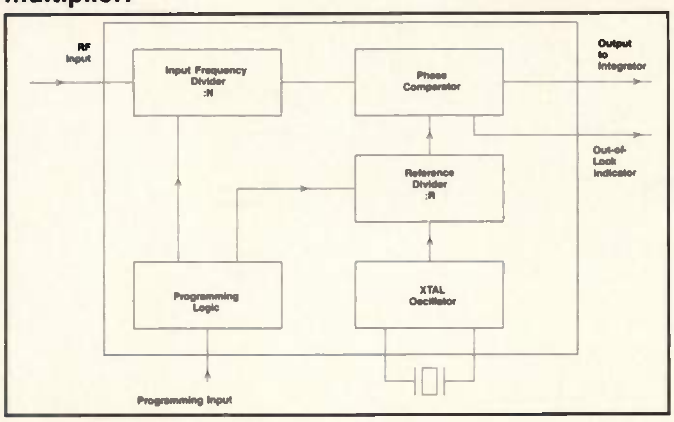


**Figure 2. Typical synthesizer IC.**

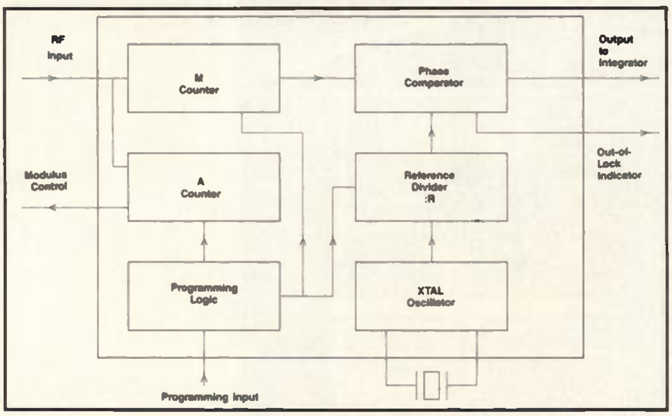


**Figure 3. Typical dual-modulus synthesizer.**

accomplished by making the comparison frequency equal to the tuning step. However, this is not always possible. Two factors may require the use of a submultiple of this frequency:

a. When the internal phase comparator cannot handle the high frequency, it may be necessary to divide both the reference and the input frequency by some factor to bring them within the phase comparator operating range. This approach increases the number of available frequencies and, if this is not desirable, provisions have to be made in the programming to eliminate them. Most of the available frequency synthesizer ICs will usually handle comparator frequencies up to about 200-400 kHz. Since this is usually an unspecified parameter, it is necessary to contact the manufacturer to obtain the approximate figure. There is one IC, now being developed by Plessey Semiconductors, the SP8850, which will reportedly handle comparator frequencies up to about 10 MHz.

b. In very high frequency PLL applications it may not be feasible to obtain the required divide by (P/P+1) prescalers. It may be necessary to use a fixed frequency divider in the input frequency or a divide by (P/P+2) prescaler. A divide by (P/P+2) prescaler is in reality a fixed divide-by-2 and a (P/P+2)/2 dual-modulus divider. In these cases it is necessary also to divide the reference frequency by the same fixed divider ratio. The frequency step then becomes the comparator frequency multiplied by the fixed divider ratio. In general, this is an undesirable solution, since it decreases the comparison frequency (for a given tuning step) with the associated shortcomings.

### Comparator Type

A very important factor in selecting a suitable IC is the phase comparator type. There are basically two distinctly different types available:

*a. Digital*

This is the most common type and is suitable for most applications. Its main advantages are the very large acquisition range (it will lock anywhere within the VTO tuning range) and the acquisition speed. No special circuits to facilitate acquisition are required. Its main disadvantages are both associated with output signal noise. Its gain is very low. Thus, the recovered signal to control the VTO is low and requires a considerable amount of amplification. The S/N ratio of this control signal is, therefore, limited, resulting in more noise in the output. Also, there is an inherent "dead zone" in the middle of its transfer characteristic, resulting in some jitter of the output signal. There are two main varieties available: the three state single output, which can be directly applied to the integrator, and the two individual outputs, which have to be differentially combined.

*b. Analog sample/hold*

For more critical applications, the analog sample/hold type of phase comparator is recommended. It is less common and typical examples are: the Signetics HEF4750, the Motorola MC145159-1 and the Plessey NJ8820. The sample/hold comparator has three

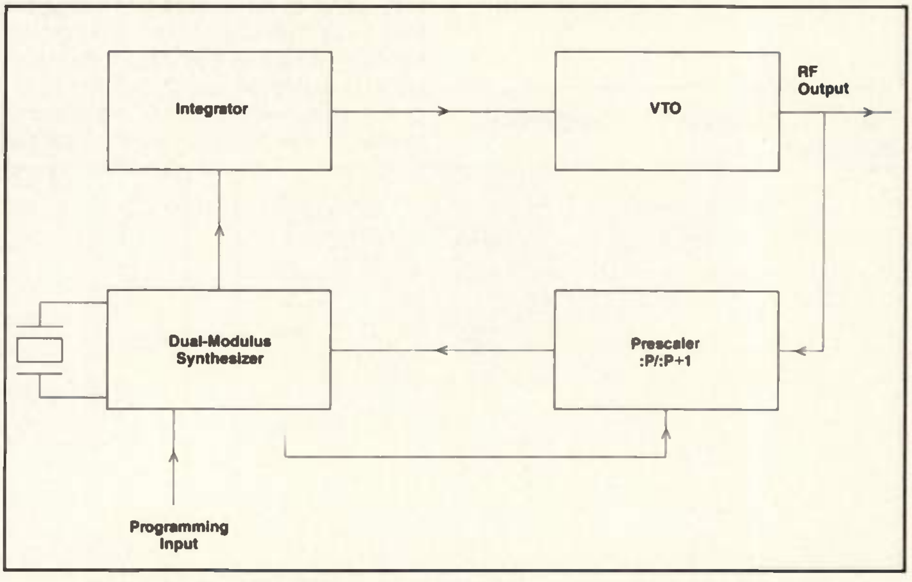


**Figure 4. Dual-modulus circuit.**

main advantages in reducing spurious outputs and noise:

— It has high gain (usually adjustable) and thus requires less amplification of the correction signal.

— The reference frequency feed-through is reduced resulting in lower output spurious sidebands.

— It does not have the "dead zone" and therefore does not produce jitter at the operating point.

It does have one major problem: its frequency acquisition characteristics are inferior. Therefore, all of the above referenced ICs employ a digital phase comparator in addition to the sample/hold. The outputs of both are summed (usually in the integrator). The digital comparator is used in the initial acquisition period. When the signal is within the sample/hold "window", the sample/hold comparator takes over. This is all accomplished automatically within the IC. The digital comparator usually goes to a high impedance state when disabled.

## Synthesizer ICs

There are four major classes of frequency synthesizer ICs:

*a. Standard low frequency*

These are single IC synthesizers usable up to about 10-15 MHz, such as the Motorola MC145145-1. To use it at a higher input frequency it is necessary to provide a fixed divider(s) to divide the frequency to the 10-15 MHz range. This automatically divides the comparison frequency by the same ratio with the disadvantages discussed previously. However, this arrangement is often used in TV sets and AM/FM/SW radios.

*b. Dual modulus*

This is a more versatile IC (Figure 3). Two internal dividers are used (:M and :A), one to control the dual modulus prescaler, as in the Motorola MC145146-1 and Plessey NJ8821. Control logic is provided internally to control these counters and also the internal reference divider, :R. A typical dual-modulus synthesizer circuit is shown in Figure 4. The total division (N) from the output frequency to the comparator frequency is:

$$N = (M)(P) + A$$

While this seems to provide a wide choice of configurations, two main restrictions apply:

$-N \geqslant A$

$-N \geqslant P^2 - P$

To program the synthesizer:

$-$ M = integer (N/P)

$-$ A = N - N(P)

The first step is to select a suitable prescaler. The value of maximum allowed P is approximately:

$$P_{max} = \frac{1 + SQR\,(1 + 4M)}{2}$$

*c. Decade synthesizer*

A typical decade synthesizer circuit is shown in Figure 5. An example of a suitable IC for this application is the Signetics HEF4750/4751. Each 10/11 prescaler is controlled by the synthesizer IC and adds another decade of control.

*d. Special synthesizer*

Several special purpose synthesizer ICs are available. Two typical examples

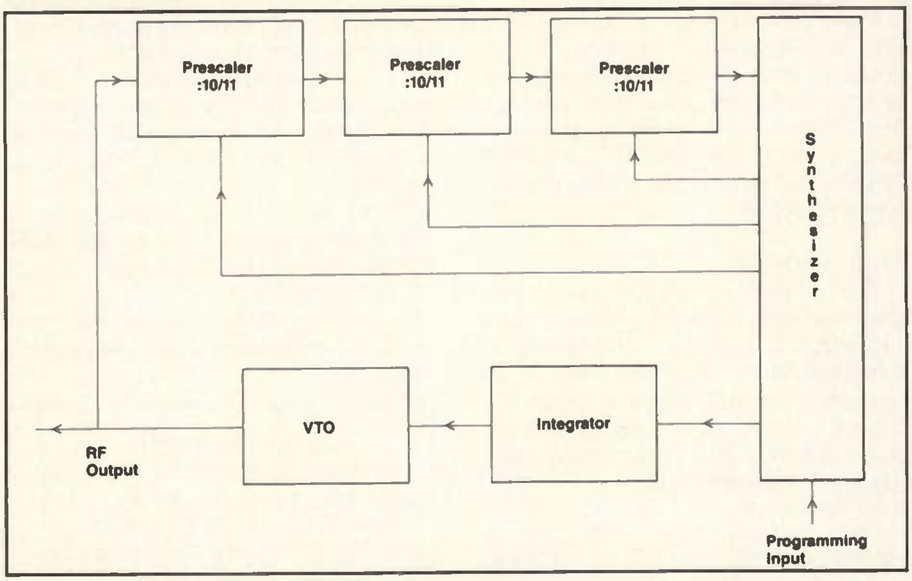


**Figure 5. Decade prescaler circuit.**

are the Motorola MC145151-1 and the Signetics HEF4751.

The Motorola device has a built-in transmit/receive offset, controlled by a single input line. Unfortunately the offset is fixed at 856, but is suitable for most standard applications.

The Signetics IC is considerably more complex. It has two programming inputs, A and B. The final division is a function of A-B. Thus, fixed and variable frequency offsets can be provided. Its companion, the HEF4750, has provisions for phase-modulating the signal and has adjustable phase comparator gain.

## Prescalers

There is a large selection of prescalers available. They are available in silicon and GaAs, and practically any logic family. Since prescalers are usually at the higher frequencies, most use ECL logic, but may have CMOS or other outputs to make them interface with other circuits.

*a. Fixed*

There are many fixed dividers available. They are used mainly in single frequency synthesizers and very high programmable synthesizers when suitable dual modulus devices are not available. Phase noise is not usually specified. The saturated logic devices usually have lower phase noise. An exception is the Anadigics ADV3040 GaAs divide-by-4 prescaler which has a specified phase noise.

Fixed GaAs dividers operating at frequencies as high as 10 GHz are available from NEC.

*b. Dual modulus*

Many dual modulus prescalers, suitable for programmable synthesizers, are available. Probably the largest selection of high frequency dividers is available from Plessey. While the dividers are ECL, some have CMOS compatible outputs to match the more popular synthesizer ICs.

## Program Input

The synthesizer IC has to be programmed to provide the desired output frequency. Usually, it is necessary to select the reference division ratio (which remains constant) and the input frequency division (which varies with selected channel). Two programming methods are usually available:

*a. Parallel input*

This is static programming and is usually the simplest. The individual control lines are biased low or high to provide the required divisions. However, it does have a disadvantage — it requires a large number of pins to provide access to all the control functions. A typical IC of this type, the Motorola MC14152-1, is a wide 28-pin DIP.

*b. Serial input*

This is a more common arrangement, since the number of IC pins can be then considerably reduced for comparable types. The programming can be truly serial (data is clocked-in one bit at a time using a single input line) as in the Motorola MC145157-1, or it can be entered by digit "words" (4 lines at a time), the digit location selected by "data select" input lines as in the Plessey NJ8820, Signetics HEF4751 and Motorola MC145145-1. These types are sometimes optimized for a PROM interface (Plessey NJ8820) or microprocessor control (Plessey NJ8821).

## References


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3. "Single Chip Synthesizer for Radio Tuning," *Electronic Components & Applications*, Philips, vol. 4, no. 3, May 1982.
4. T.G. Giles, "Designer's Guide to Frequency Synthesizers Using the HEF4750 & HEF4751 LOCMOS LSI Circuits," Mullard Limited.
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**About the Author**

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# Modeling PLL Tracking of Noisy Signals

*By G. Stephen Hatcher*
*Summit Engineering*

*In the design of receiver phase-lock loops (PLLs), the analysis of the loop performance in noise is generally limited to minimization of PLL component noise effects through analysis of the open and closed loop responses to the loop noise processes. However, transient conditions such as fading, the performance of the phase-lock loop during signal acquisition, or variable dynamics tracking during low signal-to-noise (SNR) ratios are best assessed through simulation. This article describes a simple noise model and PLL simulation which can be readily altered to model other receiver synchronization processes such as Costas loops, squaring loops and delay lock loops.*

There are two options for time domain analysis of the response of a PLL in the acquisition and tracking of noisy signals. The first requires the derivation of nonlinear stochastic differential equations which the author has found so complex that interpretation of the results is itself error-prone. The other option is a time domain simulation of the PLL process operating on a noisy signal which can yield excellent accuracy while being relatively simple. Modeling also has the advantage of allowing the designer to quickly assess "what if" scenarios. These scenarios include varying the incoming SNR and signal dynamics of doppler and doppler rate, or they can be in implementation, such as changing the phase detector or loop filter type.

Performing this modeling on a computer involves several basic steps. First, the signal plus noise must be described. The signal usually contains doppler and doppler rate terms with possibly some modulation process. The noise portion can be phase noise, amplitude noise or a combination of these. In this simulation a gaussian noise model is used. Second, the loop filtering process must be converted to a difference equation representation in order to accurately represent the filter in a discrete time simulation. This is usually a straightforward calculation using the bilinear transform. Finally, the overall simulation is created by filling in the remaining easily modeled PLL processes.

## Modeling the Signal

The modeling of the signal plus noise is separable. Each portion can be modeled and the results added together. The signal is usually sinusoidal and described as:

$$S(t) = A \sin [2\pi F(t)t + P(t)]$$

where A is the signal amplitude in volts, t is time in seconds F(t) is the frequency term which may vary with time, and P(t) is the phase term which may vary with time.

The simulation time is replaced by discrete steps. To minimize quantization errors due to time steps which are too large, use approximately 100 steps per period of the highest signal frequency expected in the model. Good results can be obtained with as few as ten steps per period but generally the simulation executes quickly and the extra steps are not prohibitive.

F(t) is the frequency term consisting of an initial offset (doppler term), a rate of change (doppler rate), and higher order terms as prescribed by the design problem. These terms are related to the line-of-sight velocity and acceleration between the transmitter and receiver by the formulas:

Doppler frequency Hz = (Vel)(Fc)/C

Doppler rate Hz/sec = (Acc)(Fc)/C

where, Vel = velocity (m/sec), Acc = acceleration (m/sec$^2$), Fc = carrier frequency (Hz), and C = speed of light (m/sec).

P(t) is the phase term which consists of an initial offset which may be random or dictated in the simulation plus a phase modulation term if required.

## Adding Noise to the Signal

In this simulation, it is assumed that the signal is corrupted by white gaussian noise. Though receiver noise processes may not be exactly gaussian in nature, systems analysis performed using a gaussian model is usually representative of the system response to the band limited noise typically found in receivers (1). If noise biases are anticipated in the noise process, this is included by offsetting the modeled gaussian noise variables. In this model the user need only specify the SNR and the RMS signal voltage out of the loop phase detector. The simulation computes the random number variables of correct RMS amplitude and distribution to accurately model the noise process.

In this simulation, the noise model recomputes the noise variable if an amplitude greater than three times the RMS value of the noise is generated. This is due to limitations in the noise

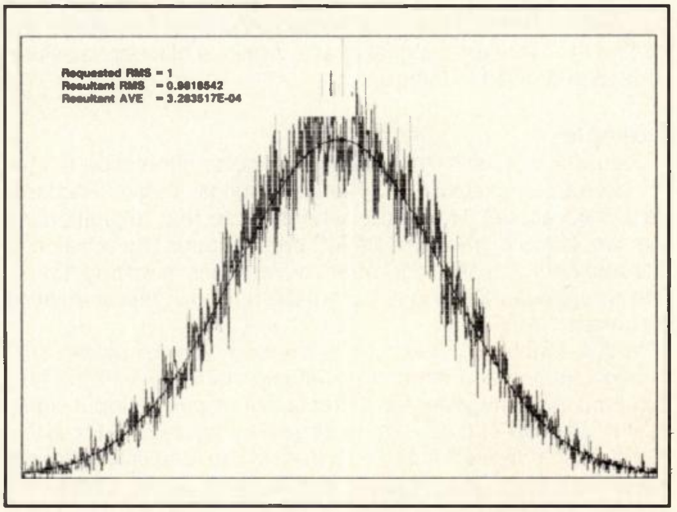


**Figure 1. Frequency distribution of simulation noise model.**

model which will degrade the overall PLL modeling accuracy if allowed to go unchecked. However, 99.8 percent of white noise has an amplitude less than 3 RMS. Hence, this limitation has virtually no effect on the model accuracy. Below 3 RMS this model produces excellent gaussian variables. Figure 1 is a plot of the frequency distribution of this model overlaid by the ideal gaussian bell curve of the same RMS and amplitude value. In this case 20,000 random numbers were generated and the resultant statistics computed to compare against the required statistics. The model produced an error of −0.16 dB and a bias of 0.33 mV on a 1 volt RMS signal. This is sufficiently accurate for nearly all design analysis.

The signal plus noise out of the phase detector is derived differently for analog and digital phase detectors. In the case of the analog phase detector it is usually simplest to compute the detector output SNR and use this in the simulation. In this way the mixer-lowpass filter model is eliminated with little loss in accuracy while expediting the simulation run time.

The expression for the SNR out of an analog phase detector is:

$$SNR_o = G\,[SNR_i + 10 \log (B_i/B_o)]$$

where, $B_i$ and $B_o$ are the input and output noise bandwidths respectively, $SNR_o$ is the detector output SNR, $SNR_i$ is the detector input SNR (use the LO SNR from the VCO if it is less than the RF input SNR), and G is the detector gain comprised of the mixer conversion loss and the lowpass filter insertion loss.

In the case of a digital phase detector, the output noise process is derived by comparing the input signal plus noise to the VCO signal in an exclusive-OR operation. Thus, the input RF SNR is used and the exclusive-OR operation performs the amplitude noise modulation to phase noise modulation conversion characteristic of a digital phase detector.

## PLL Loop Filter Modeling

The discrete time step simulation requires that the integration processes be represented by a difference equation. This is easily derived using "cookbook" techniques such as the bilinear transform. A brief tutorial on this transformation as applied to PLL filter modeling is provided in the sidebar.

The simulation also requires that the error signal phase be calculated. This is done by integrating the error frequency and adding the initial error signal phase. Again, a bilinear transform is used to derive the integral.

## Example

Figure 2 is a listing of a basic PLL noise simulation written in MicroSoft QuickBASIC, and contains those elements described above. The reader should note that implementing the simulation in standard BASIC only requires the addition of line numbers. QuickBASIC offers advantages in editing, speed and accuracy but its use is not a prerequisite in this simulation technique.

In this simulation, a second order loop with bandwidth 70.7 radians/second and damping coefficient of 0.707 using a lead compensated integrator loop filter is tested with an input signal having an SNR of 0 dB, a step frequency error of 50 Hz with a 50 Hz doppler rate. Figure 3 is a print of the results which were screen plotted. The results in this case show about 400 milliseconds were required for phase lock to be re-established. In contrast, Figure 4 shows the same test case except that the SNR is 99 dB (effectively noiseless) where phase lock was re-established in about 300 milliseconds. Keep in mind, however, that the signal acquisition time of phase-lock loops is very dependent on the loop initial conditions of signal amplitude and phase, such that noise can in some cases accelerate acquisition.

The requisite code for this screen plot is not shown as it is irrelevant to the modeling process. The user may follow the flow presented and implement the simulation in many computer languages and display the results in tabular or graphic form.

When the model is executed several times the results will vary due to the noise effects. It is often useful to nest the simulation inside a loop which repeatedly executes the case and computes the loop performance statistics for 100 cases, for example. These statistics include RMS tracking error, signal acquisition time, and probability of the loop-breaking lock.

```
'* ------------------------------------------------------------------- *
'*         MODELING PHASE-LOCKED LOOP TRACKING OF NOISY SIGNALS         *
'* ------------------------------------------------------------------- *
'*   This program performs a time domain simulation of the operational  *
'*  capabilities of a Phase Locked Loop operating in a noisy environment. *
'*                                                                      *
'*       Loop design and simulation prepared by G. Stephen Hatcher      *
'* ------------------------------------------------------------------- *

     DIM ErrorFreq(500), ErrorPhase(500), InputFreq(500)
     DIM VCOfreq(500), SimTime(500)     '** Note: arrays & # of steps can vary

     RANDOMIZE TIMER                    '** Initialize random number generator
     pi = 3.1415926#

     '** define signal conditions:
     Amp = 2.5                          '** peak voltage out of phase detector
     SNR = 0                            '** phase detector output SNR in dB
     Sigma = (Amp / SQR(2)) / 10 ^ (SNR / 20)       '** rms noise voltage
     Fdop = 50                          '** doppler frequency error in hertz
     Frate = 50                         '** doppler frequency rate in Hz/sec.
     Frate2 = 0                         '** 2nd order freq. rate in Hz/sec*sec

     '** define loop parameters:
     Gain = 64                          '** loop gain in absolute units
     Dt = 1 / 1000                      '** time step size in seconds
     Wn = 70.7                          '** loop bandwidth in radians/second
     Damp = .707                        '** loop damping coefficient
     Kphase = 1                         '** phase detector gain
     Kvco = 100                         '** VCO constant in radians/volt-sec.

     '** compute filter coefficients for difference equation representation
     '** of a lead compensated integrator H(s)=[s*Tau2+1]/[s*Tau1] where s=jw
     Tau1 = Gain * Amp * Kphase * Kvco / Wn ^ 2
     Tau2 = 2 * Damp / Wn
     A0 = Dt / (2 * Tau1) + Tau2 / Tau1  '** difference equation coefficient
     A1 = Dt / (2 * Tau1) - Tau2 / Tau1  '** difference equation coefficient

     Nsteps = 500                       '** iterations on time step simulation
     FOR N = 1 TO Nsteps
          Time = Time + Dt              '** get time & input freq for the step
          FreqIn = Fdop + Frate * Time + .5 * Frate2 * Time ^ 2
          IF N < Nsteps / 10 THEN
             FreqIn = 0                 '** input delayed step+ramp error freq
          END IF
          '** get carrier phase for this step by integrating input frequency
          Win = 2 * pi * FreqIn
          PhaseIn = .5 * Dt * Win + .5 * Dt * WinLast + PhaseInLast
          WinLast = Win
          PhaseInLast = PhaseIn
          '** get VCO output frequency and integrate to get VCO output phase
          West = Kvco * Vout
          PhaseEst = .5 * Dt * West + .5 * Dt * WestLast + PhaseEstLast
          WestLast = West
          PhaseEstLast = PhaseEst
10        RndX = RND         '** derive noise variable to add to error signal
          IF RndX < .011109 THEN
             GOTO 10
          END IF
          R = SQR(-2 * LOG(RndX))
          Noise = Sigma * R * COS(2 * pi * RND)
          PhaseErr = PhaseIn - PhaseEst       '** determine loop error signal
          FreqErr = FreqIn - West / (2 * pi)
          ErrSignal = Gain * (Amp * SIN(PhaseErr) + Noise)
          '** process error signal through loop filter difference equation
          Vout = A0 * ErrSignal + A1 * ErrSignalLast + VoutLast
          ErrSignalLast = ErrSignal
          VoutLast = Vout
          '** save desired results for output tabulation or plotting
          ErrorFreq(N) = FreqErr
          ErrorPhase(N) = PhaseErr - 2 * pi * FIX(PhaseErr / (2 * pi))
          IF ErrorPhase(N) > pi THEN
             ErrorPhase(N) = ErrorPhase(N) - 2 * pi
          END IF
          InputFreq(N) = FreqIn
          VCOfreq(N) = West / (2 * pi)
          SimTime(N) = Time
     NEXT N
```

**Figure 2. Simulation listing. (This program is available on disk — see page 127.)**

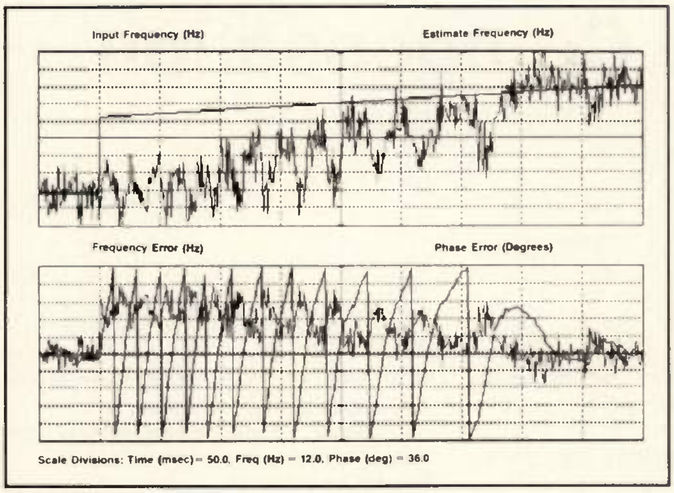


**Figure 3. Simulation with SNR = 0 dB.**

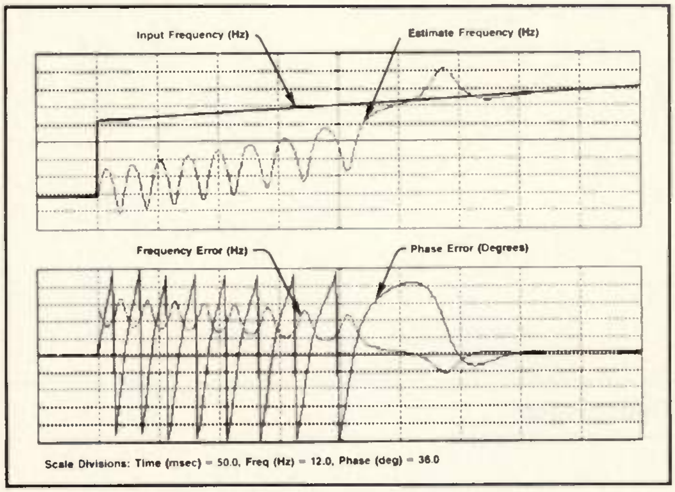


**Figure 4. Simulation with SNR = +99 dB.**

## Conclusion

The listing of Figure 2 is rudimentary. Significant details can be added by modeling loop components in greater detail. If required, the phase detector lowpass filter transfer function (amplitude and phase effects on the error signal), the VCO nonlinearity, plus adaptive bandwidth algorithms and analog-to-digital conversions, can be included. Also, this simulation is directly applicable to modeling digital signal processing when included in the loop, due to the discrete time modeling. It is very often the case that receiver synchronization algorithms are implemented in digital signal processors which must operate on noisy signals.

*The program described in this article is available on disk.* *See page 127 for details.*

## References

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### About the Author

G. Stephen Hatcher is owner of Summit Engineering and can be reached at Summit Engineering, 21827 124th Ave. SE, Kent, WA 98031; Tel: (206) 854-7027.

## The Bilinear Transform

The transformation of an analog loop filter expression to a difference equation for the purpose of discrete time simulation requires a straightforward application of the bilinear transform. The transform involves replacing each $s = j\omega$ in a Laplace transform expression of the loop filter with a mapping expression which converts the loop filter to a Z-transform for discrete time representation. This is done using the substitution:

$$\frac{2(1/Dt)(1 - z^{-1})}{1 + z^{-1}} \quad \text{into } s = j\omega$$

where Dt = the simulation time step duration in seconds.

The simplest example is the transformation of an integrator from the Laplace expression to a difference equation. Making the above substitution into $H(s) = 1/s$ results in:

$$H(z) = \frac{Y(n)}{X(n)} = \left(\frac{Dt}{2}\right)\left(\frac{1 + z^{-1}}{1 - z^{-1}}\right)$$

noting that $z^{-1}$ is a time delay of one step. The final expression is derived in terms of discrete steps (n):

$$2Y(n) - (2z^{-1})\ Y(n) = (Dt)\ X(n) + (Dt)\ X(n)\ (z^{-1})$$

$$Y(n) = (Dt/2)\ X(n) + (Dt/2)X(n - 1) + Y(n - 1)$$

This is the trapezoidal integration function where, X(n) is the present input (from the phase detector), X(n-1) is the last input value, Y(n) is the present loop filter output, and Y(n-1) is the last output value.

For a second example use the loop filter of the program listing:

$$H(s) = (sT2 + 1)/sT1$$

Replace s with $2(1/Dt)(1-z^{-1})/(1+z^{-1})$ to find H(z). After some algebraic manipulations H(z) takes the form:

$$H(z) = \frac{Y(n)}{X(n)} = \frac{A0 + A1(z^{-1})}{1 - z^{-1}}$$

$$\text{where } AO = \frac{Dt + 2(T2)}{2(T1)}$$

$$\text{and } A1 = \frac{Dt - 2(T2)}{2(T1)}$$

Noting again that $z^{-1}$ is a time delay of one step, the final expression is derived in terms of discrete steps (n):

$$Y(n) - z^{-1}Y(n) = (AO)X(n) + (A1)X(n)z^{-1}$$

$$Y(n) = (A0)X(n) + (A1)X(n - 1) + Y(n - 1)$$

More complex filters are transformed in the same manner. If the analog filters can be cascaded then the different equations can also. Using this linearity property makes the derivation of complex difference equations less tedious.

# Digital Temperature Compensation for Oscillators

*By Steven Fry*
*Murata Erie North America*

*There's no such thing as a stable crystal since the resonant frequency of all crystals varies with temperature. Although specially cut crystals can have low variation over a limited temperature range, it isn't possible to make a crystal with good stability over the temperature ranges that military and commercial equipment is subjected to. The laws of physics don't allow it.*

Physics can be controlled, however, with temperature-compensated crystal oscillators (TCXOs). There are two common forms of compensation. The first (and oldest) stabilizes the oscillator's temperature by enclosing it in a proportionally controlled oven. Such oscillators tend to be bulky, expensive and power-hungry, but have extraordinarily good stability.

A less expensive alternative uses a varactor diode to vary the crystal's frequency. A thermistor-resistor network generates a temperature-varying voltage that alters the varactor's capacitance to (almost) exactly cancel out the crystal's thermal drift (Figure 1). Almost all TCXOs use AT-cut crystals, due to their superior thermal stability over a wide temperature range. These TCXOs generally cost less than ovenized units, but are hard to manufacture consistently with a stability better than 1 ppm.

The only practical way to produce an oscillator with the stability of ovenized units and the low current drain of thermistor-compensated systems is to adopt digital compensation. Digital compensation has been known for some time but, for most users, the advantages of digitally compensated crystal oscillators (DCXOs) have not been substantial enough to outweigh the disadvantages of higher cost and larger size.

Fortunately, the falling price of LSI chips and the development of high-density assembly techniques have removed these barriers. The design described in this article is cost-competitive with some ovenized designs of comparable size and frequency stability, while offering some critically important advantages:

1. No warmup is required. The output is within 0.1 ppm of nominal frequency as soon as power is applied.
2. Power consumption is low and does not change with temperature.

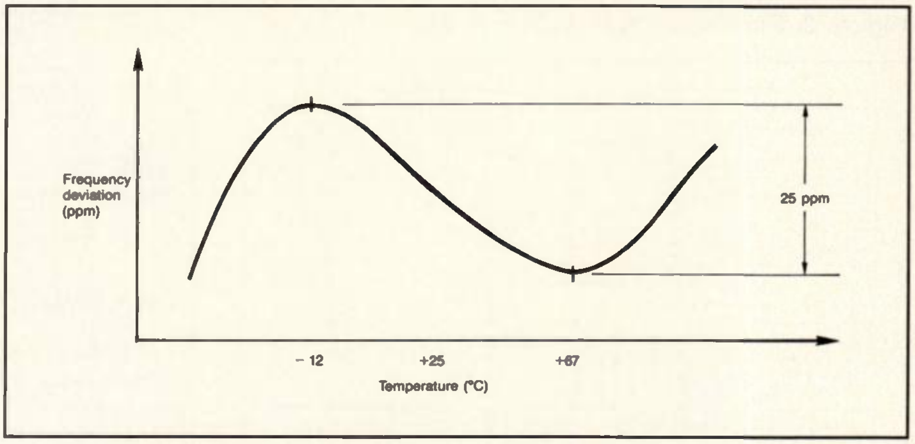



**Figure 1. Typical AT-cut crystal.**

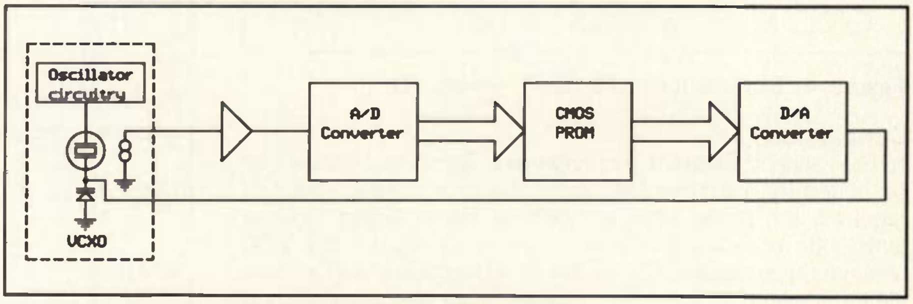



**Figure 2. DCXO block diagram (direct method).**

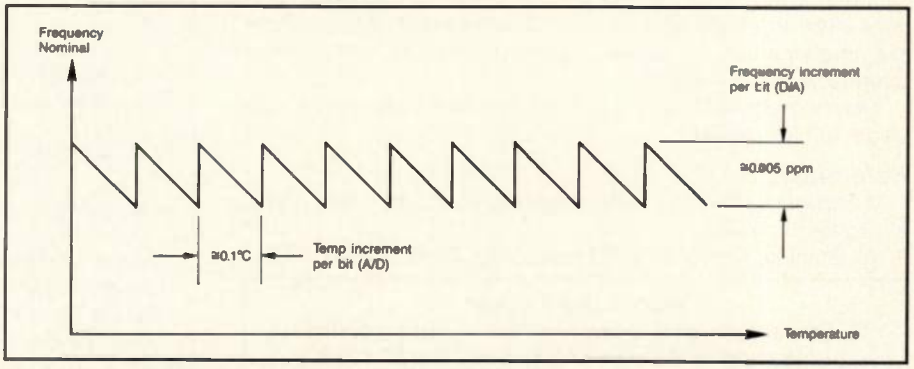



**Figure 3. Ideally compensated DCXO.**

Though these DCXOs are somewhat more expensive and larger than a conventional TCXO, their frequency-versus-temperature stability is about ten times better, over a wide temperature range.

The heart of the DCXO (and the key to its success) is a high-stability crystal oscillator. Unless the oscillator's frequency is repeatable through constant temperature cycling and very low aging, any attempt at high-tolerance compensation is futile. Hence, the designer cannot concentrate on the digital subsystem while ignoring the oscillator circuitry. In most cases, it's the crystal that limits the attainable stability and repeatability. Crystal aging and thermal hysteresis can easily be of higher magnitude than the resolution of the digital compensator.

There are a variety of digital compensation techniques, but they all fall into one of two categories. The first is direct compensation. The oscillator's frequency is electronically tunable via a varactor diode inserted in the feedback network. A compensating voltage is generated which tracks the characteristic frequency-versus-temperature drift of the crystal, pulling the oscillator back to nominal frequency over the design temperature range.

The second is indirect compensation. The oscillator is free to run at its "natural" frequency, regardless of temperature. The compensated output is derived by subtracting as many oscillator pulses as necessary to maintain a constant output frequency as the temperature changes. It is also possible to lock a phase-locked loop (PLL) to the crystal frequency and digitally vary the division rate in the PLL feedback loop.

The indirect approach yields the best medium- to long-term stability, as it permits ultra-high Q overtone crystals to be used. These crystals are too stiff to be pulled far enough to compensate the frequency shift directly. They include the highly repeatable stress-compensated SC cut. The indirect techniques are only useful, however, as time bases, or in applications where the severe phase perturbations caused by pulse swallowing or changes in the PLL division rate can be tolerated. Although the direct approaches may not perform as well over the long term, their signal is spectrally purer, and the small quantized steps in the compensating voltage may be smoothed out by filtering to maintain phase coherence.

Despite the disadvantages of direct compensation, it has been the first choice of those firms that have produced commercial DCXOs. Its most attractive point is the simplicity of the digital circuitry; indirect compensation is much more complex. Direct digital compensation is, therefore, inherently more reliable, less expensive to develop, and permits low-volume production, since "off-the-shelf" components are used for all functional blocks. It also permits using essentially the same oscillator circuits that have been optimized over years of TCXO production.

The remainder of this article describes a DCXO design currently produced at Murata Erie, State College Division, and the performance achieved by this oscillator.

As can be seen in Figure 2, this oscillator uses direct digital compensation. The crystal's temperature is detected by a sensor that is thermally coupled as tightly as possible to the crystal resonator. It is critical to minimize thermal gradients between the temperature sensor and the crystal for best thermal tracking of the compensation.

The sensor's output is conditioned and scaled to match the input range of the analog to digital (A/D) converter. The A/D effectively normalizes the unit's temperature range on a scale of "0" (minimum operating temperature) to full scale (maximum operating temperature) once it has been calibrated. The A/D's binary output then addresses the read-only memory (ROM), which contains the compensation data. (The ROM can be either "UV" EPROM or EEPROM fabricated in CMOS.)

A compensation run is performed to determine the data needed for each memory location to pull the oscillator back to nominal frequency. The unit is run slowly through the operating temperature range, while a computer interfaced with the digital circuits reads the A/D output and exercises the D/A (digital to analog) to keep the oscillator on frequency. This generates a set of data points which is smoothed and interpolated by a curve-fit algorithm to fill in the gaps between the measured temperatures. The oscillator is operational and "right-on" frequency as soon as the PROM is programmed with the data.

The resolution of the digital system is determined by several factors:

1. The number of bits of the A/D and D/A converters and the corresponding memory size.

2. The operating temperature range to be covered.

3. The amount of frequency error-over-temperature which must be corrected and the maximum slope of this error curve.

The temperature increment per bit is equal to:

$(T\ max - T\ min)/2^N$, where N equals the number of A/D bits.

The frequency increment per bit is given by:

$(F\ max - F\ min)/2^M$, where M is the number of D/A bits.

It is possible, with standard commercially available converters, to reduce these incremental errors to values better than the thermal repeatability of the crystal and oscillator circuitry.

## A Comparison

The D01775 is one of several DCXOs currently in production at Murata Erie. This unit is at 10 MHz; however, these digital compensation techniques are being applied to oscillators of many frequencies and output configurations. Figure 4 shows the results of a typical temperature run of a D01775, plotted along with a typical run of a high-quality "state-of-the-art" TCXO, for comparison.

Even with the best components, analytical testing and compensation procedures, it may still require as many as six or seven temperature runs and resistor network changes to compensate a TCXO to this degree. The DCXO, on the other hand, is usually well within its design specification on the first run after the digital compensation is applied, provided the circuitry was designed and is working properly. First verification run yields of 85 to 90 percent are common.

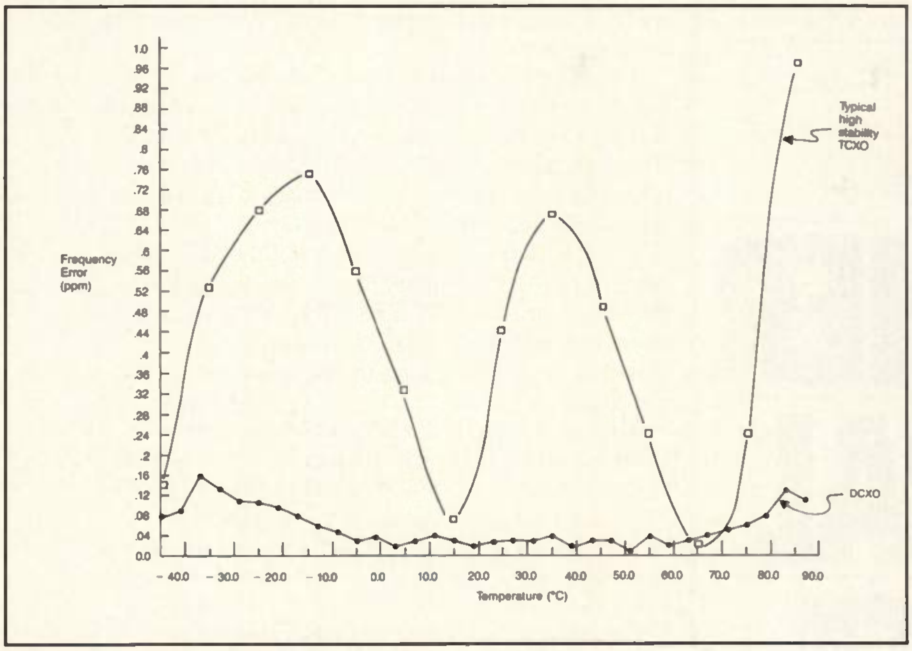



**Figure 4. 10 MHz DCXO vs. TCXO.**

These DCXOs are becoming popular as frequency references in Global Positioning System equipment, portable instruments, airborne and shipboard navigation systems, high-performance communications transceivers (including frequency-hopping types), and other telecommunications and switching equipment. As the cost of these oscillators continues to decline, the applications will continue to grow.

Research and development is ongoing to improve the electrical performance and further reduce the size and cost of these high-stability oscillators. As new semiconductor devices and other components become available, the possibilities for improvement expand. With the help of on-board processors, it will be possible to implement advanced functions, such as thermal hysteresis and thermal transient compensation, correction for aging, and self-calibration. The quest for the ideal oscillator continues.

### About the Author


Steven Fry is chief oscillator engineer at the RF and microwave division of Murata Erie North America, Inc., 1900 West College Ave., State College, PA 16801. He has been involved in the design and temperature compensation of crystal oscillators for more than 12 years. His phone number is (814) 237-1431.





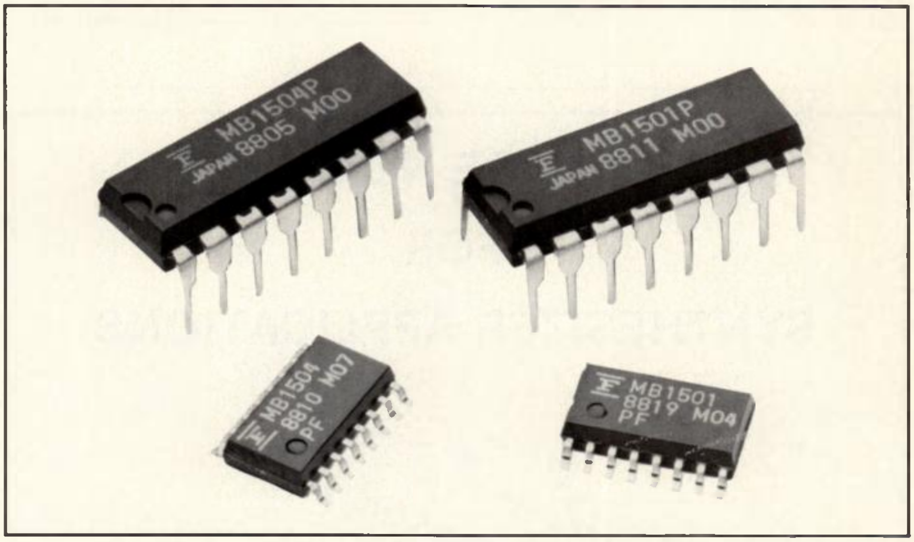




## BiCMOS Phase Lock Loop Synthesizer From Fujitsu

The Model MB1501 and 1504 phase lock loops offer combined prescalar/PLL chips fabricated in BiCMOS technology. The MB1501 operates at a maximum frequency of 1.1 GHz and draws 15 mA of current while the MB1504 operates up to 520 MHz at 10 mA. Both devices require a single power supply between 2.7 and 5.5 V and are offered in single and dual modulus with a divide ratio of 64/65 and 32/33 respectively.

Applications for the PLLs include cellular phones, radio transceivers, cordless telephones, test equipment and satellite systems. Packaging options include standard DIP and surface-mount flat packs. Price ranges from $7 to $8 when purchased in quantities of 1000. **Fujitsu Microelectronics, Inc., San Jose, CA. INFO/CARD #230.**

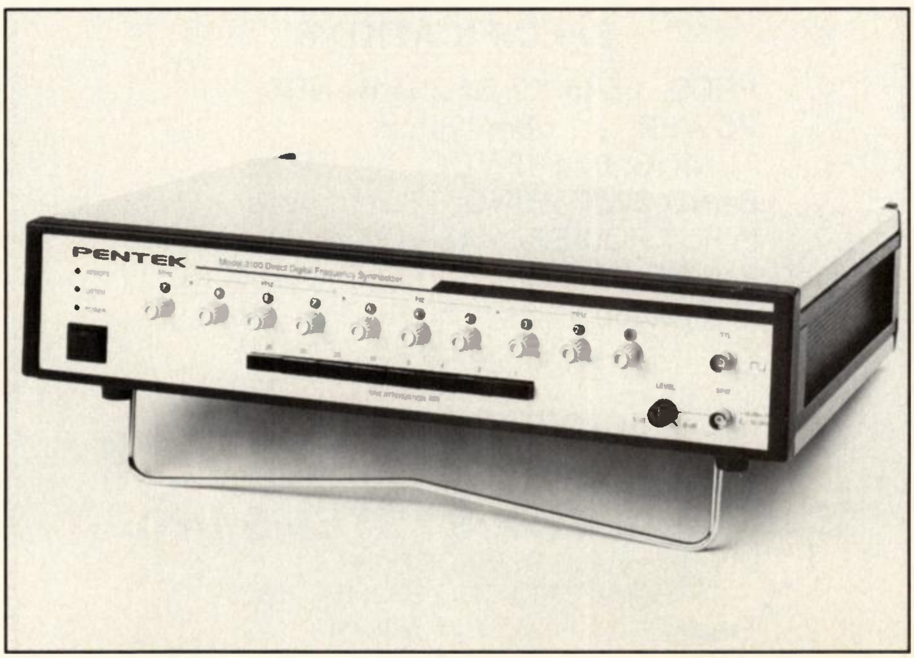



## Pentek Introduces a Frequency/Phase Synthesizer

This direct digital synthesizer features 0.001 Hz to 7.999 999 999 MHz coverage with 0.001 Hz resolution. Phase shifts throughout the 360 degree range are programmable with 0.36 degree resolution. In addition to the sine output signal with its 85 dB step attenuator, the Model 3100 features a TTL output.

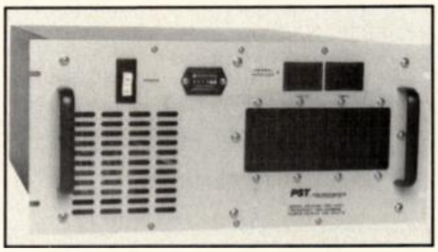
Both GPIB and BCD parallel remote interfaces are available which provide control of frequency, phase and attenuation. Submicrosecond, phase-continuous frequency switching makes the device ideal for frequency agile applications such as sweep, FSK, PSK, burst and hop patterns.

The internal 10 MHz ± 1 ppm standard frequency reference can be upgraded to an optional ovenized reference providing ± 0.01 ppm accuracy. Model 3100 also accepts a 5 MHz or 10 MHz external frequency reference. The device is priced at $2195. **Pentek, Inc., Rockleigh, NJ. INFO/CARD #229.**



### Class A Linear Power Amplifier

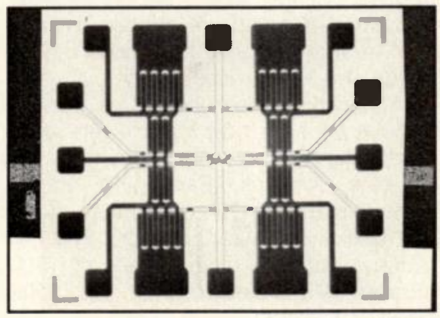
Model AR 1858-100 provides 100 watts of RF power at 1 dB compression. It operates between 100 and 500 MHz while gain is 60 dB min, gain flatness is ± 1.5 dB and harmonics are -30 dBc min. Noise figure is 11 dB max and spurious signals are measured at -60 dBc.



Model BHE 5819-100 is a 100 watt Class AB unit with a frequency range of 500 to 1000 MHz. Dynamic range is greater than 40 dB and minimum RF gain is 50 dB. Harmonics and spurious signals are -20 dBc and -60 dBc, respectively. **Power Systems Technology, Inc., Hauppauge, NY. Please circle INFO/CARD #228.**

### GaAs MMIC Transfer Switches

Tachonics introduces the TCSW-1100 GaAs MMIC DC to 6 GHz transfer switch. It features 1.7 dB insertion loss at 6 GHz, VSWR of 1.5:1 and isolation

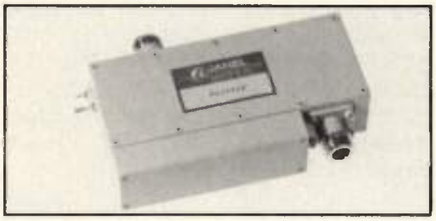


of 43 dB at 6 GHz and 60 dB at 50 MHz. Applications for the devices include digital attenuators and digital phase shifters where selection of either one of two separate externally connected RF circuits is required.

Also being introduced is the TCWP-0400 medium power amplifier chip capable of delivering 0.5 W of output (at 1 dB compression) into a 50 ohm load over the 2.2 to 6.2 GHz range. Typical small signal gain is 11 ± 1 dB from 2 to 6 GHz and input VSWR is typically 2.0:1. **Tachonics Corp., Plainsboro, NJ. INFO/CARD #227.**

### Linear Cellular Amplifier

The PA1991 has a 1 dB compression point of 44 dBm min over a 20 MHz



bandwidth in the 800 to 900 MHz band. It provides 9.5 dB of gain and a typical third-order intercept point of 52.5 dBm. This linear RF amplifier is priced at $1197 in 10- to 20-piece quantities. **Janel Laboratories, Inc., Corvallis, OR. INFO/CARD #226.**

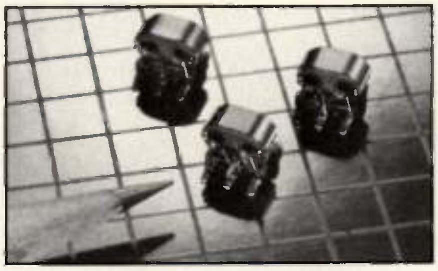
### Power Amplifiers

Witron introduces power amplifiers

with outputs of 300 W, 1 kW and 2 kW in frequency ranges from 1.5 MHz to 130 MHz. Input power for these units is 0 dBm into 50 ohms. The amplifiers are also available as fixed frequency generators. **Witron Industrie-Elektronik GmbH, Parkstein, West Germany. INFO/CARD #225.**

### Surface Mount Balun Transformers

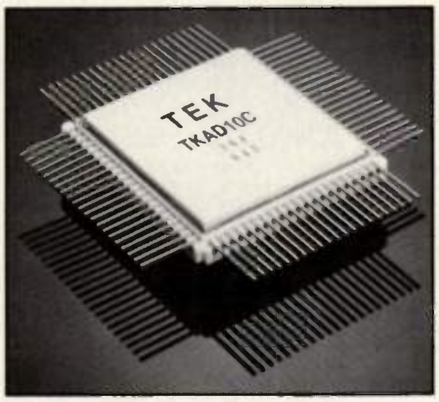
The B5F is a surface mount transformer designed for applications in impedance matching, double-balanced mix-



ers and splitter/directional couplers. The device exhibits good frequency response up to 1.3 GHz. **Toko America, Inc., Mt. Prospect, IL. INFO/CARD #224.**

### ADC Flash Converter

Tektronix introduces the TKAD10C 8-bit analog to digital converter in a ceramic surface mount package. The device exhibits 500 megasamples per second. It combines both a track-and-




hold amplifier chip and an A/D chip in an 84-pin package. Also contained is a T/H chip that contains an amplifier that reduces input-signal voltage requirements to 540 mV (peak-to-peak) and a pair of 8-bit 250 MSPS flash converters. Accuracy is 6.8 effective bits with a 250 MHz input signal. **Tektronix, Inc., Beaverton, OR. INFO/CARD #223.**

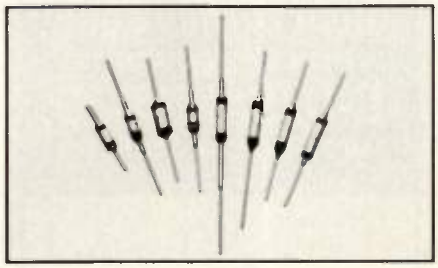
### EMI Test System

CCS-130 CAT is a computer-controlled system for emissions and susceptibility testing in the 16 Hz to 1 GHz range. It is comprised of two test receivers (EMC-11 and EMC-30) and an AT-compatible computer with a hard-drive, EGA color monitor and printer. The system includes a software package

that allows the user to develop, name and store custom tests. Also available is optional software which includes MIL-STD 461/462, MIL-STD 285A, MIL-STD 1541, CISPR, VDE and FCC. The standard system is priced at $69,900. **Electro-Metrics, Amsterdam, NY. Please circle INFO/CARD #222.**

### EMI/RFI Filters

Tusonix introduces a line of EMI/RFI filters designed for custom assemblies and interconnects. The Pi section filter



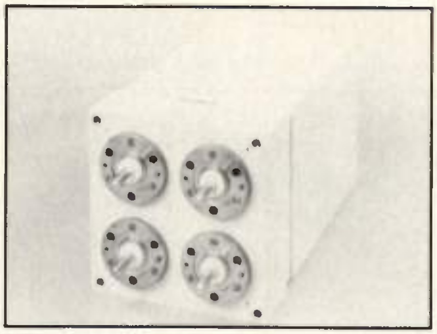
pins feature a capacitance range from 5 to 8000 pF and insertion loss up to 70 dB. The pins are tested to applicable requirements of MIL-F 15733. **Tusonix, Tucson, AZ. INFO/CARD #221.**

### Drop-In Isolator/Circulator

The TX8A1/CX8A1 operates from 860 to 900 MHz while providing 20 dB minimum isolation, 0.4 dB maximum insertion loss and 1.3:1 maximum VSWR. **Sonoma Scientific, Forestville, CA. INFO/CARD #220.**

### Transfer Switch

Model SRR-T-1 5/8-D is a remote rigid line transfer switch for use with 1 5/8 EIA equipment. Insertion loss is 0.1 dB, VSWR is 1.15:1 max and isolation is 65 dB min over the DC to 1 GHz range. Power handling capability is 2000 W CW at 1 GHz. In single quantities, the



price is $3,900. **RLC Electronics, Inc., Mt. Kisco, NY. INFO/CARD #219.**

### Voltage-Controlled Amplifier

Precision Monolithics introduces the SSM-2013 voltage-controlled amplifier that is designed for gain control of frequencies with a 800 kHz bandwidth. It has current-driven inputs and outputs and the 30 dB gain to 90 dB attenuation range is controlled by a voltage input. Packaging is 14-pin DIP and price is $2.60 in 100-piece lots. **Precision Monolithics Inc., Santa Clara, CA. Please circle INFO/CARD #218.**

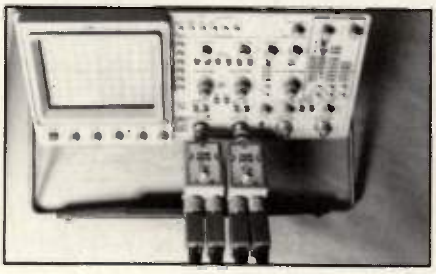
### Audio DAC

This 18-bit audio DAC is capable of 2X, 4X and 8X oversampling and accepts serial data directly from second generation digital filtering chips at rates

up to 12.7 MHz. Typical signal-to-noise ratio is 107 dB, slew rate is 9V/μs and settling time is 1.5 μs to 0.006 percent of full scale range. Model AD1860 is priced at $19.50 in 1000-piece quantity. **Analog Devices, Norwood, MA. Please circle INFO/CARD #217.**

### Oscilloscope Probe Switch

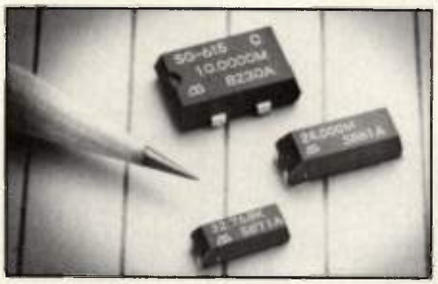
Microvolt Engineering introduces an oscilloscope probe switch that allows the selection of one of two probes to be connected to the oscilloscope input.



With the PX-1, a dual-channel instrument can accept four separate inputs. **Microvolt Engineering, Tustin, CA. INFO/CARD #216.**

### CMOS Crystal Oscillator

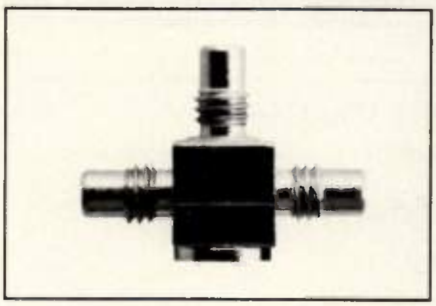
The SG-615 four-pin-type oscillators are MSO compatible with 5 ns rise and



fall times. Frequency range is 1.5 to 55 MHz. **Epson America, Inc., Torrance, CA. INFO/CARD #215.**

### Coaxial Adapter

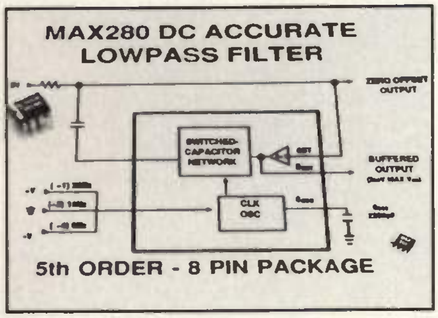
This 50 ohm coaxial adapter features a frequency range of DC to 10 GHz. Model PE9304 will mate type SMC plugs



that meet the interface requirements of MIL-39012. In 100-piece quantity, the adapter is $15.95. **Pasternack Enterprises, Irvine, CA. INFO/CARD #214.**

### Butterworth Lowpass Filter

Maxim introduces the MAX280 fifthorder Butterworth lowpass filter with a cutoff frequency that is tunable from DC to 20 kHz. This switched capacitor unit provides 1 percent cutoff frequency accuracy. The 8-pin plastic DIP pack-



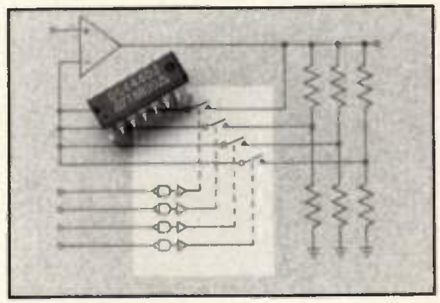

aged unit is priced at $3.55 when purchased in 100-piece quantity. A 16-pin small outline package is also available. **Maxim Integrated Products, Sunnyvale, CA. INFO/CARD #213.**

### SMA Coaxial Attenuators

Model 5510 is available for values of 3 dB, 6 dB, 10 dB, 12 dB, 14 dB and 20 dB. The attenuators have guaranteed risetimes of less than 12 ps. The standard value units are priced at $80 each. **Picosecond Pulse Labs, Inc., Boulder, CO. INFO/CARD #212.**

### Silicon Gate Analog Switches

The DG441, 442, 444 and 445 silicon gate analog switches are replacements for the metal gate DG201A, 202, 211 and 212, respectively. The DG400 family offers an on-resistance of 85 ohms, leakage current of 500 pA, power dissipation of 35 μW and 250 ns transition time. The 441 and 442 are available in 16-pin plastic and ceramic DIP, and SO packages. The 444 and 445 are available in 16-pin plastic DIP and SO packaging. Price ranges from $1.45 to $13.47 in 100-piece quantities. **Siliconix Inc., Santa Clara, CA. INFO/CARD#211.**



### Coaxial Adapter

Coaxial Components introduces a line of coaxial adapters from SMA to SMB and SMC. Also available are units between SMA, N, TNC, BNC, SMB, SMC and SSMA. **Coaxial Components Corp., Huntington, NY. Please circle INFO/CARD #210.**

# rf software

## Filter Design Software

Microwave Filter Design Kit designs microwave filter structures such as the edge coupled bandpass, the interdigital bandpass, the unfolded interdigital bandpass, and the Hi-Lo impedance lowpass filter. All distributed elements are defined by their geometrical and physical data such as line widths, lengths, substrate thickness, dielectric constant, loss tangent and ohmic conductivity. Responses such as Chebyshev and maximally flat, in addition to custom-defined ones, are provided. **Compact Software, Inc., Paterson, NJ. INFO/CARD #163.**

## Smith Chart Software

Smithsoft is a software tool that allows the placement of data directly onto a Smith chart with a mouse. The package is ideal for engineers who want to learn how to use the chart or want to improve their skills. As each component is selected from a menu, a color-coded arc is drawn on the chart showing all possible component values. The component symbol is added to the schematic window, thereby eliminating the need to remember which arc or rotation direction corresponds to each component type. Component values are calculated automatically and updated as the cursor is moved around the chart. Pull-down menus provide for drawing constant Q, constant gain, constant VSWR and constant stability circles. The software requires an IBM or compatible computer with 256k of memory, EGA and a mouse. The software is priced from $89. **Somersoft, Sebastopol, CA. INFO/CARD #162.**

## EMI Test Software

EZM-K1 is an interactive software package from Rohde & Schwarz that works in conjunction with ESH 3 and ESVP test receivers and the EZM spectrum monitor. Together they provide the capability to measure RFI voltages, currents, field strengths and power, as well as conducted and radiated interference. Frequency range is 20 Hz to 1300 MHz. Measurements are split into two stages, a preliminary measurement of the entire interference spectrum and a final measurement at critical frequencies. **Rohde & Schwarz, Lanham, MD. Please circle INFO/CARD #161.**

## RF Design Software Service

As a convenience to our readers, computer programs published in *RF Design* are now available on disk. For a minimal cost, you can avoid the time-consuming (and error-prone) task of typing program listings into your computer.

This month's disk includes programs (MS-DOS format) described in these articles:

"CAD For Lumped-Element Matching Circuits," p. 102
"Modeling PLL Tracking of Noisy Signals," p. 113
Request disk number RFD-0289

Prices are $9.00 for a 5¼ in. diskette, or $10.00 for a 3½ in. mini-floppy, postpaid. Outside of the U.S. and Canada, add $8.00 (disks will be sent airmail). Make check or money order payable to *RF Design Software Service*, and send orders to:

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**Littleton, Colorado 80161**

### 10th Anniversary Catalog

This anniversary catalog from JFW describes the company's products, capabilities and services. Introduced in the catalog are new products including an IEEE 488 interface for use with programmable attenuators and switches, attenuators and switches using thick-film hybrid techniques, and high power switches that are rated from 10 to 250 watts. Also included is a price list. **JFW Industries, Inc., Indianapolis, IN. Please circle INFO/CARD #160.**

### Tutorial on Arbitrary Function Generators

Arbitrary function generators are a class of signal generators that use digital techniques to produce custom analog waveforms. This tutorial covers the theory of operation of these instruments and shows how complex wideband waveforms can be created and generated. Also explained are the advantages of Easywave[R]. This package is LeCroy's PC-based waveform generation software which supplies all the tools to create, store and generate complex waveforms. The tutorial concludes by evaluating specifications and explaining tradeoffs between speed, resolution, memory depth and ease of programming. **LeCroy Corp., Chestnut Ridge, NY. INFO/CARD #159.**

### Coaxial Cable Catalog

Andrew announces the availability of a 112-page publication that contains a selection guide and technical resources for a wide number of cable applications including broadcast, land mobile, cellular and microwave. Featured is a broadcast systems planning guide for the selection of components for a transmission line system. Average power ratings, attenuation and system component tables for connectors and accessories are included. Prices are listed on each product page. **Andrew Corp., Orland Park, IL. INFO/CARD #158.**

### Software Data Sheet

This data sheet describes Wiltron's Option 04 dual source control software for the Model 360 vector network analyzer. Highlighted are two graphs on typical applications and an explanation of how Option 04 has the ability to provide dual source measurements. Also demonstrated is how the package allows the user to separately control up to two sources and a receiver without the need for an external controller. **Wiltron, Morgan Hill, CA. INFO/CARD #157.**

### Discrete Military Data Book

Motorola introduces a data book that includes chip qualification, process flow and packaging information, as well as a selector guide and complete data sheets for their standard MIL-tested discrete semiconductor chips. The line is made up of the more popular devices from a number of standard Motorola military product lines. **Motorola Inc., Semiconductor Products Sector, Phoenix, AZ. INFO/CARD #156.**

### Bulletin Describes Flexible Coaxial Cable

The HCF 12-50J flexible Cellflex[R] coaxial cable is described in this bulletin. It outlines electrical specifications such as maximum operating frequency, impedance, velocity of propagation, attenuation, and DC resistance. Mechanical specifications including nominal size, outer and center conductor, jacket material, maximum pulling force and minimum bending radius are highlighted. An attenuation and average power chart and a diagram of the HCF 12-50J connector type N male is included. **Cablewave Systems, North Haven, CT. Please circle INFO/CARD #155.**

### Amplifier Catalog

Detailed in this catalog are a wide range of high intercept point amplifiers covering multiple octaves from 100 kHz to 1300 MHz. Specifications of connectorized and drop-in configurations are provided. An article addressing the subject of testing amplifiers with high dynamic ranges is presented. **Advanced Milliwave Laboratories, Inc., Westlake Village, CA. INFO/CARD #154.**

### Testing and Inspection Services Literature

TUV Rheinland of North America announces the availability of literature describing their testing and inspection services. The categories described include product safety, ergonomics, medical equipment, software, RF interference and components. The types of tests that can be performed are outlined so that the reader can determine the compliances and approvals necessary to meet the mandatory requirements for all equipment bound for Germany and other European countries. **TUV Rheinland of North America, Danbury, CT. INFO/CARD #153.**

### 1989 General Catalog

Pomona introduces their 1989 general catalog of electronic test accessories that describes 900 test products. The products covered include plugs, jacks, adapters, coaxial, triaxial and audio connectors, test clips, probes, patch cords, cable assemblies, and SMD test products. Also featured are static control products, oscilloscope probe kits, SMD and VLSI test products. **Pomona Electronics Division, Pomona, CA. INFO/CARD #152.**

## Advertiser Index

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| Capacitance Cj (pF) MAX | .03 | .06 | .02 | .07 |
| Lifetime TL (ns) TYP | 50 | 50 | 150 | 150 |
| Switching Time TS (ns) TYP | 5 | 5 | 15 | 15 |

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INFO/CARD 119

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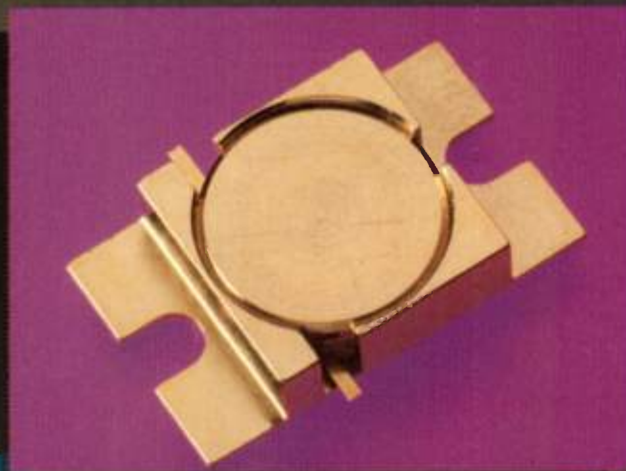

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