RF design

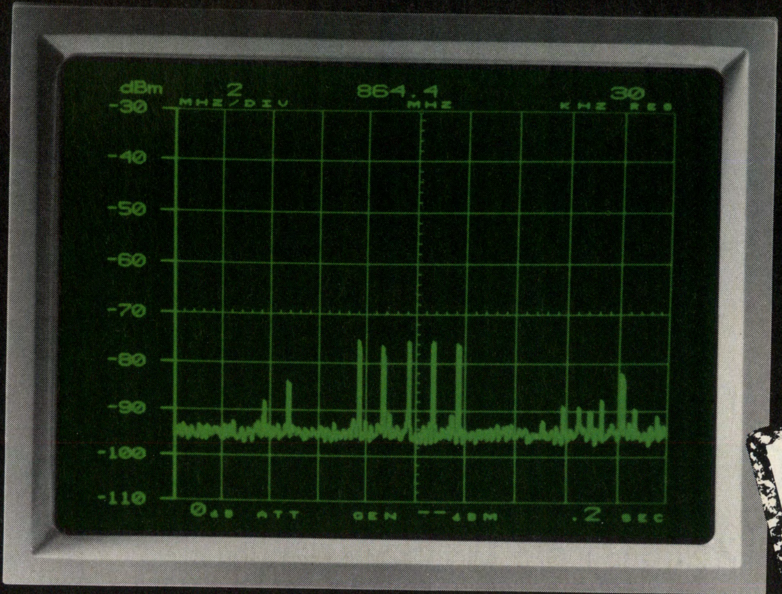
engineering principles and practices

January 1993

Cover Story
New Oscillators Reach for Lower Noise

Featured Technology
Computer Modeling Applications

If you're not using a spectrum analyzer...



Spectrum analyzer display of cellular telephone transmissions.

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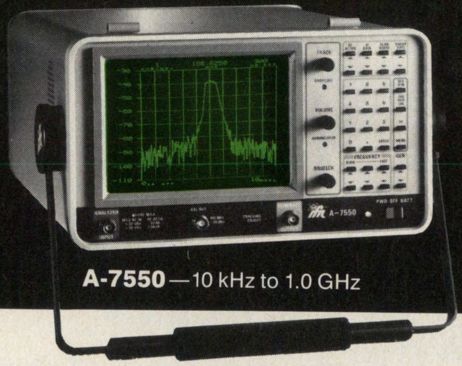
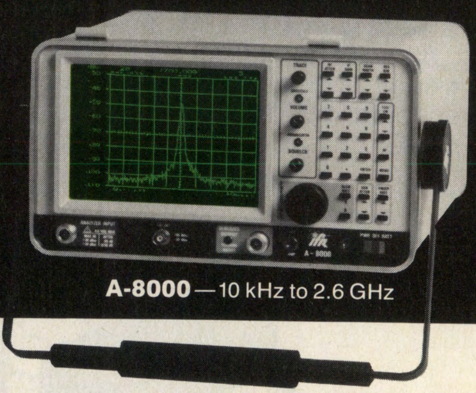
maybe it's time to take a closer look.

IFR Spectrum Analyzers



A-7550—10 kHz to 1.0 GHz



A-8000—10 kHz to 2.6 GHz

You can't hear them or see them, but there are probably thousands of electromagnetic signals being transmitted through the air around you at this very moment.

Many of these signals originate from radio and television transmission. However, many other signals are emitted from computers, CRT display terminals, and other electrical devices.

Each of these signals can be a potential source of harmful interference to sensitive electrical equipment or critical systems you are operating.

The IFR A-7550 and A-8000 Spectrum Analyzers provide a full set of features to speed the identification and measurement of signals over a wide frequency range.

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For more information or a demonstration of the A-7550 or the A-8000, contact your local IFR representative or call IFR at (316) 522-4981.

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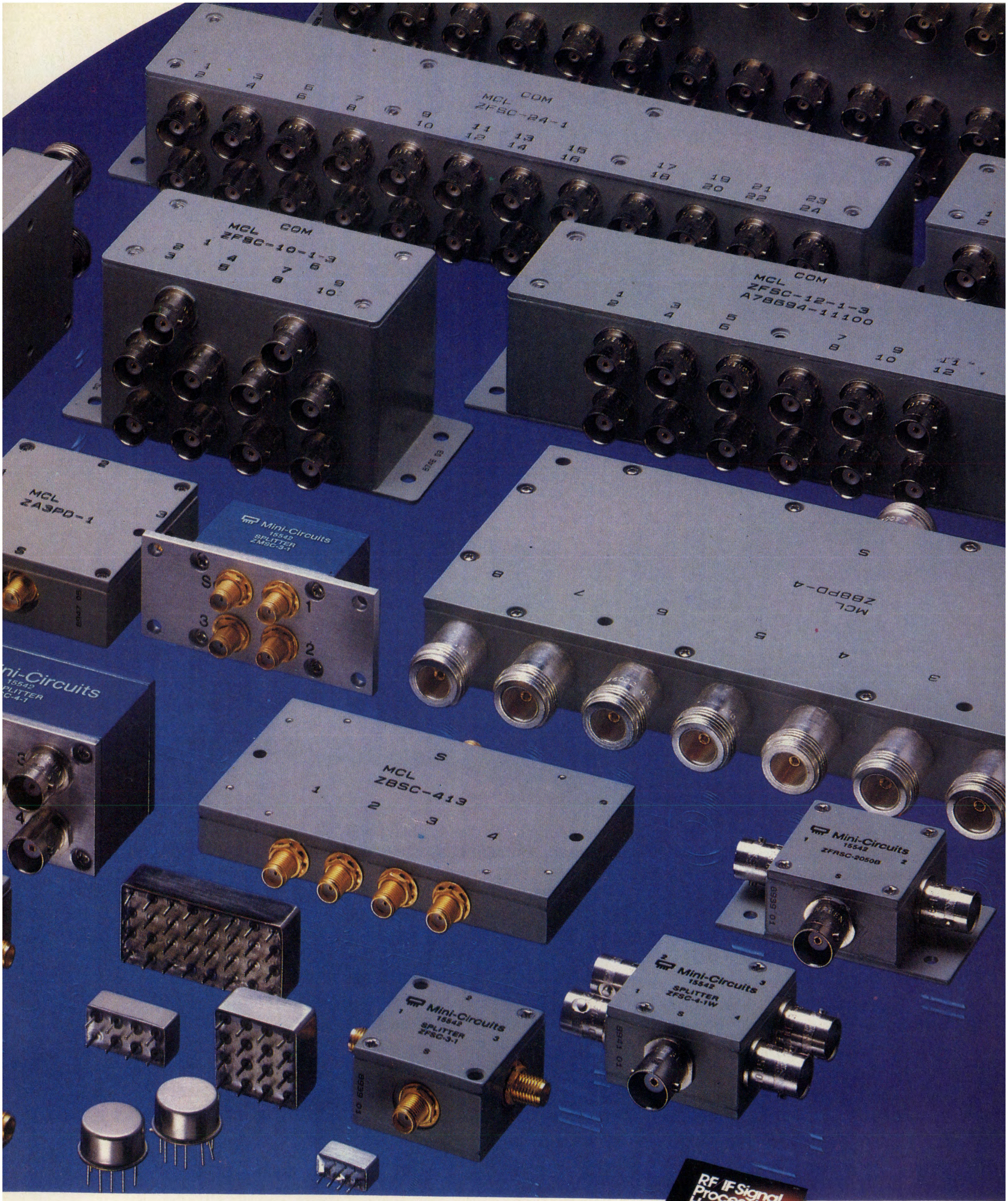
INFO/CARD 2

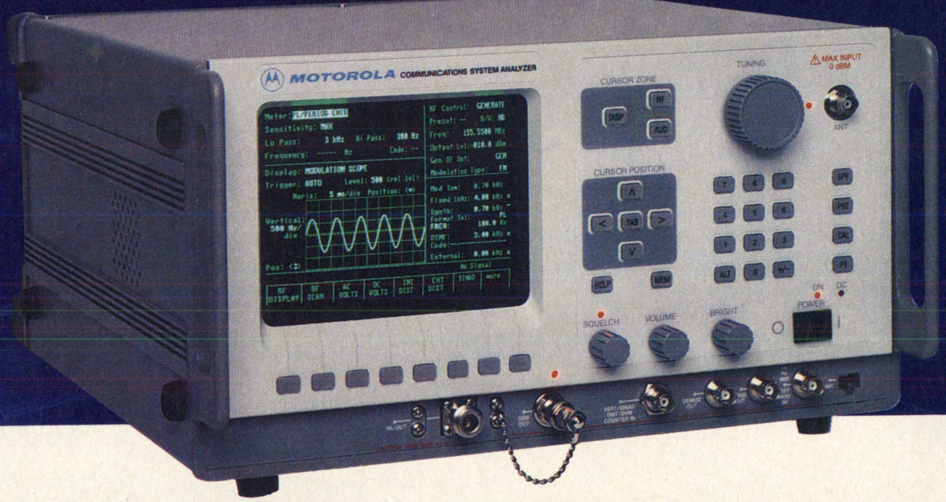


For detailed specs on all Mini-Circuits products refer to
• THOMAS REGISTER Vol. 23 • MICROWAVES PRODUCT DIRECTORY
• EEM • MINI-CIRCUITS' 740-pg HANDBOOK.

INFO/CARD 3

F134 REV. C

R2600



All that test equipment you've been using, with its endless cabling and "we mean business" knobs and switches, has served you well over the years. But someday you're going to need to replace it. That someday is today!
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MOTOROLA

INFO/CARD 4

RF design

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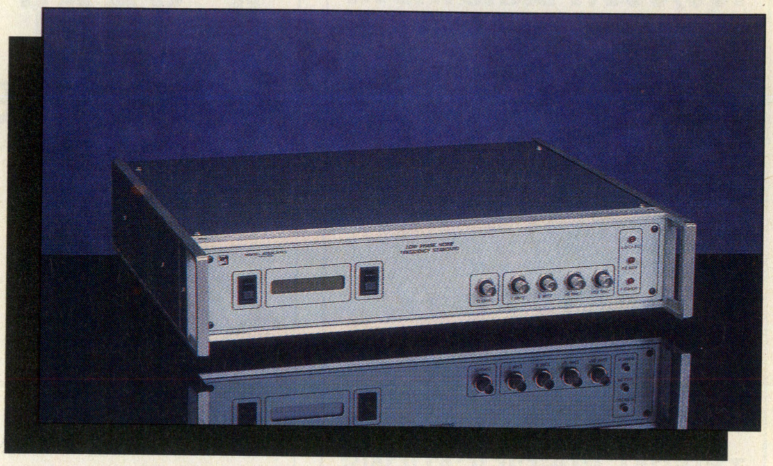
January 1993



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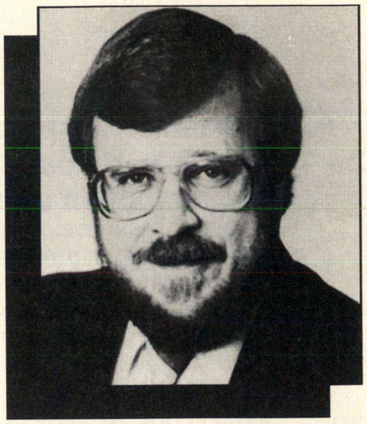
RF editorial

1993: A Year of Change



By Gary A. Breed
Editor

Change is the operative term for 1993! A new President will be inaugurated this month and one political party is in firm control of the government. Bill Clinton will be the first President who grew up after World War II, representing the "baby boom" generation that makes up a huge part of U.S. society. For better or worse, there will be change in our Federal government.

In the world of RF, personal communication equipment (PCN, PCS) is ready to take off. Low earth orbit (LEO) systems for paging, voice and data communications could see their first hardware. This could be the year that the FCC approves standards for Advanced Television, with Digital Audio Broadcast standards coming, as well. If that's not enough, most analysts see the cellular boom continuing through 1993.

There has been another major change — a change in thinking about what makes technology exciting. For over 30 years, the term "high tech" has meant pushing the laws of physics, or reaching the limits of engineering understanding. These things will always be exciting, but with a slowing world economy, there is renewed glamour in "making things that work."

Today's emphasis is on practical products that put technology to work in new ways. Engineering creativity now includes more than just translating a function into hardware. Engineers are refining existing theory and reducing complex (and expensive) concepts into inexpensive, reproducible circuitry. They are making products that are not only technically advanced, but elegantly designed to be easy to use, and at the lowest cost.

The roles and attitudes of RF engineers are changing, too. The image of an engineer stuck in his cubicle designing one perfect circuit is gone. An engineer now needs to know all the RF he or she can, then add knowledge of digital and analog control circuitry, digital signal processing, plus governmental rules and regulations — with a little promotional and marketing attitude thrown in. Instead of being reclusive, top engineers are part of a team; communicating with engineers in other departments, talking to customers, helping purchasing and sales staffs.

Here at *RF Design*, we've seen these changes happening. Over the past two or three years, the number of articles submitted for our consideration has grown rapidly as engineers recognize that exposure to the "outside world" is good for them and their companies. In the companies we talk to, more marketing and customer relations positions are being filled with engineers rather than someone with sales experience. The overall sophistication of the engineering community is advancing rapidly.

Change is not always good for everyone, however. Engineers laid off from defense-oriented companies will have a hard time finding jobs in the new cost-conscious marketplace. To make things worse, many of the RF products in development will not be manufactured in the U.S. Let's hope that today's change leads to a better combination of creative design *and* efficient manufacturing.

RF design

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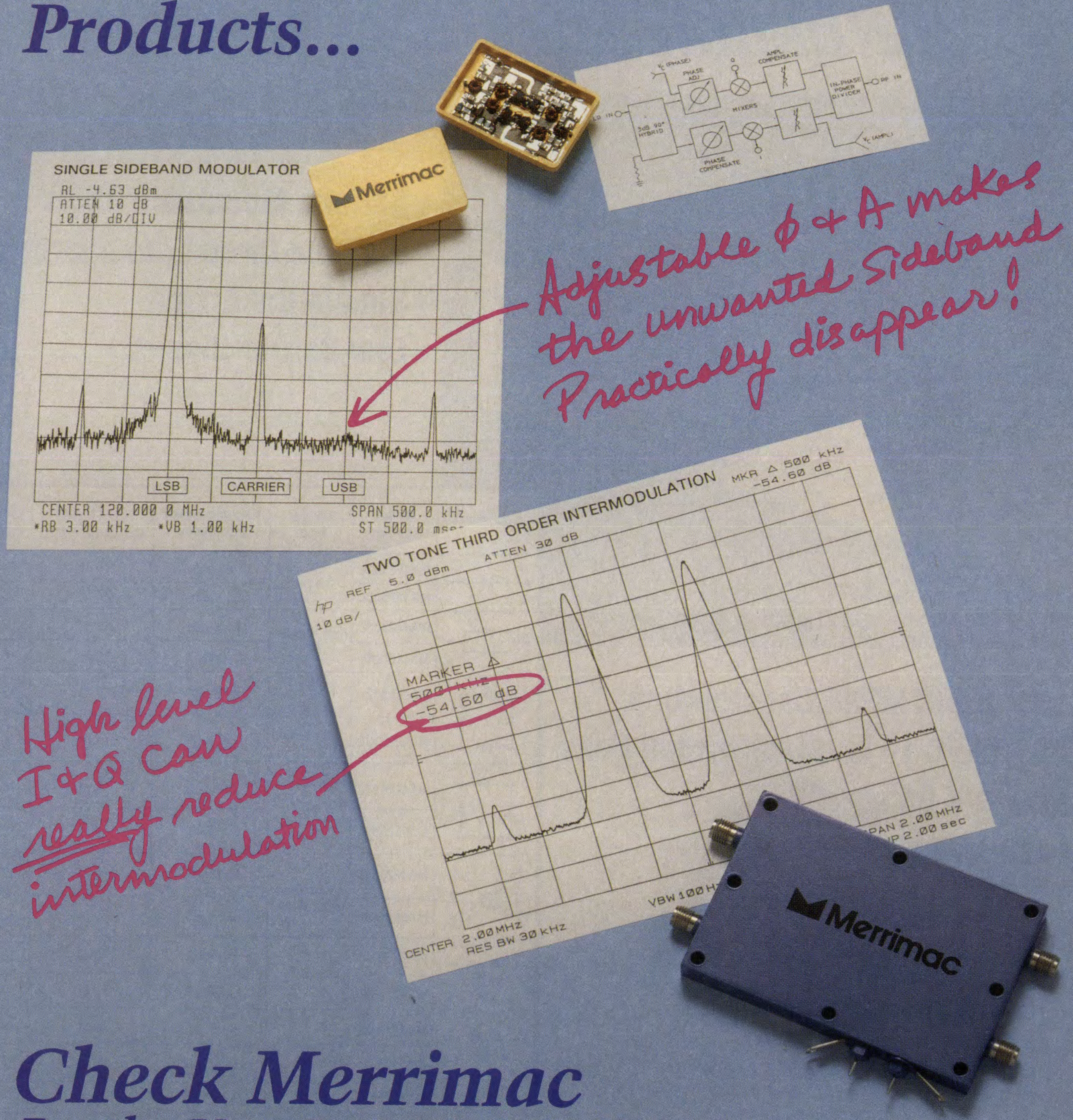

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INFO/CARD 8

RF calendar

January

11-14 **Design & Test Expo**
Anaheim, CA
Information: Miller Freeman, Inc., 13760 Noel Road, Suite 500, Dallas, TX 75240. Tel: (800) 223-7126. Fax: (214) 419-7915.

19-20 **Portable Intelligence Expo 1993**
San Francisco, CA
Information: PIX 93, Cardiff Publishing, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax:: (303) 770-0253.

February

7-11 **NEPCON West '93**
Anaheim, CA
Information: Michael Critser, Director of Conferences, Reed Exhibition Companies, 1350 E. Touhy Ave., Des Plaines, IL 60018. Tel: (708) 299-9311. Fax: (708) 635 -1571.

March

17-19 **RF Expo West**
San Jose, CA
Information: Barb Binge, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773 -9716.

30-31 **The 1993 Mid-Lantic Electronics Show**
King of Prussia, PA
Information: Mid-Lantic Electronics Show '93, Judith Ginsberg, 4113 Barberry Drive, Lafayette Hill, PA 19444. Tel: (215) 828-2271. Fax: (215) 941-6773.

30-2 **8th International Conference on Antennas and Propagation**
Edinburgh, UK
Information: IEE, Savoy Place, London WC2R 0BL. Tel: (44) 071 240 1871. Fax: (44) 071 497 3633.

April

18-21 **The 4th IEE Conference on Telecommunications**
Manchester, UK
Information: ICT 93 Secretariat, Conference Services, IEE, Savoy Place, London, WC2R 0BL, United Kingdom.

18-22 **Symposium on Ceramics for Wireless Communication**
Cincinnati, OH
Information: Henry O'Bryan, AT&T Bell Laboratories. Tel: (908) 582-6980. Fax: (908) 582-2521.

27-29 **Electro**
Edison, NJ
Information: Electronic Conventions Management, PO Box 92275, Los Angeles, CA 90009-2275. Tel: (800) 877-2668. Fax: (310) 641-5117.

RF courses

Dielectrics for Advanced Microelectronics
February 22-23, 1993, San Francisco, CA
Information: University of California-Berkeley, Continuing Education in Engineering. Tel: (510) 642-4151. Fax: (510) 643-8683.

Phased-Array Antennas: Theory, Design and Technology
January 26-29, 1993, Atlanta, GA
Antenna Engineering
February 2-5, 1993, Atlanta, GA
Principles of Pulse Doppler Radar: High, Medium and Low PRF
February 9-11, 1993, Atlanta, GA
Coherent Radar Performance Estimation
February 17-19, 1993, Atlanta, GA
Radar Signal Processing: Theory, Technology and Applications
February 22-25, 1993, Atlanta, GA
Radar Cross Section Reduction
March 23-26, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

Antennas: Principles, Design, and Measurements
March 10-13, 1993, St. Cloud, FL
Information: Southeastern Center for Electrical Engineering Education, Kelly Brown. Tel: (407) 892-6146

RF/Microwave Circuit Design II: Linear/Nonlinear Techniques and Applications
February 1-5, 1993, Los Angeles, CA
High-Resolution Microwave Imaging: Principles and Applications
February 23-26, 1993, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825-1047. Fax: (310) 206-2815.

Digital Signal Analysis
February 22-24, 1993, Tempe, AZ
Fiber Optic Communications
March 3-5, 1993, Tempe, AZ
Antenna Analysis, Design, and Measurements
March 8-11, 1993, Tempe, AZ
Information: Center for Profesional Development, College of Engineering and Applied Sciences, Arizona State Unviersity. Tel:(602) 965-1740 Fax: (602) 965-8653

EMC Measurement and Regulation
January 18, 1993, St. Petersburg, FL
January 20, 1993, Albuquerque, NM
January 22, 1993, Phoenix, AZ
February 15, 1993, San Francisco, CA
February 17, 1993, Los Angeles, CA
February 19, 1993, San Diego, CA
Information: Tektronix Inc. Tel: (800) 426-2200.

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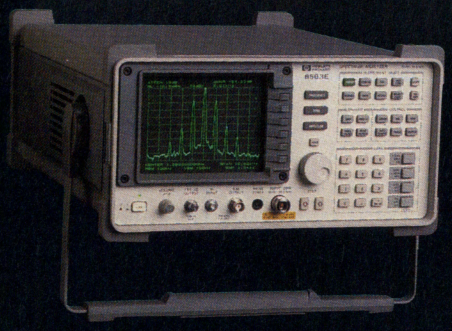

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Of course, you'll want more details than we can provide here. So call your local HP office or one of the numbers listed below to get an ordering guide with detailed product information. Plus one very important thing.

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INFO/CARD 14

GE Sells Aerospace Division to Martin Marietta

In one of the biggest deals in the defense contracting industry, Martin Marietta has agreed to purchase General Electric's aerospace division for $3 billion. The General Electric division includes: GE Aerospace, GE Government Services, Knolls Atomic Power Lab and Machinery Apparatus Operation with a 1991 revenue of $6 billion. General Electric will receive about $2 billion in cash and another $1 billion in convertable preferred stock. Martin Marietta is now ranked as the United State's largest defense firm.

Frequency Control Symposium Call for Papers — The 1993 IEEE Frequency Control Symposium has issued a call for papers for its conference to be held June 2-4, 1993, in Salt Lake City, Utah. Authors are invited to submit papers dealing with recent progress in research, development and applications in the following areas: fundamental properties of piezoelectric crystals, theory and design of piezoelectric resonators, resonator processing techniques, filters, SAW devices, quartz crystal oscillators, microwave and millimeter wave oscillators, synthesizers and other frequency control circuitry, atomic and molecular frequency standards, noise phenomena and aging, frequency and time coordination and distribution, sensors and transducers and applications of frequency control. Two copies of a 500 word summary, with a topic indicated in the upper right hand corner of the first page, together with the author's name, address and telephone number should be sent to: Mr. Jack Kusters, 52U/07, Hewlett-Packard Company, 5301 Stevens Creek Blvd., PO Box 58059, Santa Clara, CA 95052-8059. Tel: (408) 553-2041. Fax: (408) 246- 5925. The deadline for submission of summaries is January 15, 1993.

Republics Join ITU — The International Telecommunications Union has announced that the Republic of Moldova and the Republic of Bosnia and Herzegovina both joined the ITU in October, 1992. Moldova is a former USSR republic and Bosnia Herzegovina was part of Yugoslavia. This brings ITU membership to 174 countries.

Setting Standards for Optical Fibers — The telecommunications industry has teamed up with NIST's Time and Frequency Division to develop phase noise and synchronization standards for new optical fiber communication systems. Since 1991, the institute has participated in industry-wide meetings to develop standards for the proposed optical fiber network. NIST will continue to work with the industry on synchronization and timing questions, which promise to get more severe as communication data rates increase.

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	AP-400B	B	80kHz - 2.7MHz	400W	55dB (±1.5dB)
	240L	A	20kHz -10MHz	40W	50dB (±1.5dB)
	2100L	A	10kHz -12MHz	100W	50dB (±1.5dB)
	HF				
	A-150	A	300kHz -35MHz	150W	55dB (±1dB)
	A-300	A	300kHz -35MHz	300W	55dB (±1dB)
	A-500	A	300kHz -35MHz	500W	60dB (±1dB)
	A-1000	A	300kHz -35MHz	1000W	60dB (±1.5dB)
	VHF				
	325LA	A	250kHz -150MHz	25W	50dB (±1.5dB)
	3100LA	A	250kHz -150MHz	100W	55dB (±1.5dB)
	3200L	A	250kHz -120MHz	200W	55dB (±1.5dB)
	UHF				
	400AP	A	150kHz -300MHz	3W	37dB (±1dB)
	403LA	A	150kHz -300MHz	3W	37dB (±1dB)
	411LA	A	150kHz -300MHz	10W	40dB (±1.5dB)
	525LA	A	1MHz -500MHz	25W	50dB (±1.5dB)
	550L	A	1.5MHz-400MHz	50W	50dB (±1.5dB)
	5100L	A	1.5MHz-400MHz	100W	50dB (±1.5dB)
	603L	A	800kHz -1GHz	3W	40dB (±1.5dB)
	630L	A	400MHz-1GHz	30W	51dB (±2dB)
	6100L	AB	400MHz-1GHz	100W	51dB (±2dB)

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Packaging Conference Call for Papers — The International Electronics Packaging Conference to be held September 12-15,1993 in San Diego, California has issued a call for papers. Abstracts of unpublished work in recent developments in the following areas of electronic packaging are requested: device packaging, chip on board, surface mount, ceramic technology, RF/mixed signal applications, reliability, thermal management, microwave/millimeter wave packaging, high density substrates, interconnects, and materials. A 300 word abstract must be submitted before February 1, 1993. Send eight copies of the abstract to 1993 IEPS Program Committee, 114 N. Hale St., Wheaton, IL 60187-5113. Tel: (708) 260-1044. Fax: (708) 260-0867.

NIST/Industry R&D Agreement — NIST has signed a cooperative research and development agreement with Wiltron to assess the performance of commercial automatic network analyzers (ANAs) used to calibrate microwave system components. ANAs have replaced many manual systems for microwave calibrations throughout government and industry. NIST pioneered the development of six-port ANAs, which enhanced ANA technology. While the six-port device is still available for NIST calibrations, it is costly for the user. NIST wants to determine whether it can achieve calibration integrity by using commercial instruments at a lower cost to customers. During the three-year agreement with Wiltron, NIST will explore the feasibility of using commercial ANAs for NIST calibration services and develop procedures for validating their uses in routine calibrations. For more information, contact Bob Judish, Div. 813.01, NIST, Boulder, CO 80303. Tel: (303) 497-3380.

Stetco Sells Thirty Percent of Stock — Stetco, Inc. recently announced the sale of thirty percent of its corporate stock to Fastron GmbH. Stetco will continue to serve as Fastron's agent in the United States and Mexico, and Fastron Malaysia will represent Stetco's sales interests in Southeast Asia. The two companies will also work closely together in developing new products for the electronic marketplace.

Harris Wins Naval Contract — The U.S. Navy Space and Naval Warfare Systems Command has awarded Harris Corporation RF Communications Division a contract to supply high frequency radio communications systems for the Navy's surface ships. The contract is for $180 million over a six-year period, with an initial award of $33.5 million. Harris will supply state-of-the-art, solid-state shipboard communications systems, based on a newly developed approach to broadband HF radio architecture. The new design permits simultaneous operation of multiple transmitters and receivers in a collocated environment while minimizing mutual interference.

BFGoodrich Awarded TCAS Contract — BFGoodrich Flight Systems was recently awarded its first major contract for its Traffic Alert and Collision Avoidance System. Mesa Airlines selected BFGoodrich to provide the TCAS I for all seven divisions of the airline. Terms of the contract were not released.

Imaging Technology Center Receives NSF Grant — The National Science Foundation has awarded a $1.2 million grant to finance a new technology center for electronic imaging. The Center for Electronic Imaging Systems was established by Harris Corporation and other Rochester, New York based high-tech corporations and universities. The NSF grant is being matched by contributions from New York State and private industry, giving the center a total of $3.6 million over four years.

Milcom Awarded Amplifier Contract — Milcom International has received a $200,000 contract from a major Far East electronics manufacturer. The contract is for the development of an 800 MHz, 1000 Watt ultra linear power amplifier for digital cellular telephone applications. Prototypes will be delivered in early 1993 with project completion in the spring of 1993.

Retliff Receives Laboratory Approval — Retliff Testing Laboratories has been granted "Approved Laboratory Status" by Interference Technology International, a United Kingdom-

based, European Community-designated Competent Body. To receive the certification, a one week on-site review of Retliff's Quality Assurance, Calibration and Administrative Procedures as well as its Training and Technical Capabilities, was conducted. As a result, Retliff's test data can now be accepted for use in the Technical Construction Files ITI generates to attest to the compliance of products to the EC EMC Directive.

Education Foundation Endowed by Wayne Kay Estate — The estate of E. Wayne Key has endowed $1.5 million to the Manufacturing Engineering Education Foundation of the Society of Manufacturing Engineer's in Kay's memory. The monies will be used exclusively for scholarships provided by the SME Foundation. For information regarding the SME Foundation and the Wayne Kay Scholarship Fund, contact the SME Foundation, One SME Drive, PO Box 930, Dearborn, MI 48121. Tel: (313) 271-1500, ext. 511 or 512.

Cubic Defense Systems Awarded $12.8 Million Contract — Cubic Defense Systems was recently awarded a $12.8 million contract for their personnel locator systems by the U.S. Army's Communications and Electronics Command. The contract is a follow-on to an existing one and calls for $9 million with an option for an additional $3.8 million. Deliveries are scheduled during fiscal years 1993-1995.

Teklogix and Atlet AB Sign Strategic Agreement — Teklogix International and Atlet AB recently announced the signing of an agreement for the distribution of Teklogix radio frequency data communications (RF/DC) systems in Europe. Atlet and Teklogix already had a business alliance in the United Kingdom and will now distribute Teklogix RF/DC products and systems in Sweden and Holland. Teklogix's new European subsidiary, Teklogix GmbH will take over all of the Atlet Mobile Terminal Systems staff in Germany.

Malaysian Radio System Contract — The Government of Malaysia has signed an $81 million contract with Motorola's Land Mobile Products Sector to provide the Royal Malaysian Police with a public safety radio communication system. The contract was awarded in an open tender for the supply, delivery and commissioning of an integrated VHF telecommunication system. Under the terms of the contract, Motorola will provide the Royal Malaysian Police with its Advanced Securenet System™. The system offers communication security through digital voice encryption. The new system will be commissioned by early 1994.

Webb Laboratories Relocates — Webb Labs has announced its relocation to a new facility. Their new address is 13731 W. Capitol Drive, Suite 260, Brookfield, WI 53005. Their phone and facsimile numbers are unchanged.

Test Equipment Trends — Processing Power and Automation

By Gary A. Breed
Editor

Recent demonstrations of new test equipment products make two points clear: 1) more power is being put into each box, board or module; and 2) greater access to the units for control and communication is deemed necessary. These enhancements to performance and productivity offer design, test and service personnel greater capabilities, but they require special attention to "human engineering" to avoid making instruments difficult to use due to overcomplexity.

Control and Communications

Microprocessor and standard communications buses make test instruments more accessible to the rest of the engineering environment. Extensions to the IEEE-488/GPIB bus have increased its capability, and the continuing growth in VXI bus instrumentation systems is an example of reliance on communications among the various test, analysis and record-keeping functions.

On a smaller scale, local setups can be controlled by the internal processor that may be in one or more of the instruments used. For example, Hewlett-Packard has optional IBASIC controller software, external keyboards, and simplified programming capabilities to eliminate the need for a dedicated controller. Tests can be coordinated among various instruments and collected for network collection or diskette storage.

At the single-box level, screen display interfaces to plotters and printers allow hard-copy communications with greater ease than ever. The IFR 8000 Spectrum Analyzer, for example, has an internal controller that is essentially a PC, allowing standard interfaces to peripheral devices. Preset test sequences let engineers run standard tests without having to set them up each time. Anritsu-Wiltron has also done extensive work in developing test and analysis software for key applications for their 360 network analyzer.

Processing Power

Inside the latest generation of high performance test instruments is real computing power using both microprocessor and dedicated digital signal processing (DSP) components. Mathematical manipulation of measured data allows quick decisions instead of post-test analysis. Limit lines, frequency and bandwidth computations and searches for predictable harmonic and spurious responses save considerable operator time. Fast Fourier Transform (FFT) analysis in DSP allows precise time- and frequency-domain analysis. The availability of computing power has created fast frequency counters, which measure the period of each cycle of a waveform instead of the conventional many cycles per unit time.

In signal generating equipment, processing power allows for complex modulation that is completely arbitrary (within the limits of the digital architecture used). Spread spectrum signals, complex vector modulation, video, chirp and pulse formats are all available in a single instrument. Multiple tests can be performed under computer control with fewer instruments.

High precision digital processing requires analog circuitry that is better than ever. High dynamic range front ends, low distortion amplification, low noise, fast frequency synthesizers and optimal filtering are all essential in the analog portion of digitally- enhanced test equipment. This combination of technologies has led to rapid advances in the performance of RF test equipment.

A logical follow-on is greater integration of instruments, computers, controllers and accessories. This is especially important in automated test systems, where instruments must be compatible with a common controller. Development of instruments for the VXI bus is primarily aimed at automated test systems. VXI bus products are generally smaller, and can be combined into a compact test package, saving room on the manufacturing floor when used in production testing. Racal-Dana, Hewlett-Packard, EIP Microwave, Gigatronics and others continue to invest in VXI bus products.

PC-based instrumentation is beginning to increase. General purpose instruments such as voltmeters, data acquisition systems and low frequency oscilloscopes have been around for some time, but few RF products have been available until recently. Among the most complex PC-card instruments is the Morrow Technologies 9052 spectrum analyzer that has high performance specifications and plenty of operating features. Inline Components offers an HF receiver card with full programmability for surveillance and other applications. Also offering PC-card products are Pole-Zero with hopping filters, and Guide Technology, which offers fast frequency counters.

Equipment Cost Trends

Competition and customer demand has led to an interesting situation — many features and processing power are being added without increasing the cost over previous models. In some cases, prices are being reduced as performance edges ahead. Some of this new-found value is in response to a slow economy, but it is also partly a result of competition. Economy instruments have improved, closing the gap between themselves and the bottom of the premium instrument lines. At the same time, the state-of-the-art has been advancing for the most critical applications. This overall upward trend in performance is helping instrumentation customers keep their costs under control. The instrumentation industry has been significantly affected by the economic downturn of the past few years. Test equipment is part of a customer's capital budget, which has been cut drastically at most firms. Only by offering more performance at the same or lower cost can instrument makers entice their customers to fight for extra test equipment budget dollars in a tight economy. *RF*

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WAVETEK

Simulating Lossy Transmission Lines With PSpice®

By Walter Banzhaf, P.E.
Ward College, University of Hartford

MicroSim Corp. has added a lossy transmission line element to its analog circuit simulator PSpice, allowing attenuation and dispersion to be included in circuit simulations. This now makes PSpice, an established tool for the analog circuit designer, a powerful software package for the RF circuit designer.

The Berkeley SPICE analog circuit simulator, in addition to basic circuit elements (independent and dependent voltage and current sources, resistors, capacitors, inductors and semiconductors) includes a transmission line circuit element which is lossless. While this is a reasonable approximation for physically short lines, such as those that might be found within an integrated circuit or on a small circuit board, for many other applications the transmission line loss and dispersion must be included if circuit behavior is to be predicted accurately. MicroSim Corporation, which developed PSpice, included the capability for describing a lossy transmission line in a recent release (1) (Version 5.1, January 1992).

When a circuit simulation is done, an input file is submitted to SPICE or PSpice which describes all circuit elements and the kind of analysis desired. The location of circuit elements is done by specifying the numbers of the circuit nodes to which it is connected. The type of element is indicated by the first letter of the element name: resistors begin with R, capacitors with C, and transmission lines with T. In order to describe a lossless transmission line to SPICE (or PSpice), an element line must be added to the input file. The element line would look like:

```
T_STUB 5 0 12 26 Z0=72 F=50MEG NL=0.5
T_STUB 5 0 12 26 Z0=72 TD=10E-9
```

The two element lines above are equivalent and describe the same line; the first locates the input of lossless transmission line T_STUB between nodes 5 and 0, the output between nodes 12 and 26, and describes a line with a characteristic impedance of 72 ohms which at 50 MHz has an electrical length of 0.5 wavelength. The second element line expresses the line length by using its delay time of 10 ns.

A transmission line with loss can be described to PSpice in a similar way; however, instead of specifying the characteristic impedance, the primary line constants, R (loss per meter), L (inductance per meter), C (capacitance per meter) and G (conductance per meter) must be given along with the physical (as contrasted with electrical) length of the line. An element line for a lossy transmission line would look like:

```
T_STUB 5 0 12 26 LEN=3.048 R=2 L=.3U
+ G=50U C=120P
```

This line is 3.048 meters (10 feet) long, and has primary line constants of 1.665 ohms, 0.3uH, 50 uS and 120pF per meter. The plus symbol (+) in the second line makes it a continuation of the first line. Notice that the characteristic impedance, attenuation and electrical length are not explicitly stated. The next section will illustrate how to derive primary line constants from transmission line data supplied by manufacturers.

Data about a transmission line obtained from manufacturers' literature (2) often includes the characteristic impedance, the capacitance per foot, and the attenuation in dB per 100 feet. For example, for an RG-58C coaxial cable, Z_0 is 50 ohms, C is 30.8 pF/ft, and at 100 MHz, loss is 4.9 dB/100 ft. In order to describe a lossy transmission line to PSpice, the primary line constants (R, L, G and C) must be extracted from the data above. The first step is to convert C to pF/meter by multiplying by 3.2808, resulting in C = 101 pF/m. Since for high frequencies (3) $Z_0 = \sqrt{(L/C)}$, L can be found from:

$$L = Z_0^2(C) = 50^2(101\times10^{-12}) = 0.2526\ \text{nH/m} \qquad (1)$$

The loss of 4.9 dB/100 feet can easily be converted to 0.1608 dB/m (by dividing by 30.48), which in turn is 0.0185 neper/m (1 neper = 8.686 dB). When a transmission line is operated at high frequencies, (valid when $\omega L >> R$ and $\omega C >> G$) an approximate expression for transmission line loss, α, in neper/m, is: (4)

$$\alpha \approx R/(2Z_0) + G(Z_0/2) \qquad (2)$$

In the absence of detailed knowledge about skin effect losses and dielectric losses, one way to convert this loss to the primary line constants R and G involves making an educated guess. Since most of the loss of a coaxial transmission line at high frequencies comes from the resistance of the center conductor (R) (5), as an approximation 90 percent of the loss can be assigned to R. This is reasonable, as the dielectric losses are small compared to the conductor losses for most transmission lines up through microwave frequencies (5). Thus, $0.9(\alpha) \approx R/(2Z_0)$, and R can be found from:

$$R \approx 0.9(\alpha)(2Z_0) \approx 0.9(0.0185)(2)(50) \approx 1.665\ \Omega/\text{m} \qquad (3)$$

Similarly, if the remaining 10 percent of the loss is assigned to G, $0.1(\alpha) \approx G(Z_0)/2$, and

$$G \approx 0.1(2)(\alpha)/Z_0 \approx 0.1(0.0185)(2)/(50) \approx 74\ \mu\text{S/m} \qquad (4)$$

A more analytical method which will give the R and G constants for a transmission line requires that the loss tangent of the dielectric, δ, be known at the operating frequency. For high frequency operation, the relationship between δ and G is (6):

$$G = \omega C \tan\delta \qquad (5)$$

For RG-58C, which has a solid polyethylene dielectric, the loss tangent is approximately 0.0002 at 100 MHz (7).

Once δ is known, G can easily be found, and then R can be determined. To illustrate how to use this method, at 100 MHz:

$$G = 2\pi(100\times10^{6})(101\times10^{-12})\tan(2\times10^{-4}) = 12.7\ \mu S/m \qquad (6)$$

Since $\alpha \approx R/(2Z_0) + G(Z_0/2)$, and α was found previously to be 0.0185 neper/m, the equation can be rearranged to give R:

$$R = [\alpha - G(Z_0)/2](2Z_0) = 2Z_0\alpha - GZ_0^{2}$$

$$R = 2(50)(.0185) - 12.7\times10^{-6}(50)^{2} = 1.82\ \Omega/m \qquad (7)$$

The results for R are close for the two methods, while the results for G differ considerably. However, since the effect of the G parameter is quite small compared to the R parameter, often the simpler method where 90 percent of the loss is attributed to R and 10 percent is attributed to G gives satisfactory results. Once the primary line constants have been determined, the lossy transmission line, at 100 MHz, can now be described to PSpice. This will be illustrated in Example 1, below, using the parameters obtained with the simpler method.

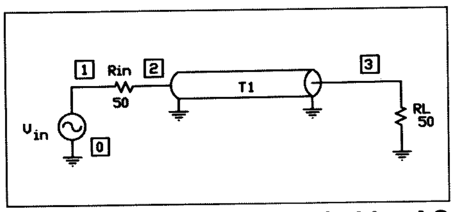

Figure 1a. Matched line fed by AC source.

```
EXAM1.CIR        Lossy RG-58C, Sinusoidal Input
VIN  1  0  AC  2
RIN  1  2  50
*        Units below are per meter,
*        determined from manufacturer's data.
T1  2  0  3  0   LEN=30.48  R=1.665
+ L=.2526U    G=74U     C=101P
RL  3  0  50
.AC  LIN  21   80MEG   120MEG
.PROBE
.END
```

Figure 1b. PSpice input file EXAM1.CIR.

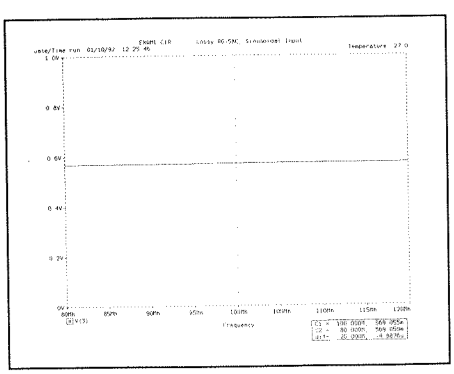

Figure 1c. PSpice analysis results for EXAM1.CIR.

Example 1: Matched Line At 100 MHz

A 100-foot length of RG-58C will be described to PSpice as a lossy transmission line, terminated by a matched load, connected to a sinusoidal source which is varied from 80 MHz to 120 MHz. Figure 1a is the diagram of the circuit, with node numbers shown. Note that the generator has a 50 ohm output impedance and an open-circuit voltage of 2 V. The input file which describes the circuit is shown in Figure 1b, where the primary line constants (R, L, G and C) are given along with the length (30.48 meters, or 100 feet).

The PSpice analysis results, Figure 1c, show that the load voltage, V(3), is constant at 569 mV as the frequency varies from 80 MHz to 120 MHz. If the line were lossless, the load voltage would be 1 V. The line loss, given by 20(log(1V/.569V), is 4.90 dB, which agrees with the original manufacturer's data at 100 MHz. It should be noted here that the parameters R and G, which determine the loss of the transmission line, are constants to PSpice, whereas in reality they increase with frequency (R varies with the square root of frequency, and G varies linearly with frequency) (8).

Example 2: Pulses On A Matched Line

A 100 MHz pulse train is fed into a ten-foot length of RG-58C which is terminated by a matched load. The circuit is shown in Figure 2a, and Figure 2b is the input file describing this circuit. Notice that the length parameter, LEN, is 3.048 (understood to be in meters), equivalent to ten feet. Figure 2c shows the pulses at the transmission line input on top, and the pulses across the load resistor on the bottom. A ten-foot length of RG-58C has a propagation time of 15.4 ns, which can be seen as the time delay between the first pulse occurring at the input and the output. The output pulses are slightly lower in amplitude than the input pulses, due to line attenuation.

Example 3: Z_{in} of Unmatched Line at 100 MHz

Figure 3a shows a one amp sinusoidal current source feeding a ten-foot lossy RG-58C transmission line which is terminated by a five ohm resistor. A one amp current source is used so that the voltage at the transmission line input is the same as the input impedance of the line. The load has a VSWR of 10:1, which means that as frequency varies the input impedance of the transmission line can vary between a minimum of 5+j0 ohms and a maximum of 500+j0 ohms (if the line were lossless). The impedances between 5+j0 ohms and 500+j0 ohms will be reactive, with a non-zero imaginary component. The input file describing this circuit is shown in Figure 3b, and specifies that PSpice perform an AC analysis at 201 frequencies linearly spaced between 80 MHz and 120 MHz.

The input impedance is presented two

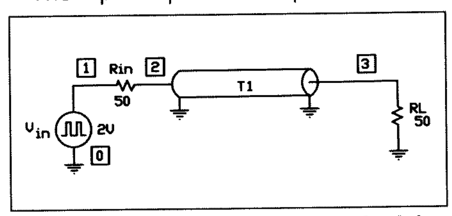

Figure 2a. Matched line fed by pulse source.

```
EXAM2.CIR        Lossy RG-58C, Pulse Input
VIN  1  0  PULSE(0  2  2N  .1N  .1N  1N  10N)
RIN  1  2  50
*        Units below are per meter, determined
*        from manufacturer's data, at 100 MHz.
T1  2  0  3  0   LEN=3.048  R=1.665
+ L=.2526U    G=74U     C=101P
RL  3  0  50
.TRAN .1N 30N
.PROBE
.END
```

Figure 2b. PSpice input file EXAM2.CIR.

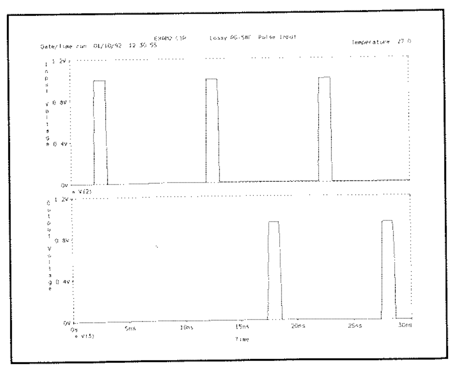

Figure 2c. PSpice analysis results for EXAM2.CIR.

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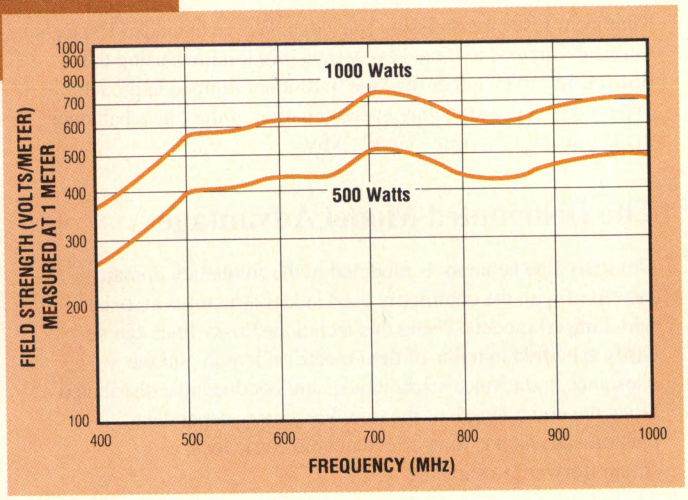

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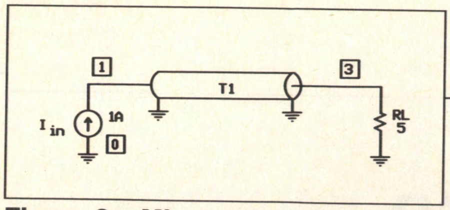

Figure 3a. Mismatched line fed by sinusoidal current source.

```
EXAM3.CIR          Lossy RG-58C, Sinusoidal Input
*         Current Input Gives Zin Directly
IIN  0  1  AC  1
*         Units below are per meter,
*         determined from manufacturer's data.
T1  1  0  3  0  LEN=3.048  R=1.665
+ L=.2526U    G=74U     C=101P
RL  3  0  5
.AC  LIN  201  80MEG   120MEG
.PROBE
.END
```

Figure 3b. PSpice input file EXAM3.CIR.

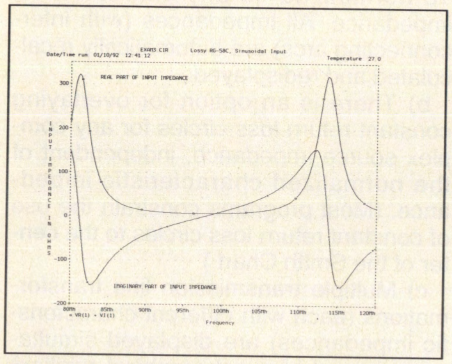

Figure 3c. Real and imaginary input impedance vs. frequency.

different ways: Figure 3c is a graph of the real and imaginary parts of the input impedance of the transmission line versus frequency, while Figure 3d plots the imaginary impedance versus the real impedance. Figure 3c shows that the real part of the input impedance reaches a maximum of 321 ohms at about 81.2 MHz, and a minimum of 7.8 ohms around 100 MHz; at both these frequencies, since the imaginary component is zero, the input impedance is purely real. Due to the attenuation of the transmission line, the maximum input impedance is less than 500 ohms and the minimum is more than 5 ohms. The cyclical nature of the input impedance can be seen also in Figure 3d, whereas frequency varies the plot of the impedance imaginary part versus the impedance real part describes a circle. If the PSpice internal transmission line model allowed loss to increase with frequency, Figure 3d would be a spiral whose radius decreased as frequency increased.

Conclusion

PSpice now has the capability of simulating circuits containing lossy transmission lines. The transmission line is described to PSpice by its primary line constants (R, L, G and C) and its physical length. The primary line constants for any transmission line can easily be derived from manufacturers' data. Information on PSpice, including an evaluation version, may be obtained from MicroSim Corporation, 20 Fairbanks, Irvine, CA 92718, (714) 770-3022. *RF*

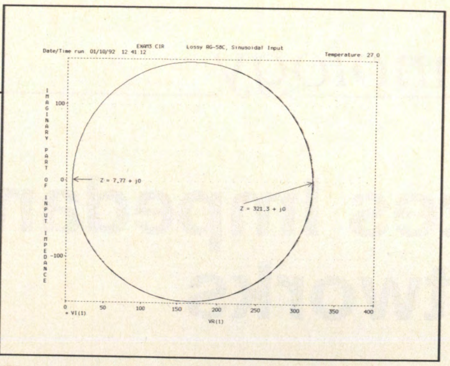

Figure 3d. Imaginary input impedance vs. real input impedance.

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8. Stephen F. Adam, *Microwave Theory and Applications*, Prentice Hall, Inc., 1969, p. 41.

About the Author

Walter Banzhaf is professor and department chair at Ward College of Technology, University of Hartford. He has BEE and MEE degrees from RPI, and previously worked on antenna systems at the Naval Underwater Systems Center. He is the author of the Prentice Hall book *Computer-Aided Circuit Analysis Using PSpice*, and can be reached at: U of Hartford, West Hartford, CT 06117, or by telephone at (203) 768-4764.

MIMP Analyzes Impedance Matching Networks

By Dan Moline
Motorola Semiconductor Products Sector

This is the age of computer aided design. Most engineers have sophisticated tools at their disposal to synthesize, analyze, optimize and help fabricate RF circuits. These programs are typically priced in the thousands of dollars. Their degree of complexity and versatility generally warrants this price. So, what earthly good is a "freeware" program that can only do simple impedance matching?

Motorola's Impedance Matching Program(1) (MIMP, programmed using the Microsoft® C Professional Development System, Version 6.0) is a specialized form of CAD specifically developed for RF power amplifier circuit design. It provides a simple environment for entering and analyzing impedance matching circuitry. Commercially available programs include a multitude of circuit elements and provide numerous analytical capabilities. However, MIMP focuses only on impedance transformations. This type of analysis is more important than others for most RF power amplifier design problems, since data sheets for RF devices present only large signal impedances, representing a single combination of frequency, voltage, power level and power dissipation. These impedances must generally be transformed to another set of impedances (50 ohms or the input/output impedance of another device). To do this, MIMP includes a standard library of passive circuit elements, including various combinations of capacitors, inductors and transmission lines in both series and shunt configurations. It also contains a unique, distributed capacitance element that models a capacitor distributed along a transmission line.

The real nucleus of MIMP is its computer aided Smith™(2) Chart. It is uncommon for CAD programs to incorporate the benefits of manipulating actual impedance transformations on a Smith Chart. If commercially available programs were used for merely impedance matching, a typical final result of a computer run would be: S11 vs. frequency. To supplement this, many RF designers still keep a Smith Chart, compass, straight edge and pencil handy so they can pictorially represent each circuit component's contribution to the total transformation. MIMP's Smith Chart facility provides this service — electronically. It also displays each circuit element's contribution to the total impedance transformations. Simply stated, it is an electronic form of a standard, printed Smith Chart. Some of the unique features of the Smith Chart display screen include:

a) The Smith Chart can be instantly 're-normalized' to any characteristic impedance. All impedances (with interconnecting arcs) are automatically recalculated and redisplayed.

b) There is an option for overlaying constant return loss circles for any complex source impedance, independent of the normalized characteristic impedance. (Most programs constrain the use of constant return loss circles to the center of the Smith Chart.)

c) Multiple transmission line transformations (each with different characteristic impedances) are displayed simultaneously and in exact graphical relationships to each other, independently of the Smith Chart's normalized impedance. (Drawing transmission line transformations by hand requires an iterative denormalize/renormalize/replot/redraw procedure.)

d) A tabular impedance display is provided to view the impedance at any 'node'.

e) Constant Q arcs can be added to

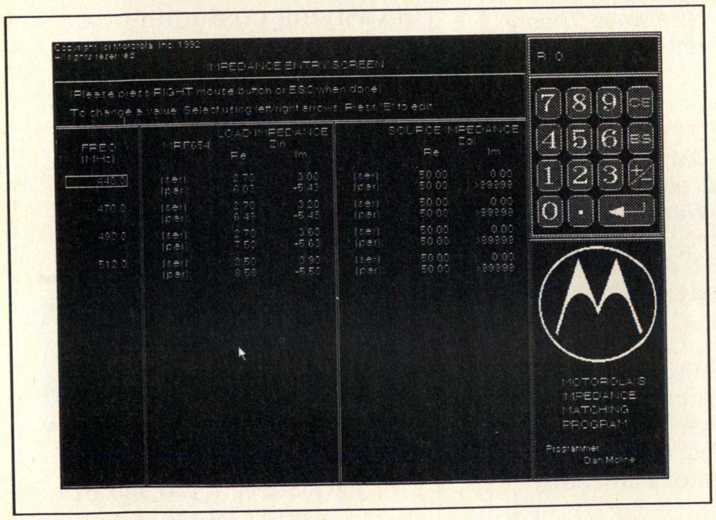

Figure 1. Impedance entry screen.

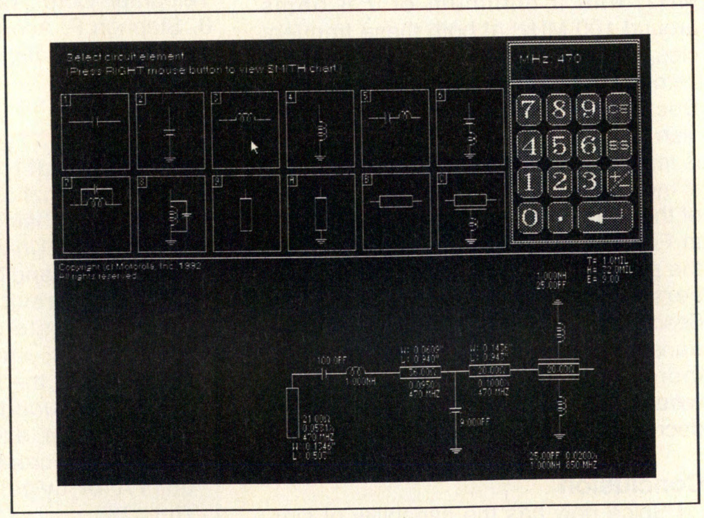

Figure 2. Circuit entry screen.

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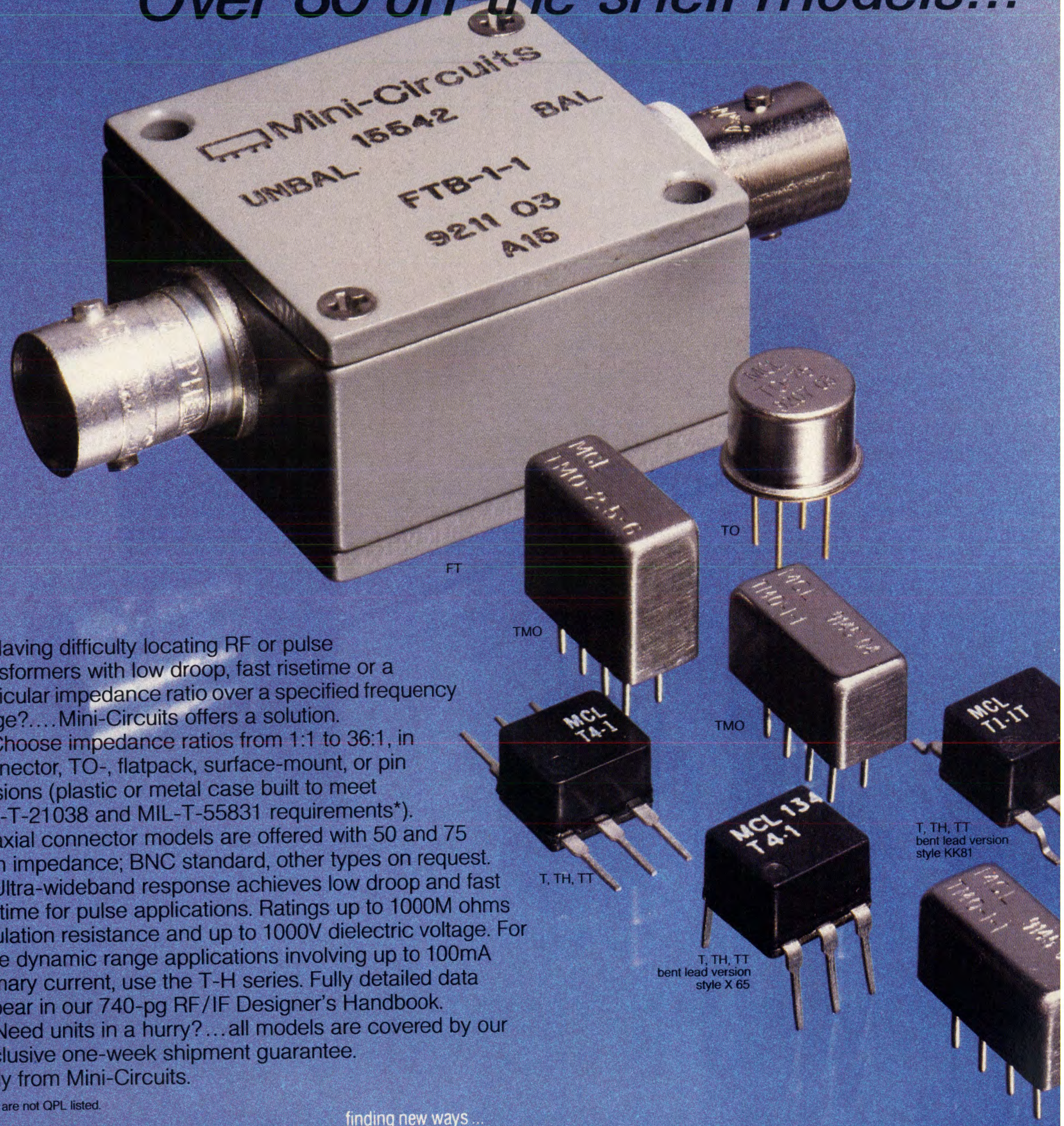

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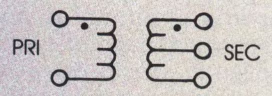
A* 	T	T1-1T	1	.05-200	.05-200	.08-150	.2-80	4.45
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		T2.5-6T	2.5	.01-100	.01-100	.02-50	.50-20	4.95
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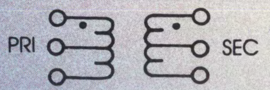
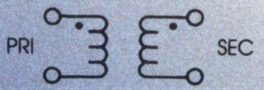
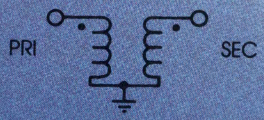
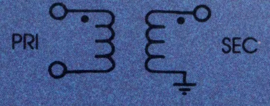
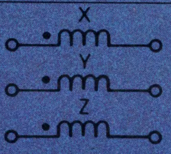
		T8-1T	8	0.3-140	0.3-140	0.7-90	1-60	7.95
		T13-1T	13	.3-120	.3-120	.7-80	5-20	4.95
		T16-6T	16	.03-75	.03-75	.06-30	.1-20	5.65
	TH	T4-1H	4	10-350	10-350	15-300	25-200	5.95
	TMO	TMO1-1T	1	.05-200	.05-200	.08-150	.2-80	7.95
		TMO2-1T	2	.07-200	.07-200	.1-100	.5-50	8.45
		†TMO2.5-6T	2.5	.01-100	.01-100	.02-50	.05-20	8.45
		†TMO3-1T	3	.05-250	.05-250	.1-200	.5-70	7.95
		TMO4-1	4	.2-350	.2-350	.35-300	2-100	6.25
		TMO5-1T	5	.3-300	.3-300	.6-200	5-100	8.45
		TMO13-1T	13	.3-120	.3-120	.7-80	5-20	8.45
B* 	TT	TT1-6	1	.004-500	.004-500	.02-200	.1-50	6.95
		TT1.5-1	1.5	.075-500	.075-500	.2-100	1-50	5.95
		TT2.5-6	2.5	.01-50	.01-50	.025-25	.05-10	6.45
		TT4-1	3	.05-200	.2-50	.2-50	1-30	5.95
		TT4-1A	4	0.1-300	0.1-300	0.2-250	0.3-180	6.95
		TT25-1	25	.02-30	.02-30	.05-20	.1-10	9.95
	TTMO	TTMO25-1	25	.02-30	.02-30	.05-20	.1-10	11.95
		TTMO1-1	1	.005-100	.005-100	.01-75	.05-40	11.45
		TTMO4-1A	4	0.1-300	0.1-300	0.2-250	0.3-180	13.95
C 	T	T1-1	1	.15-400	.15-400	.35-200	2-50	3.25
		T1.18-3	1.18	0.01-250	0.01-250	0.02-200	0.03-50	5.65
		T1-6	1	.01-150	.01-150	.02-100	.05-50	5.65
		T1.5-1	1.5	.1-300	.1-300	.2-150	.5-80	4.45
		T1.5-6	1.5	.02-100	.02-100	.05-50	0.1-25	5.65
		T2.5-6	2.5	.01-100	.01-100	.02-50	.05-20	4.45
		T4-6	4	.02-200	.02-200	.05-150	.1-100	4.45
		T9-1	9	.15-200	.15-200	.3-150	2-40	3.95
		T16-1	16	.3-120	.3-120	.7-80	5-20	4.45
		T36-1	36	.03-20	.03-20	.05-10	.1-5	6.95
	TO	TO-75	1	10-500	—	10-500	40-250	6.95
	TH	T1-1H	1	8-300	8-300	10-200	25-100	5.95
		T9-1H	9	2-90	2-90	3-75	6-50	6.45
		T16-1H	16	7-85	7-85	10-65	15-40	6.45
	TMO	TMO1-02	1	1-800	1-800	2-500	—	9.45
		TMO1-1	1	.15-400	.15-400	.35-200	2-50	6.25
		TMO1.5-1	1.5	.1-300	.1-300	.2-150	.5-8	8.45
		†TMO2.5-6	2.5	.01-100	.01-100	.02-50	.05-20	7.95
		†TMO4-6	4	.02-200	.02-200	.05-150	.1-100	7.95
		TMO6-1	6	.3-200	.3-200	.5-150	5-50	7.95
		TMO9-1	9	.15-200	.15-200	.3-150	2-40	7.95
		TMO16-1	16	.3-120	.3-120	.7-80	5-20	7.95
D 	T	T2-1	2	.050-600	.050-600	.1-400	.5-200	3.95
		T3-1	3	.5-800	.5-800	2-400	—	4.45
		T4-2	4	.2-600	.2-600	.5-500	2-250	3.95
		T8-1	8	.15-250	.15-250	.25-200	2-100	3.95
		T14-1	14	.2-150	.2-150	.5-100	2-50	4.95
	TMO	TMO2-1	2	.050-600	.050-600	.1-400	.5-200	7.95
		TMO3-1	3	.5-800	.5-800	2-400	—	8.45
		TMO4-2	4	.2-600	.2-600	.5-500	2-250	7.95
		TMO8-1	8	.15-250	.15-250	.25-200	2-100	7.95
		TMO14-1	14	.2-150	.2-150	.5-100	2-50	8.45
	FT	FT1.22-1	1.22	.005-100	.005-100	.01-50	.05-25	35.95
		FT1.5-1	1.5	.1-400	.1-400	5-200	1-100	35.95
E 	FTB	FTB-1	1	.2-500	.2-500	.5-300	1-100	36.95
		FTB1-6	1	.01-125	.01-125	.05-50	.1-25	36.95
		■FTB-1-75	1	.5-500	.5-500	5-300	10-100	36.95
F 	T	T-622	1	0.1-200	0.1-200	0.5-100	5-80	3.25
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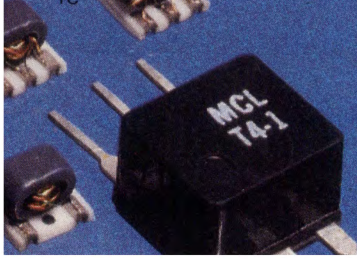

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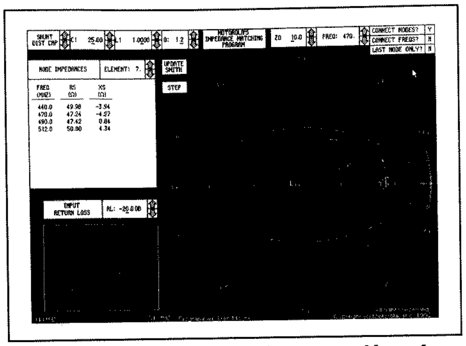

Figure 3. Smith Chart display screen.

the Smith Chart.

f) Real time changes in the impedance transformation are displayed while individual circuit elements are tuned. This utility is provided to perform manual circuit optimization. A scalar display of Input Return Loss is updated simultaneously as an additional tool for optimization.

RF engineers, who choose to use Motorola's RF Power Transistors in their designs, can access the auxiliary database. Included with MIMP are the input and output impedances for many of the RF power devices contained in Motorola's *RF Data Book*.

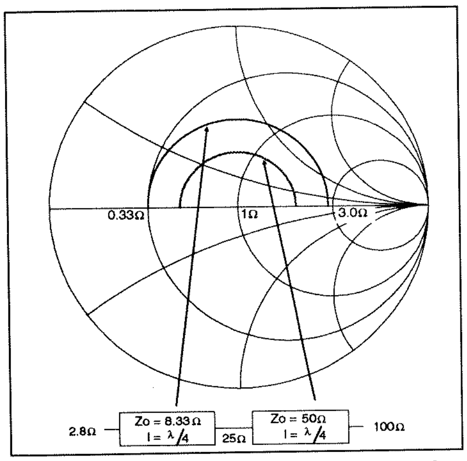

Figure 4. Two different transmission lines normalized and plotted on same Smith Chart.

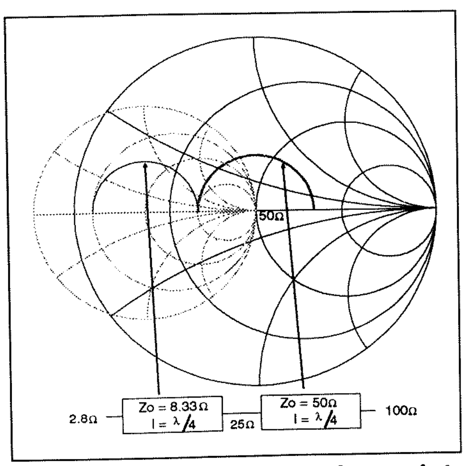

Figure 5. Transmission lines plotted on separately renormalized reflection coefficient planes.

Program Description

Motorola's Impedance Matching Program (MIMP) is divided into three screens: the impedance entry screen, the circuit entry screen and the Smith Chart display screen. A mouse is recommended for easy entry and manipulation of data. However, there are keyboard equivalents for most of the mouse functions. Pressing the ESC key or the right mouse button cycles through the impedance entry screen, circuit entry screen and Smith Chart display screen (after minimum data entry has been completed). While in the Smith Chart display screen, pressing both left and right mouse buttons returns the operator immediately to the circuit entry screen.

Hardware requirements are an IBM compatible with 640k, an 80286 or higher and a VGA graphics adapter. A mouse is recommended.

Impedance Entry Screen

To begin the program, change to the directory in which MIMP.EXE resides. This directory must also contain the following three files:

DEVFILE.ASC (This file is a listing of the MRF devices and the location of its associated impedance data in the datafile.) DATAFILE.BIN (This file contains the impedance information for each device included in DEVFILE.ASC.) HELVB.FON (This is a Microsoft C font file.)

The program begins in the impedance entry screen, which is separated into four basic sections: the frequency table, load impedance table, source impedance table and data entry keypad. The first prompt requests the user to specify the number of frequencies at which source and load impedances will be entered. Up to 11 frequencies can be entered. If a standard Motorola device is to be specified, the user is prompted to press ENTER. (MIMP includes a database of input/output impedances for MRF, 2N, JO and TP devices.) Mouse users may enter numeric data by pointing and clicking on the keypad display, or numeric data may be entered from the keyboard.

User Entry

The program first prompts the user for the number of frequencies to be entered. The program will only accept values less than or equal to 11. (If zero is entered, the program advances to an standard device entry sequence; as described later.) After the number of frequencies has been supplied, each frequency should be entered sequentially starting with the lowest value. When the last frequency has been entered, the user is prompted to supply the load impedance data for each frequency. An option to specify 50 ohms is supported by simply pressing ENTER. After all the load impedance data is furnished, the source impedance data is requested. Again, there is an option to specify 50 ohms by pressing ENTER. (Parallel equivalents are calculated and displayed for all impedances.) After the

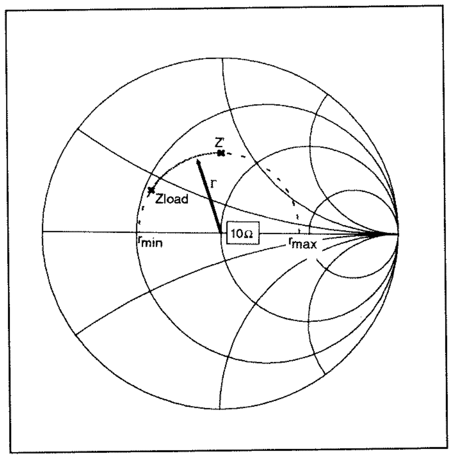

Figure 6. Load and transformed impedances plotted and real axis impedances interpolated.

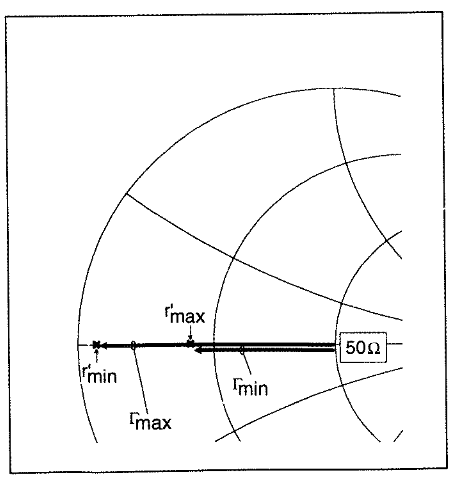

Figure 7. $\Gamma_{max/min}$ determined via renormalized $r_{min/max}$.

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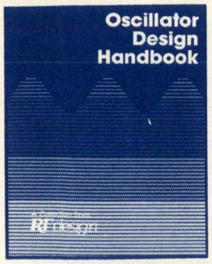

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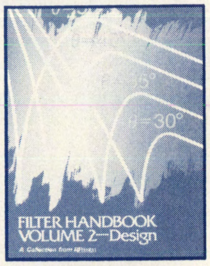

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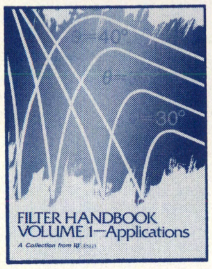

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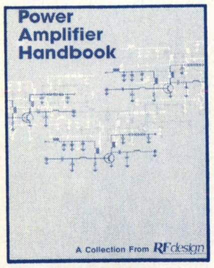

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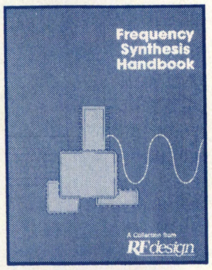

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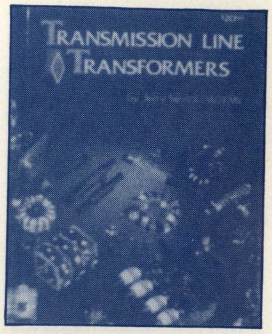

This book is becoming widely recognized as the best text on broadband transformers in the HF range. After an introduction that covers the classic Ruthroff and Guanella approaches, the author proceeds to describe and analyze well over a hundred different designs, including detailed construction information and extensive measured frequency response and loss data. It is one of the few technical books that combines rigorous engineering accuracy with down-to-earth practical information.

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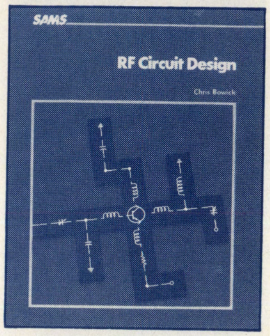
By Chris Bowick



The author introduces RF circuit design at the most basic level, starting with individual components and proceeding through resonant circuits, transmission lines, filters, the Smith chart, impedance matching, small-signal amplifiers and power amplifiers. It is an excellent teaching text for beginning engineers, and a valuable reference for fundamental RF principles.

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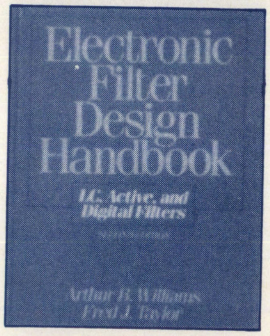

A modern classic for active, passive and digital filter design, this book is an indispensible engineering reference. Basic information is provided on filter types, passband and stopband characteristics, mathematical analysis and various topologies. Extensive tables of filter parameters are included for ease of design, and notes on component selection assist in construction. Completeness is enhanced with chapters on phase shift networks and delay equalizers.

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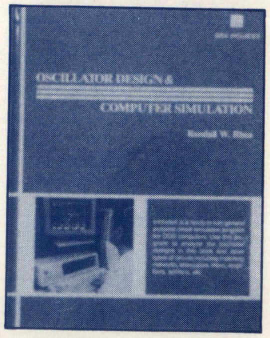

This unified approach to oscillator design is of particular help to engineers new to these circuits, but serves as an excellent reference for even the most experienced designers. The book covers a wide range of resonator types and active devices, and is applicable to both RF and microwave frequency ranges. Included with the text is a copy of Eagleware's Star 2.0 circuit analysis program, which is intended to reduce the tedium in analyzing oscillator circuits as various parameters are modified.

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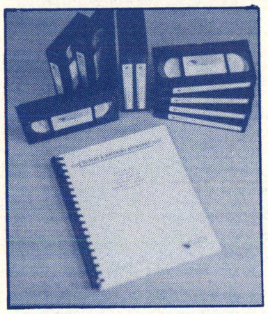

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data is entered, the user may proceed to the circuit entry screen or edit any of the frequency or impedance data. The ESC key or right mouse button may be pressed at anytime to proceed to the circuit entry screen.

Standard Device Selection

If a standard device is to be selected as the load impedance, the ENTER key is pressed at the prompt for 'Number of frequencies to be entered'. The following selections are provided: 2N, MRF, JO and TP. The user selects the prefix for the device he wishes to use, and the prefix is immediately displayed in the numeric entry window, allowing the user to complete the device number followed by pressing ENTER. Next the program prompts the user to select either the device's input impedance or output impedance as the load. If the requested device is included in the database, the impedance information is displayed at their corresponding frequencies. The source impedance is entered manually as described above in the user entry section. Editing can be performed on any frequency or impedance before proceeding to the circuit entry screen.

The program will accept only positive numbers for the real portion of the load or source impedances. The ESC key or right mouse button may be pressed at anytime to proceed to the circuit entry screen.

Circuit Entry Screen

The circuit entry screen is separated into three sections: the component library, data entry keypad and circuit display area.

A component is selected by clicking on it with a mouse (using the left mouse button). Keyboard users should press the number or letter displayed in the upper left hand corner of the box that outlines the component. Immediately after a component is selected, the numeric keypad is activated and the user is prompted to enter the appropriate component values. Inductors are recognized in nanohenries; capacitors in picofarads. If the user prefers to specify the capacitive/inductive reactance, press ENTER at the prompt. The program then expects a reactance to be provided (in ohms) at a specific frequency.

A transmission line is defined by its characteristic impedance (in ohms) and its electrical length (in fractions of a wavelength). The electrical length needs to be referenced to a specific frequency (in MHz). Transmission lines may also be classified in physical terms. Pressing ENTER at the prompt causes the program to request the conductor's width and length. Whenever the first transmission line is selected (whether it is defined in electrical or physical properties), the following basic microstrip information is also requested: relative dielectric constant, dielectric thickness and conductor thickness (3). This information is assumed to be the same for all subsequent transmission lines and is displayed in the upper right hand corner.

Most CAD programs assume that a capacitor has *no* width; i.e., it contacts the circuit at a single point. As frequencies increase, this assumption introduces a significant error in circuit analysis, particularly at low impedances. Since a capacitor is typically mounted on a transmission line, a significant phase shift can occur across its width at higher frequencies. (A 100 mil capacitor can have an electrical width of 0.02 λ at 1 GHz if mounted on Al_2O_3). The error can be reduced by modeling a capacitor as a 'distributed' component. On most CAD programs this involves subdividing the capacitor and transmission line into several smaller sections to comprehend the collective capacitive effects and transmission line transformation. MIMP provides a component, called the distributed capacitor, which first prompts the user for a capacitor value along with any accompanying series lead inductance. Next, it asks for the characteristic impedance of the transmission line, on which the capacitor is mounted. Finally, the transmission line's electrical length (in fractional wavelengths) is entered for that portion of the transmission line on which the capacitor is mounted. MIMP then calculates the combined effect of the two.

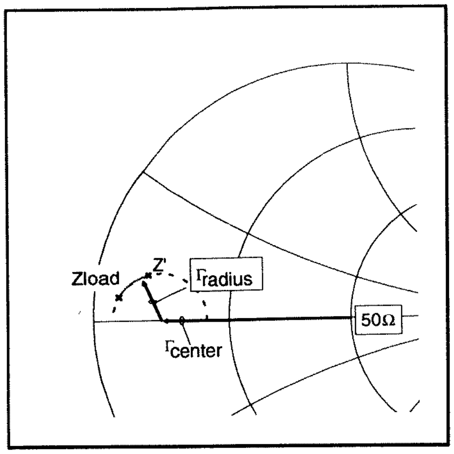

Figure 8. Center and radius of rescaled transmission line impedance.

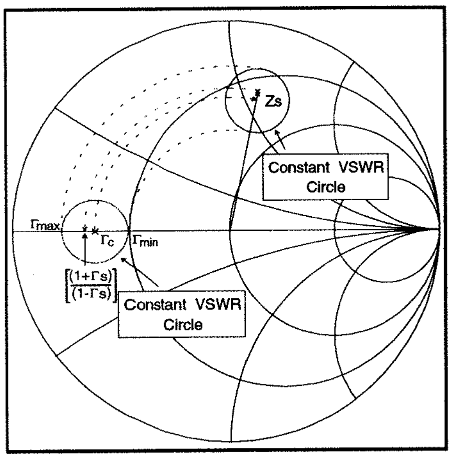

Figure 9. Remapping of constant reflection coefficient circle from one impedance to another (possibly complex) impedance.

After the circuit has been entered, the user may proceed to the Smith Chart display screen by pressing ESC or the right mouse button. If the circuit requires editing, elements may be removed in a LIFO (Last In First Out) sequence and new elements may then be entered if desired.

Smith Chart Display Screen

The Smith Chart display screen is separated into four sections: the Smith Chart display, the menu bar, the nodal impedance display and the scalar input return loss graph.

The Smith Chart graphically displays the impedances transformed by each shunt or series element. These impedances are represented by small X's. Each frequency is depicted by a different color. If there are multiple series or shunt elements, the combined effect of all the elements is lumped together as one element. Upon entry to the Smith Chart display screen, the following default conditions are set:

- The impedances at the lowest frequency are interconnected.
- The conjugate of the source impedance is shown as a yellow X encircled by a –20 dB return loss circle. (The conjugate is the desired transformed impedance.)
- The Smith Chart is initially normalized to 10 ohms .
- The constant Q arcs are set to 0.
- The nodal impedances are listed for the last node.
- The first circuit element is selected for tuning.

The menu bar allows the user to select and fine tune circuit elements and change display parameters. These changes can be made with a mouse or via the keyboard.

The nodal impedance display shows the actual transformed impedance produced by each circuit element. Multiple series or shunt elements are lumped together. (NOTE: These are not really nodal impedances. It is the transformed circuit impedance, starting from the load up to (and including) the selected element.)

The return loss display shows (on a scalar chart) how well the transformed impedance matches the source impedance. The reference return loss is indicated by the yellow line on the scalar display and the yellow circle on the Smith Chart display. If the source impedance is frequency dependent, the circles on the Smith Chart will be relocated for each frequency. To eliminate the circle; change return loss to a very large negative number.

Pressing ESC or the right mouse button returns the user to the impedance entry screen. Mouse users can press left and right buttons simultaneously to return directly to the circuit entry screen.

MIMP's Smith Chart Plotting

The accepted practice, for plotting impedance transformations on Smith charts, requires that each circuit impedance be first normalized to its respective transmission line's characteristic impedance. Once normalized, the impedances can be plotted (and transformed) on a

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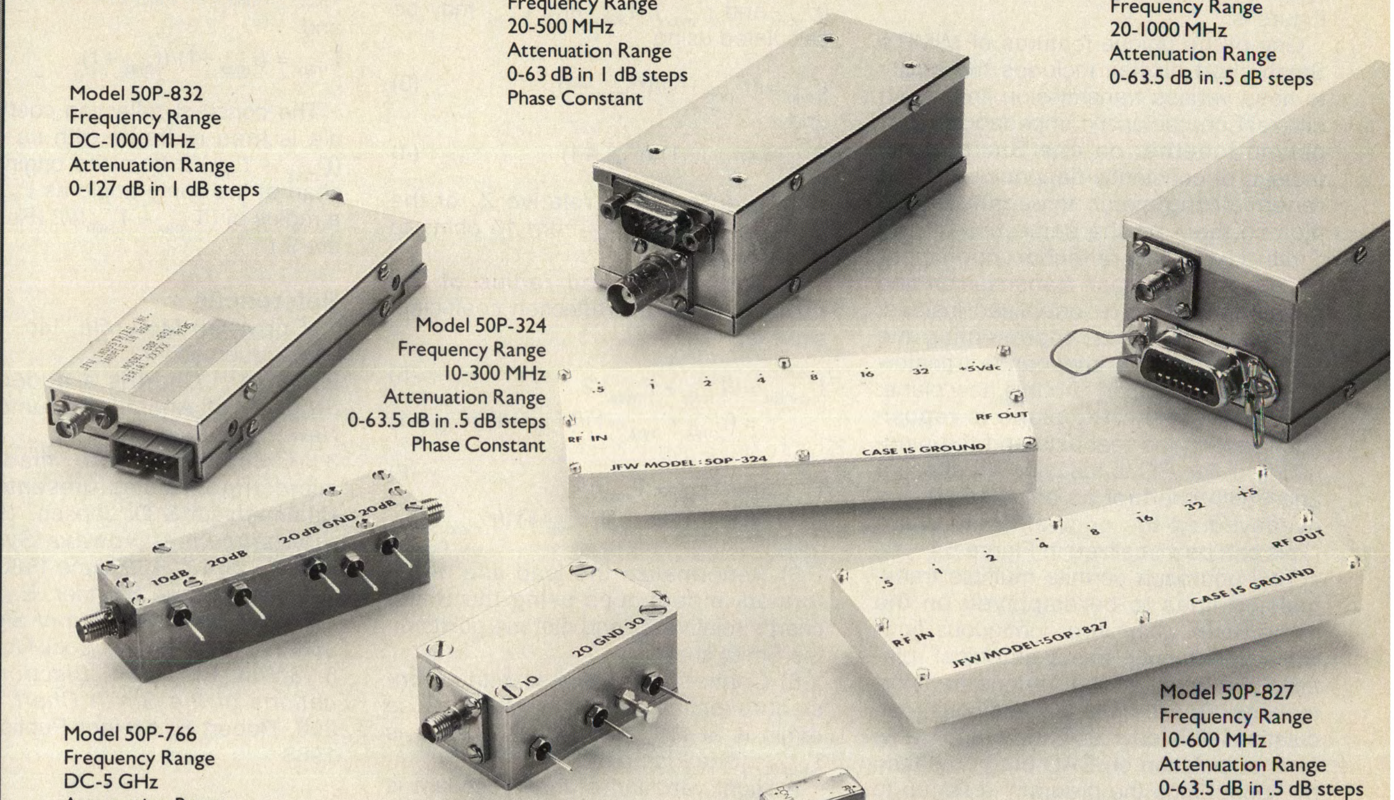

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similarly normalized Smith chart. Each time a new transmission line is encountered in the circuit, the impedances must then be denormalized using the old Z_0 and renormalized with the new Z_0, before additional graphical manipulations can be accomplished. The normalized Smith chart (typically 1 ohm) remains unchanged with each new transmission line transformation. This results in a series of non-contiguous impedance transforms whose visual relationships have little or no value. See Figure 4.

One of the unique features of MIMP's Smith chart display includes the ability to have various transmission lines (with different characteristic impedances) displayed together on one Smith chart. Instead of constantly denormalizing and renormalizing circuit impedances and plotting them on the same normalized Smith chart, the reflection coefficient (RC) plane (for each transmission line transformation) is renormalized instead. The locus of points, representing the transmission line's impedance transformation, is remapped into this new plane. The origin of the RC plane is repositioned along the real axis and the magnitude of the RC is rescaled. In effect, a 2nd Smith chart (of an adjusted size) is overlayed on the original Smith chart. (This can be visualized in Figure 5.)

This approach permits multiple transmission lines to be displayed on the same Smith chart in a contiguous flow, while maintaining exact graphical relationships between the various transformations. Because of the additional calculations, it is obviously more applicable for manipulation on CAD programs than by hand. Once the program is set up to handle these calculations, all impedance transformations maintain their relative positions as the relative characteristic impedance of the Smith chart is changed. (MIMP also allows the user to specify any relative Z_0 for the Smith chart.)

Here is the basic procedure (4,5) used to remap a transmission line transformation (whose Z_0 differs from the Smith chart's relative Z_0):

1) The load and transformed impedances are determined for a particular transmission line. (The load impedance should already be known. The transformed impedance (Z') is calculated based on the transmission line's Z_0 and electrical length.)

2) Next, two hypothetical impedances are calculated (Figure 6). These are the real axis impedances; assuming the reflection coefficient is extended to 0 degrees and 180 degrees. r_{min} and r_{max} are calculated using:

$$r_{min} = (1 - \Gamma)/(1 + \Gamma) \qquad (1)$$

and

$$r_{max} = (1 + \Gamma)/(1 - \Gamma) \qquad (2)$$

3) r_{min} and r_{max} are denormalized using the respective transmission line's Z_0 (ZTx) and then renormalized to the relative Z_0 of the Smith chart (Figure 7). Using the new values of r_{min} and r_{max} (r'_{min} and r'_{max}); Γ_{min} and Γ_{max} may be calculated using:

$$\Gamma_{min} = (r'_{max} - 1)/(r'_{max} + 1) \qquad (3)$$

and

$$\Gamma_{max} = (r'_{min} - 1)/(r'_{min} + 1) \qquad (4)$$

(In Figure 7, the relative Z_0 of the Smith chart changes from 10 ohms to 50 ohms.)

4) The center and radius of the rescaled constant reflection coefficient circle is:

$$\Gamma_{center} = (\Gamma_{min} + \Gamma_{max})/2 = (r'_{min}\, r'_{max} - 1)/(r'_{min} + 1)(r'_{max} + 1) \qquad (5)$$

$$\Gamma_{radius} = (|\Gamma_{max} + \Gamma_{min}|)/2 = (r'_{max} - r'_{min})/(r'_{min} + 1)(r'_{max} + 1) \qquad (6)$$

5) Renormalize the load and transformed impedances using the Smith chart's relative Z_0 and plot the points on the Smith chart.

6) Connect the 2 points with a constant reflection coefficient circle whose origin is at Γ_{center} and whose radius is Γ_{radius}, (See Figure 8).

A slight variation of this procedure is also used for plotting a constant reflection coefficient circle for any impedance on the Smith chart. Many programs plot constant RC circles about the center of the Smith chart (when the network's source impedance is also assumed to be in the center of the Smith chart). However, if the source impedance is different from the Smith chart's Z_0 (it can even be complex); its constant reflection coefficient circle can also be remapped to a different location on the Smith chart. When remapping to a new real value source impedance, the procedure is the same as described above. Plotting the constant reflection coefficient circle, for a complex source impedance, requires some additional steps. The center and radius for the newly mapped constant reflection coefficient circle are determined by first translating Γ_{Zs} to either 0 degrees or 180 degrees. From this, R is calculated using :

$$R + (1 + \Gamma_{Zs})/(1 - \Gamma_{Zs}) \qquad (7)$$

Followed by calculating r_{min}, r_{max}, Γ_{min} and Γ_{max}.

$$r_{min} = (1 - \Gamma_{VSWR})/(1 + \Gamma_{VSWR}) \bullet R \qquad (8)$$

and

$$r_{max} = (1 + \Gamma_{VSWR})/(1 - \Gamma_{VSWR}) \bullet R \qquad (9)$$

$$\Gamma_{min} = (r_{min} - 1)/(r_{min} + 1) \qquad (10)$$

and

$$\Gamma_{max} = (r_{max} - 1)/(r_{max} + 1). \qquad (11)$$

The constant reflection coefficient circle is then redrawn with its center at $(\Gamma_{max} + \Gamma_{min})/2$ from the origin of Smith chart at the same angle as Γ_{Zs} and with a radius of $(|\Gamma_{max} - \Gamma_{min}|)/2$ (Refer to Figure 9.) ***RF***

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1. Copyright Motorola, Inc. 1992, All rights reserved.
2. Smith™ Chart is a trademark and property of Analog Instruments Co., New Providence, NJ.
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Microstrip Antenna Modeling

By Andre Boulouard, Michel Le Rouzic and Jean Paul Castelletto
Centre National d'Etudes Des Télécommunications

Microstrip antennas play an important role in the design of radio communication systems as both feed and array elements. They have found increasing use because they can be fabricated at low cost by lithographic techniques. As the demand for commercial products containing RF antennas continues to grow, the availability of good microstrip antenna models is required. This article describes accurate transmission-line antenna models suited for commercially available CAD packages.

Microstrip antenna design requires good quality models for low and high permittivity substrates. As the need for increased bandwidth leads to thicker substrates, the models used for thin, low dielectric constant substrates must be improved (1). Although analytical models such as the loaded transmission-line resonator (2) and the cavity models (3) involve rather simplified approximations, they provide sufficient information for a first pass design. More rigorous solutions may be obtained by finite difference techniques or moment solutions, although the computational cost is high (4).

The transmission-line model is both accurate and numerically efficient. It can be represented by a transmission line section terminated at both ends by radiation admittances, surface wave loss conductances and coupled current sources as shown in Figure 1. The radiation conductance is expressed as:

$$G_r = \left[(\omega \int_0^{\omega} \frac{\sin(x)}{x} dx + \sin(\omega)/\omega + \cos(\omega) - 2)(1 - s^2/24) + (1/3 + \cos(\omega)/\omega^2 - \sin(\omega)/\omega^3)s^2/12 \right] / (n_0 \pi$$

where:

c_0 = 299,792,500 m/s
n_0 = 376.467 ohms
$k_0 = 2\pi$ frequency/c_0
$\omega = k_0 W_e$
$s = k_0 D_l$
W_e = antenna effective width
D_l = open end equivalent line length

The surface wave loss equivalent conductance is obtained from Reference 5 as:

$L_0 = c_0$/frequency

$$H_s\omega = (H/L_0)\sqrt{(\varepsilon_r - 1)}$$

$$E_s\omega = 1 - 3.4H_s\omega + 1600(H_s\omega^3 - 100H_s\omega^5 6)\varepsilon_r^3$$

where:

H = substrate height
ε_r = substrate dielectric constant

The self susceptance, B_r, is obtained from the open end equivalent transmission line length as:

$$B_r = \tan(k_0 D_l)\sqrt{(\varepsilon_e)} / Z_c$$

where:

ε_e = transmission line effective dielectric constant
Z_c = transmission line characteristic impedance

The real and imaginary parts of the mutual admittance, Y_m, are expressed as:

$x = k_0(L + D_l)$
$F_g = J_0(x) + s^2 J_2(x)/(24 - s^2)$
$N_b = (\pi/2)(Y_0(x) + s^2 Y_2(x)/(24 - s^2))$
$F_b = N_b/(\ln(s/2) - 0.9228 + (s^2/12)/(24 - s^2))$
$G_m = G_r F_g$
$B_m = B_r F_b(1 - \exp(-0.21\omega))$
$Y_m = G_m + jB_m$

where J_0, J_2 and Y_0, Y_2 are Bessel functions of the first and second kind respectively.

In order to improve the accuracy of the resonant frequency compared to the results published in Reference 1, the physical antenna length should be corrected as follows:

$$L_c = L + (-0.12 + 0.043\varepsilon_r)H$$

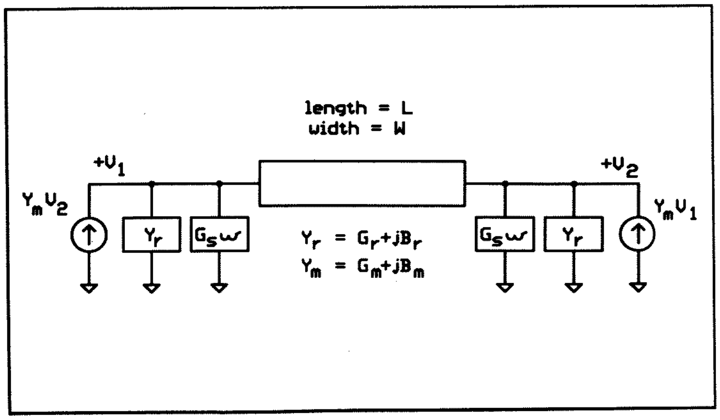


Figure 1. Antenna transmission line model.

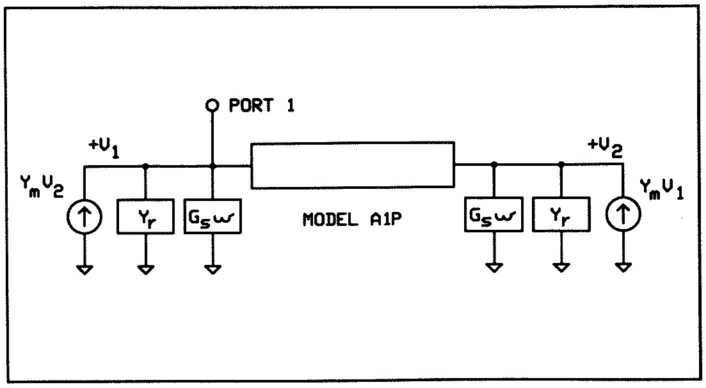


Figure 2. One port antenna fed along a radiating edge.

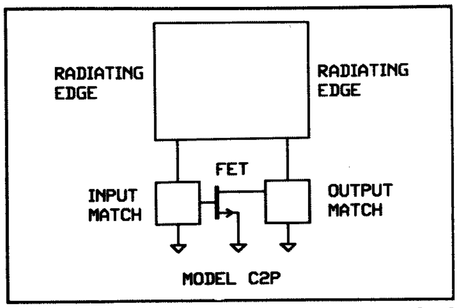


Figure 3. Non-radiating edge fed oscillator-antenna.

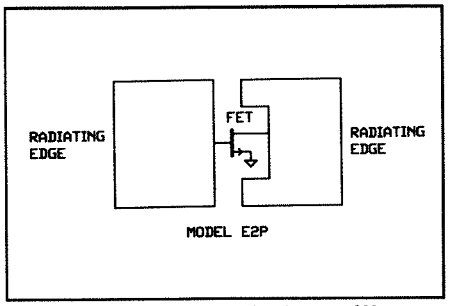


Figure 4. Center fed oscillator-antenna.

Different antenna types have been modeled after this basic theory. They differ by the position(s) of the feeding line(s), relative to the antenna as seen in Table 1. The simplest one port model shown in Figure 2 is obtained when the antenna is fed by a high impedance microstrip transmission line along a radiating edge (model AIP).

As the antenna input impedance is the highest when fed along a radiating edge, a lower impedance is obtained by etching an insert at the center of a radiating edge (model BIP) or by feeding the antenna on a non-radiating edge (model CIP), thus facilitating the matching procedure. Two-port antennas are used to build linear arrays (model A2P) (6) or active radiating antennas using FETs (models A2P, C2P and E2P) as shown in Figures 3 and 4 (7,8).

Using a commercial CAD software (9), fewer than 100 equations were necessary to define each of these models inside a circuit page. A special Microstrip Antenna Components Library file has been built upon these models and can be made available to interested readers for comparison purposes. Contact the authors at the address given below. *RF*

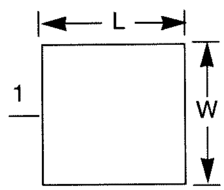
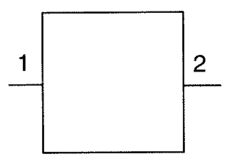
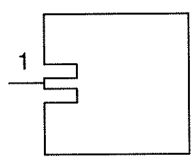
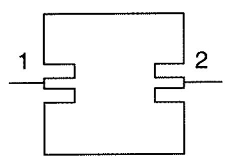
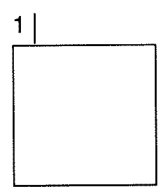
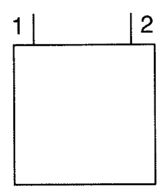
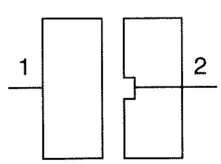
Antenna features	One-port	Two-port
• High input impendance • Fed along a radiating edge	 Model AIP	 Model A2P
• Low input impedance • Fed along a radiating edge (insert)	 Model BIP	 Model B2P
• Low input impedance • Fed along a non-radiating edge	 Model CIP	 Model C2P
• Very low input impedance • Center fed • Active antenna (oscillator)		 Model E2P

Table 1. Different types of microstrip antennas.

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Andre Boulouard is a senior engineer; and Michel Le Rouzic and Jean Paul Castelletto are chief technicians at Centre National d'Etudes des Télécommunications. They are currently involved in MMIC, MHMIC, MIC and subsystem designs for high-speed optical and millimeter-wave communications systems. They can be reached at: CNET/LAB/OCM/MLS, CNET - Route de Tregastel, Lannion, 22301, France. Fax: (33) 96 05 10 08.

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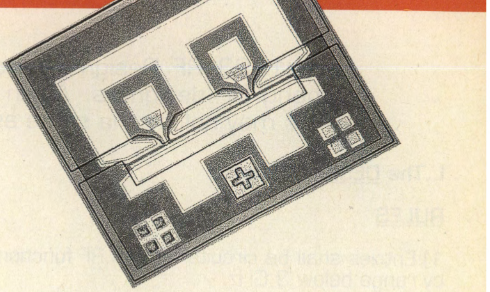

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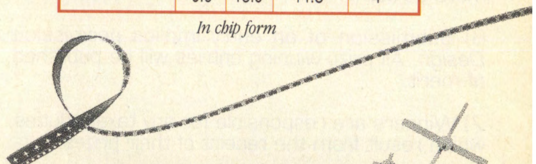

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Noise Figures (NF) in dB
Gain (Ga) in dB
Power Out (P1dB) in dBM
Frequency in GHz

California Eastern Laboratories

CEL Headquarters, 4590 Patrick Henry Drive, Santa Clara, CA 95056-0964; (408) 988-3500 FAX (408) 988-0279 □ Santa Clara, CA (408) 988-7846 □ Los Angeles, CA (310) 645-0985 San Diego, CA (619) 467-6727 □ Bellevue, WA (206) 455-1101 □ Richardson, TX (214) 437-5487 □ Shawnee, KS (913) 962-2161 □ Woodridge, IL (708) 241-3040 □ Cockeysville, MD (410) 667-1310 Peabody, MA (508) 535-2885 □ Hackensack, NJ (201) 487-1155 or 487-1160 □ Palm Bay, FL (407) 727-8045 □ Snellville, GA (404) 978-4443 □ Nepean, Ontario, Canada (613) 726-0626

1993 RF DESIGN AWARDS CONTEST

CALL FOR ENTRIES

1993 RF Design Awards Contest
Official Rules

The 1993 RF Design Awards Contest provides recognition for innovation and engineering excellence among RF designers. Again this year, there are two separate entry categories. Please note that the rules may not be the same as previous years' contests.

I. The DESIGN Contest

RULES

1) Entries shall be circuits with an RF function, operating in the frequency range below 3 GHz.

2) Circuits entered shall have a complexity equivalent to that of a circuit using 8-10 discrete active devices, or 6-8 integrated circuits. This rule is for ease of judging, to have all entries be of a similar scope. The entry can be a portion of a larger system.

3) Entries may represent design or test methods. Design method entries should include an example circuit. Test method entries should include a description of the device or system under test.

4) The entries shall be the original work of the entrant, not previously published. If developed as part of the entrant's employment, entries must have the employer's approval for submission.

5) Only one entry per person is permitted. An entry may have two or more co-authors.

6) Submission of an entry implies permission for publication by *RF Design*. All prize-winning entries will be published, plus additional entries of merit.

7) Winners are responsible for any taxes, duties, or other assessments which result from the receipt of their prizes.

8) Entries must be postmarked by March 19, 1993 and received no later than March 26, 1993.

9) All entries will remain confidential until the publication of the July 1993 issue of *RF Design*.

JUDGING CRITERIA

Originality — Each design will be evaluated for similarity to work by others, and other judgements of its unique contributions.

Engineering — Entries should clearly identify how the entry was created in response to a need.

Documentation — A complete description of the circuit or technique is required, including sufficient theoretical background, description of circuit operation, and performance data.

II. The PC SOFTWARE Contest

RULES

1) Each entry shall be a computer program which assists in the design, test, or control of RF circuits or systems.

2) Programs must operate on computers compatible with MS-DOS/PC-DOS or Apple Macintosh operating systems. Any special hardware requirements should be noted (memory, graphics, etc.).

3) Programs should be provided in a form that can be run without special support software; programs should be provided in compiled, directly executable form. Programs using the BASICA or GWBASIC interpreters are acceptable. Programs that require spreadsheet or mathematics software packages cannot be accepted.

4) Programs entered must be submitted on disk. Supporting documentation on theory of operation, references, and operating instructions must be supplied in printed form. Source code must be supplied, either on disk or in printed form.

5) The entries shall be the original work of the entrant, not previously published or distributed (including public BBS or shareware). If developed as part of the entrant's employment, entries must have the employer's approval for submission.

6) Only one entry per person is permitted. An entry may have two or more co-authors.

7) Submission of an entry implies permission for publication by RF Design and distribution by the RF Design Software Service. All prize-winning entries will be published, plus additional entries of merit. Some restrictions on publication and distribution of source code may be acceptable.

8) Winners are responsible for any taxes, duties, or other assessments which result from the receipt of their prizes.

9) Entries must be postmarked by March 19, 1993 and received no later than March 26, 1993.

10) All entries will remain confidential until the publication of the July 1993 issue of *RF Design*.

JUDGING CRITERIA

Technical Merit — Computer programs will be compared to accepted standards for accuracy, and will be judged for their achievement in translating RF theory into software tools.

Usefulness — The value of the software to the RF engineering community will be evaluated. Both the nature of the computations performed and ease of operation of the program will be considered.

DEADLINE FOR ENTRIES:
POSTMARKED BY MARCH 19, 1993 — RECEIVED BY MARCH 26, 1993

Send entries to:

RF Design Awards Contest
***RF Design* magazine**
6300 S. Syracuse Way, Suite 650
Englewood, CO 80111

1993 RF DESIGN AWARDS CONTEST

GRAND PRIZE ANNOUNCEMENT

Continuing our tradition of offering prizes valued in the tens of thousands of dollars, here are the Grand Prizes to be awarded to the winners of the Circuit Design and PC Software categories of the 1993 RF Design Awards Contest —

THE DESIGN GRAND PRIZE

The NOISE COM model UFX-BER Noise Generator for Bit Error Rate Testing

This instrument offers advanced testing using calibrated injection of White Gaussian noise into the IF or RF of a tranmission system. It displays the actual E_b/N_o or (C+N)/N for testing of satellite communications or digital radio systems. Center frequencies are available for all common IFs, to be selected by the winner. A powerful test solution for today's communications!

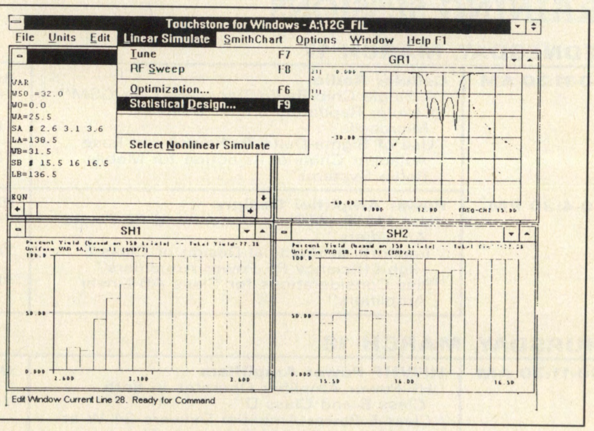
provided by NOISE/COM

THE PC SOFTWARE GRAND PRIZE

EEsof's new Windows-based software — Libra®, Touchstone® and LineCalc®

Version 3.5 software offers advanced circuit simulation in the PC environment under the Microsoft Windows operating system. Linear analysis is provided by Touchstone for classic amplifier, filter and other linear circuits. Libra offers nonlinear analysis for such circuits as power amplifiers, oscillators and mixers. LineCalc adds the capability for physical design of transmission line circuits to the package. The highly visual Windows operating system allows flexibility in graphical and physical representations.

provided by EEsof®



Watch for more prizes in the next issue of *RF Design*!

New Oscillators Advance the Art of Low Noise Performance

By Charles Wenzel
Wenzel Associates

Over the last two decades, tremendous improvements in telecommunications, radar systems, and instrumentation have driven phase noise specifications from near obscurity to prime importance. Quartz crystal reference oscillators are especially crucial since oscillator noise often limits the channel capacity of communication systems, degrades the resolution of radar and timing instruments, gives synthesizers wide skirts and restricts the resolution of spectrum analyzers.

Wenzel Associates' new reference standard offers extremely low phase noise performance.

Wenzel Associates has developed new low noise oscillators with significant improvements in phase noise. In addition to a line of component VHF oscillators, a high stability frequency standard is available that employs phase locking to achieve excellent noise and long term aging. The phase noise performance of three, 100 MHz oscillators is shown in Figure 1.

Noise floor measurements near –180 dBc/Hz push the limits of ordinary phase noise measurement systems, so a phase noise calibration was requested of Fred Walls at the National Institute of Standards and Technology (NIST). His system uses a cross correlation technique to greatly reduce the system noise floor for measuring the higher Fourier frequencies. [*This measurement technique will be described in the March 1993 issue of* RF Design — *Editor*] Fourier frequencies below 20 Hz were made using a multiplied 5 MHz reference. Each oscillator's noise data was fit to a mathematical model included in Figure 1. Overall measurement accuracy was stated to be ±1 dB between 1 Hz and 50 kHz.

Previous attempts to achieve low noise floors have employed drive levels reaching 50 mW in a brute force signal to noise battle with a corresponding degradation in close-in noise and aging. The unusual combination of a –180 dBc/Hz noise floor with a –77 dBc/Hz flicker frequency level is achieved by using low crystal dissipation (1 mw) in specially designed low noise circuitry. Careful attention was paid to voltage regulator noise and to oven current noise, two common sources of degraded performance. The oscillators' output level is set to 1 VRMS so the sideband noise at the floor is about 1.4 nanovolts/$\sqrt{\text{Hz}}$ or about as much noise as exhibited by a 120 ohm resistor. Taking full advantage of such low noise levels poses a significant challenge to the systems engineer.

The model 500-03475 low phase noise frequency standard combines this new VHF oscillator with a similarly improved 5 MHz reference in a phase locked configuration. Figure 2 shows the phase noise for the standard's various output frequencies. The 5 MHz oscillator noise drops below –150 dBc/Hz 10 Hz away from the carrier and continues down to a floor of –180 dBc/Hz. The bandwidth of the internal phase locked loop is set to about 300 Hz where the 100 MHz oscillator begins to out perform the multiplied 5 MHz reference. The model 500-03475 employs a shielded

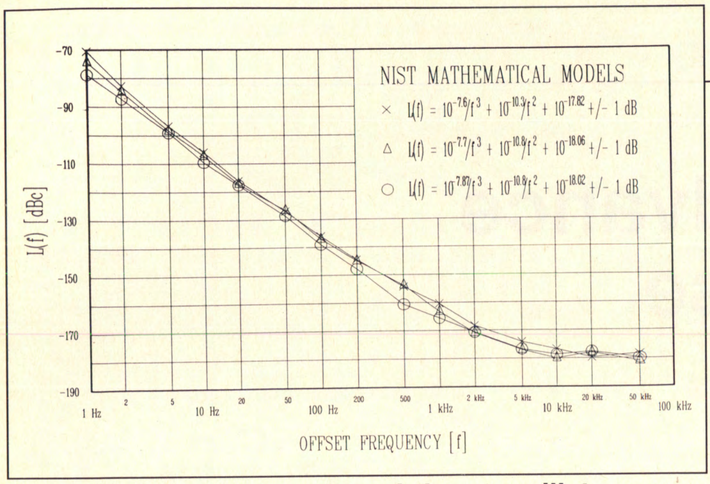


Figure 1. The phase noise of three oscillators measured by NIST.

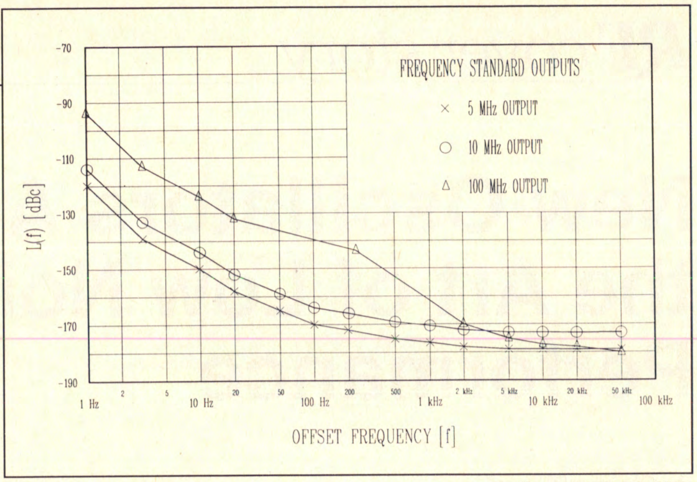


Figure 2. Phase noise performance of the 500-03475 Frequency Standard.

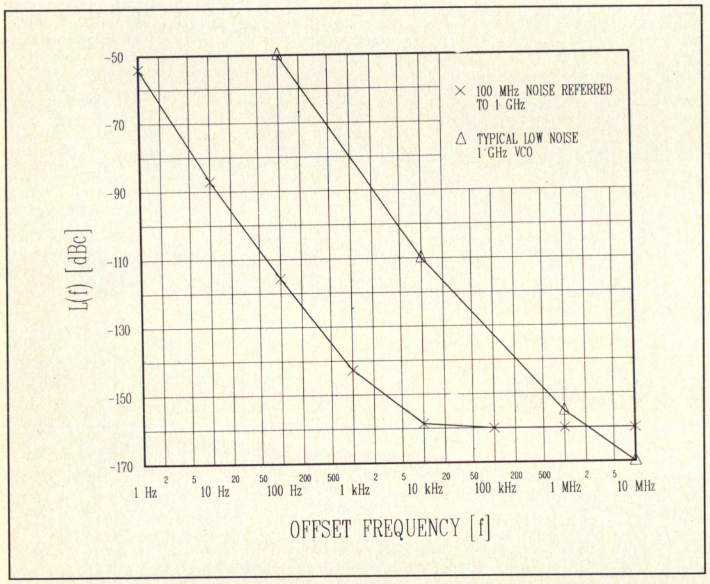


Figure 3. Comparison of a low noise VCO at 1 GHz to the multiplied 100 MHz oscillator.

line transformer to reduce line related noise in addition to a 10 hour rechargeable battery. Other features include digital monitoring of output levels and battery condition and digital control of tuning and aging rate.

High Performance Applications

Crystal oscillators with significantly improved phase noise can enhance a variety of systems. Microwave systems are particularly susceptible to reference oscillator phase noise because the process of frequency multiplication increases the power in the sidebands by the square of the multiplication factor. Phase locked loops or filters are often used to clean up the multiplied reference. In a typical system the filtering bandwidth is set near the point where the microwave oscillator's flicker noise drops below the reference oscillator's noise floor. With a reference noise floor of –180 dBc at 100 MHz an ideal x10 frequency multiplier would exhibit a noise floor of –160 dBc at 1 GHz, which rivals the noise floor of many microwave sources. Even after multiplication to 10 GHz the resulting –140 dBc floor would remain below the flicker noise of most VCOs out to several MHz. In some applications the VCO might be replaced by a simple filter to remove the subharmonics without any phase noise penalty. Figure 3 compares the new oscillator's noise, after multiplication, to an "ultra low noise" cavity oscillator operating at 1 GHz.

The combination of low flicker and low noise floor improves the bit error rate of a digital communication system for a given modulation scheme since the BER increases with the area under the phase noise curve. This small integrated noise or phase jitter similarly improves the resolution and probability of detection of radars and enhances the accuracy of distance measuring devices.

Modern spectrum analyzers using low noise synthesized local oscillators have improved sufficiently to allow for the direct observation of sideband noise of fairly good sources. A higher performance reference would lower the local oscillator noise even further, making smaller measurement bandwidths feasible for direct measurement of all but the best sources.

The electronics in these new oscillators is beginning to approach theoretical limits so future improvements will concentrate on the crystal resonators with a goal of lowering the flicker frequency level. Of course, improvements in temperature stability, aging, and other critical parameters will receive continued attention.

For more information on these low noise oscillators and frequency standard, contact the author at the address below, or circle Info/Card #160. *RF*

References

1. Fred L. Walls, "Calibration of 100 MHz crystal oscillators Model 500-02268B Serial #s 0578-9041 and 2110-9220 for Wenzel Associates," report by the National Measurement Laboratory, National Institute of Standards and Technology, Boulder, Colorado, 25 September 1992.
2. Addendum to above report, 6 November 1992.
3. Warren F. Walls, "Cross-Correlation Phase Noise Measurements," 46th Annual Symposium on Frequency Control, Hershey, PA, 1992.

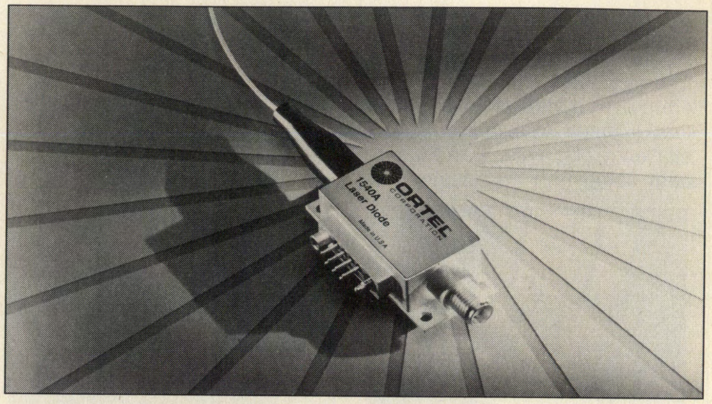
About the Author

Charles Wenzel is president and founder of Wenzel Associates, a manufacturer of high performance frequency standards and oscillator modules since 1978. He is a graduate of the University of Texas at Austin, and a past Grand Prize winner in the RF Design Award Contest. He can be reached at Wenzel Associates, 14050 Summit Drive, Austin, TX 78728; telephone (512) 244-7741.

Communications Link Laser

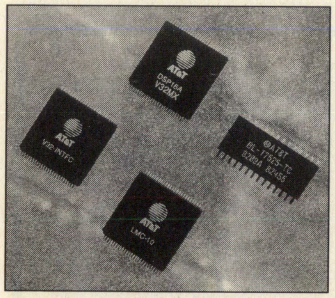
Ortel has introduced its Model 1541A, a 10 GHz microwave distributed feedback (DFB) laser, designed for transmission distances longer than 5 to 10 km. The single-mode spectrum of DFB lasers offer dramatic improvements over multi-mode lasers in long fiber-path transmission performance. The DFB laser incorporates an optical isolator mounted with the laser diode on the thermo-electric cooler to stabilize the temperature for a more stable optical output over wide operating temperature ranges. Benefits of the new laser and isolator include: higher spur free dynamic range of 115 dB/Hz$^{2/3}$, transmission distances of 30 to 40 km and reduced link loss of 25 to 30 dB. RF parameters of the model 1541A include: frequency range of 0.1 to 10 GHz, input impedance of 50 ohms and EIN at 2 GHz of less than −125 dBm/Hz. With a guaranteed frequency range of 10 GHz, the laser can be used with C-band satellite up-links, military satellite communications and airborne radar bands.

Ortel Corporation
INFO/CARD #250



Landline/Wireless Modem Chipset

AT&T Microelectronics plans to incorporate the MOBITEX two-way wireless data-transmission architecture in its V.32bis and V.17 Complete Modem Chipset, making these modem chips the first to support both high-speed landline and wireless communi-

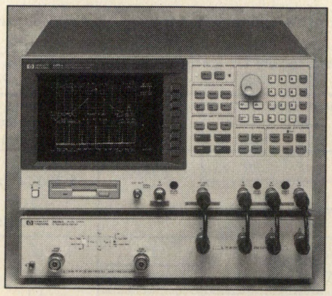


cations. AT&T's DSP-based modem technology uses the AT&T DSP16A digital-signal processor for transmission and signal recovery of the GMSK-encoded baseband signal used in RAM Mobile Data's Mobitex networks. The DSP16A works with the T7572 codec, the V32INTFC interface device and the LMC10 modem controller. The V32MX Complete Modem Chipset achieves a 8000 bit/s data rate over the air and interfaces directly to third party radio transceivers. The chip set can be sampled in early 1993 and will be in full production by mid 1993. In quantities of 10,000, the chips will be priced at $109, packaged in plastic quad flat packs. In small-outline QFPs suitable for use in the PCMCIA form factor, the chips will be priced at $129.

AT&T Microelectronics
INFO/CARD #249

Spectrum/Vector Network Analyzer

Hewlett-Packard has introduced the HP 4396A, a 1.8 GHz combination spectrum and vector network analyzer with an optional built-in instrument controller (Option 1C2). The HP 4396A combines dual measurement and control functions in one instrument, allowing engineers to test a wide variety of RF devices or circuits on one test station. A stepped FFT provides fast spectrum analysis, while excellent noise performance and narrow filter-shape factors improve close-in and small-signal spectrum analysis results. An optional time-gated spectrum analysis capability allows measurement of repetitive signals. Full S-parameter measurements with full error correction can be made within the

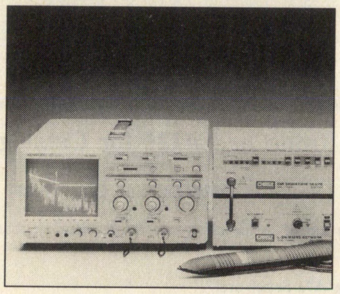


vector network mode. Fast sweep times allow operators to see real-time gain and phase-trace updates. Option 1C2 consists of an instrument-controller and HP Instrument BASIC (IBASIC). Base price for the HP 4396A network/spectrum analyzer is $33,900 with delivery at four weeks ARO.

Hewlett-Packard Company
INFO/CARD #248

EMC Pre-qualification Equipment

The Chase ESS7500 displays the spectral signature of an item under test and can compare this signature to either limit lines or against previously stored signa-

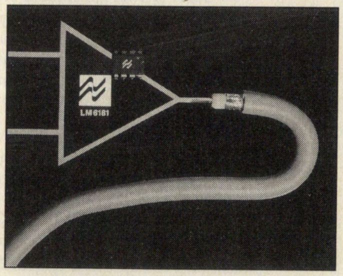


tures. The ESS7500 works with any oscilloscope having a bandwidth of 30 MHz or more. The flicker-free display can show either the complete frequency range (from 3 kHz to 30 MHz) or can inspect smaller ranges individually for a more detailed display. Up to 15 pre-programmed limit-lines (for international or in-house standards) can be stored in ROM. Two non-volatile signature memories are available. Before/after comparisons are possible with this feature, as are comparisons between "golden samples" and production samples. A magnetic near- field probe (MFP9150) and an electric near-field probe (EFP9152) are available for radiated emissions testing, and a 6 Amp mains network (CLN2060) is available for measurement of conducted emissions. The ESS7500 is priced at $4500, with delivery in two weeks.

Ibex Group, Inc.
INFO/CARD #247

Current Feedback Amplifier

National Semiconductor Corporation has introduced a current feedback amplifier, the LM6181, offering a combination of bandwidth (100 MHz), slew-rate (2000 V/us) and output current (100 mA) which makes it ideal for video and other high-speed amplifier applications. Because the LM6181 is a current feedback amplifier, bandwidth is relatively independent of closed loop gain. The amplifier has differential gain of 0.05 percent and a differential phase of 0.04 degrees, key parameters in video applications. The high power output stage enables



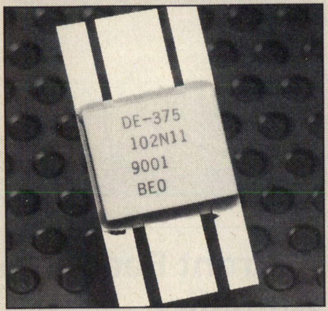
the LM6181 to directly drive a 10 V signal into 50 ohm or 75 ohm back terminated coaxial cable systems over the full industrial temperature range. A free LM6181 evaluation kit, including a fully assembled evaluation board and SPICE micromodel, is available. The LM6181 is fully characterized and specified at both ±5 V and ±15 V. Available in both 8-pin plastic DIP and 16-pin SO packages, The LM6181's suggested 1000-piece resale pricing is $2.45 for both package styles.

National Semiconductor
INFO/CARD #246

Product Spotlight: Transistors

Power MOSFETs

The DE-Series from Directed Energy makes it feasible to generate over a kilowatt of power in the HF band, with a conversion efficiency of greater than 85 percent using only two devices in

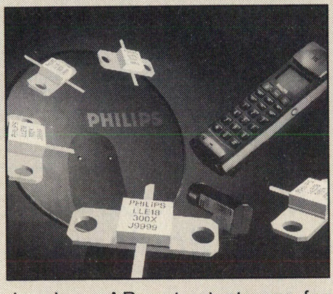


push pull configuration. Designs based on these devices, at less than $0.50/watt, are a fraction of the cost of conventional bipolar and low voltage MOSFET approaches.

Directed Energy, Inc.
INFO/CARD #245

Microwave Power Transistors

New NPN silicon microwave power transistors from Philips Semiconductors reduce the number of components required



in class AB output stages for transmitters operating in the 1.5 to 2 GHz band. The LLE18100X and LLE18300X deliver respectively 10 and 30 W and are optimized for operation in the 1.7 to 2.0 GHz range. LLE16120X and LLE16350X deliver 12 and 35 W, respectively, and are optimized for the 1.5 to 1.7 GHz band.
Philips Semiconductors
INFO/CARD #244

SMT Small Signal Transistors

Motorola has introduced its lowest noise small signal amplifier transistors in a miniature surface-mount package. The MRF947 and MRF957 small signal transistors are state-of-the-art low noise NPN bipolar devices with the same die as Motorola's MRF941 and MRF951 transistors, mounted in the SC-70 surface mount package. In quantities less than 1000, the MRF947 and MRF957 are priced at $0.43 and $0.49, respectively.
Motorola Semiconductor
INFO/CARD #243

SMT Military Transistors

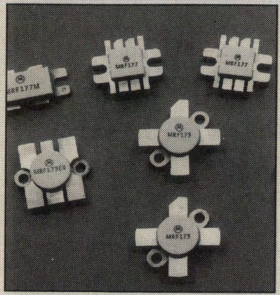
Central Semiconductor offer four device types in the new CERSOT- 23 case: CHT918, a NPN RF oscillator; CHT2222A, a NPN amplifier/switch; CHT-2369A, a NPN saturated/switch; CHT2907A, a PNP amplifier/switch. The CERSOT-23 case is

a hermetically sealed, leadless chip carrier which mounts on SOT-23 pads. JANTXV construction includes pre-cap visual inspection.
Central Semiconductor Corp.
INFO/CARD #242

INFO/CARD 125

Broadband, N-Channel MOS-FETs

Two transistors from Motorola feature high gain



and high power capabilities, along with broadband operation. The MRF173 is designed for broadband commercial and military applications in the 2 to 200 MHz range. At 150 MHz and 28 V, the MRF173 has 11 dB gain and an output power of 80 W. The MRF177 is designed for broadband applications in the 2 to 400 MHz range. At 400 MHz and 28 V, the MRF177 has 12 dB gain and an output power of 100 W.
Motorola Semiconductor
INFO/CARD #241

HJ FET

The NE32684A from CEL is a pseudomorphic HJ FET designed for low noise figure requirements in the 0.5 to 30.0 GHz range. Noise figure is 0.5 dB typ. with gain of 11.5 dB typ. at 12 GHz. Gate length is less than 0.2 microns. The NE-32684A is in an epoxy sealed metal/ceramic package and is available on tape and reel.
California Eastern Laboratories, Inc.
INFO/CARD #240

AMPLIFIERS

Gain Adjustable Amplifier

MITEQ introduces a new series of ultra-wideband, low-noise amplifiers with variable gain option. Designated as AVG6-00102100-8, the amplifier covers the 100 MHz to 21 GHz frequency range with 24 dB gain. Gain can be continuously adjusted for 0-15 dB attenuation. The amplifier input/output ports are matched to 10 dB min. return loss.
MITEQ
INFO/CARD #239

Low Distortion Amplifiers

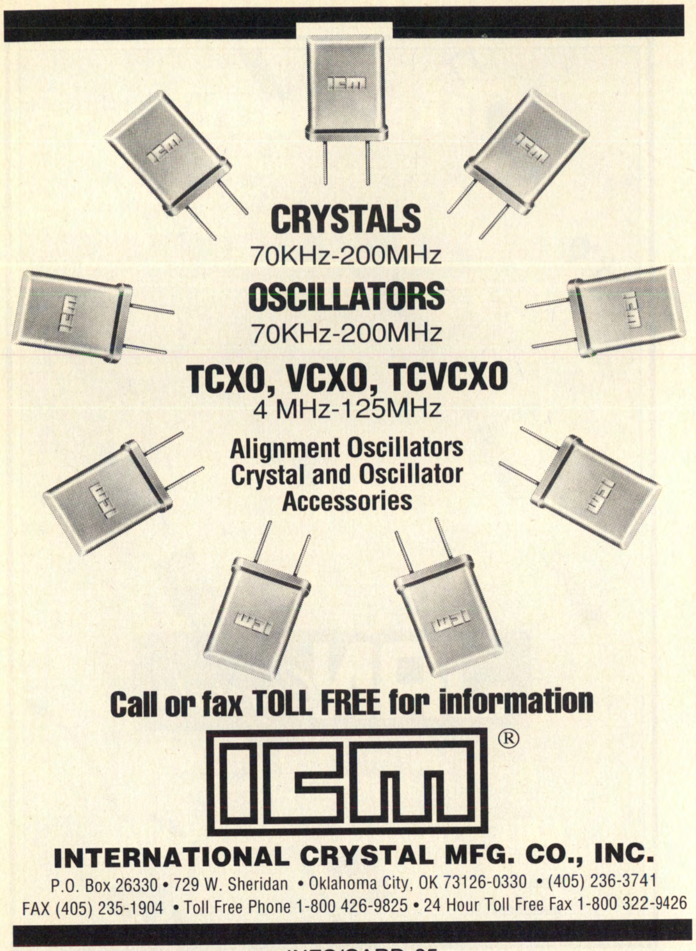
Phoenix Microwave announces a new class of low distortion power hybrid amplifiers for the communications industry. Model PA996 offers linear output power of 20 dBm over the 50 to 500 MHz frequency range with typical third and second order intercepts of 35 and 64 dB, respectively. The unit is offered in a hermetic dual in-line or flatpack package configuration. Availability is stock to 45 days.
Phoenix Microwave Corporation
INFO/CARD #238

100 W, UHF Amplifier

Model 400-220-100-35A UHF power amplifier has an output power level of 100 W CW from 220 to 440 MHz or can be customized to deliver that power over the 400 to 500 MHz band. The amplifier has over 45 percent efficiency with 35 dB of gain and operates from a 28 V supply. This model is available as a module (measuring 4.84 x 2 x 1 inches) or as a rack-mountable system with power and cooling.
LCF Enterprises
INFO/CARD #237

Integrated Power Amplifiers

Modularized, integrated designs are employed in amplifiers producing RF output powers to 10 kW for select frequency bands between 2 and 135 MHz. Efficient RF designs, coupled with switching regulator and power factor correction pre-regulation modules, result in minimum AC line current with low distortion. Typical ratio of RF power output to AC line VA input is 0.50.
Brounley Engineering
INFO/CARD #236



INFO/CARD 35

SEMI-CONDUCTORS

12-bit DAC for DDS

Burr Brown's DAC600 is a 12-bit, monolithic D/A converter featuring the best harmonic distortion and DC linearity in its speed range. The DAC600 has a 256 MHz clock rate, 5 ns 12-bit settling time, 900 mW power dissipation and 74 dB SFDR (f_{clk}=100 MHz, f_0=10 MHz). It is ECL compatible and priced from $79 each in quantities in the 100's. Delivery is from stock.
Burr-Brown Corp.
INFO/CARD #235

Feedback-less Differential Amplifier

Maxim announces the new MAX435/MAX436 wideband (275 MHz) transconductance amplifiers with a unique architecture that provides extremely high common mode rejection (53 dB @ 10 MHz) and accurate gain (±2.5 percent) without feedback. The commercial, plastic DIP and SO versions of the MAX435 and MAX436 are priced at $2.75 in quantities of 1000.
Maxim Integrated Products
INFO/CARD #234

Low Cost GaAs VCO

The ADC20010, by Anadigics is a GaAs VCO chip costing only $7 in quantities of 100. The VCO has output power of 5 dBm and phase noise of –70 dBc/Hz at 10 kHz offset. Tuning range is 1430 to 2600 MHz with a control voltage from 1 to 20 V. The chip comes in a surface mount package with power dissipation typically less than 65 mW. Small quantities are available from stock; larger quantities can be shipped in six to eight weeks.
Anadigics, Inc.
INFO/CARD #233

High Speed RSSI IF System

The NE/SA624/625/627 FM IF system ICs from Philips Semiconductors feature received signal strength indicators (RSSIs) which respond up to eight times faster than those in previously available devices. The 624 incorporates two limiting IF amplifiers, a quadrature detector, muting circuit, logarithmic RSSI and a voltage regulator. The 625 adds a mixer oscillator, and the 627 adds a mixer oscillator and an additional limiter amplifier.
Philips Semiconductor
INFO/CARD #232

50 MHz, 80 dB RSSI

Designed as a demodulating logarithmic amplifier for received signal strength indicator (RSSI) functions in mobile phones and receivers, Analog Devices' AD606 is also effective as a high performance limiting amplifier in FM demodulators and wireless LAN equipment. As a log amp, overall dynamic range is –75 to +5 dBm. As a limiter, it provides a hard-limited signal output as a differential current of ±1.2 mA from open-collector outputs.
Analog Devices, Inc.
INFO/CARD #231

LNA MMIC

TriQuint has announced the TQ9121N, a surface mount,low noise amplifier (LNA) for GPS and other narrow band applications in the 1.2 to 1.6 GHz. The TQ9121N supplies typically 16 dB gain at the GPS frequency band. The device has a 1.25 dB noise figure and third order intercept point, with respect to the output, of +11 dBm. It operates from a +5 V supply. The TQ9121N is $5.40 in quantities of 500 to 999.
TriQuint Semiconductor
INFO/CARD #230

CABLES & CONNECTORS

Phase-Matched Cables

Available on a quick delivery basis from MAST Microwave is a full line of custom phase-matched microwave cable assembly sets. Both flexible and semi-rigid cables are offered in a wide variety of cable and connector combinations. The MAST line covers frequencies from DC to 26.5 GHz. Typically, the cable assembly sets hold absolute phase

deviation within ±5 degrees.
MAST Microwave
INFO/CARD #229

SIGNAL SOURCES

Miniature Phase Locked Sources

EM Research Engineering has introduced a line of miniature phase locked signal sources covering 100 to 2500 MHz. S packages come as small as 1.5 x 1.5 x 0.75 inches and are PC mountable. The A package is a standalone 2 x 2 x 0.75 inch aluminum housing with standard SMA female connectors. These sources can be internally or externally referenced.
EM Research Engineering, Inc.
INFO/CARD #228

TCXO

Piezo Technology has developed a new temperature compensated crystal oscillator. Model XO3010C offers a standard ±0.3 ppm stability over the temperature range of 0 to +70 degrees C. The standard frequency is 10.0 MHz, with optional frequencies between 8.0 and 15.0 MHz. Output is 7 ±2 dBm sinewave into 50 ohms with a supply of +12 V.
Piezo Technology, Inc.
INFO/CARD #227

Low Profile OCXO

Model TK802, hybrid ovenized oscillator, operates over 50 to 100 MHz with temperature stabilities of ±0.05 ppm over –20 to +70 degrees C. With warm-ups less than five minutes, the OCXO achieves 5×10^{-8} stability with +28 V at 250 mA heater current. Phase noise is –135 dBc/Hz at 1 kHz and –155 dBc/Hz at 10 kHz. The package measures 2 x 2 x 0.75 inches. Price is $790 in quantities less than 10.
ST Microsonics Corp.
INFO/CARD #226

Half-Size ACMOS

M-tron's MAH series oscillators offer a state-of-the-art, TTL and CMOS compatible, high frequency clock oscillator. The MAH oscillator is available in the 70-135 MHz range. Its leads are on 0.300 x 0.300 inch centers (8-pin DIP compatible), and the overall dimensions are 0.520 x 0.520 inches maximum. Surface mount and extended temperature and stability options are available.
M-tron Industries, Inc.
INFO/CARD #225

TEST EQUIPMENT

HF Channel Simulator

Signatron's model S-251A Real-Time HF Channel Simulator permits comprehensive modem testing over a very wide range of HF channel conditions. The simulator comprises a card to be installed in an IBM compatible computer and software. The simulator operates on a 4 kHz bandwidth channel and allows the user to independently vary the delay, Doppler shift, Doppler spread and amplitude for up to six multipath components.
Signatron Corp.
INFO/CARD #224

Analog Network Analyzer

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AP Instruments
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Switching Speed: Definition and Measurement

By Ra'anan Sover
Microkim Ltd.

RF switches have been around for quite some time. Whether they are electro-mechanical or solid state, switching speed is still one of the most important characteristics. A general comparison of different types of switches and their performance has already been conducted, see References 1 and 2. Even so, there are several different definitions and methods of measurement of the switching speed (3,4,5,6,7,8). In this paper, we will take a closer look at the definition of switching speed and several methods of measurement.

The exact definition of switching speed must be related to the specific switch being tested and the system requirements. Since most "off-the-shelf" switch vendors do not know where their switch will end up, a more common definition is required. This definition is what we call the classic switching speed definition (Figure 1). It can be seen that the switching speed is made up of two basic time intervals: the driving circuit delay and the transition speed of the RF components. In some cases, the transition time (rise/fall time) is mistaken for the switching speed. The switching speed is the total time interval between the crossing of the 50 percent level of the control signal and the 10 percent or 90 percent level of the RF signal. The 100 percent level corresponds to the steady state insertion loss and the zero percent corresponds to the ideal isolation state. A true zero percent RF (infinite dB isolation) is not physically achievable. We must then relate the 10 percent and 90 percent RF to the insertion loss level. Therefore, the 90 percent RF level can be seen as the insertion loss level less 0.45 dB, and the 10 percent RF would be the insertion loss level less 10 dB. Since most of the RF switches have an isolation much greater than 10 dB, the classic definition takes into consideration only a small part of the isolation. This may be a good enough definition for general switching speed measurement, but it leaves a lot of uncertainty about what happens at higher isolation levels.

For example, let us look at two solid state PIN diode switches (Figure 2). Both switches were designed to give at least 80 dB isolation within the 2 to 6 GHz region. One was designed using only series PIN diodes while the other uses both series and shunt PIN diodes. Using a test set-up, which will be described later, and a suitable driver, the switching behavior can be seen in Figure 3. The switch with the series/shunt diodes reaches full isolation fairly quickly. The switch with all series diodes initially starts to "turn off" quickly but then decays slowly to the full isolation state after a much longer period of time. This behavior is due to the fact that the diodes are in series and the stored charge in the I-layer of the diodes must recombine through the other diodes while the stored charge of the series/shunt diodes can recombine directly through the driver. Using the

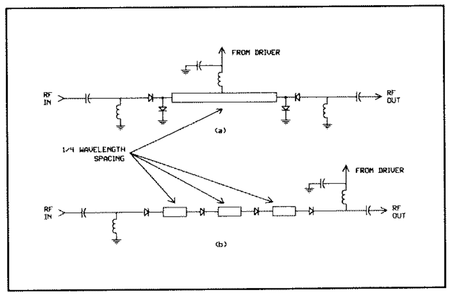

Figure 2. Tested switches' circuit diagrams, a) Series/shunt PIN diode switch, b) All series PIN diode switch.

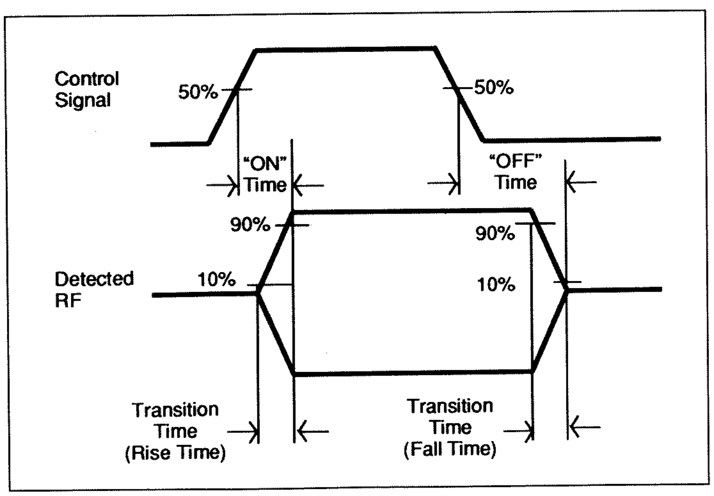

Figure 1. The classic switching speed definition.

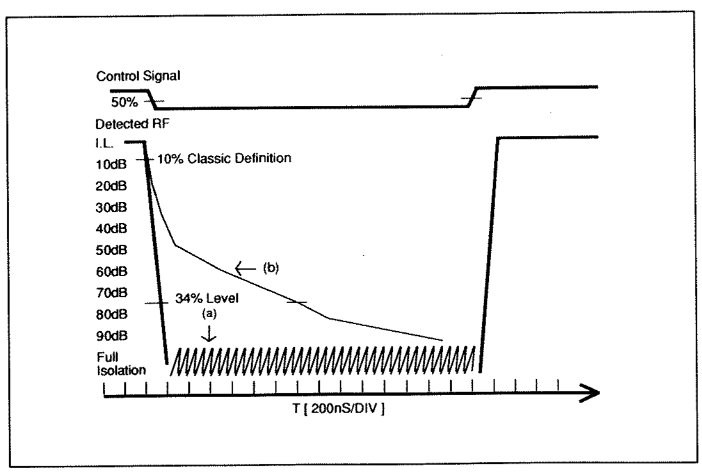

Figure 3. Switching behavior of the two switches, a) Series/shunt switch, b) Series only switch.

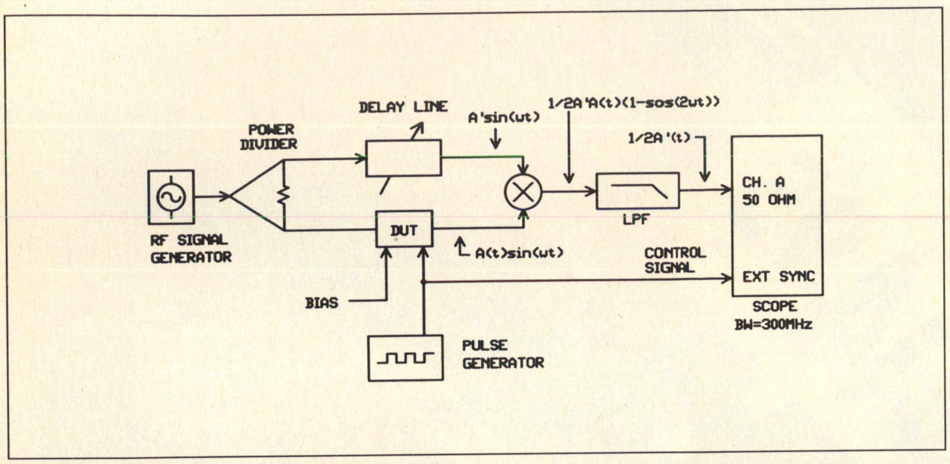

Figure 4. Test set-up using mixer.

classic definition, both these switches have about the same switching speed even though only one reaches full isolation within a short period of time. Similar examples can even be seen from vendors' advertisements and photographs of switching times. One only has to take a closer look at the photographs of the switching behavior to see a slight ringing effect or sloping-off at the trailing edge of the RF envelope. They may seem slight, but the dynamic range of the scope is only 20 to 40 dB, so those "bumps" can be as high 20 to 30 dB.

These examples show that the classic definition of switching speed is sometimes incomplete. To get a more precise definition we have to transform the classic 10 percent RF level to a much lower level within the range of the required isolation. If, for instance, the required isolation of the switch is at least 80 dB then we can define the "off time" as the time the RF level reaches a level within 5 dB of the required isolation (75 dB, 3 x 10^{-6} percent) or any other criterion as required. This will then give us a true picture of how the switch behaves. This new criterion ultimately changes the definition of the transition time as well. Therefore, it would be measured now from the crossing of the 0.5 dB from insertion loss level (90 percent) to the crossing of the 75 dB isolation level and vice versa.

Until now we have dealt with RF component delays. However, the driving circuit also has a large effect on the overall switching speed due to the internal delay of the circuit itself. A large number of drivers comprise some sort of TTL/ECL input circuit which has a propagation delay of several nanoseconds (depending on the technology) and an analog driving circuit with a delay of several nanoseconds to milliseconds. High switching speed hybrid drivers switch states within 6 to 12 ns depending on the circuit load and ambient temperature. Direct driving circuits have even faster switching speeds. Most monolithic driver circuits are somewhat slower; they are in the 50 to 200 ns range. Choosing the right driver to fit the load of the RF circuit can also determine the overall behavior. You can connect a super fast driver to a switching element presenting a heavy load and end up with a much longer switching speed than you expected; whereas a slower driver capable of driving the heavy switch would yield better results.

Methods of Measurement

There are several methods of measuring the switching speed of an RF switch. They can range from very simple to very complex as we shall see. Perhaps the most common and simplest is the method using an AM detector circuit (Schottky or tunnel diode) to convert the RF peak voltage into a DC level, which can then be seen on a scope. An example of such a test set-up can be seen in Reference 9. The scope should have a low input impedance (50 ohms) in order that the time constant of the AM detector be small enough to track the fast transients and a large enough bandwidth to see those transients. Typical detectors have PT_{SS} in the vicinity of –40 dBm and a dynamic range of 40 dB. Since the detected voltage follows the square law of the input power, it is very difficult to see the entire dynamic range on the scope at a specific V/div setting. Using an attenuator, one can see and calibrate at least a 20 dB dynamic range (in most cases). If one uses the classic definition, where at least 10 dB dynamic range is required, this method of measurement is quite satisfactory. Most of the detectors of this type have a high VSWR at the input, therefore it is recommended that an isolator be placed between the DUT and the detector in order to prevent ringing due to mismatching.

Another method of measurement is the down conversion method whereby the modulated RF signal is downconverted to DC with the help of a mixer

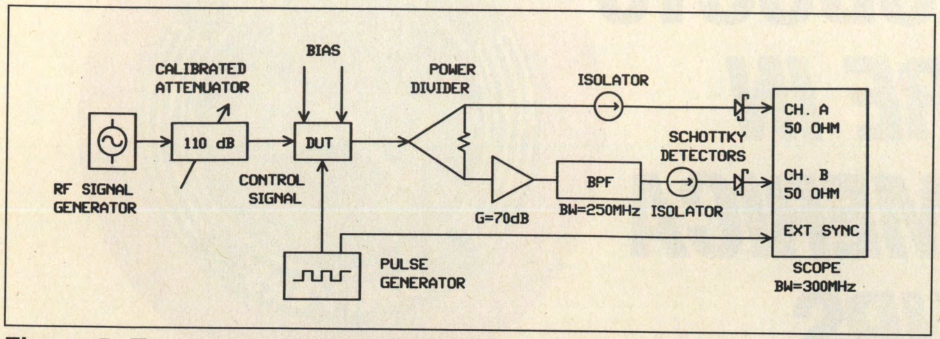

Figure 5. Test set-up measuring the first and last 20 dB of the switch dynamic range.

(Figure 4). A CW RF signal is passed through the DUT which modulates the signal. This signal is mixed with the original CW RF signal to produce two new byproducts, one at twice the original frequency of the RF signal and one at DC. This signal at DC is actually the RF envelope of the modulated RF signal from which the switching signal can be obtained. A low pass filter is used to remove the high frequency components and the phase shifter is used to obtain a maximum RF envelope amplitude from the mixer by multiplying the modulated signal and the CW RF signal in phase. The RF envelope can be seen on an oscilloscope as described earlier, but this method lacks dynamic range due to the limitations of the oscilloscope.

With present sampling oscilloscopes, the DUT can be directly connected to the scope and the RF signal envelope seen on the display. This method of measurement also lacks dynamic range as does the previous method because, here too, a specific V/div is usually set, which limits the range one can observe. In some cases, the size and resolution of the display is enough to see a 40 dB dynamic range (200mV/2mV ratio) making this method slightly better than the previous one. Again, this method is quite satisfactory when using the classic definition. The only disadvantage of this method is the extremely high price of very high bandwidth sampling oscilloscopes.

As we have noted, the problem is: how can we observe the full dynamic range at one time? A possible solution would be to use a device called a DLVA detector (Detector Log Video Amplifier) or SDLA (Successive Detector Log Amplifier) (10,11) which consists of three basic components: 1) an AM detector of the RF signal 2) a wide band RF/video amplifier and 3) performance of LOG of the detected and amplified signal. These enable the detection of weak signals since the PT_{SS} of these devices is between –40 dBm and –70 dBm with a dynamic range from 40 dB to 70 dB. The output signal voltage is presented in a dB scale (mV/dBm) enabling us to see simultaneously a much larger dynamic range on the oscilloscope. This would seem to have solved our problems, but some of the other characteristics of these devices prevent us from using them to measure fast switches (sub-microsecond switching speeds). First, these devices have a propagation delay of their own, which may be within the same time range as the switching speed of the DUT. Secondly, the rise and fall times are limited by the slew rate of the video amplifier and could be mistaken as the rise/fall time of the DUT. And lastly, they usually have a very long recovery time which may be interpreted as part of the "off time" of the DUT.

Since it is impossible to see a dynamic range of 60 to 100 dB on one oscilloscope display, let us look at the two extremes: the first 20 dB measured from the insertion loss level and the last 20 dB measured to the isolation state. Using the test set-up in Figure 4, the first and last 20 dB can be seen simultaneously on the scope. The RF signal from the switch is split into two paths: channel A displays the first 20 dB since the input power is high enough to be detected directly by the detector, and channel B displays the last 20 dB since the extremely low power near the isolation is amplified to a level high enough for the detector. Each channel compliments the other, what can not be seen at a specific time on one channel can be clearly seen on the other. For example, during insertion loss the detector of channel B is "blinded" by the saturation of the amplifier, but the detector of channel A detects the signal properly. Care should be taken when choosing the amplifier: 1) the gain should be in the

vicinity of the isolation of the DUT 2) an amp with a very short delay should be chosen 3) the amp saturation recovery time should be very fast and 4) the output saturated power must be low so as to not burn out the detector. With the use of a step attenuator, one is able to find the –0.5 dB level on channel A and the 5 dB from isolation (for example) on channel B. The filter is used to remove any spurs produced by the amplifier due to the RF switching. This method ensures that no unwanted behavior occurs in the entire dynamic range of the switch being tested. As described earlier, the slow decay of the series switch would be immediately observed since the switch will settle in the full isolation state only after a long period of time.

This method can be used for any dynamic range required. So if the switch being tested has an isolation of 60 dB, one only has to chose an amp with approximately 50 dB gain and calibrate using the step attenuator to the –55 dB level. In all cases one must be careful with the construction of the switching speed test set-up. Cable lengths must be taken into account. Isolator, filter and amplifier propagation delays must be compensated. In some cases the delays of the above components are within the same time range of the switching speed being tested. These propagation delays become very important when measuring very fast switches. A 10 ns propagation delay due to a test set-up is negligible when measuring a switching speed in the millisecond range, but it gives 100 percent error for a device switching in about 10 ns.

Summary

We now know that the classic switching speed definition may still be valid for general switching speed measurement. Although it does not always give us a complete picture of the behavior of the switch being tested. In those cases where high isolation is required, one has to use a different definition and perhaps a different test set-up with which to measure it. One of the most common errors of definition is equating transition time with switching speed. In most cases the transition time is much shorter than the overall switching speed. *RF*

Acknowledgement

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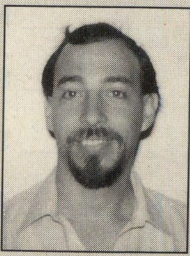
About the Author



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The Design of Constant Phase Difference Networks

By Robert J. Dehoney
Consultant

This article presents a unified method for calculating the pole frequencies and element values for a variety of constant phase difference network configurations. The accompanying BASIC program, PHASEDIF, implements the equations.

Phase difference networks are linear circuits with one input and two outputs. With a constant input amplitude, the two output amplitudes are constant with frequency, but their phases shift in such a way as to keep their phase difference constant over some bandwidth. When the phase difference is 90 degrees, the circuits are useful for single sideband generation, antenna pattern and polarization control, signal processing, FSK systems, CRT display generators, and solid state power module combining. With 60 or 120 degree shift, three-phase power amplifier combining is possible. This allows cancellation of odd order harmonics as push-pull does for even order harmonics.

The design of constant phase difference networks has been the subject of many papers and articles since Darlington's classic work in 1950 (1). In spite of this wealth of information, the practicing engineer still cannot find a simple straightforward method of designing phase difference networks of any bandwidth, any number of poles, even or odd, and any phase difference. In addition, there is no unified method of calculating the element values for the various possible all-pass circuits. This article starts to fill this need. It is based on the work of Darlington, Beauregard (2), and Lloyd (3).

In contrast to some other papers, no attempt is made here to summarize the theory; it is well covered in the literature and is too complicated to be easily described. The interested reader can consult the references included at the end of this article.

The basic configuration consists of two all-pass networks, driven in parallel, which shift the phase by different amounts. The difference in the phases of the two networks remains constant within some prescribed error over the design bandwidth.

Required is knowledge of the highest and lowest frequencies between which the phase difference is to be controlled. Given these frequency limits and the desired phase difference, a table of phase error vs. number of poles, N, can be calculated, enabling the engineer to decide on the configuration. With N known, the pole frequencies can be calculated, and after the network types are chosen, the circuit component values can be determined.

The formulas used in PHASEDIF are presented here in algebraic form. We proceed as follows:

- Enter the lowest frequency f_l and the highest frequency f_h.
- Calculate r and f_0.

$$r = \frac{f_l}{f_h} \tag{1}$$

$$f_0 = \sqrt{f_l f_h} \tag{2}$$

- Using the iterative procedure from Lloyd's program, calculate y, then q.

$$y(j) = \frac{r^2}{j}\left(j - \frac{5}{4}\right)(1 + y(j+1)) \qquad j = 12, 11, \cdots 1 \tag{3}$$

$$h = -\frac{y(1)}{4 + 2y(1)} \tag{4}$$

$$\text{If } h > 10^{-6} \text{ then } \quad q1 = h + 2h^5 + 15h^9 + 150h^{13} \tag{5}$$

$$\text{If } h < 10^{-6} \text{ then } q1 = h \tag{6}$$

$$q = \exp\left(\frac{\pi^2}{\ln(q1)}\right) \tag{7}$$

- Enter the desired phase difference d, degrees.
- Convert d to radians.
- Calculate phase error vs. number of poles.

$$e = 4q^n \sin(d) \text{ for } n = 1, 2, \cdots P \quad (d \text{ in radians}) \tag{8}$$

PHASEDIF.BAS limits P to 20 by a DIM statement in line 10.

- Display e in degrees vs. n for errors greater than, say, 0.1 degrees.
- If the networks are to be used for sideband canceling, calculate sideband suppression = 20 log{tan(e/2)}, dB.
- Enter number of poles needed, N
- Find normalized pole frequencies p(k) from p(1) to p(N).

If N is even, then

$$x = \frac{(k-1)\pi - \frac{d}{2}}{N} \tag{9}$$

If N is odd, then

$$x = \frac{(k-1)\pi - \frac{\pi - d}{2}}{N} \tag{10}$$

$$p1(k) = \cos x + q^2 \cos 3x + \cdots + q^{(m^2-m)} \cos((2m-1)x) \tag{11}$$

$$p2(k) = \sin x - q^2 \sin 3x + \cdots + (-1)^m q^{(m^2-m)} \sin((2m-1)x) \tag{12}$$

$$\text{Finally, } p(k) = \frac{p1(k)}{p2(k)} \tag{13}$$

- To get actual pole frequencies, de-normalize.

$$p1 = p(1)f_0 \tag{14}$$

$$p2 = p(2)f_0, \text{etc.} \tag{15}$$

When all the pole frequencies have been obtained, it will be seen that some are negative and the rest are positive. If N is even, there will be N/2 of each. The positive poles pertain to the A network and the negative poles to the B network.

Performance

At this point, the performance of the system can be calculated. Each pole contributes a phase shift, thus the total phase difference is simply the sum of all the individual shifts.

$$\frac{A}{2} = \tan^{-1}\left(\frac{f}{p1}\right) + \tan^{-1}\left(\frac{f}{p2}\right) + \cdots + \tan^{-1}\left(\frac{f}{p_N}\right) \qquad (16)$$

If this calculation is made, it will be seen that, from f_l to f_h, the phase difference will be as chosen with equal positive and negative error, e.

Circuit Values

Circuit files and schematics are provided in file "PHASEDIF.BAS"; the schematics are seen in Figure 1. The simplest implementation is to provide as many single pole circuits as there are poles. We will consider balanced and unbalanced RC and LC circuits and active RC circuits. In what follows, ω_k represents $2\pi p_k$, a pole frequency in radians with the sign ignored.

RC Lattice — For design, choose either R or C and either R1 or R2.

$$RC = \frac{1}{\omega_k} \qquad (17a)$$

$$R1 \cdot R2 = R^2 \qquad (17b)$$

Isolation is required between stages since the input impedance is not constant.

$$\frac{E_{out}}{E_{in}} = \frac{1}{\left(\frac{R}{R2}+1\right)^2} \qquad (17c)$$

$$\text{Phase Shift} = -2\tan^{-1}(\omega RC) \qquad (17d)$$

LC Lattice — Choose R. Then,

$$C = \frac{1}{R\omega_k}, \quad L = \frac{R}{\omega_k} \qquad (18a,b)$$

Isolation is not required between stages since the input impedance is constant and equal to R.

$$\text{Phase Shift} = -2\tan^{-1}(\omega RC) \qquad (18c)$$

LC Tee — Choose R. Then,

$$C = \frac{2}{R\omega_k}, \quad L = \frac{R}{2\omega_k} \qquad (19a,b)$$

$$M = L \text{ (very tight coupling)} \qquad (19c)$$

This is a difficult circuit to realize, because of the unavoidable stray capacitance between primary and secondary of the transformer and the high degree of coupling that is required.

$$\text{Phase Shift} = -2\tan^{-1}(\omega\sqrt{LC}) \qquad (19d)$$

Active RC — For design, choose either R or C

$$RC = \frac{1}{\omega_k} \qquad (20a)$$

$$\text{Phase Shift} = -2\tan^{-1}(\omega RC) \qquad (20b)$$

Two Pole Circuits

These circuits generate a phase shift equivalent to two single pole circuits in tandem. Designed with the same pole frequencies, they will produce the same phase shift vs. frequency. We will consider RC active and LC passive types. Information on RC passive multipole networks can be found in References 4-6.

For these circuits, the two pole frequencies are pj and pk or in radians, ω_j and ω_k. Then:

$$\omega_0 = \sqrt{\omega_j\omega_k} \qquad (21)$$

$$S = \frac{(\omega_j + \omega_k)}{\omega_0} \qquad (22)$$

$$x = \frac{\omega}{\omega_0} \qquad (23)$$

$$\text{Phase Shift} = -2\tan^{-1}\left(\frac{Sx}{1-x^2}\right) \qquad (24)$$

Wien Bridge Active All Pass Type 1 and Type 2 — These circuits have been extensively analyzed, but usually with restrictions on the relations between R1C1 and R2C2 or R1C2 and R2C1. See References 7-9. Here, we relax the restrictions and allow the designer to select C1 and C2. Since 1 percent resistors are cheaper and more common than 1 percent capacitors, this seems like a sensible scheme. As a result, the design equations are as follows:

Choose C1 and C2. Two sets of values for R1, R2 and M will result:

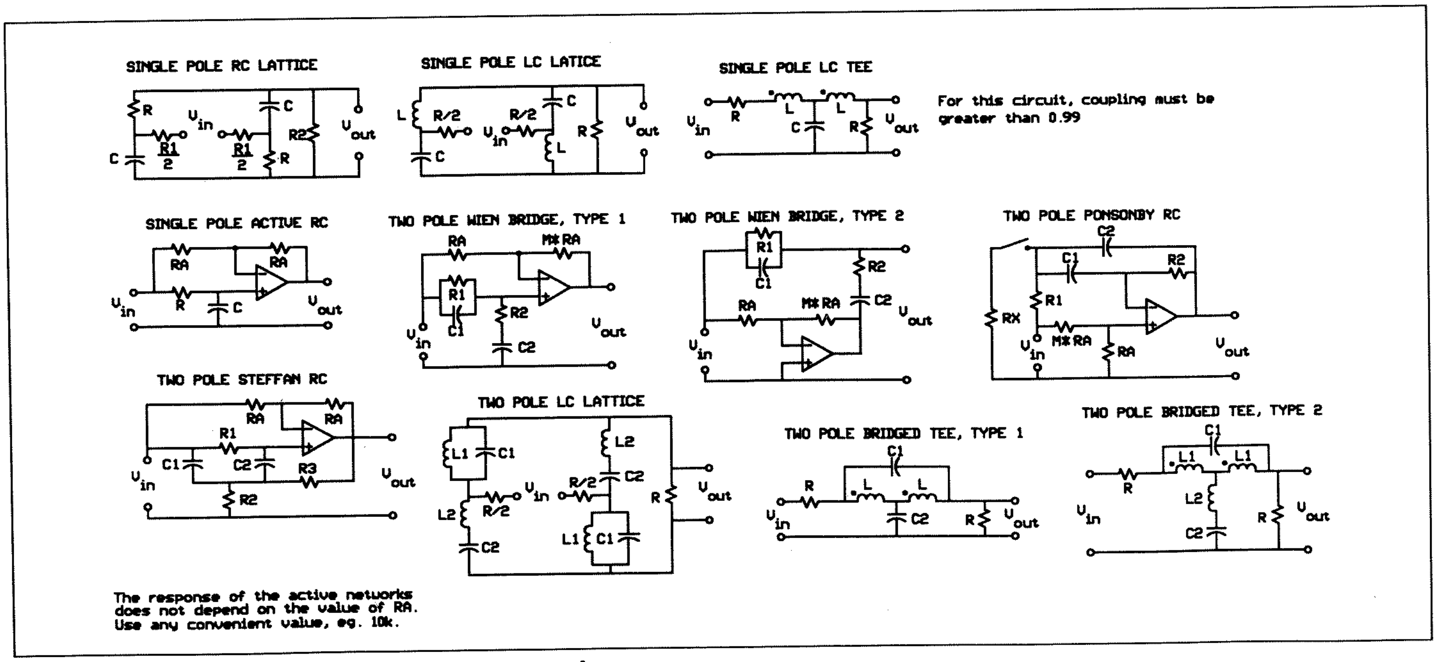

Figure 1. Constant phase difference networks.

$$R1=\frac{S+\sqrt{S^2-\frac{4(C1+C2)}{C1}}}{2\omega_0(C1+C2)} \quad (25a)$$

$$R2=(\omega_0^{\,2}\cdot C1\cdot C2\cdot R1)^{-1} \quad (25b)$$

$$M=\frac{2S-\omega_0 R1\cdot C2}{\omega_0 R1\cdot C2} \quad (25c)$$

and

$$R1=\frac{S-\sqrt{S^2-4\frac{(C1+C2)}{C1}}}{2\omega_0(C1+C2)} \quad (25d)$$

$$R2=(\omega_0^{\,2}\cdot C1\cdot C2\cdot R1)^{-1} \quad (25e)$$

$$M=\frac{2S-\omega_0 R1\cdot C2}{\omega_0 R1\cdot C2} \quad (25f)$$

Obviously, not all values of C1 and C2 will provide real values of R1, but the calculation is so trivial on a programmable calculator or computer that satisfactory values can quickly be found.

Ponsonby Active RC All Pass — This two pole circuit has the interesting feature that the phase shift can be disabled by switching in a resistor. See References 10 and 11. As with the Wien Bridge types, the circuit values can be derived in a number of ways, depending on the initial assumptions, but as before, we will assume values for C1 and C2. First, calculate K,

$$K=\frac{C2+C1}{C2\cdot S} \quad (26a)$$

Then,

$$R1=(C2\cdot\omega_0 K)^{-1} \quad (26b)$$

$$R2=(\omega_0^{\,2}\cdot C1\cdot C2\cdot R1)^{-1} \quad (26c)$$

$$M=\frac{2S}{K},\quad Rx=\frac{R1}{M} \quad (26d,e)$$

This circuit does not have unity gain, as do the others. Its gain magnitude is:

$$G=\frac{1}{1+M} \quad (26f)$$

As the C1/C2 ratio increases, the gain approaches unity.

Steffan Two Pole RC — Enter C1 and C2

$$CP=\frac{C1\cdot C2}{(C1+C2)} \quad (27a)$$

$$R1=\frac{S+\sqrt{\frac{S^2+4C2}{CP}}}{2C2\,\omega_0} \quad (27b)$$

$$RP=(\omega_0^{\,2}\cdot R1\cdot C1\cdot C2)^{-1} \quad (27c)$$

$$K=\frac{RP\cdot C1}{R1\cdot CP} \quad (27d)$$

$$R2=(S\omega_0 C1)^{-1},\quad R3=\frac{RP}{K} \quad (27e,f)$$

LC Two Pole Lattice — Choose R. Then:

$$C1=(R\omega_0 S)^{-1},\quad L1=\frac{SR}{\omega_0} \quad (28a,b)$$

$$L2=\frac{R}{\omega_0 S},\quad C2=\frac{S}{\omega_0 R} \quad (28c,d)$$

As with the single pole lattice, the input impedance is R, so isolation is not required between stages.

Bridged Tee, Type 1 — First, solve for K, the coefficient of coupling:

$$K=\frac{S^2-1}{S^2+1} \quad (29a)$$

Choose R. Then:

$$L1=\frac{S\cdot R}{\omega_0(1+K)} \quad (29b)$$

$$C1=(2\omega_0 RS)^{-1},\quad C2=\frac{2S}{\omega_0 R} \quad (29c,d)$$

This circuit is seldom used since it requires a specific coefficient of coupling K. The next circuit helps this problem.

Bridged Tee, Type 2 — Solve for K_{min}:

$$K_{min}=\frac{S^2-1}{S^2+1} \quad (30a)$$

Choose a value of K ($>K_{min}$) appropriate to the transformer construction technique. Choose R. Then:

$$L1=\frac{S\cdot R}{\omega_0(1+K)} \quad (30b)$$

$$L2=\frac{R}{2\omega_0 S}-L1\cdot\frac{1-K}{2} \quad (30c)$$

$$C1=(2R\omega_0 S)^{-1},\quad C2=\frac{2S}{\omega_0 R} \quad (30d,e)$$

Adjust L2 to compensate for errors coupling.

This article has presented, in one place, the equations used in the design of several types of constant phase difference networks. While specific operating instructions for PHASEDIF were not given here, the program follows the design procedure outlined in this article.

PHASEDIF is available on disk from the RF Design Software Service. Please see page 10 for ordering information. **RF**

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About the Author

Robert Dehoney received his BSEE from MIT in 1950. After graduation, he worked for the Allen B. DuMont Labs designing RF equipment for UHF TV. He retired from Fairchild Weston in 1987 as a Technical Director in ECM. He is now a private consultant. He can be reached at 4602 Palm Blvd., Isle of Palms, SC, 29451, or by phone at (803) 886-5785.

A Logic-Compatible Mixer

By Thomas P. Hack
Comlinear Corp.

RF circuit designers increasingly must drive analog mixers and modulators with inherently digital signals. Digital circuits form the bulk of frequency division and fractional-N circuitry in frequency synthesizers. BPSK and QPSK modulators (by definition) work with digital signals. Direct-sequence spread-spectrum systems use digitally-produced spreading codes. In all of these examples the signal that eventually makes it to the mixer or modulator starts with conventional logic levels. A mixer/modulator that works with existing logic levels (TTL, ECL, ect.) could simplify all of these designs. A unique circuit based on a high frequency multiplexer will be described which does just that.

Besides working with TTL and ECL signals, the circuit has a number of other advantages. It is small in size and uses only eight parts. It can provide lower conversion loss (with proper transformer selection) than passive approaches. All ports are well isolated from one another. DC to VHF operation is possible on all three ports by adding one high frequency op amp and two resistors to the basic design.

Circuit Operation

Although this circuit actually uses a high-speed two-input multiplexer, operation is conceptually the same as a conventional passive double-balanced mixer. With the proper local oscillator applied (large enough to provide good diode switching action) a positive LO signal makes D2 and D4 conduct, connecting the lower half of T2's secondary (Figure 1a). D1 and D3 are reverse biased. The RF signal is coupled to the IF port with an in-phase relationship to the RF port. A negative LO excursion makes D1 and D3 conduct (D2 and D4 are off) and an out-of-phase RF signal shows up at the IF port (Figure 1b).

Compare this with an active mixer based on a multiplexer (Figure 1c). T1 acts as a phase splitter and the LO port once again determines whether the RF signal is routed to the IF port in an in-phase or out-of-phase relationship. The multiplexer replaces the diodes in routing the RF signal to the IF port.

Although conceptually the same, the circuit differences between the two approaches make for a number of interesting performance differences. The active approach has good isolation. LO to RF isolation is good since the inputs of the multiplexer are buffered. LO to IF isolation is also very good in the active approach because only one transformer and the multiplexer determine mixer balance. The multiplexer has high channel-to-channel matching (because in part it is a monolithic device and the two signal paths are laid out for high channel-to-channel matching). Compare this with a passive mixer which must match four diodes and have two well balanced transformers in order to get reasonable isolation.

When it comes to distortion the active mixer is extremely good at low frequencies, but the passive approach is better at high frequencies. The active mixer's distortion performance is less affected by improper IF port terminations (such as those caused by IF filters).

A Complete Circuit

Figure 2 is the complete schematic for the mixer. T1, a 1:4 impedance ratio transformer delivers anti-phase voltages to input A (pin 2) and input B (pin 4) of multiplexer U1. With select (pin 7) high, channel A is routed to the output (pin 11) of the multiplexer. With select low, channel B is routed to the output. C2 and C3 provide active circuit compensation. C1 and C4 are power supply bypasses. R1 in conjunction with T1 sets the RF port input impedance. R2 sets the output impedance. Pin 7 is a logic-compatible LO port. The circuit as shown (with pin 6 open) works with ECL logic levels. Connecting pin 6 to Vcc makes pin 7 TTL compatible.

Mixer Test Results

Figure 3 shows the block diagram of the test setup used in evaluating the performance in a low frequency mixer application. An HP 3326A synthesizer provides an 8 MHz RF signal and a HP 8662A synthesizer provides a 29.45 MHz LO signal. Low pass filters are used after each synthesizer to remove harmonics which might otherwise corrupt the measurement. In addition, a squaring circuit based on an MC10114

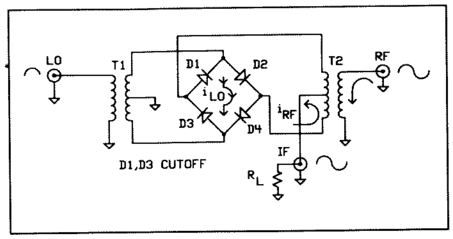

Figure 1a. Passive double-balanced mixer, positive-going LO.

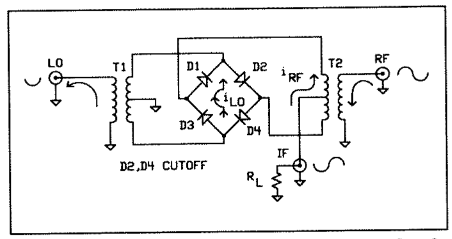

Figure 1b. Passive double balanced mixer, negative going LO.

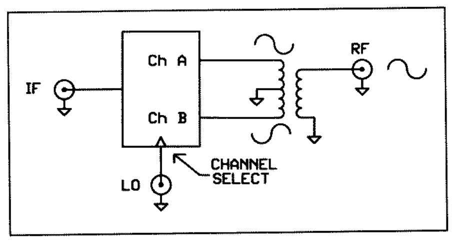

Figure 1c. Active double-balanced mixer.

is used to convert the LO to ECL logic levels. At this point you might be asking "Why filter the LO signal when it is only going to be squared up?" The answer is that the LO filter's primary purpose is to remove even order harmonics. The squaring circuit will not generate significant even-order terms if set up properly.

With the signals fed to the mixer, the mixer not only produces the desired IF (which in this case is at approximately 21.45 MHz) but undesired signals due to feedthrough, higher-order mixing, and so on. A HP 3588A spectrum analyzer measures all of these products (up to 150 MHz).

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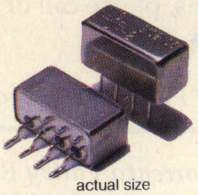

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Model	LO Power (dBm)	Freq. LO/RF (MHz)	■ Conv. Loss (dB) $\overline{X}$	■ Conv. Loss (dB) δ	Isol. L-R (dB)	Price,$ Ea. 10 qty
TUF-3	7	0.15-400	4.98	0.34	46	5.95
TUF-3LH	10		4.8	0.37	51	7.95
TUF-3MH	13		5.0	0.33	46	8.95
TUF-3H	17		5.0	0.33	50	10.95
TUF-1	7	2-600	5.82	0.19	42	3.95
TUF-1LH	10		6.0	0.17	50	5.95
TUF-1MH	13		6.3	0.12	50	6.95
TUF-1H	17		5.9	0.18	50	8.95
TUF-2	7	50-1000	5.73	0.30	47	4.95
TUF-2LH	10		5.2	0.3	44	6.95
TUF-2MH	13		6.0	0.25	47	7.95
TUF-2H	17		6.2	0.22	47	9..95
TUF-5	7	20-1500	6.58	0.40	42	8.95
TUF-5LH	10		6.9	0.27	42	10.95
TUF-5MH	13		7.0	0.25	41	11.95
TUF-5H	17		7.5	0.17	50	13.95
TUF-860	7	860-1050	6.2	0.37	35	8.95
TUF-860LH	10		6.3	0.27	35	10.95
TUF-860MH	13		6.8	0.32	35	11.95
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TUF-11A	7	1400-1900	6.83	0.30	33	14.95
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TUF-11AH	17		7.3	0.28	35	19.95

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■ $\overline{X}$ = Average conversion loss at upper end of midband ($f_U/2$)

δ = Sigma or standard deviation

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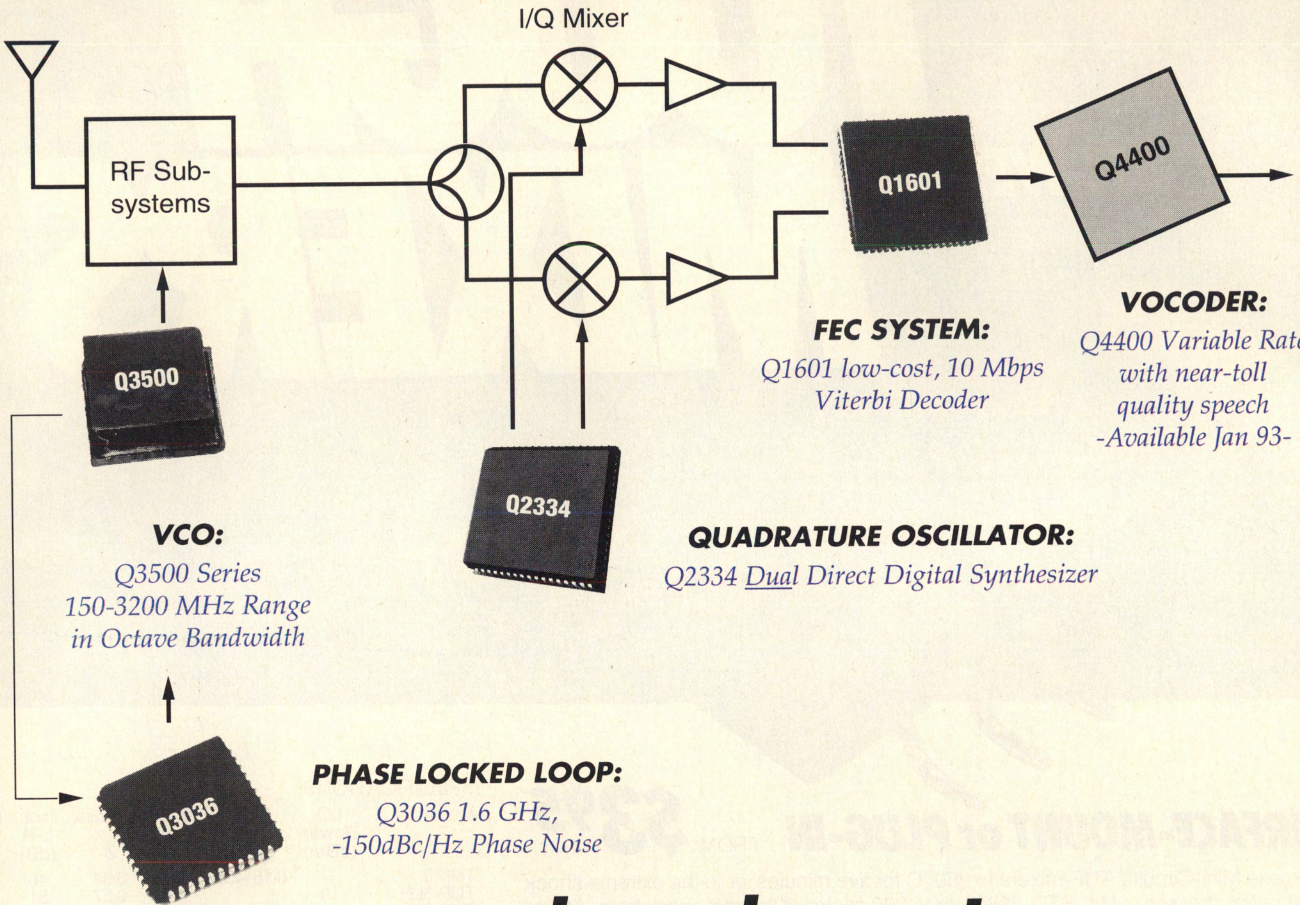

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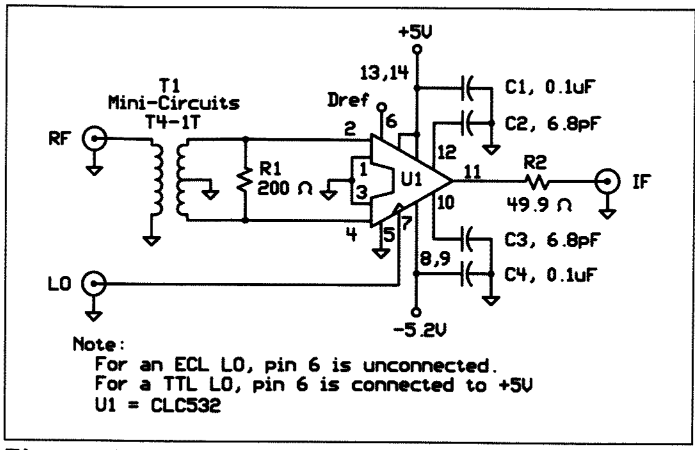

Figure 2. The logic compatible active double-balanced mixer.

Mixer harmonic intermodulation (in dBc to desired IF)

RF order	LO 0	LO 1	LO 2	LO 3	LO 4
0	—	−48.3	−36.8	−61.8	−35.3
1	−45.7	0.0	−42.3	− 9.2	−39.0
2	−52.6	−49.5	−50.4	−48.5	−48.2
3	−62.4	−39.5	−59.9	−39.4	−58.1
4	−63.5	−64.8	−61.8	−64.4	−61.2

Harmonic LO order

LO = 29.45 MHz, 4dBm
RF = 8.00 MHz, −4dBm

Figure 4. Mixer test results

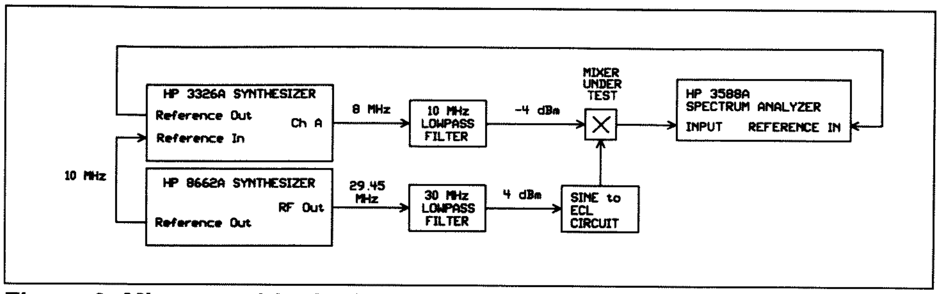

Figure 3. Mixer test block diagram

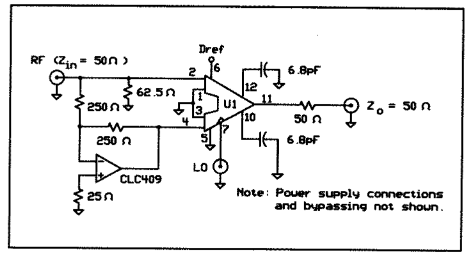

Figure 5. A fully DC-coupled mixer.

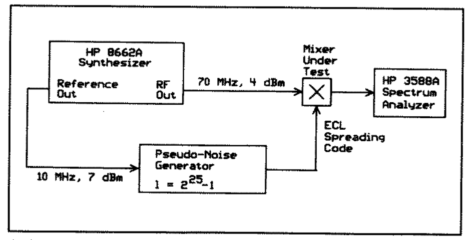

Figure 6. BPSK and direct-sequence spread-spectrum test circuit.

where any given mixer product will show up on the spectrum analyzer. The spectrum analyzer can use a zero frequency span setting and still provide accurate measurements. Zero frequency span was used producing very fast sweeps. A 10 sweep video average was used to reduce measurement uncertainty.

The results are given in Figure 4. At these frequencies the mixer is roughly comparable to a standard, type-1 (+7 dBm LO), passive double-balanced mixer. At lower frequencies it does even better. With a 1:4 impedance ratio transformer for T1, single sideband conversion loss is approximately −9 dB. Substituting a 1:16 transformer should drop the conversion loss to approximately −3 dB.

A DC-Coupled Mixer

We have seen that this part makes a reasonable low-frequency mixer. Higher- order mixer products are reasonable and the LO and IF ports are DC coupled. The RF port can be DC coupled by removing the transformer and substituting an active phase splitter (Figure 5). Now all ports are DC coupled — something not found in most mixers. This might be useful in products such as low frequency spectrum analyzers.

Modulators

Probably the best application of this circuit is in BPSK, QPSK, and direct sequence spread spectrum modulators. To test the circuit, a 10 Mchips/sec (suitable for the 915 MHz ISM band) linear maximal PN spreading code was applied to the LO port and a 70 MHz pure tone was applied to the RF port of the circuit, (Figure 6). A long spreading code ($2^{25}-1$) was used so that the spreading code itself did not degrade the observed carrier suppression (The frequency of occurence of ones and zeros in a linear maximal sequence is slightly imbalanced but gets better as the sequence length is increased.) ECLinPS logic was used to generate the PN signal (Figure 7) because this logic family has excellent edge rates (giving the best possible code balance). For most consistent performance, the ECL device that drives the modulator and the modulator should share the same −5.2 V supply.

Carrier balance is very good in this application (no visible spike at the center frequency, Figure 8). The results shown are without any adjustments. Code balance is imperfect (spikes are visible in the nulls in the spectrum) but can be improved with tweaking.

Phase Detector Tests

In addition to modulator and mixer applications, this circuit also can be used as a phase detector. To test the circuit in phase detector applications, two squarewaves, one at 1 MHz and another at 1.0001 MHz were applied to the circuit (Figure 9). The phase detector produces an output proportional to the difference in instantaneous phase between the two signals (the difference product) as well as a "sum" product which is filtered before the mixer output is sent to a digitizing oscilloscope. The difference frequency is 100 Hz so the instantaneous phase between the two input signals slips 2π radians every 10 milliseconds. With a two volt p-p RF input signal, the phase detector output amplitude is 0.8 Vpp with a double terminated output (Figure 10). This corresponds to a phase detector constant of 0.25 V/radian. Phase detector linearity is very high with no obvious flat spots or other problem areas.

Summary

A digitally compatible mixer has been described with a number of desirable attributes:

1) The LO port is ECL and TTL compati-

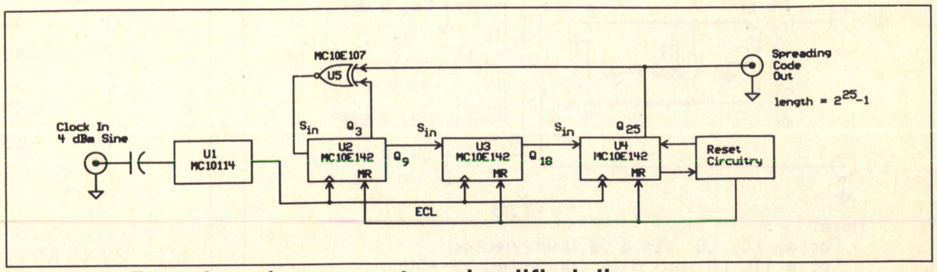

Figure 7. Psuedo noise generator, simplified diagram.

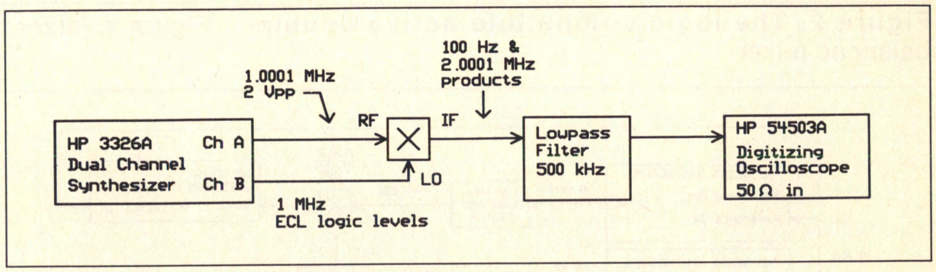

Figure 9. Phase detector test circuit.

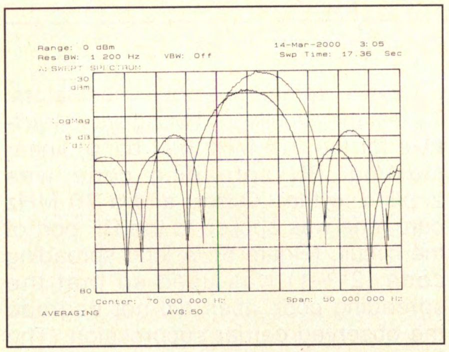

Figure 8. A 70 MHz carrier with 10 Mchip/sec pseudo noise modulation. An offset trace is also shown to display null spikes.

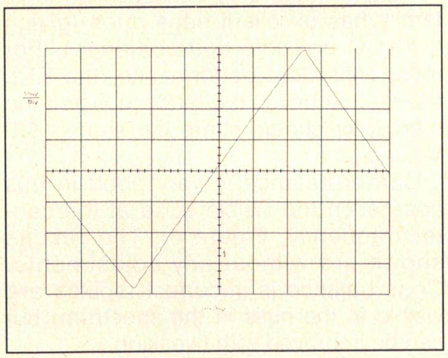

Figure 10. Phase detector output.

ble making it suitable for digital modulators, phase detectors, and spread spectrum applications.
2) All three ports can be DC coupled. This may be useful in low frequency spectrum analyzers.
3) It can have lower conversion loss than passive double-balanced approaches.
4) Performance is high at low frequencies.

This circuit provides another alternative in several thorny mixer applications. *RF*

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2. *RF/IF Signal Processing Handbook*, Volumes 1 and 2, Mini-Circuits.

About the Author

Until recently, Tom Hack was an applications engineer at Comlinear Corporation, He received his BSEE from Cooper Union, his MSEE from Renesselaer Polttechnic Institute and his MBA from the University of Colorado. Questions regarding the CLC532 should be addresses to Comlinear Corp. at 4800 Wheaton Dr., Ft. Collins, Co. 80525. Other questions should be directed to Mr. Hack at (303)229-0601.

Mixers: An RF Balancing Act

By Andy Kellett
Technical Editor

It is hard to find a block diagram for an RF communications system that doesn't contain a mixer. Up conversion, down conversion, product detection, IQ modulation, phase locked loops, all these functions are implemented with mixers. As a result, mixer use is expanding with the RF market.

Mostly found at the front end of communications systems, mixers have the difficult task of taking a weak signal and translating it in frequency, an inherently non-linear function. In the process, spurious products must be kept small while the resulting signal is kept out of the noise. It is no wonder that designers look to mixer manufacturers for what might seem to be an easy module to homebrew.

Some of the key specifications for mixers (apart from price, size, reliability and other such generic qualities) are isolation, intercept point (third order and others), conversion gain, LO power and dynamic range. These qualities are all linked; improving one can improve others — and degrade yet others. So what specifications are manufacturers working to improve? The answer depends on the application the mixer will go into.

High signal purity is the most important quality in some systems. For these, mixers with high third order intercept point and high isolation are required. The mixer most commonly used to meet these requirements is the double balanced diode ring mixer. Standard lines of these mixers are available from a number of companies: Synergy Microwave, Tele-Tech, RF Prime, Lorch Vernitron and Mini-Circuits, among others, but very high performance requirements more often lead to custom orders.

"We have a catalog which represents what we are capable of doing, but we make a number of custom mixers for different people. That's how we've been able to grow — by supplying the market with special mixers," says Shankar Joshi, engineering manager for Synergy Microwave Corp. Similarly, Lorch Vernitron offers custom high performance mixers, "Our customers come to us saying for instance, 'Lorch, can you build me something that instead of 12 dB isolation has 40?'; I tell our mixer engineer to make something 5 dB better and he comes back a week later with something 6 dB better," says Mark McWhorter, Lorch's director of marketing and product development.

While signal performance is still an issue for higher volume applications, power consumption and cost are also important considerations. The much ballyhooed emergence of wireless communications has made battery operation and affordability new goals in communications designs. "Local oscillator power has become more important with all the handheld applications, so for these high volume [applications] LO power is more of an issue. That, of course, is then related to cost because high volumes require low cost, and finally, low cost ties into surface mount and tape and reel," says RF Prime's President, Steve Markoe Sr.

Active mixers have also found use in cost conscious applications. Whereas passive mixers exhibit a conversion loss, active mixers exhibit conversion gain, meaning an active mixer can eliminate a stage of amplification after conversion. However, active mixers pay for their gain with narrower bandwidth, higher noise figure and smaller dynamic range than passives.

Discrete active mixers can be implemented with just a FET, however, according to the President of Tele-Tech Corp., John Duncan, "I don't think there's as big a market for active mixers because the people who use them build them themselves." Active mixers as integrated circuits might be used by a larger market says Gary LaBelle, Product Marketing Manager for Hewlett-Packard Company's Wireless Component Division, "Really, it's the digital cellular and cordless stuff that's pushing active mixers, because they are much more complex. Instead of just having a simple modulator in a system, now you need a vector modulator, which also requires some phase shifting and maybe an upconversion. The demand is to integrate all this into one package."

What things are mixer manufacturers working on? "I think packaging is going to be the big push," says Tele-Tech's Duncan, "there really aren't any good surface mount packages for mixers." According to Duncan, conventional gull-wing SMT packages have too much parasitic inductance for high performance, and substrate-on-board packaging results in poor production yields because of poor bonding.

Synergy Microwave has also recognized a need for improved surface mount packaging and has introduced a metalized, surface mountable package, "Ordinary plastic packages lack EMI/RFI shielding, and because of a lack of common ground, they lack in isolation," says Synergy's Joshi.

"We're seeing the frequency move from the dominant area which was 900 MHz, which is where your cellular phones are today, to 1.8, 2.4, and up to 4.8 and 5.2 GHz," says RF Prime's Markoe, "As the frequency goes up we see a lot of people in military microwaves going into the commercial sector.

The use of mixers will grow with the RF market. Balancing customer's electrical specifications and price requirements has been and will be the task that mixer manufacturers face. Says Mini-Circuits' President and CEO, Harvey Kaylie, "When I got into the mixer market it was small and it required a much smaller investment than it does today, so it represented a good oportunity, and fortunately we gave the customers what they wanted." Though the mixer market today is large, making mixers that meet customers' individual needs still seems to be the way to succeed. *RF*

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INFO/CARD #173

VCO Specification Guide

M/A-COM Semiconductor Division offers a selection guide for specifying their voltage controlled oscillators. The VCOs are available in TO-8 packages and in frequency bands between 1 and 10 GHz, with bandwidths from 5 to 40 percent.
M/A-COM Semiconductor Division
INFO/CARD #172

RFI/EMI Shielding Products

A four-color brochure from Tech-Etch describes their line of RFI/EMI shielding products. The fold-out brochure contains a product selection matrix, displaying products across the top and parameters and descriptions going down the page. Among the products are finger stock, knitted wire mesh strips, flat gaskets, shield-wrap, and vent and filter shields.
Tech-Etch, Inc.
INFO/CARD #171

EMI Groove Gasket Mounting Guidelines

A designer's guide from Spira discusses five different ways to employ their high shielding, long wearing EMI gaskets. One of the groove designs highlighted is the new "pinch boss" groove which offers a very inexpensive, reliable solution for high volume commercial applications. The pros and cons of each method are examined, including: shielding benefits, cost considerations, and manufacturability.
Spira Mfg. Corp.
INFO/CARD #170

RF design

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